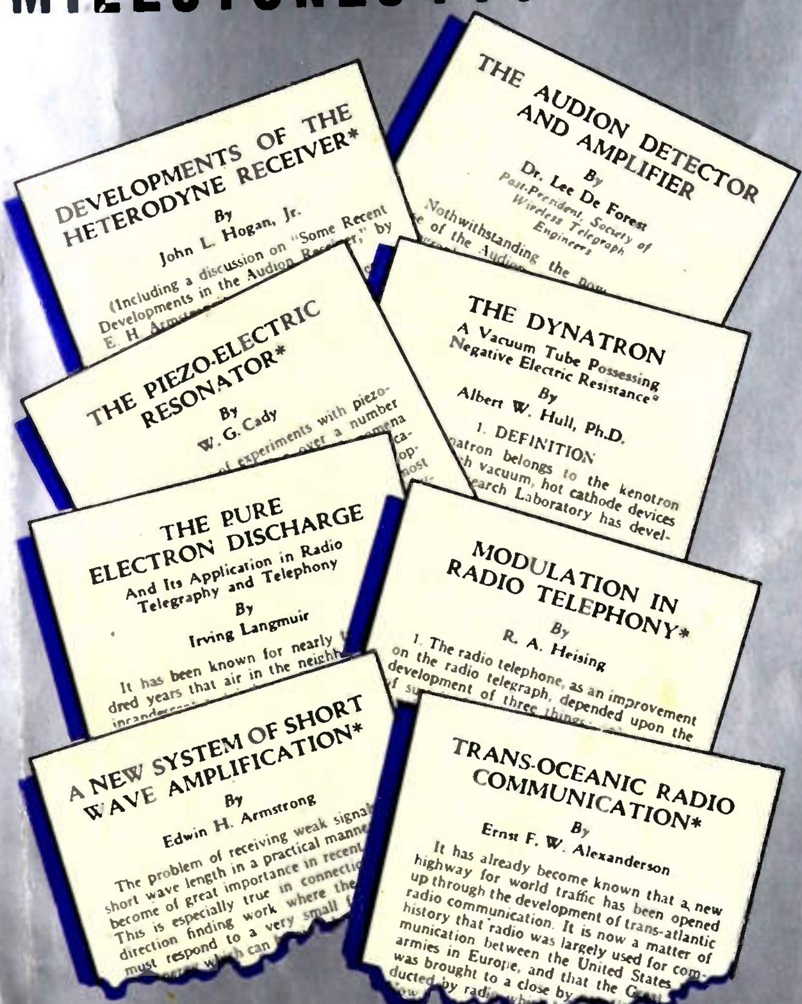
Vol. 40 May 1952 No. 5

Proceedings of the I·R·E

40th Anniversary

EARLY MILESTONES . . .



PROCEEDINGS OF THE I.R.E.

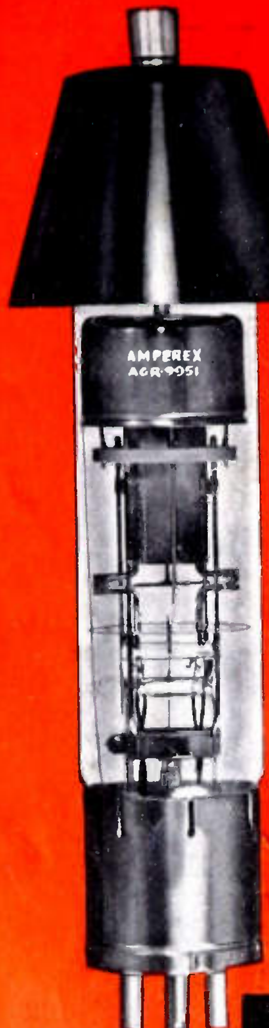
Table of Contents, Indicated by Black-and-White Margin, Follows Page 64A

The IRE Standards on Pulses and Standards on Magnetrons, Definitions of Terms, appear in this issue.

The Institute of Radio Engineers

New AMPEREX tubes

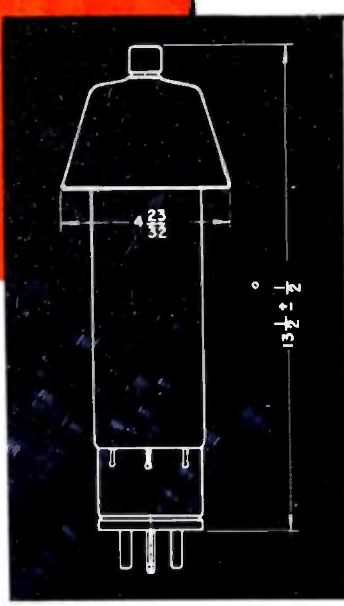
Specifically designed for grid-control operation at peak anode voltages as high as **27,000 v.** for heavy duty INDUSTRIAL uses and high power TRANSMITTERS with outputs to **150 KW.** (3 phase full wave)



	AGR-9951/5870		AGR-9950/5869	
CATHODE Directly Heated, Oxide Coated				
MAXIMUM PEAK ANODE VOLTAGE				
Inverse	27,000	10,000	13,000	10,000
Forward	27,000	10,000	13,000	10,000
CONDENSED MERCURY TEMPERATURE LIMITS (centigrade)	+30° to +40°	+25° to +60°	+25° to +55°	+25° to +60°
MAXIMUM PLATE CURRENT (Amperes)				
Peak	10		4	
Average	2.5		1	
FREQUENCY RANGE (cps)	25 to 150		25 to 150	
FILAMENT VOLTAGE	5.0		5.0	
FILAMENT CURRENT (amperes)	15		6.5	
TUBE VOLTAGE DROP (volts, approx.)	14 (Ib = 10 amperes)		15 (Ib = 4 amperes)	

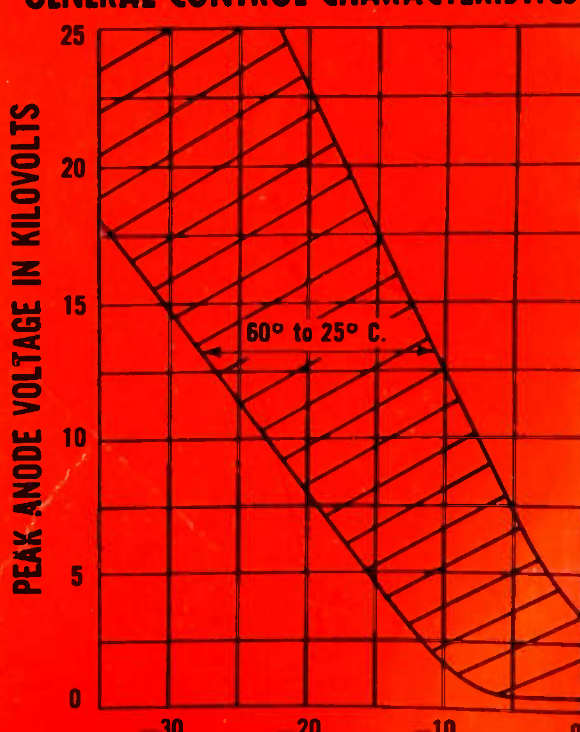
PROVEN LIFE

AGR-9951/5870
$110.00



Re-tube with AMPEREX

GENERAL CONTROL CHARACTERISTICS



Data sheets and charts available on request

THREE-ELECTRODE, MERCURY VAPOR

RECTIFYING TUBES

with NEGATIVE CONTROL characteristics

AMPEREX ELECTRONIC CORP.
25 WASHINGTON STREET, BROOKLYN 1, N. Y.
In Canada and Newfoundland: Rogers Majestic Limited
11-19 Brentcliffe Road, Leaside, Toronto, Ontario, Canada

PROCEEDINGS OF THE I.R.E. May, 1952, Vol. 40, No. 5. 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. Printed in U.S.A.

Table of Contents will be found following page 64A

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

NEREM, Saturday, May 10, 1952
Copley Plaza Hotel, Boston, Mass.
New England Radio Engineering Meeting
Gen. Chairman: Alfred J. Pote
71 West Squantum
N. Quincy, Mass.

Δ

National Conference on Airborne Electronics
May 12, 13 & 14, 1952
Hotel Biltmore, Dayton, Ohio
Exhibits: Paul D. Hauser
1430 Gascho Drive, Dayton 3

Δ

4th Southwestern IRE Conference
May 16, 17, 1952
Rice Hotel, Houston, Texas
Exhibits: Gerald L. K. Miller
1622 W. Alabama
Houston 6, Texas

●

Radio Parts Show
Chicago
May 19-24, 1952

Δ

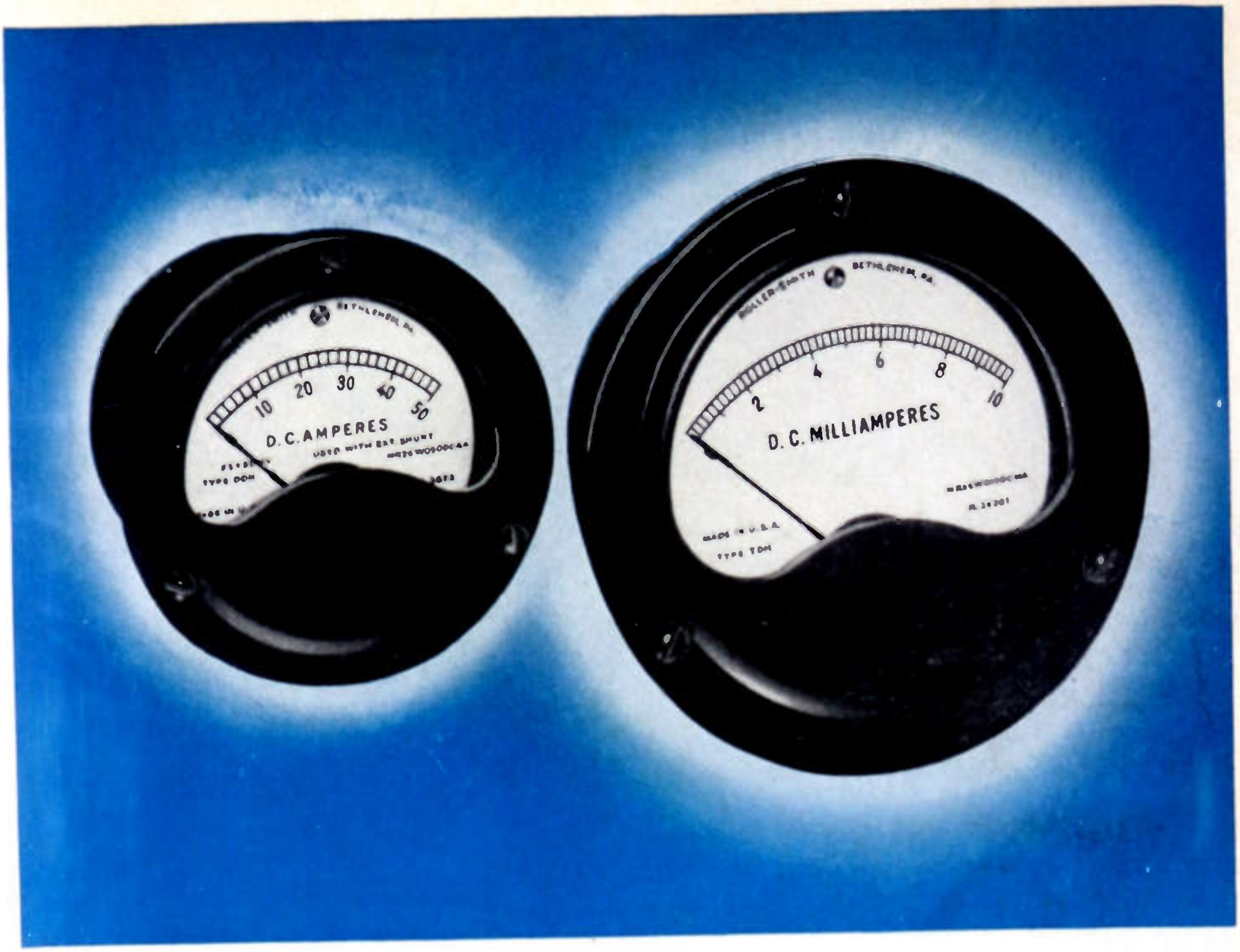
Western Electronic Show and IRE Regional Convention
August 27, 28 & 29, 1952
Municipal Auditorium
Exhibits: Heckert Parker
215 American Avenue
Long Beach, Calif.

●

I.S.A.
Seventh National Instrument Exhibit and Instrument Society of America Conference
September 8-12, 1952
Cleveland Municipal Auditorium
Exhibits: Mr. Richard Rimback, Mgr.
921 Ridge Avenue
Pittsburgh 12, Pa.

Δ

National Electronic Conference
Sept. 29, 30, Oct. 1, 1952
Hotel Sherman, Chicago, Ill.
Exhibits Manager: Mr. R. M. Krueger,
c/o Amphenol, 1830 South 54th Ave.,
Chicago 50, Ill.


10 20 30 40 50
D.C. AMPERES
2 4 6 8 10
D.C. MILLIAMPERES

Bendix...

...is Prepared to Design Dynamotors to MIL-D-24 for Quantity Production...

● The design and manufacture of dynamotors for military service has been Red Bank's business for over ten years. The requirements of the new dynamotor specification MIL-D-24 therefore include many of the features that are incorporated in all Bendix dynamotors. When compliance with MIL-D-24 is required, Bendix engineers will work with you to design a unit exactly fitting your needs and will prepare the detailed supplementary specifications covering your model as required by MIL-D-24. Following approval and assignment of a military designation, Bendix production will be geared to your schedule. Write direct to:

RED BANK DIVISION OF BENDIX AVIATION CORPORATION
RED BANK, NEW JERSEY
Export Sales: Bendix International Division, 72 Fifth Avenue, New York 11, N. Y.

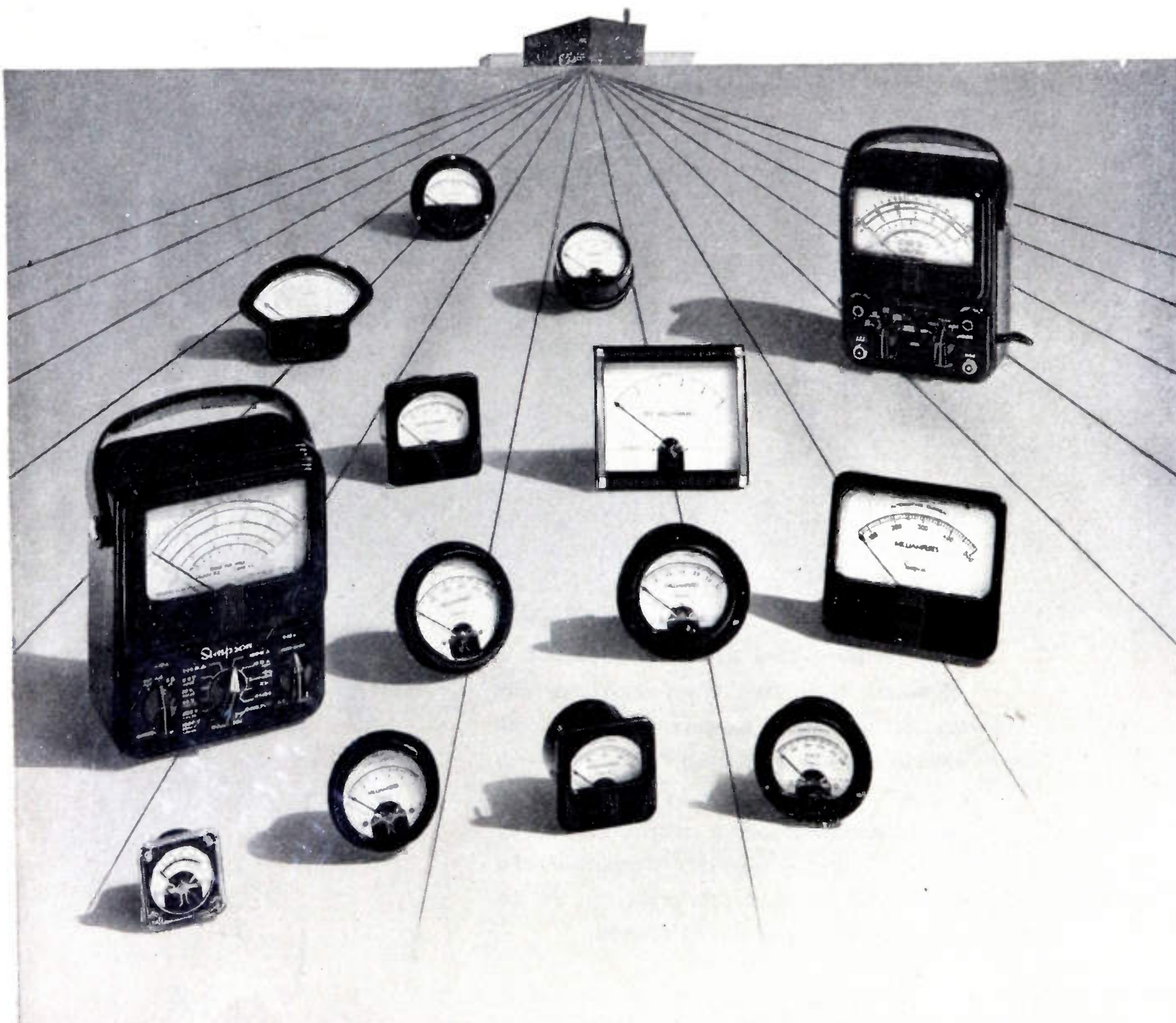
RED BANK DIVISION

Bendix
AVIATION CORPORATION

SPECIFY Simpson

for Accurate, Dependable Electrical Measurements

BECAUSE—Simpson has developed quality control to a new modern high with *this* successful production policy. Design everything that goes into an instrument—make everything that goes into an instrument—keep designing for the future—keep quality steadily higher—keep prices consistent with material and labor costs without exploitation. This quality control is evident in every Simpson instrument whether panel or switchboard, custom-built or stock. The instruments illustrated in panel below are only a few of the wide variety of instruments in the complete Simpson line. Let Simpson engineers help you solve your panel instrument problems—and for your standard instrument requirements take advantage of our large stock.



Simpson ELECTRIC COMPANY

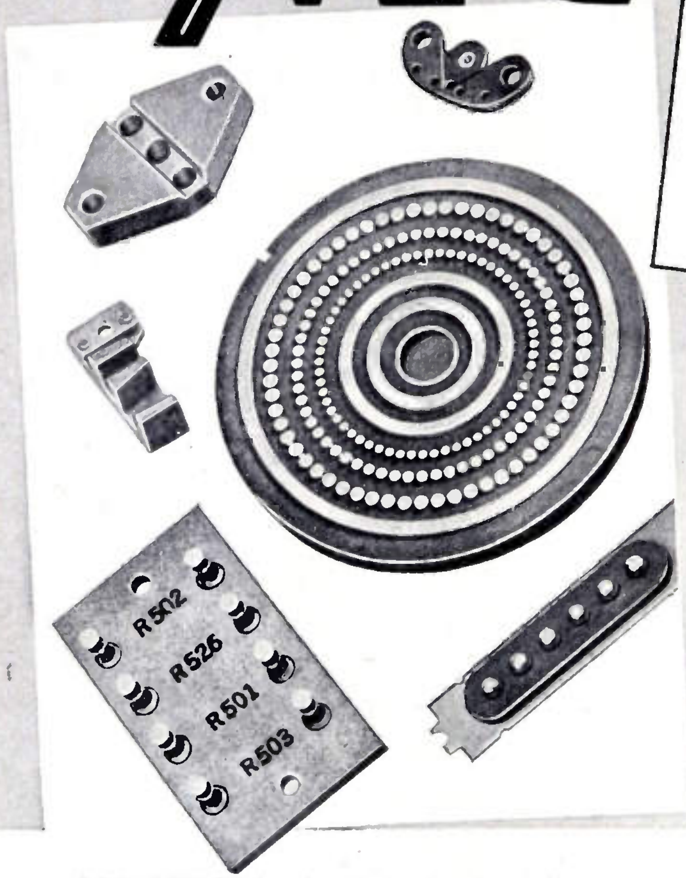
CONSERVE CRITICAL MATERIALS

5200 W. KINZIE ST., CHICAGO 44 • COlumbus 1-1221 • In Canada: BACH-SIMPSON, LTD., LONDON, ONT.

For better products at lower cost-

SPECIFY **MYCALEX**

Glass-Bonded Mica INSULATION



- MOLDS AND MACHINES TO CLOSE TOLERANCES
- MOLDABLE WITH METAL INSERTS
- CAN BE TAPPED, THREADED, SLOTTED
- AVAILABLE IN RODS, SHEETS, SPECIAL SHAPES
- MOLDED IN PRACTICALLY ANY SHAPE OR SIZE
- LOW-LOSS FROM 60 CPS TO 24,000 MCS

MYCALEX glass-bonded mica insulation is the one highly adaptable, versatile insulating material that combines every desirable characteristic required in a modern dielectric. Although far superior to lower cost dielectrics, MYCALEX offers considerable advantages over many materials costing several times as much. MYCALEX is available in various grades, each featuring specific characteristics to meet particular needs. Since proper application of the right grade of MYCALEX has resulted in simultaneous product improvement and lower cost in hundreds of instances, it's good business to check with MYCALEX before specifying sheet, rod, fabricated or molded insulation.

JAN APPROVED

MYCALEX 410 is approved fully as Grade L-4B under National Military Establishment Specification JAN-I-10, "Insulating Materials, Ceramic, Radio, Class L."

MYCALEX 400 is approved fully as Grade L-4A under National Military Establishment Specification JAN-I-10, "Insulating Materials, Ceramic, Radio, Class L."

Write for 20-Page Catalog Today!

A valuable compilation of engineering data and manufacturing information on electrical insulation that you'll surely want for your technical file. Request it today—no obligation.

CHARACTERISTICS

MYCALEX GRADE	400	410	410X
POWER FACTOR, 1 MC	0.0018	0.0015	0.012
DIELECTRIC CONSTANT, 1 MC	7.4	9.2	6.9
LOSS FACTOR, 1 MC	0.013	0.014	0.084
DIELECTRIC STRENGTH, volt/mil	500	400	400
VOLUME RESISTIVITY, ohm-cm	2×10^{15}	1×10^{15}	5×10^{14}
ARC RESISTANCE, seconds	300	250	250
MAX. SAFE OPER. TEMP., °C	370	350	350
WATER ABSORPTION % 24 hrs.	NIL	NIL	NIL

SINCE 1919 MYCALEX THE INSULATOR TRADE MARK REG US PAT OFF

MYCALEX CORPORATION OF AMERICA

Owners of 'MYCALEX' Patents and Trade-Marks

Executive Offices: 30 ROCKEFELLER PLAZA, NEW YORK 20 — Plant & General Offices: CLIFTON, N.J.

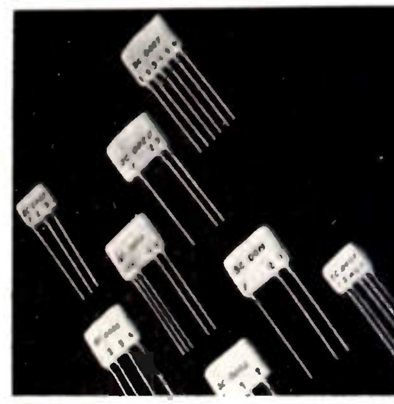

FREED Instruments & Transformers
"PRODUCTS OF EXTENSIVE RESEARCH"
MINIATURE TRANSFORMERS
MINIATURE TOROID INDUCTORS
PULSE MODULATORS
No. 1030 LOW FREQUENCY "Q" INDICATOR
No. 1150 UNIVERSAL BRIDGE
SLUG TUNED DISCRIMINATORS
HIGH FIDELITY TRANSFORMERS
No. 1010 COMPARISON BRIDGE
FREEDSEAL TREATMENT
FILTERS
No. 1020B MEGOHMMETER
No. 1180 A.C. SUPPLY
A.C. VOLTS
A.C.-V.T. VOLTMETER MODEL 1040 FREED TRANSFORMER CO
NO. 1040 VOLTMETER
VOLTAGE RANGES: .001 volts to 100 volts in five ranges (.01, .1, 1, 10, and 100 volts full scale).
ACCURACY: 2% on full scale on all five ranges, on sinusoidal voltages.
FREQUENCY RANGES: 10 to 200,000 cycles, .1 db. variation from 20 cycles to 150,000 cycles; .50 db. variation from 10 cycles to 200,000 cycles.
INPUT IMPEDANCE: Equivalent to 500,000 ohm resistance in parallel with a 15 MMF. condenser.
STABILITY: Effect of variation in line voltage from 100 volts to 125 volts is 1%. Effect in changes of tubes is less than .5%.
METER: 4" suppressed zero 1 MA meter protected against overloads.
POWER SUPPLY: The instrument is entirely self-contained and operates on 100-125 volts, 50-60 cycles. Total consumption, 40 Watts.
DIMENSIONS: 4⅞" High, 5⅝" Wide, 9⅞" Long.
WEIGHT: 12 pounds.
No. 1110A INCREMENTAL INDUCTANCE BRIDGE
No. 1170 D.C. SUPPLY
SEND FOR COMPLETE TRANSFORMER AND INSTRUMENT CATALOGS
FREED TRANSFORMER CO., INC.
1720-B WEIRFIELD ST., BROOKLYN (RIDGWOOD) 27, N. Y.
EXPORT DIVISION:—458 BROADWAY N.Y.C. 13, N.Y.

News—New Products

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

Printed Circuits

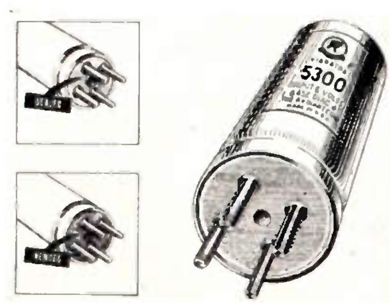
Stupakoff Ceramic and Manufacturing Co., Latrobe, Pa., announces the development of a series of printed electrical circuits. A number of standard circuits are available, some of which incorporate as many as six separate resistors and capacitors in a permanent circuit. Special circuits can be made to meet individual requirements.



In the production of these circuits, patterns for resistors, capacitors and conductors are "printed" on vitreous, high-dielectric ceramic plates by a silk screen process. The dielectric properties of the ceramic are used for the capacitors, while silver is used for conductors and carbon graphite or other resistance materials for resistors. After the patterns are printed they are bonded permanently to the ceramic surface by controlled curing; then are protected from abrasion and humidity by the application of a plastic covering.

New Type Vibrator

Milton S. Roth, Jobber Sales Manager of the **Radiart Corp.**, 3571 W. 62 St., Cleveland 2, Ohio, announces an improvement in the firm's line of replacement auto-radio vibrators.



An automatic vent has been incorporated in the "Red Seal" base. This vent is wax sealed at the factory. When the vibrator is put into use, the temperature rise inside the vibrator melts out the wax and permits air circulation for greater performance and longer life.

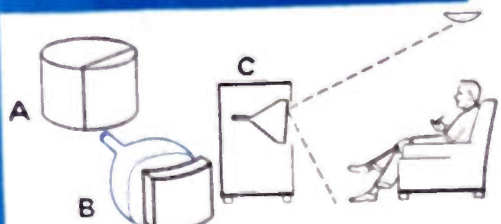
(Continued on page 18A)

New CBS-Hytron cylindricals
17LP4 and 21FP4A low-voltage electrostatics
17QP4 and 21EP4A electromagnetics

Facts

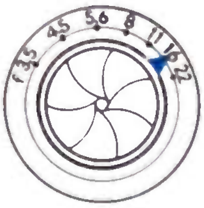
YOU'LL WANT TO KNOW ABOUT NEW

CBS-HYTRON Cylindricals



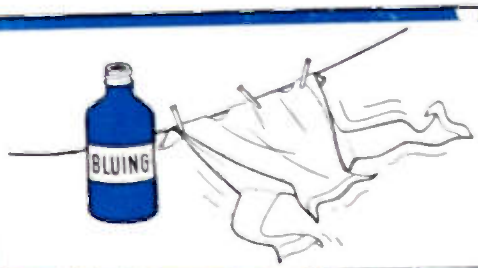
WHY CBS-HYTRON CYLINDRICAL?

To eliminate reflected glare? How? Simple as ABC: A. Imagine a cylinder; slice it vertically. B. You now have the shape of the face plate of a cylindrical tube: curved horizontally; straight, vertically. C. Light falling on this surface at an angle from above is reflected at the same angle... downward. Tilting the tube directs glare downward even more, away from the viewer's eyes.



WHY CBS-HYTRON SHIELDED LENS?

With this shielded lens in the electron gun, greater depth of field and better definition are achieved. Just as when you stop down the diaphragm of a large, fast camera lens (f/3.5) to a small aperture (f/16). Distortion caused by interaction of external electrostatic fields used to focus and accelerate the electron beam is avoided. Focusing is easier, less critical. Slight changes in voltages and currents do not cause drift.



WHY CBS-HYTRON BLUE-WHITE SCREEN?

Ever notice how a shirt laundered with bluing appears whiter? With the CBS-Hytron blue-white screen, whites appear whiter; blacks, blacker. Picture definition is crisper. In fringe areas, the expanded gray scale of the blue-white screen gives noticeably clearer pictures. No wonder CBS-Hytron's original blue-white screen is fast becoming the standard preferred by consumers for best definition.

CBS HYTRON RADIO TUBES

MANUFACTURERS OF RECEIVING TUBES SINCE 1921
HYTRON RADIO AND ELECTRONICS CO.
A DIVISION OF COLUMBIA BROADCASTING SYSTEM, INC.

These are just a few reasons why it's smart to demand CBS-Hytron... original studio-matched rectangulars. Try the new CBS-Hytron cylindricals yourself. Discover for yourself why 9 out of 10 leading set manufacturers pick CBS-Hytron.

MAIN OFFICE: SALEM, MASSACHUSETTS

HI-Q SERVES NATIONAL DEFENSE

Wherever Electronics Play Tag with a Plane

Guided missiles that can chase an enemy plane for miles . . . and eventually catch and destroy it . . . are just one of the many "fantastic weapons" which electronics have contributed to the defense of our nation. And here, as in all other phases of this great new science, you'll find **HI-Q** components valued for their dependable performance, long life and rigid adherence to specifications. Whether it be disk capacitors . . . tubulars, plates or plate assemblies . . . high voltage slug types . . . trimmers, wire wound resistors or choke coils . . . you can count on the **HI-Q** trade mark as a guarantee of quality in ceramic units. And you can likewise count on **HI-Q** engineers for skilled cooperation in the design and production of new components to meet specialized or unusual needs.

HI-Q TUBULAR CAPACITORS

. . . may be had with axial leads and a specially developed endseal as shown above, or with conventional leads. **HI-Q** tubulars are available in a complete range of by-pass, coupling and temperature compensating types as well as in an HVT line developed specifically for use on the relatively high pulse voltages encountered in the horizontal sweep and deflection sections of television circuits. Whatever your needs for tubular capacitors or other ceramic components, you are invited to consult **HI-Q**.

HI-Q* DIVISION

AEROVOX CORPORATION

OLEAN, NEW YORK, U. S. A.

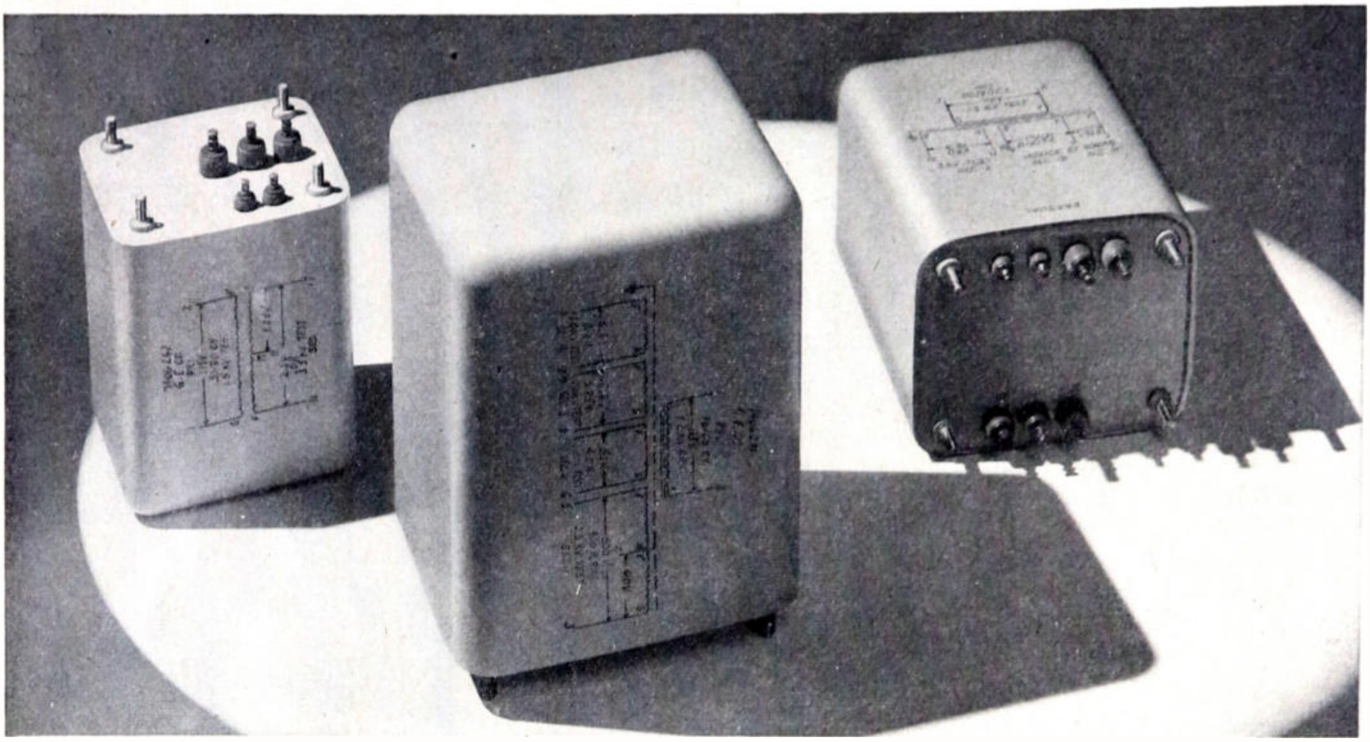
*HI-Q is a registered trademark

Export: 41 E. 42nd St., New York 17, N. Y. • Cable: AEROCAP, N. Y. • In Canada: AEROVOX CANADA LTD., Hamilton, Ont.

JOBBER ADDRESS: 740 Belleville Ave., New Bedford, Mass.

SALES OFFICES IN ALL PRINCIPAL CITIES

GE DESIGNER'S



New silicone bushings and drawn-steel case mean longer life, better seal.

New G-E hermetic transformers available for immediate shipment

Transformer covers are press-fitted to case for strength, then solder-sealed against dust and moisture on this induction heater.

Enlarged production facilities, rigid quality control mean more units built to MIL-T-27 specs

Uninterrupted supplies of General Electric's new hermetically sealed MIL-TEE transformers are helping speed production of electronic equipment to meet record military demands. These compact, newly designed units withstand extreme operating conditions. Streamlined drawn-steel cases have only one soldered seam. Tough, shockproof silicone rubber bushings effectively resist corrosion and temperature excesses.

To simplify equipment design and to reduce costs, this new line is standardized in 11 case sizes. Your G-E representative can give you full details. And to learn why these transformers more than meet MIL-T-27 Grade 1 performance requirements, send for new Bulletin GEA-5778. *General Electric Company, Schenectady 5, New York.*

GENERAL ELECTRIC

DIGEST

TIMELY HIGHLIGHTS ON G-E COMPONENTS

Rectifiers

Reactors

Transformers

Compact high-voltage components offered in wide range of ratings

G-E high-voltage components—designed for applications 5000 volts and higher where corona must be kept to a minimum—are available tailored to meet your needs.

All are oil filled and hermetically sealed to resist moisture, dirt, and dust. Conforming to MIL specs for military electronic equipment, these components are sturdily designed for reliable service under severe operating conditions, including mechanical shocks and widely varying temperatures.

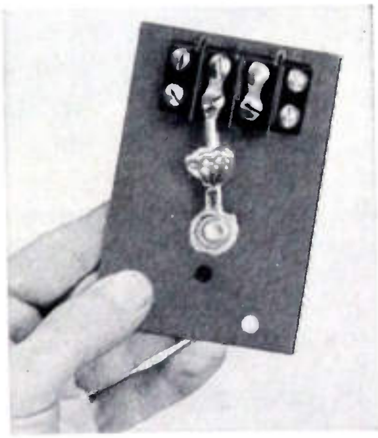
In sending your design inquiries, include all functional requirements, limiting dimensions, and expected quantities. Write to *General Electric Co., Sect. 667–20, Schenectady 5, N. Y.*

Germanium rectifiers in industrial ratings!

For use where size and weight are important, new G-E industrial germanium rectifiers offer:

- lowest forward drop per amp—for best regulation
- highest output voltage per cell
- best current output
- smallest size per watt output
- lightest weight per watt output
- instantaneous rectification

For ratings and operating characteristics, see new Bulletin GEA-5773.



New hermetically sealed relay resists breakdown

G.E.'s new hermetically sealed aircraft relay for use in exposed locations has extra protection against permanent breakdown due to voltage surges. Special polyester compound used to mold contact arms into stack insulation is non tracking, provides greater arc resistance. More powerful magnet structure yields higher tip pressures for surety of make. Rated 28 volts d-c, 3 amp. See Bulletin GEA-5729.

EQUIPMENT FOR ELECTRONIC MANUFACTURERS

A partial list of the thousands of items in the complete G-E line. We'll tell you about them each month on these pages.

Components

Meters and instruments	Timers
Capacitors	Indicating lights
Transformers	Control switches
Pulse-forming networks	Generators
Delay lines	Selsyns
Reactors	Relays
*Thyrite	Amplidynes
Motor-generator sets	Amplistats
Inductrols	Terminal boards
Resistors	Push buttons
Voltage stabilizers	Photovoltaic cells
Fractional-hp motors	Glass bushings
Rectifiers	Dynamotors

Development and Production Equipment

Soldering irons
Resistance-welding control
Current-limited high-potential tester
Insulation testers
Vacuum-tube voltmeter
Photoelectric recorders
Demagnetizers

*Reg. trade-mark of General Electric Co.

General Electric Company, Section B 667-20
Schenectady 5, New York

Please send me the following bulletins:

Indicate: √ for reference only
X for planning an immediate project

☐ GEA-5729 Hermetically sealed Relays
☐ GEA-5773 Germanium Rectifiers
☐ GEA-5778 MIL-T-27 Transformers

Name ....
Company ....
City .... State ....


STACKPOLE
Fixed Composition Resistors in accordance with
JAN-R-11
specifications
Electronic Components Division
STACKPOLE CARBON COMPANY, St. Marys, Pa.
RC10
RC20
RC21
RC30
RC31
RC41
RC42
Insulated types. Write for Bulletin for complete details.
A DEPENDABLE SOURCE OF RESISTOR SUPPLY for over 20 YEARS

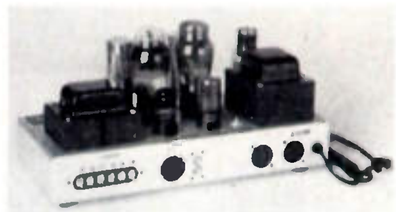
News—New Products

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

(Continued from page 10A)

Power Amplifier

The new Type 220-A laboratory power amplifier, a versatile and inexpensive amplifier for both laboratory and high-fidelity music reproduction operation is available from **Hermon Hosmer Scott, Inc.,** 385 Putnam Ave., Cambridge 39, Mass. Among its features are a rated power output of 20 watts, and a frequency response flat from 12 cps to 55,000 cps



The specifications are: Harmonic distortion less than 0.5 per cent at full 20 watts output; first order difference-tone intermodulcation component less than 0.1 per cent at full rated peak output; hum level minus 90 db below full output; input for full rated 20 watt output, 0.5 volts on low level input, 1.5 volts on high level input; input impedance 0.5 megohms for low level input, 1.5 megohms for high level input. The unit also has an input level adjustment. Free bulletin on request.

Hand Calculator

The Curta Calculator, a new portable calculator for men who work with figures on formulae is announced by **Curta Calculator Co.,** 5543 S. Ashland Ave., Chicago, Ill.

Precision-made by Swiss watchmakers, the calculator will add, subtract, multiply, divide, figure square roots, factors, cubes and percentages, and is as easy to operate as a slide rule. Quite accurate, the new device carries to five decimal places and totals to 99 billion. Figures can be checked and rechecked on 3 sets of dials.

It weighs 8 ounces, and is claimed to do as much as electric desk calculators costing up to $800.00. The price of the device is $129.00 which includes a rubber-lined metal case.

(Continued on page 22A)

News—New Products

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

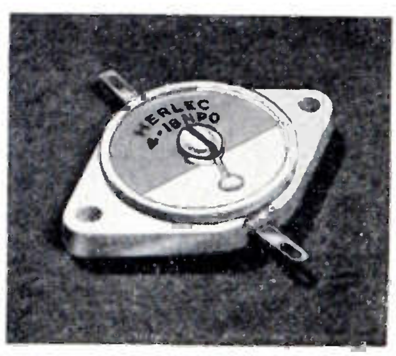
(Continued from page 18A)

Deposited Carbon and Boron Carbon Resistors

International Resistance Co., 401 N. Broad St., Philadelphia 8, Pa., has announced a new product to be produced on the nation's first mechanized assembly line for deposited-carbon and boron-carbon resistors. Physically alike except in color, both types of resistors provide higher resistance values in less space, and at a lower cost than wire wound precision types.

As an added feature, the boron-carbon resistor provides a greater degree of temperature stability. Both units are conservatively rated at ½ watt, with a body length of 9/16 inch. Diameter of the outside caps is 5/32 inch. Applications: military electronic equipment, radar, fire control instruments, and meter multipliers. Write to International Resistance Co.; for catalog bulletins B-6 (BOC) and B-7 (DCC).

Trimmer Capacitors



Ceramic trimmer capacitors for use in circuit applications where stability of capacitor characteristics is of importance are a new product of the **Herlec Corp.,** 422 N. Fifth St., Milwaukee 3, Wis., a wholly-owned subsidiary of the **Sprague Electric Co.** Full details on Sprague-Herlec Type AO8 capacitors are available on business letter head request for Engineering Bulletin 604 to either Sprague or to Herlec.

(Continued on page 118A)

IT'S PENNY WISE

and

POUND WISE

to use

ROBINSON ENGINEERED (MET·L·FLEX) MOUNTING SYSTEMS FOR VIBRATION CONTROL

Savings *start* with the *design*. A Robinson engineered mounting system is designed for a specific piece of equipment *and* the conditions under which it must operate. It is not just a combination of a standard tray suspended on stock unit mounts, with potential misalignment and attachment problems.

Savings add up through model, prototype test and production stages. Since you are assured of permanent protection, your engineers can use less rugged components—often saving up to 20% of equipment weight and cost—yet gain better equipment performance.

You save even more directly. The cost of a Robinson engineered system is often less than the total cost of unit mounts plus attachment tray—even when the extra assembly costs they entail are ignored.

During the last ten years every major electronic company, airframe manufacturer, airline and branch of the military service has called on Robinson to help solve some complex problem of vibration control. From these cooperative efforts have come many design "firsts" and basic shock mount improvements. Engineers everywhere have found that when *ounces* and *dollars* count—it pays to call on Robinson; to use the experience that comes only from years of specialized engineering.

Robinson engineered mounting systems, with their exclusive MET-L-FLEX elements, provide maximum vibration and shock protection at any altitude, in any part of the world—and do it *permanently*.

Start saving time and money today. Call on your nearest Robinson engineering representative.

Save on Design

Give your mounting problems to experts in vibration control. Since 1942, Robinson has pioneered many new and effective air-borne mounts, including the first all metal design. Their efficient, production wise designs cut your development costs.

Save on Installation

Use complete MET-L-FLEX mounting bases to prevent misalignment and possible malfunctioning. Instead of the 16 mounting holes and drilling template required by unit mounting bases, Robinson engineered bases have 4 holes, all accurately prespaced.

Save Replacement Costs

Eliminate the usual servicing and replacement expenses. Robinson mounting systems never wear out; never rust; never weaken. They always deliver the same unvarying performance, regardless of environmental or operating conditions; aging or extremes of temperature.

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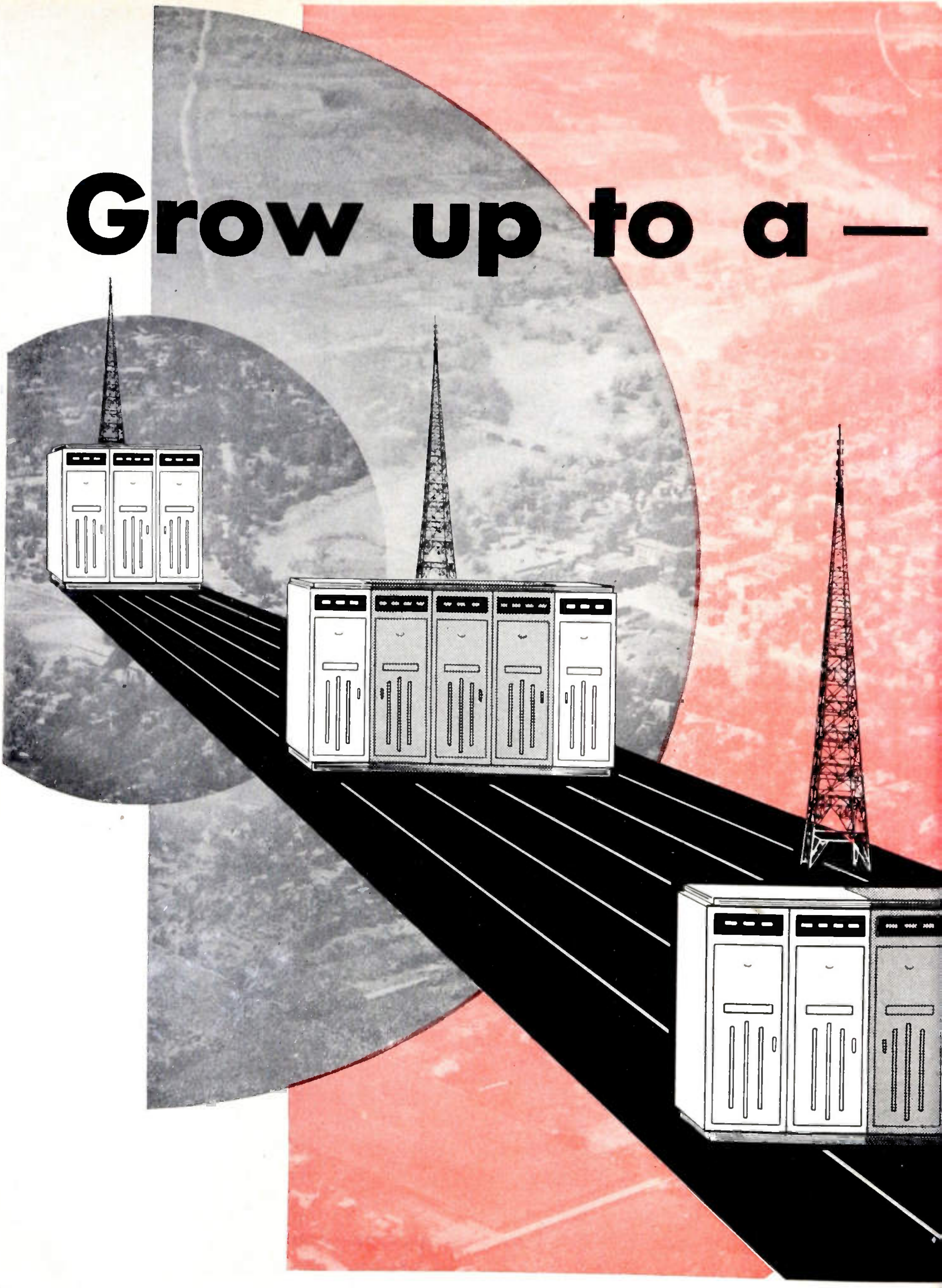
ROBINSON AVIATION INC.
TETERBORO, NEW JERSEY
Vibration Control Engineers

"SEA LEVEL PERFORMANCE AT ANY ALTITUDE"


Grow up to a —

NEW! High-level, direct-reading portable test sets simplify laboratory and field SHF work

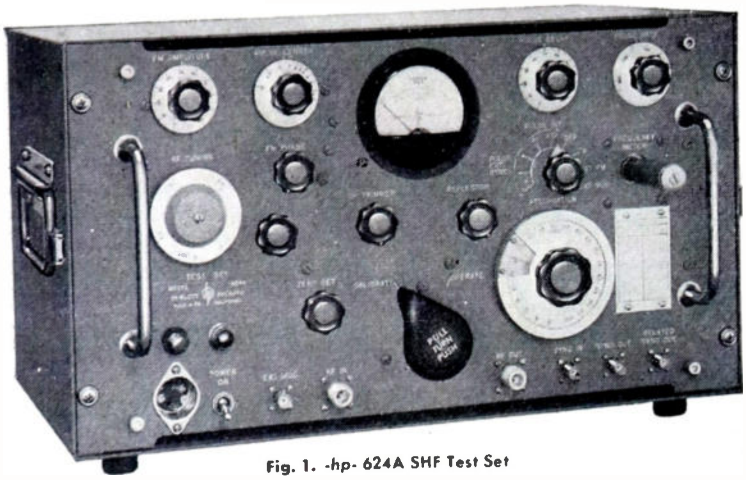


Fig. 1. -hp- 624A SHF Test Set

.223 v. maximum rf output
Direct tuning, reading
Pulse and fm modulated
Stable, accurate 100 db attenuator
Measures external rf power
Measures external frequency
Compact, sturdy, portable

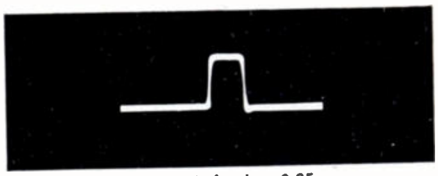


Fig. 3. Typical rf pulse, 0.25 μsec.

SPECIFICATIONS

-hp- 624A SHF Test Set

RANGE: 8,500 to 10,000 mc.

OUTPUT: 0 dbm (1 mw) to −100 dbm into 50-ohm load. Type N jack.

OUTPUT ACCURACY: Within 2 db, −10 to −100 dbm into matched load.

INTERNAL MODULATION: Pulsed or fm.

PULSE MODULATION: Length variable from approx. .25 to 10 μsec. Rise and fall times, each, 0.05 μsec. Rate variable 35 to 3,500 pps.

EXTERNAL SYNC: Internal pulser operates free-running or in sync with external 5-v. peak pulse, pos. or neg., or 5-v. rms. sine waves. May be externally square-wave modulated. BNC jack.

FM: Internal fm at power line frequency. ±7.5 mc deviation max. Also fm modulation by external 35 to 3,500 cps voltages.

TRIGGER PULSES: (a) Coincident with start of output rf pulse; (b) 3 to 250 μsec ahead of output rf pulse.

POWER METER: 2 mw full scale. Accurate within 1 db.

FREQUENCY METER: Full range, accurate within 0.03% at 25°C ambient.

PRICE: $2,250.00 f.o.b. factory.

-hp- 623B SHF Test Set

TOTAL FREQUENCY RANGE: 5,925 to 7,725 mc.

INDIVIDUAL KLYSTRON RANGES:

5,925–6,225	6,575–6,875	7,125–7,425
6,125–6,425	6,850–7,150	7,425–7,725

OUTPUT: 0 dbm (1 mw) to −70 dbm into 50-ohm load. Direct-reading control.

OUTPUT ACCURACY: Within 2 db, 0 to −70 db, into matched load.

INTERNAL MODULATION: FM from 1,000 cps internal source; phase, deviation adjustable; max. deviation ±15 mc.

EXTERNAL MODULATION: FM, 50 cps to 10 kc. May be pulsed or square-waved externally.

DETECTOR OUTPUT: Xtal detector to provide rectified output when fm or pulsed power applied.

(Other specifications similar to 624A)

PRICE: $1,500.00 (includes one klystron)

Data subject to change without notice

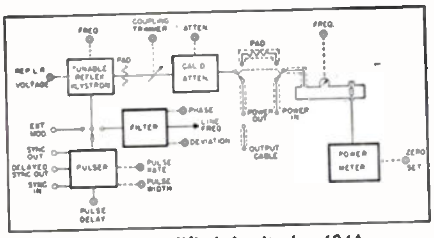


Fig. 2. Simplified circuit, -hp- 624A

-hp- 624A SHF Test Set is a high-level, accurate, multi-purpose instrument designed to speed and simplify a wide variety of tests between 8,500 and 10,000 mc. It is an ideal one-piece unit for measuring receiver sensitivity or selectivity, transmitter tuning or power level, and is particularly adapted to testing complete radar or gunfire control systems or beacon equipment. The instrument includes pulsing circuitry providing a variety of high-quality rf pulses.

-hp- 624A consists of a signal generator and a power and frequency meter section. The generator includes a modern klystron generator with excellent frequency stability and an output attenuator of the waveguide-beyond-cutoff type, insuring high accuracy and stability. The attenuator is not subject to temperature, humidity or age changes. The power and frequency meter section can be used to adjust the signal generator's frequency and level as well as measure external rf energy. The instrument employs 50-ohm Type N coaxial connectors, but for maximum versatility includes adaptors for waveguide connection.

-hp- 623B Test Set is designed for operation at frequencies between 5,925 mc and 7,725 mc. This overall frequency range is covered in six bands, each of which is a full 300 mc wide. Bands are selected by installation of the proper klystron tube (see specifications). The instrument is particularly useful in field-testing SHF radio relay stations and communications equipment as well as general tests involving fm modulated equipment. It includes a 1,000 cps modulator and may also be square-waved or pulsed by external sources with frequencies ranging from 60 cps to 100 kc.

Both *-hp-* 624A and 623B weigh less than 60 pounds, are of extra-sturdy construction and are equipped with carrying handles and snap-on cover. Sets also fit standard relay racks.

See your -hp- engineer-salesman or write direct for complete data.

ELECTRONIC MEASURING INSTRUMENTS

HEWLETT-PACKARD COMPANY 2444-D PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U.S.A.
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LITTON INDUSTRIES NEWS

Litton Model 3900 Thermopile

LITTON THERMOPILE WITH STANDARD METER FORMS ACCURATE, LOW COST INDICATOR FOR SMALL DIFFERENTIAL TEMPERATURES

Engineers in increasing numbers are using Litton Model 3900 Thermopile in conjunction with microwave water loads to measure rf power, and in cooling systems to monitor temperature changes.

The Thermopile has 30 pairs of copper-constantan junctions, tapped at 10 and 20 pairs. Junctions protrude into a fluid flow channel milled in a plastic block to which water fittings are mechanically attached. The plastic block is encased in a cast aluminum housing. Binding posts are provided for electrical connection, and ¼" Uniflare fittings for water connection. Internal resistance is approximately 6 ohms.

With rf water loads using appropriate water flow, meter sensitivity and number of junctions, average powers from 10 watts to several kilowatts can be measured conveniently and accurately. For lower power levels, several thermopiles can be used in series.

The 30-junction thermopile generates approximately 1 millivolt per °C differential temperature. To determine water flow rate and indicating meter, the following formula is useful:

(P = power dissipated in watts; Q = water flow in gals. per minute; R = meter internal resistance in ohms; M = meter sensitivity in millivolts for full-scale deflection.)

For full-scale meter deflection, approximately:

$$250M\frac{(R+6)}{R}=\frac{P}{Q}$$

Also, to avoid excessive heat losses, differential temperature should not exceed 20°C, where for pure water

$$T=\frac{P}{246Q}$$

(T being temperature differential in °C.)

Because of stray losses in plumbing and the load, the system is best calibrated by direct dissipation of metered power in a water-cooled resistor in series with the water load.

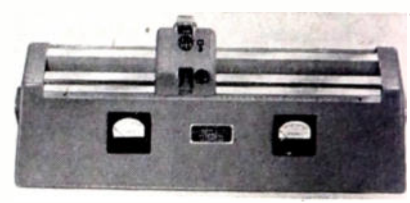
Time of response in minutes is determined by the volume of the system V in gallons divided by Q. (Time constant of thermopile is negligible.) For a typical installation of Litton Model 4000 U-Line, Model 4100 Water Load and Model 3900 Thermopile, operating at the kilowatt level, using a meter with M = 7 millivolts, R = 71 ohms, time of response is approximately 20 seconds. Litton Model 3900 Thermopile, price $75.

WATER LOAD

Litton Model 4100 Water Load is a termination for 1⅝", 50-ohm coaxial lines, and is particularly useful in high-power applications where power output must be accurately measured. The Load is conservatively rated at 2 kilowatts capacity, 950 to 3,000 mc/sec. VSWR is less than 1.2 over full range, less than 1.1 above 2,000 mc/sec. The equipment includes two adjustable-depth probes for sampling rf power. Model 3900 Thermopile is recommended for use with this load. Model 4100 Water Load, price $425.

U-LINE AND STUB COMBINATION

Litton Model 4000 U-Line offers convenience and accuracy in quickly determining VSWR in high- or low-power coaxial lines. The equipment transduces power from a 1⅝" coaxial line to a U-shaped configuration with a rigid central and outer conductor. A traveling probe moves on a precision carriage through the open end of the "U." A 500-millimeter scale with vernier indicates probe position.



Litton Model 4000 U-Line

Model 4000 U-Line offers continuous frequency coverage from 450 to 2,750 mc/sec. with insertion VSWR of less than 1.05. Teflon bead supports permit a CW power rating of 2 kilowatts. Mounting holes are provided for meters. Price $700.

STUB

For use with Model 4000 U-Line. Permits rapid insertion, variation of phase position and withdrawal of mismatch of known VSWR in the U-Line. Calibrated scale permits insertion of known VSWR up to 2.0, at frequencies 950 to 2,750 mc/sec. Thus, equipment may be used as a calibrated mismatch or matching device.

Insertion at any phase position is possible with relative phase readable on millimeter scale on the U-Line.

Model 4200 Stub is a metallically-loaded Teflon rod contoured to fit the U-Line. The stub is suspended from a carriage riding on the U-Line. Price, $100.

Data subject to change without notice. All prices f.o.b. San Carlos, Calif.

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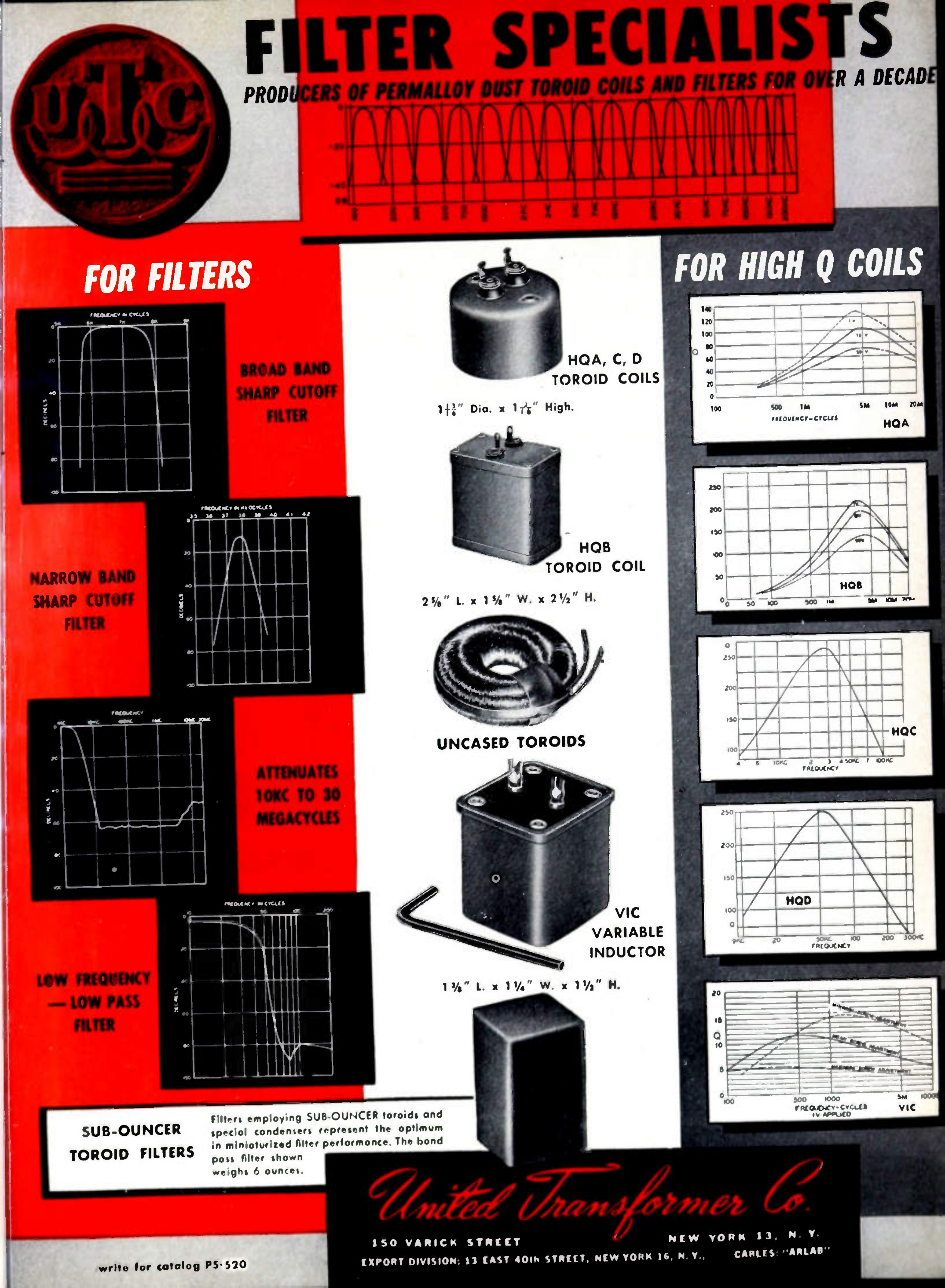
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EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y., CABLES: "ARLAB"
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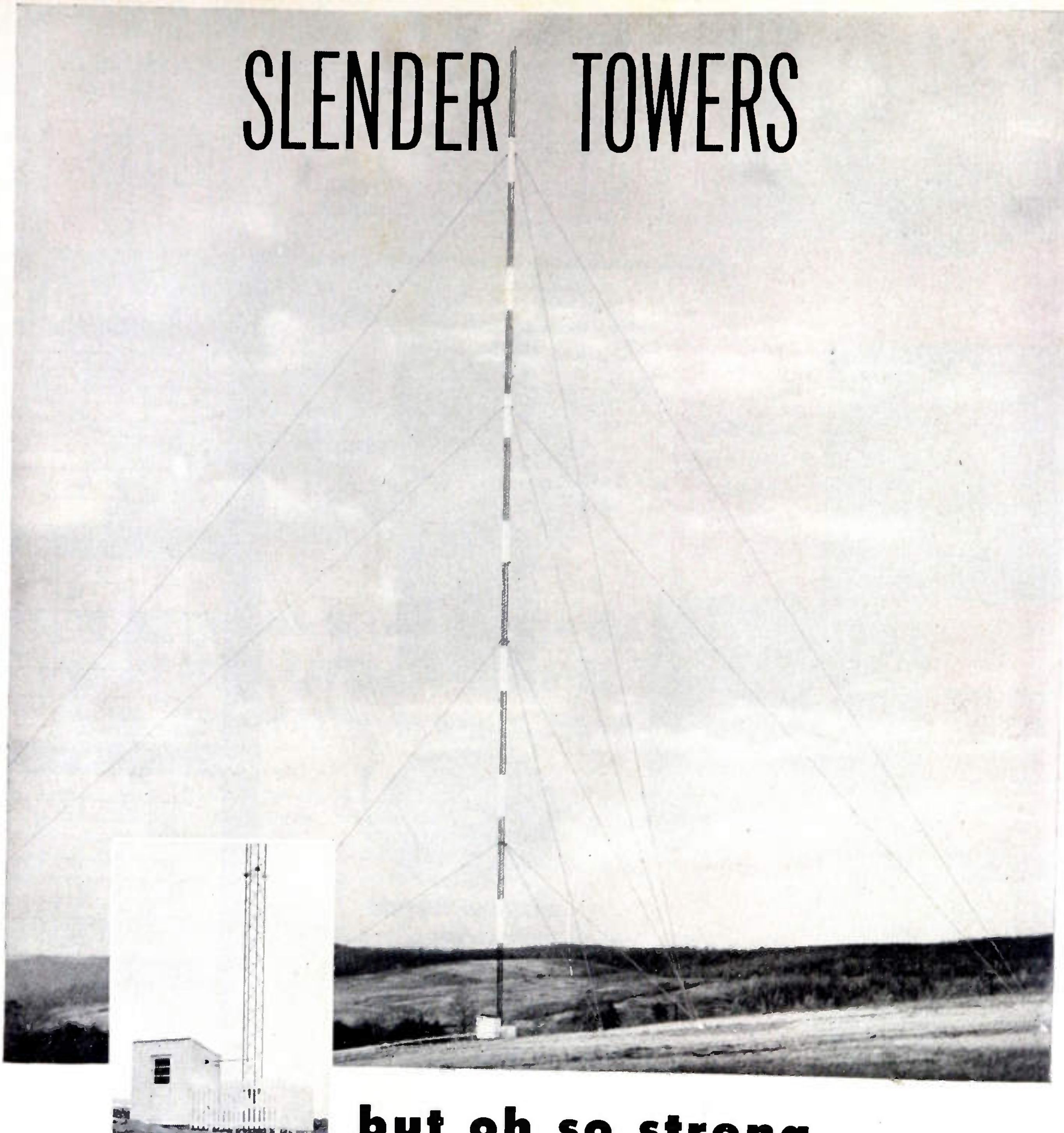
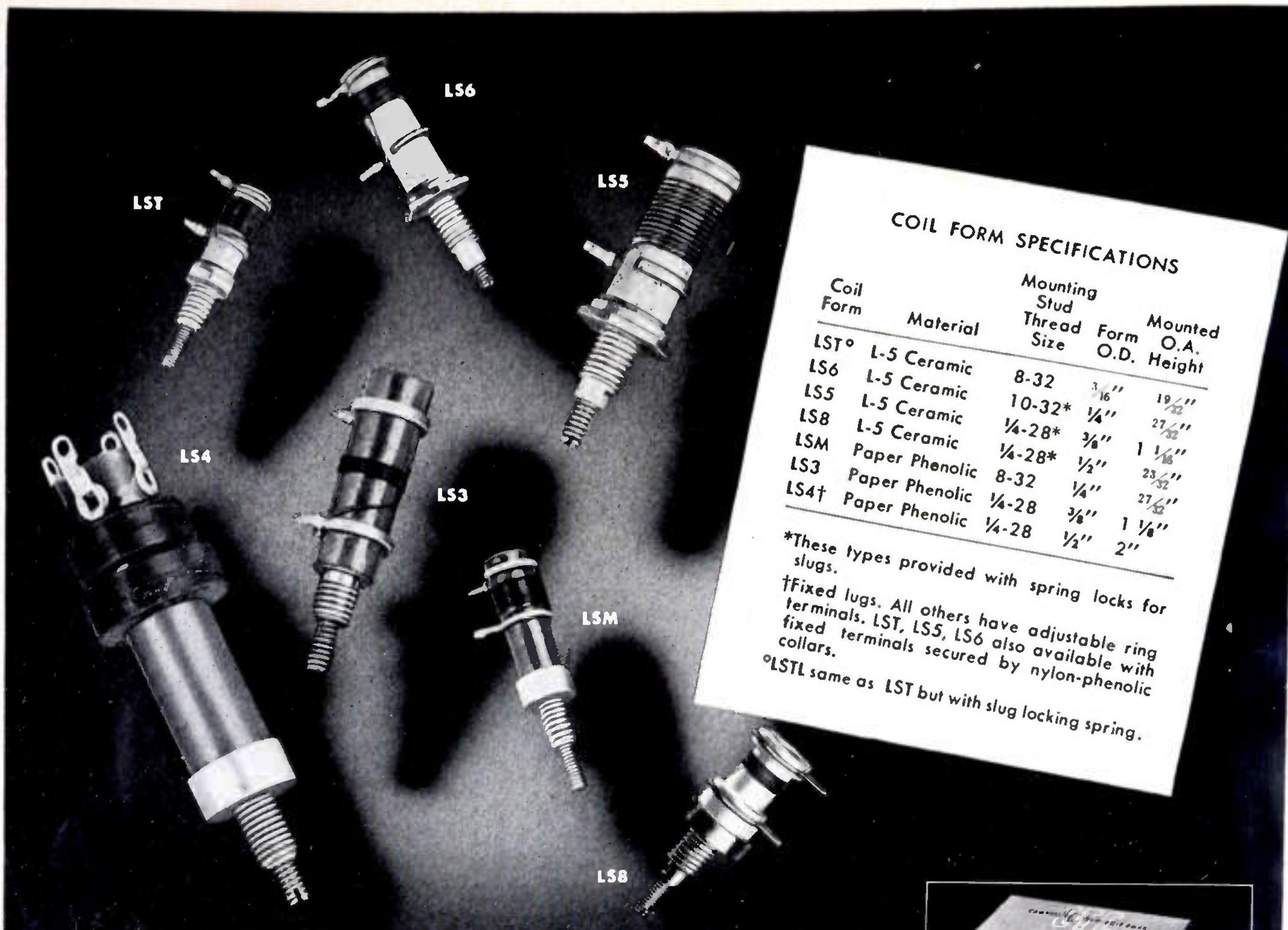

SLENDER TOWERS



COIL FORM SPECIFICATIONS

Coil Form	Material	Mounting Stud Thread Size	Form O.D.	Mounted O.A. Height
LST°	L-5 Ceramic	8-32	3/16"	19/32"
LS6	L-5 Ceramic	10-32*	1/4"	27/32"
LS5	L-5 Ceramic	1/4-28*	3/8"	1 1/16"
LS8	L-5 Ceramic	1/4-28*	1/2"	23/32"
LSM	Paper Phenolic	8-32	1/4"	27/32"
LS3	Paper Phenolic	1/4-28	3/8"	1 1/8"
LS4†	Paper Phenolic	1/4-28	1/2"	2"

*These types provided with spring locks for slugs.

†Fixed lugs. All others have adjustable ring terminals. LST, LS5, LS6 also available with fixed terminals secured by nylon-phenolic collars.

°LSTL same as LST but with slug locking spring.

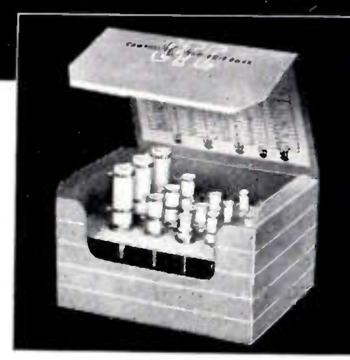
Here are the coils you want ...*the way you want them!*

Take advantage of one of C.T.C.'s most popular and useful services . . . the winding of slug tuned coils to exact specifications. Single layer or pie types furnished. You can be sure your specs—military or personal—will be faithfully followed to the last detail of materials and methods, and with expert workmanship.

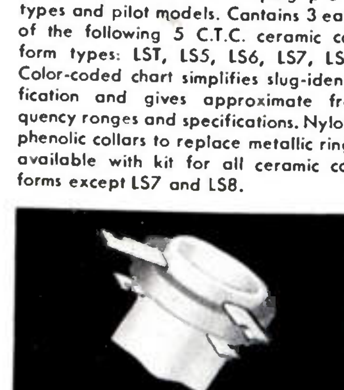
C.T.C. coil forms are made of quality paper base phenolic or grade L-5 silicone impregnated ceramic. Mounting bushings are cadmium plated brass and ring type terminals are silver plated brass. Terminal retaining collars of nylon-phenolic also available in types LST, LS5, LS6.

Wound units can be coated with durable resin varnish, wax or lacquer. Both coils and coil forms are furnished with slugs and mounting hardware — and are obtainable in large or small production quantities. Be sure to send complete specifications for specially wound coils.

All C.T.C. materials, methods, and processes meet applicable government specifications. For further information on coils, coil forms or C.T.C.'s special consulting service, write us direct. *This service is available to you without extra cost.* Cambridge Thermionic Corporation, 456 Concord Avenue, Cambridge 38, Mass. West Coast manufacturers, contact: E. V. Roberts, 5068 W. Washington Blvd., Los Angeles 16, and 988 Market Street, San Francisco, California.



NEW CERAMIC COIL FORM KIT. Helps you spark ideas in designing electronic equipment or developing prototypes and pilot models. Contains 3 each of the following 5 C.T.C. ceramic coil form types: LST, LS5, LS6, LS7, LS8. Color-coded chart simplifies slug-identification and gives approximate frequency ranges and specifications. Nylon-phenolic collars to replace metallic rings available with kit for all ceramic coil forms except LS7 and LS8.



NEW NYLON-PHENOLIC COLLARS. Terminals held securely; soldering spaces doubled; excellent for both bifilar and single pie windings. Show an increase in Q and many new benefits over metallic rings — without impairing in any way the moisture- and fungus-resistant qualities of coil form assemblies.

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CTC

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Erie High Voltage TV By-Pass Capacitors

ERIE STYLES 412 and 414 High Voltage Ceramicons are designed primarily to supply high voltage power supply filtering for television receivers. Conservative designing has been followed by months of proving-in tests before placing these new styles on the market.

Style 412, rated at 20 KV, and Style 414, rated at 10 KV, are both available in various combinations of terminals, threaded internally, externally, and plain; and are made to individual manufacturer's requirements.

Case insulation, of low loss molded thermosetting plastic, provides a moisture seal of proven superiority. Ring convolutions, molded into the surface of Style 412, assure a positive check against surface leakage resulting from conductive deposits in ordinary handling. Effective creepage path is at the same time increased in length by more than 14%.

SPECIFICATIONS

	Style 412	Style 414
Capacitance: at 1KC, 1 to 5 volts rms.	500 mmf +50%; −20%	500 mmf +50%; −20%
Power Factor: at 1KC, 1 to 5 volts rms.	1.5% max.	1.5% max.
Leakage Resistance:	50,000 meg.Ω min.	50,000 meg.Ω min.
Dielectric Strength:	30,000 D.C.	18,000 D.C.
Life Test, 1000 hours at 85°C.	21,000 D.C.	12,000 D.C.
Rated Voltage:	20,000 D.C.	10,000 D.C.
External Creepage Path:	2 Inches	1 1/4 Inches

10 KV

ACTUAL SIZE

Style 414
10 KV
500 MMF

20 KV

Style 412
20 KV
500 MMF

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Sensitive Wide Band Electronic Voltmeter

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 . . .
6 mmfds shunted by 11 megs

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

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- Same accuracy at *ALL* points on a logarithmic voltage scale and a 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.
- Provides a 60 DB amplifier flat within 1 DB from 50 cycles to 6 MC.

Specifications on other Ballantine Electronic Voltmeters

MODEL	FREQUENCY RANGE	VOLTAGE RANGE	INPUT IMPEDANCE	ACCURACY	PRICE
300	10 to 150,000 cycles	1 millivolt to 100 volts	1/2 meg. shunted by 30 mmfds.	2% up to 100 KC 3% above 100 KC	$210.
302B Battery Operated	2 to 150,000 cycles	100 microvolts to 100 volts	2 megs. shunted by 8 mmfds. on high ranges and 15 mmfds. on low ranges	3% from 5 to 100,000 cycles; 5% elsewhere	$225.
305	Measures peak values of pulses as short as 3 microseconds with a repetition rate as low as 20 per sec. Also measures peak values for sine waves from 10 to 150,000 cps.	1 millivolt to 1000 volts Peak to Peak	Same as Model 302B	3% on sine waves 5% on pulses	$280.
310A	10 cycles to 2 megacycles	100 microvolts to 100 volts	Same as Model 302B	3% below 1 MC 5% above 1 MC	$235.

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

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

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The following types are available in production quantities at Newton and Chicago, and in smaller quantities at our 400 Special Tube Distributors.

DIODES SHOWN TWICE SIZE

	CK705 General Purpose	CK706 Video Detector	CK707 50 V. dc Restorer	CK708 100 V. dc Restorer	CK709 Bridge Rectifier	CK710 UHF Mixer	CK711 Bridge Rectifier	CK712 200 V. dc Restorer	CK713 Computer Diode	CK715 Frequency Multiplier	1N67 High Back Resistance
MAXIMUM RATINGS (at 25°C.)											
DC Inverse Voltage (volts)	60	40	80	100		5		200	75		80
Average Rectified Current (ma.)	50	35	35	35		50		22.5	50		35
Peak Rectified Current (ma.)	150	125	100	100		150		70	150		100
Surge Current (for 1 sec.) (ma.)	500	300	500	500		500		250	500		500
Ambient Temperature for all types	–50°C to +100°C for all types										
CHARACTERISTICS (at 25°C.)					Four matched type CK705 diodes		Four matched type CK707 diodes			Low impedance frequency multiplier for UHF television	
Max. Inverse Current at –0.6 volts (ma.)						0.2					
Max. Inverse Current at –5 volts (ma.)			0.008								0.005
Max. Inverse Current at –10 volts (ma.)	0.05										
Max. Inverse Current at –40 volts (ma.)									0.25‡		
Max. Inverse Current at –50 volts (ma.)	0.8		0.10								0.05
Max. Inverse Current at –100 volts (ma.)				0.625							
Max. Inverse Current at –200 volts (ma.)								0.8			
Min. Forward Current at +0.5 volts (ma.)						3.0					
Min. Forward Current at +1 volt (ma.)	5.0		3.5	3.0				2.0			4.0
Min. Forward Current at +2 volts (ma.)									21.0‡		
Min. DC Reverse Voltage for Zero Dynamic Resistance (volts)	70.0	50	100.0	120.0		10.0		225.0	75.0‡		100.0
Shunt Capacitance (uuf), average	1.0		1.0	1.0		1.7		1.0	1.0		1.0
Rectification Efficiency at 54 mc (approx.%)		60									
Oscillator injection current (ma.)						0.75*					

*Conversion loss at 500 mc. and noise factor comparable with 1N21B. ‡at 50° C.

*For several years, Raytheon Germanium Diodes have been made from "Single Crystal" germanium.

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TRANSISTORS

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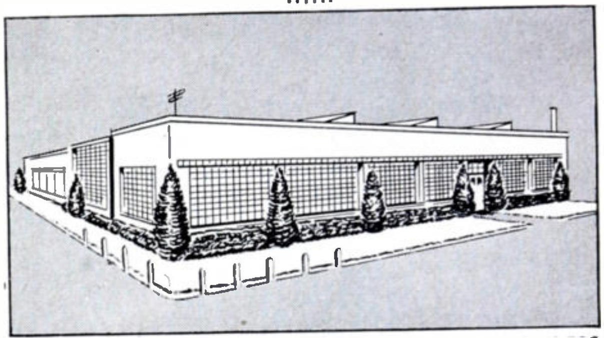
Transistors have several advantages over other components. These include small size, no cathode power or warm-up time required, very high efficiencies, long life, ruggedness, stability.

Uses are limited today, however, by factors like frequency response (usually below 1 megacycle) and temperature effects (usable at temperatures only slightly above normal ambients at present). Both of these problems are being actively studied.



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GENERAL ELECTRIC

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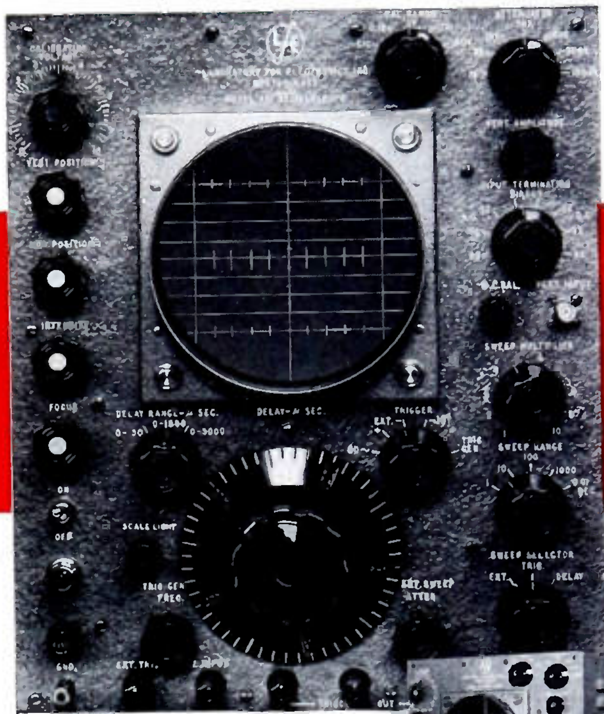
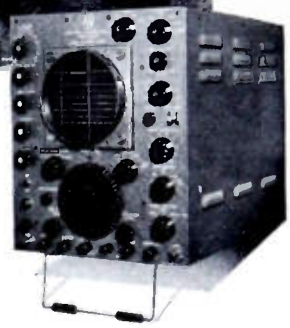
for precise, quantitative studies of pulse waveforms, transients and other high or low speed electrical phenomena



LFE Model 401 Oscilloscope . . .
A high gain, wide band, versatile, general purpose instrument

Advances in electronics have placed greater demands on the time, frequency, and amplitude measuring capabilities of laboratory oscilloscopes. LABORATORY FOR ELECTRONICS, INC., recognizing the ever-increasing requirements of the rapidly expanding electronics industry, and using specifications set forth by electronic engineers, has developed the Model 401 oscilloscope to provide the features and conveniences required in a medium price, general purpose instrument.



SPECIFICATIONS

Y-Axis

Deflection Sensitivity — 15 millivolts peak-to-peak/cm

Frequency Response — DC to 10Mc

Transient Response — Rise Time — 0.035 microseconds

Signal Delay — 0.25 microseconds

Input line terminations — 52, 72, or 93 ohms, or no termination, for either AC or DC input

Calibrating Voltage — 60 cycle square wave.

Input Imp. — 1 megohm, 30 mmf.

X-Axis

Sweep Range — 0.01 sec/cm to 0.1 microseconds/cm

Delay Sweep Range — 5-5000 microseconds in three ranges — continuously adjustable

Triggers — Internal or External, + and −, or 60 cycles, or delayed trigger outputs are available at suitable binding posts.

Built-in trigger generator for triggering external circuits and sweeps.

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Low capacity probe

Functionally colored control knobs conveniently grouped

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Facilities for mounting oscilloscope cameras

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New

A High Quality Line of CERAMIC CAPACITORS

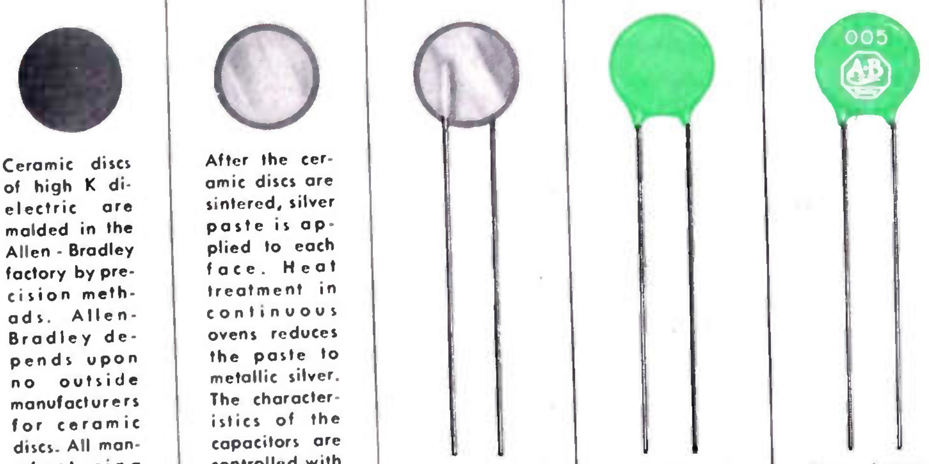
After long research Allen-Bradley has developed a high quality line of ceramic capacitors. Every step in the manufacturing process from making the high K dielectric discs to the final impregnation of the finished capacitor is performed in the Allen-Bradley plant by expert operators using highly specialized production equipment.

These disc-type ceramic capacitors are available in capacities ranging from 0.001 to 0.01 microfarads.

Allen-Bradley ceramic capacitors have been approved by the engineering departments of the largest electronic, electrical, and telephone laboratories.

Samples for qualification tests and type approval will be supplied upon request.

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



Ceramic discs of high K dielectric are molded in the Allen-Bradley factory by precision methods. Allen-Bradley depends upon no outside manufacturers for ceramic discs. All manufacturing processes are A-B controlled throughout.

After the ceramic discs are sintered, silver paste is applied to each face. Heat treatment in continuous ovens reduces the paste to metallic silver. The characteristics of the capacitors are controlled with great accuracy during their manufacture.

Leads are soldered to silver surfaces.

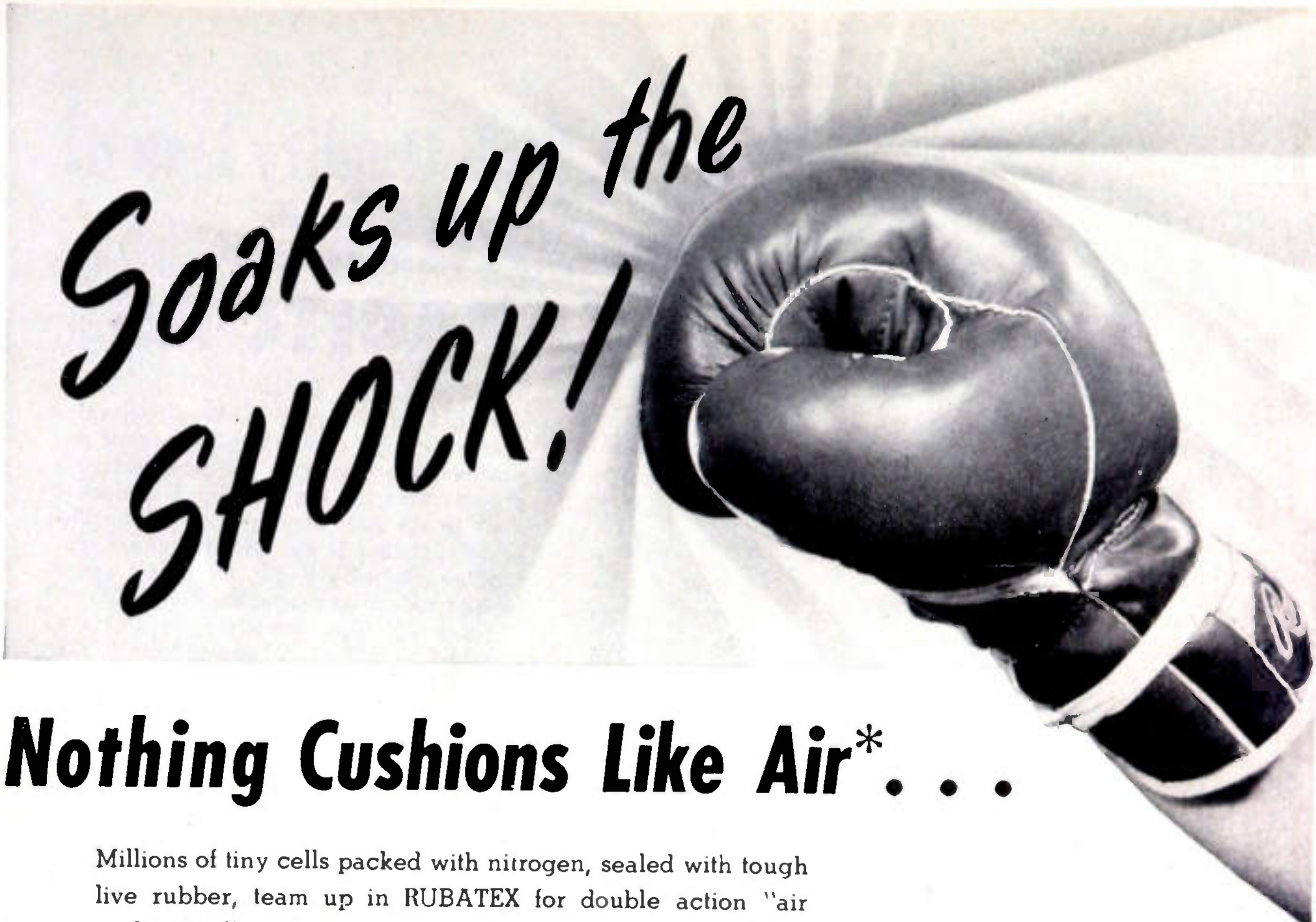
Capacitor is insulated with phenolic resin.

Wax impregnated to resist moisture.

The above five panels show the successive steps in the manufacture of Allen-Bradley high quality ceramic capacitors.

ALLEN-BRADLEY
RADIO & TELEVISION COMPONENTS

QUALITY



Nothing Cushions Like Air*. . .

Millions of tiny cells packed with nitrogen, sealed with tough live rubber, team up in RUBATEX for double action "air cushioning" — quick recovery when compressed — ability to "soak up the shock" again and again!

It's the reason why shippers of fragile goods and manufacturers of delicate scientific instruments pack their products in RUBATEX — protective insurance against any breakage or damage in transit.

If your product has an "air cushioning", shock absorbing, or vibration damping application — check the double action advantage of RUBATEX!

* Pure air oxidizes rubber. Rubatex uses nitrogen — all the advantages of air without its harmful effect.

ADDITIONAL RUBATEX QUALITIES

- SOFT
- LIGHT WEIGHT
- WATERPROOF
- BUOYANT
- INSULATOR
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- LONG LIFE

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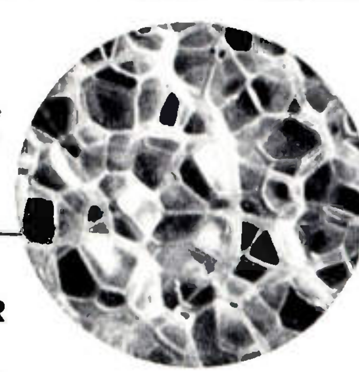
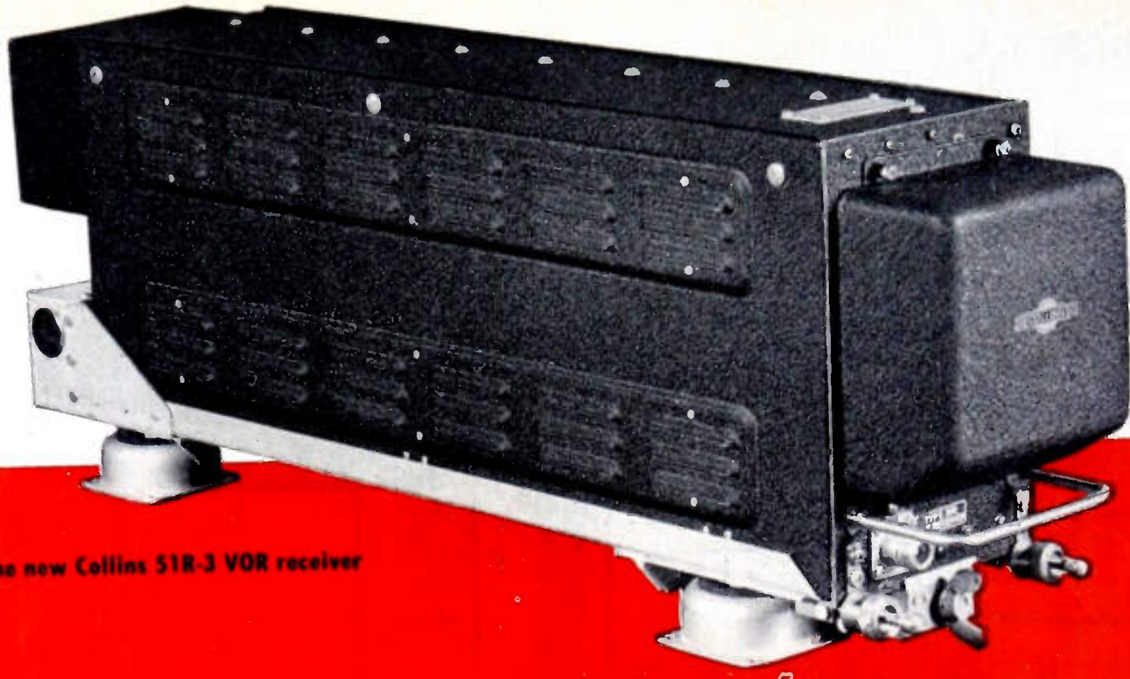
Closed Cells (magnified section at right) are responsible for Rubatex's special properties which are not possessed by ordinary sponge rubber having open cells.



RUBATEX CLOSED CELL RUBBER



The new Collins 51R-3 VOR receiver

The leader in VOR* Leads again!

It was 1946 when Arinc tossed the omnirange ball into the radio industry's court. Collins has been throwing baskets ever since.

First demonstrations of the Collins 51R airborne VOR receiver and instrumentation were made in January, 1947, for Arinc, commercial airlines engineers, and the Air Transport Association's research group. Collins designs were approved.

Within a year, orders for 51R equipment were received from American, Chicago & Southern, Northwest, Pan American and United. Since then almost every leading airline of the United States (and, most recently, Air France), as well as many users of executive aircraft, have adopted the 51R as standard.

Collins has earned and is widely accorded the leadership in the VOR field and today, by a wide margin, is the largest producer of airborne VOR receivers and accessories.

* Visual omnirange VHF aircraft navigation system.

The 51R research-engineering team has never stopped playing ball — studying users' servicing and performance reports — engineering refinements and incorporating them in 51R design. The 51R-2 was a prime example of the value of this continuing effort toward the ultimate.

We now announce the 51R-3, which represents the latest advances in performance and ease of servicing.

In general, the changes include, a: greatly improved stability over wider ranges of temperature and climatic conditions, and b: potentiometer type adjustments for all major instrument circuits, and component location changes which save much bench test time.

There is a 51R-3 illustrated descriptive bulletin waiting for you. Please ask us for it on your business letterhead.

In radio navigation equipment, it's . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

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And now **ANTARA CHEMICALS** presents

FERROMAGNETIC POWDER "J"

for HIGHER frequencies

This powder is made from a new alloy — by the same carbonyl process which has already furnished a number of widely used ferromagnetic powders.

"J" Powder was developed in our laboratories — designed for high Q cored coils at VHF. It has the lowest losses for its relatively high permeability. Its properties compare favorably with those for the long-established Type SF. (Note the graphs on the left-hand page. These are not included in the Manual described beneath the graphs.)

Here are approximate comparisons between "J" Powder and Type SF . . . . .

Permeability: same as SF (packing fraction being equal) or 6% higher than SF (densities being equal). *Q Values:* above 30 mc: equal or better than SF. *Loss factors:* eddy current — lower than SF: after-effect and hysteresis — higher than SF, TH or E. *Particle density:* slightly lower than SF. *Apparent density:* slightly lower than SF. *Compressibility:* same as SF. *Density ratio:* same as SF. *Stabilities* against temperature changes, humidity, long time periods, magnetic shock and chemicals: excellent, as with all G A & F Carbonyl Iron Powders.

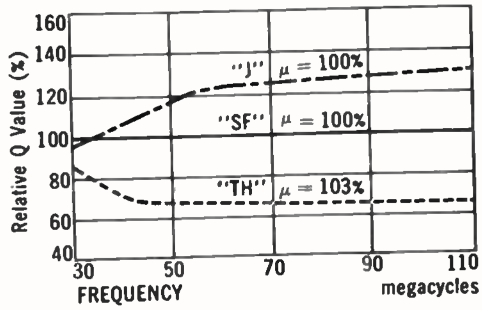
"J" Powder is now available in quantity. We invite you to write for further details and samples — and to test it for new applications.

* * *

Collectively, G A & F Carbonyl Iron Powders blanket a wide range of applications — in electronic cores over the whole frequency spectrum, in metallurgy, in chemistry, in pharmacy and in magnetic fluids. The particles may be large, soft crystals — or extremely small, hard crystals arranged in concentric spherical-shell layers. The surfaces are free and active. The purity is invariably high, with non-ferrous metals in traces only; some grades contain beneficial small amounts of carbon, nitrogen and oxygen.

HIGH-FREQUENCY G. A. & F. CARBONYL IRON POWDERS—RELATIVE Q vs. FREQUENCY

Form Factor—6. 8



We urge you to ask your core maker, your coil winder, your industrial designer, how G A & F Carbonyl Iron Powders can increase the efficiency and performance of the equipment or product you make, while reducing both the cost and the weight. Let us send you the book described on the left-hand page.

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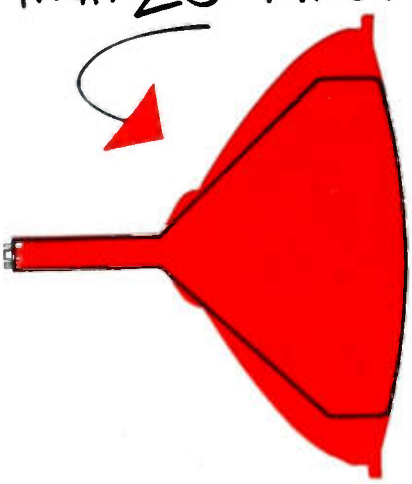
Rauland Announces 27" Rectangular Tube!

Actually shorter than 20" tube!



On March 3, Rauland unveiled the first "giant-screen" tube that makes attractive cabinetry possible.

This new 27" tube, with 390 square inch picture area, minimizes cabinet problems in two ways. First, it has the compactness of rectangular rather than round cone and face. Second, by means of 90° deflection, depth has actually been held slightly shorter than present 20" tubes!

The tube employs Rauland's usual "reflection-proof" filter glass face plate with maximum reflection of only 2½% of incident light. It uses the Rauland tilted offset gun with indicator ion trap. It is offered with either magnetic or low-focus-voltage electrostatic focus. Weight is held at minimum by use of a metal cone.

If you want a picture of really spectacular size that can be housed in acceptable furniture, here is your answer.

A picture actually more than 70 sq. in. larger than the center spread of a tabloid newspaper. Rectangular for minimum cabinet height and width. And actually permitting a small reduction in depth from today's 20" cabinets!

THE RAULAND CORPORATION

Perfection Through Research

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TUNER-TUBE PROBLEM *Solved* FOR TV DESIGNERS!

- The kit of new General Electric tubes at right is *your* answer, Mr. Designer, to the $64 TV-tuner question on which the success of your set may depend a year from now!
- Usable at all frequencies from 45 mc to 870 mc, these G-E tubes make possible a single, combined tuner circuit that (1) is simple in layout, (2) *saves* components, (3) gives one-dial tuning without the need to switch tubes between low and high bands.
- To a far simpler tuning circuit, add low noise level and freedom from snow. Add the big advantages of less radiation interference, greater selectivity!
- Investigate this up-to-the-minute tube group! Wire or write today for facts on the characteristics and performance of G.E.'s new tuner "4"s! Or if you wish, a G-E tube engineer will be glad to call on you. *Tube Department, Section 15, General Electric Company, Schenectady 5, New York.*

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Grounded-grid r-f triode (2 stages)

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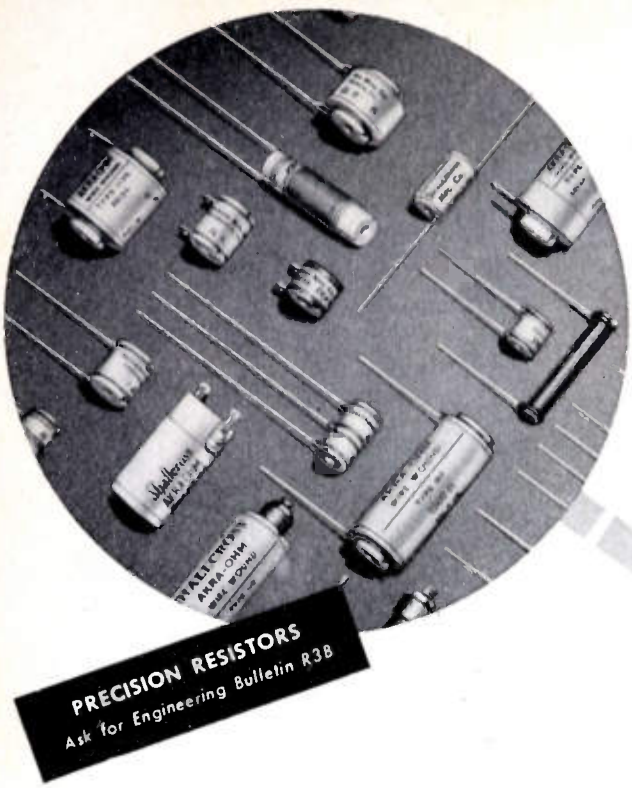
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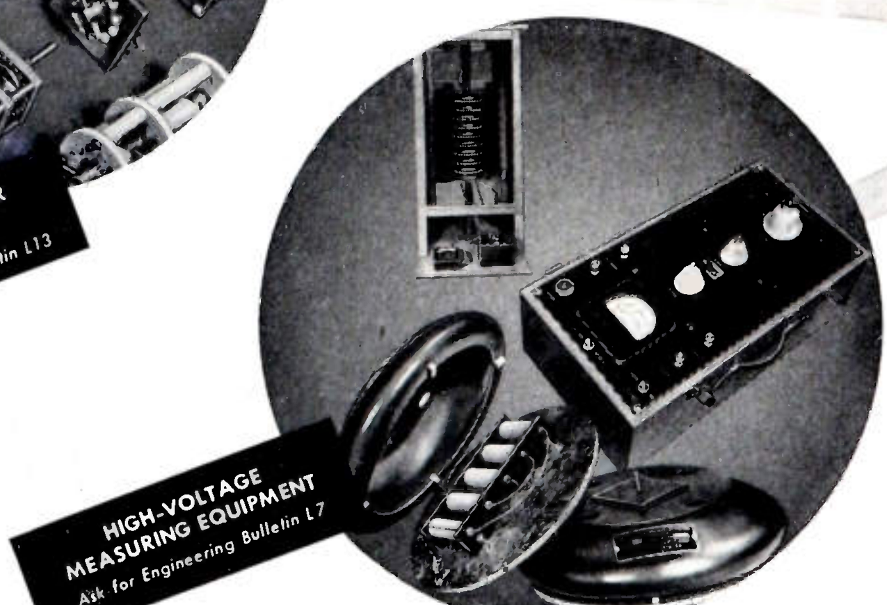
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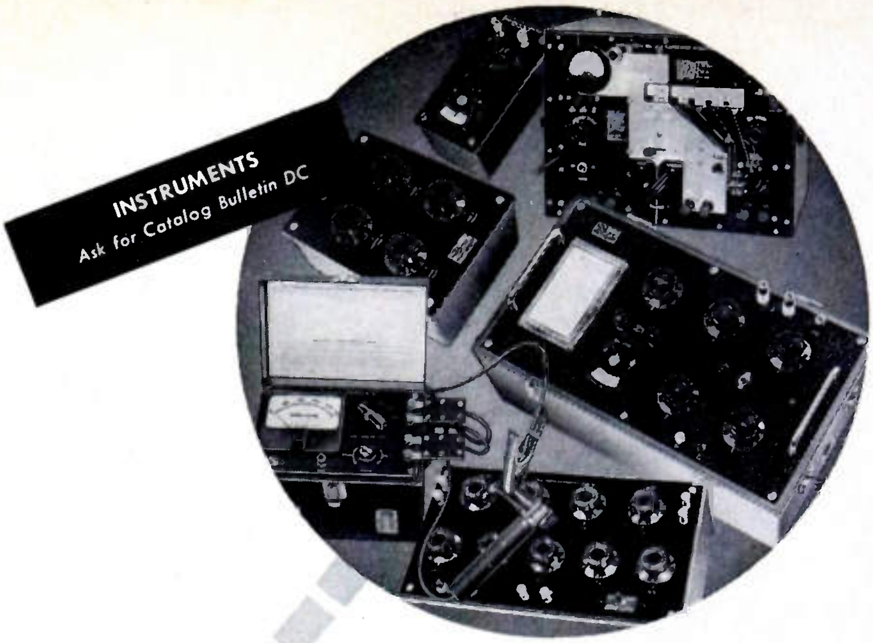

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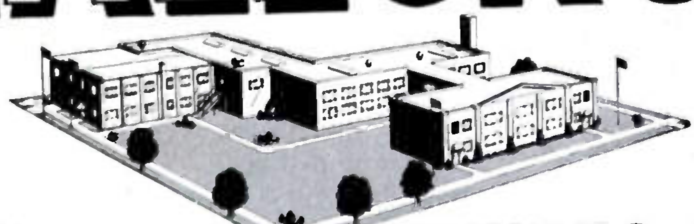

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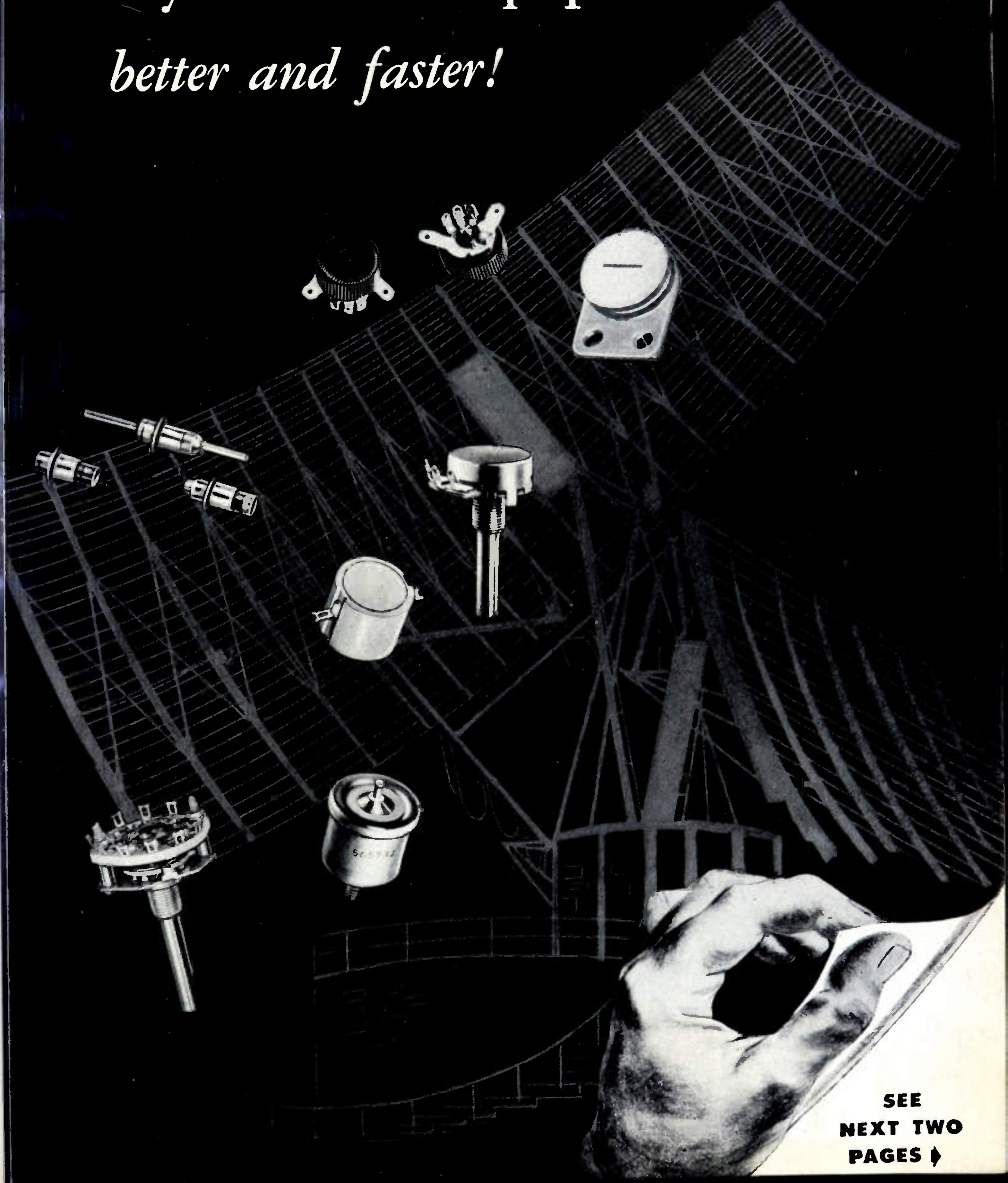

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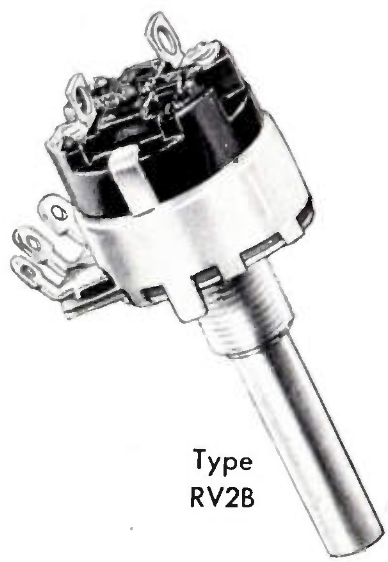
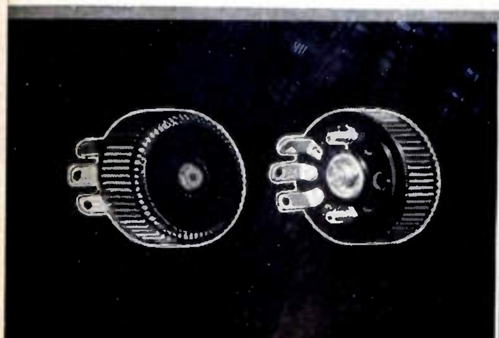
RV2A



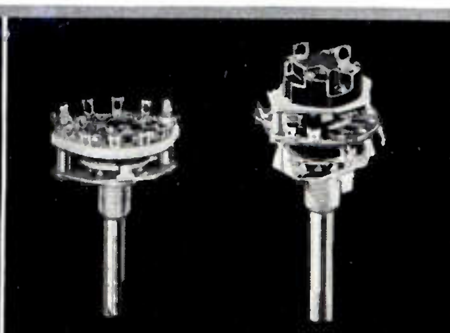
Type
RV2B

CENTRALAB MODEL 2 VARIABLE RESISTORS

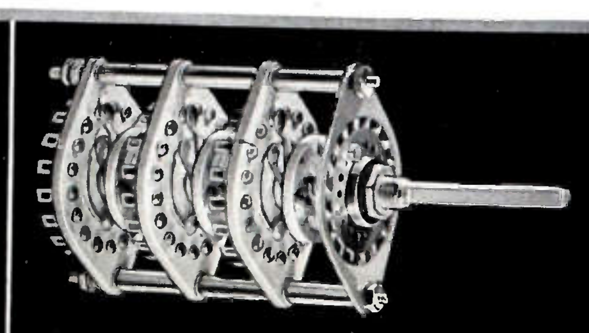
There's no prior contract approval or waivers required if you specify Centralab's Model 2 variable resistors on your next military order. They meet JAN R94, characteristic U requirements. Two types available — RV2A and RV2B — plain or with attached switches. Ratings from 2000 ohms to one megohm. For complete engineering data, check Bulletin No. 42-85 in coupon below.



Model 1, miniature variable resistors . . . no bigger than a dime . . . available in Standard or Hi-torque types. Either with or without on-off switch. Also available with slot — front or rear — for screw-driver adjustment. Hi-torque units hold settings under conditions of vibration or shock. For complete data check No. 42-158 in coupon below.

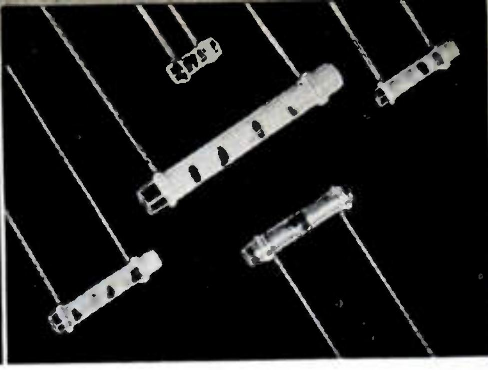


For miniature switches — specify Centralab's Series 20 with Steatite or Phenolic sections. Steatite is Grade L5. Meets JAN I-8 specs. Phenolic sections conform to JAN P-13 . . . Grade LTSE4. Available in 2 to 11 positions with stops, or 12 positions, continuous rotation—single or multiple sections—with or without attached on-off switch. Check No. 42-156.

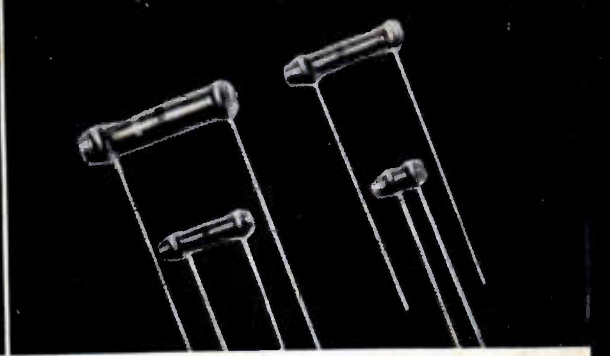


Centralab's Medium-Duty Power Switches. Use for R. F. or 110-115 V. application . . . 7½ amps. Voltage breakdown to ground — 3000 volts — RMS 60 cycles. Available with Grade L5 (JAN I-8) Steatite sections — shorting or non-shorting contacts. Models in 1, 2 or 3 poles, 18 contacts per section with adjustable stops, can be furnished up to 20 sections per shaft. Contacts and collector rings are coin silver. For complete data, check No. 42-136 in coupon.

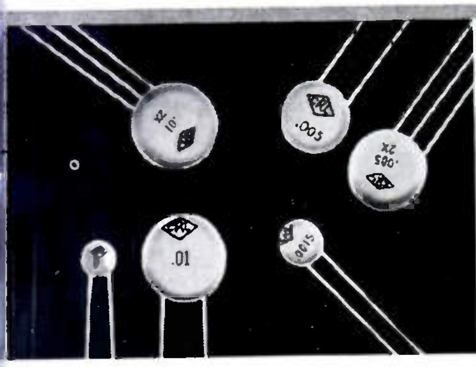
Centralab's Type 850 high voltage ceramic capacitors are especially designed for high voltage, high frequency circuits. Centralab's Type 950 high accuracy ceramic capacitors are especially developed for exacting electronic applications. Check bulletin No.'s 42-102 and 42-123.



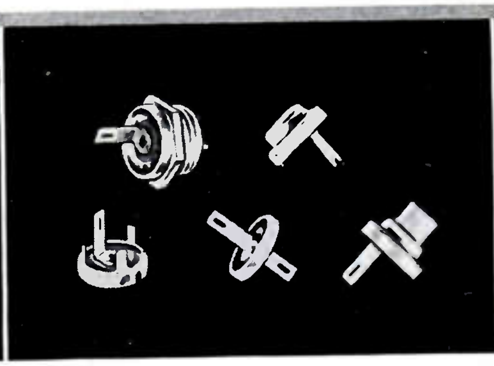
TC (Temperature Compensating) Tubulars — No prior contract approval or waiver necessary. Meet JAN-C-20A requirements. Type TCZ shows no capacitance change over wide range of temperature. Type TCN has special ceramic body to vary capacitance according to temperature. Bulletin No. 42-18.



BC (Bypass Coupling) Tubulars — Recommended for bypass coupling. Well suited to general circuit use. Centralab's own Ceramic X body provides imperviousness to moisture and low power factor. Easily withstands temperatures normally encountered in most electronic equipment. Bulletin No. 42-3.

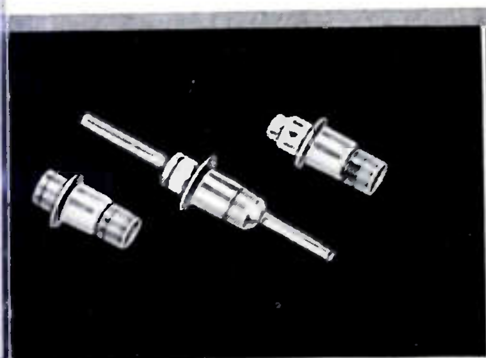


Ceramic Disc Hi-Kap Capacitors hold thickness to a minimum . . . have very high capacity in extremely small size. Use in h.f. circuits for bypass and coupling. Ceramic body assures low inductance. Other characteristics— humidity resistance, power factor, etc. — similar to BC Tubulars, Bulletin No. 42-4R.

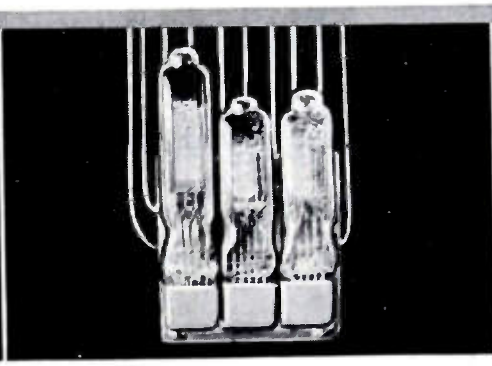


Something new in miniature ceramic capacitors! These "button types" are available in 5 different styles. Used for bypassing in low-power, high-frequency applications where small size, low inductance and light weight are essential. Check Bulletin No. 42-122 in coupon for more information.

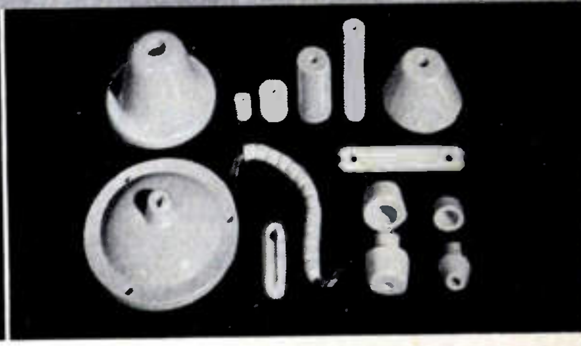
Centralab Ceramic Trimmers meet applicable portions of JAN-C-81. Very small size. Screw driver adjustment over full capacity range (180° rotation). Maintain stability in any position and under vibration. Spring pressure contact for rotor and stator. Bulletin No. 42-101.



Centralab's New Eyelet-Mounted Feed-Through Ceramic Capacitors are smallest available. They meet applicable portions of JAN-C-20A specifications. Capacities range from 10 to 3000 mmf . . . the widest range on the market. Voltage rating. 500 V.D.C.W. Check No. EP-15 in coupon.



New Sub-miniature Model III Ampec — a full three-stage speech amplifier of remarkably small dimensions — approximately 1 1/32" x 15/16" x 11/32" (barely larger than a postage stamp!). Excellent for microphone preamplifiers and similar applications. Check No. 42-130 on coupon for complete information.



Centralab standard and custom-molded Steatite ceramics plain or metallized . . . fully comply with JAN I-8. Steatite is Grade L5 for military use. Characteristics — high dielectric strength, low loss at high frequencies, high mechanical strength. For data on standard parts or custom molding, check No. 720.

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PROCEEDINGS OF THE I.R.E.

Published Monthly by

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VOLUME 40 — *May, 1952* — NUMBER 5

Changes of address (with advance notice of fifteen days) and communications regarding subscriptions and payments should be mailed to the Secretary of the Institute, at 450 Ahnaip St., Menasha, Wisconsin, or 1 East 79 Street, New York 21, N. Y.

* Numerals in parentheses following Directors' names designate Region number.

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

Responsibility for the contents of papers published in the PROCEEDINGS OF THE I.R.E. rests upon the authors. Statements made in papers are not binding on the Institute or its members.

The Founders of the IRE

Robert H. Marriott formed The Wireless Institute in 1909, and served as its President until it merged with the Society of Wireless Telegraph Engineers to form The Institute of Radio Engineers, May 13, 1912. He became the first President of the IRE for the remainder of the year 1912. In 1913, he became the Vice President of the IRE, and served as a member of the IRE Board of Directors in 1914–1916, 1920–1922, and 1926–1932. He then became a Fellow of the IRE in 1915.

Mr. Marriott received his technical training at Ohio State University where he started his radio experimental work, and later became well-known among important early pioneering radio concerns. He was the first man to put into use in America the telephone and detector method of radio reception, a system that was a forerunner of the vacuum tube. He was a radio aide of the United States Navy and a consultant engineer of the Federal Radio Commission. Mr. Marriott retired from private practice as consultant engineer in 1943. He passed away October 31, 1951.

ROBERT H. MARRIOTT

JOHN V. L. HOGAN

John V. L. Hogan, who was a member of The Society of Wireless Telegraph Engineers, became a Director of The Institute of Radio Engineers upon its formation and served in this capacity continuously until 1936, and again from 1948–1950. He was the IRE Vice President from 1916–1919, and has been the only person to hold this office for more than one year. He then served as the IRE President in 1920. Mr. Hogan became a Fellow of the IRE in 1915.

Mr. Hogan studied electrical engineering at the Sheffield Scientific School and was associated with a number of early research projects. From 1914–1921 he was a member of the staff of the National Electric Signaling Company which became the International Radio Telegraph Company. He then became a consulting engineer in New York City. He is responsible for many inventions in radio, television, and facsimile fields, and was founder and owner of station WQXR, until it was acquired by the *New York Times*.

Mr. Hogan is president of Hogan Laboratories, Inc., New York City.

Alfred N. Goldsmith was a member of The Wireless Institute from its inception in 1909. He was serving as its Editor when it became part of The Institute of Radio Engineers, and has continued in that post for 40 years, except when he served as IRE President in 1928. He was Secretary from 1918–1927, and has served continuously on the IRE Board of Directors since 1912. He became an IRE Fellow in 1915.

Dr. Goldsmith received his B.Sc. degree from the College of the City of New York, where he later became associate professor of electrical engineering. He obtained his Ph.D. from Columbia and Sc.D. from Lawrence College. He has been a consulting engineer for General Electric Company, director of research for Marconi Wireless Telegraph Company, and also was a Vice President of RCA. At present Dr. Goldsmith is a consulting engineer in New York City. He has made a number of inventions in the radio, phonograph, TV, color TV, and motion picture fields. He has received the IRE Medal of Honor and many other awards.

ALFRED N. GOLDSMITH

"Life Begins at Forty"

To paraphrase Lincoln, three decades and ten years ago, our engineering forefathers brought forth in this Western Continent a new Institute, founded on the theories that the dissemination of knowledge was the basis of progress, that solidarity of engineering effort was a source of achievement, that long-term ideals and hard work far excelled political expediency and mere verbalisms in their value to humanity, and that the engineers of those days and of later decades were and would remain sufficiently imbued with worthy aims and capable of such continued effort as would ensure the-success of any organization founded by them with these sincere purposes.

Thus, a small group of enthusiasts with a vision met on May 13, 1912, to give birth to their infant Institute of Radio Engineers, an infant which promptly began to grow into a healthy and strong childhood.

At present, on its fortieth birthday, our Institute of Radio Engineers is a vigorous youth looking forward to a productive and prolonged maturity. On a merely numerical basis, the IRE of 1952 is, of course, vastly different from that of 1912. Tens of members have grown into tens of thousands. Tens of pages published annually have similarly grown into thousands of pages. A single meeting group has changed into more than half a hundred Sections scattered over the Western Hemisphere from Montreal to Buenos Aires, and from Boston to Hawaii. These Sections have, in fact, been well advised to form themselves into Regional associations for better and more democratic representation and operation. Almost twenty Professional Groups now exist. Some of these are already flourishing organizations with many hundreds of members and with published TRANSACTIONS of high caliber. In fact, some of them are approaching the stage of regularly scheduled and extensive publication of their TRANSACTIONS. The Technical Committees of the Institute are almost legion. And the devoted service given to communications and electronic technology, and to the corresponding industries, by the members of these Committees is a monument to engineering idealism and far-sightedness.

But more important than the quantitative expansion of IRE in the last forty years is the aspect of qualitative growth. It is incumbent upon us at all times to ask ourselves the question: Are the quality, the aims, and the significance of our labors always improving? The answer to that question is the true measure of the extent to which we have succeeded in fulfilling the vision of the founders of the Institute.

As we pause at this milestone to plot the curve of growth from 1912 to the present, one cannot help but conclude that the IRE is still young and vigorous. And as we extrapolate this curve into the future, it becomes apparent that at forty life is just beginning.

—*The Editor.*

The Genesis of IRE*

Inasmuch as only nine of the present thirty thousand members of the Institute participated in the formation of IRE, the following account of how the Institute was formed forty years ago will be of interest to many readers.—*The Editor*

Introduction

IN THE YEAR 640 B.C., a Greek philosopher named Thales found to his astonishment that when he rubbed a piece of amber, or "elektron" as it was then called, it became capable of attracting straws and other light objects. From this humble beginning man went on to scientific and technological advances which were to change the course of civilization more than twenty centuries later. He discovered electricity, investigated its properties, determined its relationship to magnetism, and penetrated the basic secrets of the atom. With each new step he set about to apply his new-found knowledge so that he might harness this undreamed-of power to increase his comfort, safeguard his health, and improve his prosperity.

But such a task could not be accomplished by any one man or single group of men. By 1912 it became evident with the rapid advancement of the communications field that if a full measure of success were to be achieved in radio, each must share his knowledge with all; if advances were to be made, the efforts of all must have a common purpose and direction.

It was to this end that The Institute of Radio Engineers was formed on May 13 of that year.

The Dawn of Radio

Since earliest times man has sought to devise means by which he could communicate over long distances. The African drum, the Indian's smoke signal, even the lantern in the tower of the Old North Church in Boston which sent Paul Revere on his famous ride—all were evidences of man's never-ending struggle to extend his range of direct communication.

From the beginning, experimenters in electricity and magnetism were intrigued by the idea that their investigations might provide them with a solution to this problem. Their researches led them to experiment with the conductive properties of water, with wire, and with magnetic induction. In 1867 the most startling answer was provided by James Clark Maxwell when he predicted theoretically the existence and behavior of electromagnetic waves. It remained for Heinrich Hertz 19 years later to demonstrate experimentally for the first time the transmission and reception of electromagnetic waves generated by an oscillating circuit.

Alert to the significance of these developments, Marconi, Poulsen, Fessenden, Lodge, and other engineers developed equipment by which messages in code could be transmitted without wires and over long distances. To this achievement was added the transmission of speech by Fessenden, Flemming's electronic valve detector, and de Forest's 3-element electron tube, so that by 1907 the infant "radio" was beginning to emerge.

The Society of Wireless Telegraph Engineers

On February 25th of that year in Boston, Mass., John Stone Stone, renowned radio pioneer, formed the first radio engineering society, the Society of Wireless Telegraph Engineers (SWTE). An outgrowth of seminars held by engineers of the Stone Wireless Telegraph Company, the SWTE started with only 11 members, all recruited from the Stone staff. In fact membership in SWTE was restricted solely to Stone engineers for a time, but eventually its rolls were opened to men from the Fessenden Company and other organizations.

Stone was elected the first President and served for the term 1907–1908. Lee de Forest took over the helm in 1909–1910, followed by Fritz Lowenstein in 1911–1912. Already suffering from a limited source of new members, the SWTE was dealt an almost fatal double blow when, first, the Fessenden Company moved from Brant Rock, Mass., to Brooklyn, N. Y., taking with it a number of members, and secondly, when the Stone Wireless Telegraph Company ceased operation in 1910. Nevertheless, the SWTE made an important contribution to the new field by laying the foundation for the more successful venture to follow.

The Wireless Institute

Nothing concrete had been accomplished, however, to overcome the barriers of super-secrecy and friction which existed between rival wireless companies. It was an era of jealously guarded secrets and ruthless competition. Men were dismissed for associating with engineers of rival firms. Ingenious methods of manufacturing and assembling wireless apparatus were devised to make it as difficult as possible for anyone to learn how the equipment operated by taking it apart. Competition was often devoid of ethics as we know them today. And yet despite, or perhaps because of, this atmosphere of intense rivalry, there was a growing realization among a number of engineers that many of the technical problems encountered by one company were common to all, and that these problems might be solved more readily if engineers from all companies could get together to discuss them.

It was under these circumstances that Robert H. Marriott made the first specific attempt to form a radio engineering society composed of members from any and all companies. On May 14, 1908, he sent a circular letter to some two hundred persons interested in wireless asking their opinions regarding the formation of such a society. It is rather remarkable that after 44 years the embryo of IRE can still be clearly distinguished in Marriott's letter, which reads as follows:

"Dear Sir:

"You have often thought no doubt that Wireless Telegraphy would be developed faster if those engaged in it would work together more.

"The Electrical Engineers have come together in the United States by forming the American Institute of Electrical Engineers. This institution has helped to make better Electrical Engineering, better Electrical Engineers, and better feeling between competitive firms.

"Why should not we form the Institute of Wireless Engineers and pattern it after the American Institute of Electrical Engineers. The American Institute of Electrical Engineers' plan as applied to Wireless people would be briefly as follows:

First: Any person interested in Wireless with proper recommendations, etc., would be eligible to associate membership.

Second: Any person having done valuable, original work in Wireless would be eligible to full membership.

Third: Any person whom the Society, by vote, should decide upon would be eligible to honorary membership.

Fourth: Meetings would be held once a month, at which papers on Wireless subjects would be read and criticized.

Fifth: Every member and associate would receive a copy of the paper read, together with criticisms, thus giving absent members the same information as those present.

Sixth: A library of Wireless publications would be accumulated as rapidly as the funds of the Institute would permit. Each member or associate member would have access to this library.

Seventh: The Officers and Committees would be about as follows: President, Vice-President, Manager, Treasurer, and Secretary.

Committees: Executive Committee, Committee on Finances, Committee on Papers, Board of Examiners, Library Committee, Editing Committee, and necessary special committees appointed from time to time.

Eighth: The dues would be possibly about $10.00 per year.

"I believe an organization formed on a plan similar to the above would materially improve Wireless, increase the knowledge and ability of members, avoid friction between employees, between employees and employers, and to some extent between Wireless companies.

"Would you join such an organization as outlined? If so, please write me and give full expression of your views in regard to the matter in order that an organization may be formed on the right lines. Also such an organization might contemplate the establishment of a beneficiary association in connection with the Institute.

Yours very truly,
R. H. Marriott
Ass't., Scientific Manager
United Wireless Telegraph Co.
42 Broadway, New York"

Marriott's letter bore fruit. He received about 60 replies, and with one or two exceptions they were favorable to forming an institute on the lines he indicated. On January 23, 1909, a temporary organization was formed to draw up a constitution, and on March 10 of that year the first meeting was held at the United Engineers Building (now the Engineering Societies Building) in New York City at which the constitution was adopted and officers were elected. The name of the new society was "The Wireless Institute."

Robert Marriott was elected first President of the Wireless Institute, and served in that capacity during the three years of its existence. The Vice-Presidents were Harry Shoemaker and Greenleaf W. Pickard, who is at present a very active member and Fellow of IRE. Pickard was at this time also a member of the Society of Wireless Telegraph Engineers and was the only person belonging to both pre-IRE societies. The remaining officers, mostly wireless operators for the United Wireless Telegraph Co. in New York, were Sidney L. Williams, Secretary; Eugene M. Thurston, Treasurer; and John S. Murphy and Richard A. Sommerville, Directors.

The Wireless Institute (TWI) was somewhat more active than its fellow society in Boston, SWTE, because of the greater number of wireless operators and engineers in the New York area. It started with only 14 members, but by 1911 had grown to a more impressive total of 99. TWI held meetings every month except July and August, and unlike the SWTE, was able to publish the papers presented at these meetings in a *Proceedings* for distribution to members. Alfred N. Goldsmith took on the task of Editor of *The Wireless Institute Proceedings*, little realizing that more than four decades later he would still be serving in the same capacity for its successor, the PROCEEDINGS OF THE I.R.E.

The Birth of IRE

Despite an encouraging start, The Wireless Institute, like the SWTE, found that its membership soon reached a saturation point. In fact, by 1912 it had dropped from a high of 99 to 27 and was struggling to keep out of debt. It was at this time that Robert H. Marriott, Alfred N. Goldsmith, representing the Wireless Institute, and John V. L. Hogan, who was very active in the Society of Wireless Telegraph Engineers, held an informal meeting to discuss the plights of both societies. Out of this meeting came a plan to consolidate the two societies in order that the influence exerted by each might thereby gain greater effectiveness. On April 5, 1912, the Wireless Institute held a meeting in room 304 of Fayerweather Hall, Columbia University, with members of SWTE attending, at which the proposed consolidation was agreed upon. The Wireless Institute held its final meeting on May 6 at which time the members were advised that the new society would probably be called "The Institute of Radio Engineers."

On the night of May 13, 1912, members of the erstwhile TWI and SWTE gathered once again at Columbia University. A constitution was approved and an election for officers was held with the following results: Robert H. Marriott, President; Fritz Lowenstein, Vice-President; E. D. Forbes, Treasurer; E. J. Simon, Secretary; Alfred N. Goldsmith, Editor; and Lloyd Espenshied, Frank Fay, J. H. Hammond, Jr., John V. L. Hogan, and J. S. Stone, Managers.

The original membership roster of the Institute consisted of 46 members, 22 from SWTE and 25 from TWI (one member, Greenleaf W. Pickard, belonged to both societies). Because of its historical interest, the list of Charter Members is here presented.

Charter Members of the IRE

From the Society of Wireless Telegraph Engineers

J. C. Armor
Sewell Cabot
W. E. Chadbourne
George H. Clark
Thomas E. Clark
Ernest R. Cram
George S. Davis
†*Lee de Forest
E. D. Forbes
V. Ford Greaves
Guy Hill
*John L. Hogan, Jr.
Comm. W. S. Hogg
F. A. Knowlton
F. H. Kroger
Fritz Lowenstein
Walter W. Massie
†*Greenleaf W. Pickard
Col. Samuel Reber
Oscar C. Roos
†John Stone Stone
†*A. F. Van Dyck

From the Wireless Institute

William F. Bissing
A. B. Cole
*P. B. Collison
James N. Dages
*Lloyd Espenschied
Phillip Farnsworth
Frank Fay
Edward G. Gage
†*Alfred N. Goldsmith
Francis A. Hart
Robert L. Hatfield
Arthur A. Hebert
*Frank Hinners
Joseph M. Hoffman
†Robert H. Marriott
A. F. Parkhurst
†*Greenleaf W. Pickard
H. S. Price
Adolph Rau
Harry Shoemaker
*Emil J. Simon
A. Kellogg Sloan
Clark H. Sphar
Floyd Vanderpoel
Roy A. Weagant

Examination of the records of the Institute show that 9 of the above individuals (indicated by asterisks) are still active members of the IRE after forty years. It is also of interest that 6 of the Presidents of the Institute (indicated by daggers) were Charter Members of the IRE.

By the end of 1912, the Institute's membership had risen to 109 and during the succeeding year more than doubled. The rapid increase in membership after consolidation, compared with the slow rate of growth of SWTE and TWI (see Table I), bore out the wisdom of the founders who suggested the merger.

TABLE I
Membership Growth of SWTE, TWI, and IRE

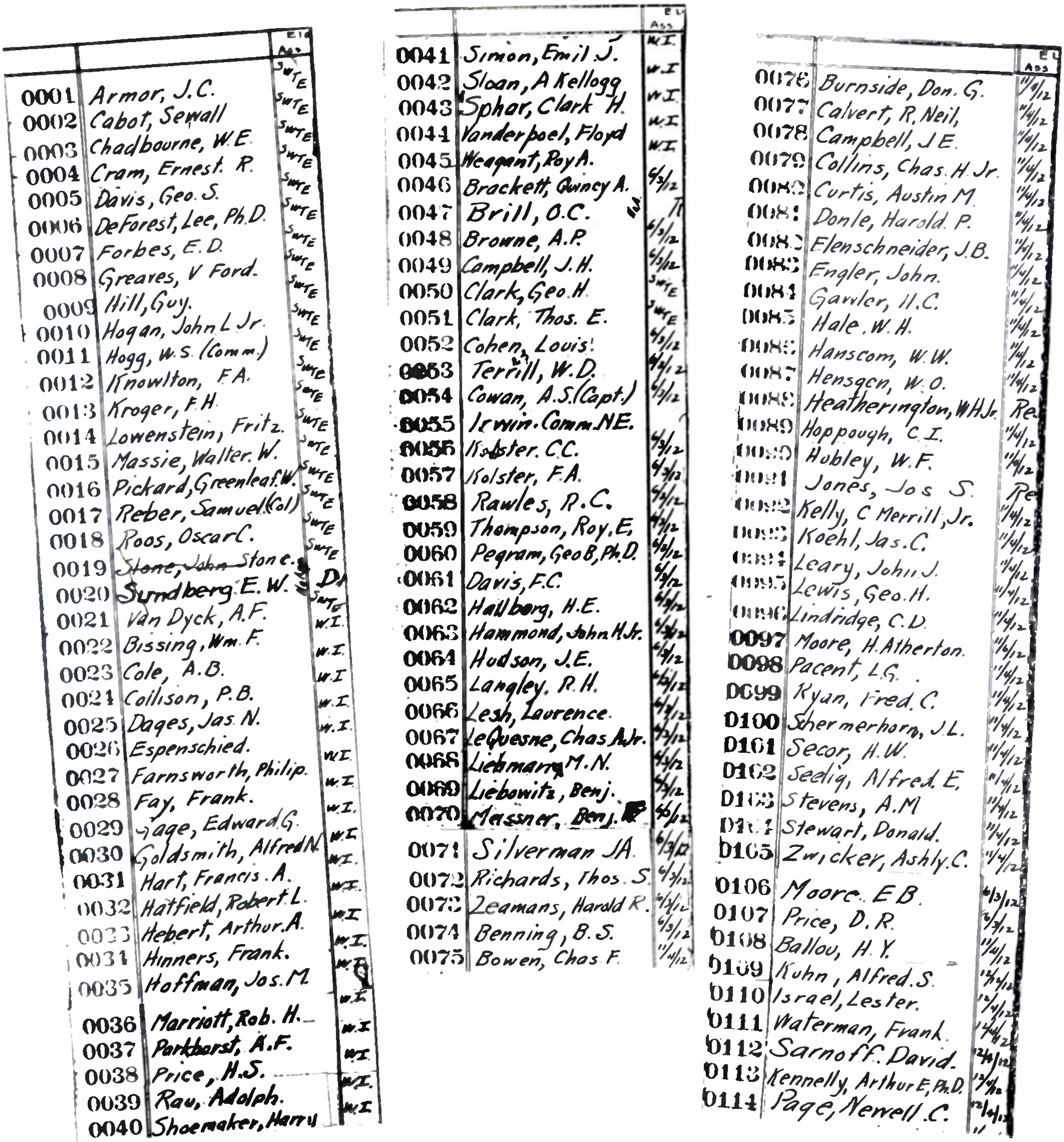
	SWTE	*TWI*	*IRE*
Feb. 25, 1907	11		
Jan. 1, 1908	17		
Jan. 1, 1909	27		
March 10, 1909		14	
Jan. 1, 1910	36	81	
Jan. 1, 1911	36	99	
Jan. 1, 1912	43	27	
May 13, 1912	(22)	(25)	46
Jan. 1, 1913			109
Jan. 1, 1914			231
March 1, 1914			271

The original ledger book of the Institute, in which the names and dues payments of early members were recorded, constitutes a veritable *Who's Who* in the history of radio. The names of many radio pioneers can be seen in the accompanying illustration (at right) of the first few pages covering those members who joined the Institute in 1912.

The careful planning and foresight with which the IRE founders organized a lasting society is reflected in the fact that they gave to the Institute a name and triangular symbol which has withstood the test of forty years. A keen appreciation of long-term values is clearly demonstrated by the manner in which they arrived at a suitable name and symbol.

In considering a title for the Institute, the founders felt that something should be preserved from the names of the two component societies. The word "Institute" was borrowed from the Wireless Institute, and "Engineers" from the Society of Wireless Telegraph Engineers. Because the word "radio" was gradually supplanting "wireless," the title "The Institute of Radio Engineers" suggested itself. There was considerable temptation to add "American," particularly since TWI and IRE were modeled after the American Institute of Electrical Engineers in certain other respects. However, the temptation was resisted because it was expected that the IRE, as the only radio engineering society in existence, would be international in scope, an expectation which was promptly realized.

In considering the design of a suitable symbol for IRE, the founders first examined the emblems of SWTE and TWI. The SWTE symbol was diamond shaped and showed a Hertz dumbbell oscillator. The TWI emblem was circular and depicted a circular Hertz receiving loop. A Hertz oscillator was later added in the center. It was evident to the IRE founders that the Hertz loop and oscillator, or for that matter any current day de-



No.	Name	El. / Ass.
0001	Armor, J.C.	SWTE
0002	Cabot, Sewall	SWTE
0003	Chadbourne, W.E.	SWTE
0004	Cram, Ernest. R.	SWTE
0005	Davis, Geo. S.	SWTE
0006	DeForest, Lee, Ph.D.	SWTE
0007	Forbes, E.D.	SWTE
0008	Greaves, V Ford.	SWTE
0009	Hill, Guy.	SWTE
0010	Hogan, John L Jr.	SWTE
0011	Hogg, W.S. (Comm.)	SWTE
0012	Knowlton, F.A.	SWTE
0013	Kroger, F.H.	SWTE
0014	Lowenstein, Fritz.	SWTE
0015	Massie, Walter. W.	SWTE
0016	Pickard, Greenleaf W.	SWTE
0017	Reber, Samuel (Col)	SWTE
0018	Roos, Oscar C.	SWTE
0019	~~Stone, John~~ Stone.	D
0020	Sundberg E.W.	SWTE
0021	Van Dyck, A.F.	W.I.
0022	Bissing, Wm. F.	W.I.
0023	Cole, A.B.	W.I.
0024	Collison, P.B.	W.I.
0025	Dages, Jas. N.	W.I.
0026	Espenschied.	W.I.
0027	Farnsworth, Philip.	W.I.
0028	Fay, Frank.	W.I.
0029	Gage, Edward G.	W.I.
0030	Goldsmith, Alfred N.	W.I.
0031	Hart, Francis. A.	W.I.
0032	Hatfield, Robert L.	W.I.
0033	Hebert, Arthur. A.	W.I.
0034	Hinners, Frank.	W.I.
0035	Hoffman, Jos. M.	W.I.
0036	Marriott, Rob. H.	W.I.
0037	Parkhurst, A.F.	W.I.
0038	Price, H.S.	W.I.
0039	Rau, Adolph.	W.I.
0040	Shoemaker, Harry	W.I.
0041	Simon, Emil J.	W.I.
0042	Sloan, A Kellogg	W.I.
0043	Sphor, Clark H.	W.I.
0044	Vanderpoel, Floyd	W.I.
0045	Weagant, Roy A.	W.I.
0046	Brackett, Quincy A.	6/3/12
0047	Brill, O.C.	Tr
0048	Browne, A.P.	6/3/12
0049	Campbell, J.H.	6/3/12
0050	Clark, Geo H.	SWTE
0051	Clark, Thos. E.	SWTE
0052	Cohen, Louis.	6/3/12
0053	Terrill, W.D.	4/3/12
0054	Cowan, A.S. (Capt.)	6/3/12
0055	Irwin. Comm. N.E.	
0056	Kolster. C.C.	4/3/12
0057	Kolster, F.A.	4/3/12
0058	Rawles, R.C.	4/3/12
0059	Thompson, Roy. E.	4/3/12
0060	Pegram, Geo B, Ph.D.	4/3/12
0061	Davis, F.C.	4/3/12
0062	Hallborg, H.E.	6/3/12
0063	Hammond, John H. Jr.	6/3/12
0064	Hudson, J.E.	6/3/12
0065	Langley. R.H.	6/3/12
0066	Lesh, Laurence.	6/3/12
0067	LeQuesne, Chas A. Jr.	4/3/12
0068	Liebmann M.N.	4/3/12
0069	Liebowitz, Benj.	4/3/12
0070	Messner, Benj.	4/3/12
0071	Silverman J.A.	6/3/12
0072	Richards, Thos. S.	6/3/12
0073	Zeamans, Harold R.	6/3/12
0074	Benning, B.S.	6/3/12
0075	Bowen, Chas. F.	11/4/12
0076	Burnside, Don. G.	11/4/12
0077	Calvert, R. Neil,	11/4/12
0078	Campbell, J.E.	11/4/12
0079	Collins, Chas. H. Jr.	11/4/12
0080	Curtis, Austin M.	11/4/12
0081	Donle, Harold. P.	11/4/12
0082	Elenschneider, J.B.	11/4/12
0083	Engler, John.	11/4/12
0084	Gawler, H.C.	11/4/12
0085	Hale. W.H.	11/4/12
0086	Hanscom, W.W.	11/4/12
0087	Hensgen, W.O.	11/4/12
0088	Heatherington, W.H. Jr.	Re
0089	Hoppough, C.I.	11/4/12
0090	Hubley, W.F.	11/4/12
0091	Jones, Jos. S.	Re
0092	Kelly, C Merrill, Jr.	11/4/12
0093	Koehl, Jas. C.	11/4/12
0094	Leary, John J.	11/4/12
0095	Lewis, Geo. H.	11/4/12
0096	Lindridge, C.D.	11/4/12
0097	Moore, H. Atherton.	11/4/12
0098	Pacent, L.G.	11/4/12
0099	Ryan, Fred. C.	11/4/12
0100	Shermerhorn, J.L.	11/4/12
0101	Secor, H.W.	11/4/12
0102	Seelig, Alfred. E.	11/4/12
0103	Stevens, A.M.	11/4/12
0104	Stewart, Donald.	11/4/12
0105	Zwicker, Ashly. C.	11/4/12
0106	Moore. E.B.	6/3/12
0107	Price, D.R.	6/3/12
0108	Ballou, H.Y.	11/4/12
0109	Kuhn, Alfred. S.	12/4/12
0110	Israel, Lester.	12/4/12
0111	Waterman, Frank.	12/4/12
0112	Sarnoff. David.	12/4/12
0113	Kennelly, Arthur E, Ph.D.	12/4/12
0114	Page, Newell. C.	12/4/12

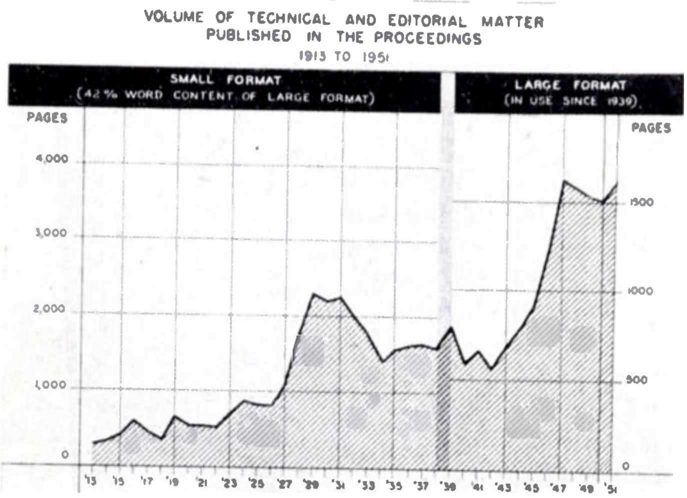
Pages from the first IRE record book listing members who joined during 1912.

vice, would eventually become obsolete as new techniques and equipment were developed. The one thing that seemed permanent in the art was the familiar electrical and magnetic forces and their relationship. Accordingly the electrical force was represented by a vertical arrow, and the magnetic force by a circular arrow surrounding the electric force and in the conventional relationship to it. The shape of the resulting drawing obviously lent itself to a triangular placement of the letters I.R.E. This in turn led to the selection of an attractive triangular emblem.

The IRE Expands

On the firm foundation laid forty years ago, the structure of the Institute has steadily risen. The Institute after its formation continued to hold its meetings at 304 Fayerweather Hall, Columbia University, for several years. The papers and discussions presented at these meetings were published in the Proceedings of the I.R.E., which began publication in 1913 under the editorship of Alfred N. Goldsmith. The Proceedings quickly established itself as one of the leading publications in the world devoted to the radio engineering field.

Beginning as a small quarterly publication of about 50 pages per issue, PROCEEDINGS now publishes monthly and contains a total of more than 3,000 pages each year, with occasional special issues totalling as high as 400 pages each, an expansion which has paralleled closely the rapid growth of membership.



Graph showing PROCEEDINGS rise to big-publication class.

In addition to its meetings and publication activities, the Institute promptly began work in the standardization field. A Standardization Committee was formed during the first year and issued a tenative report on September 10, 1913. The membership of this first committee on Standards comprised the following: Robert H. Marriott, Alfred N. Goldsmith, John V. L. Hogan, A. E. Kennelly, Roy A. Weagant, Greenleaf W. Pickard.

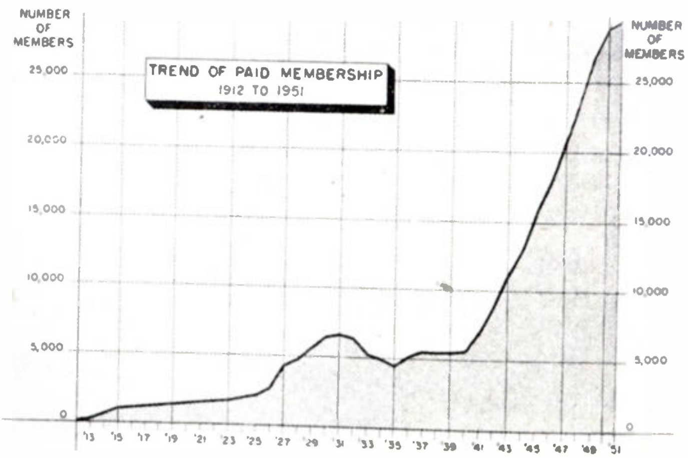
Since that time, standardization activities have played a permanent and prominent role in Institute affairs, resulting in the eventual formation of more than twenty technical committees which originate the standards which now appear at frequent intervals in the PROCEEDINGS. These activities have contributed appreciably to the advancement of the electronic and communication art, bringing conformity and clarity to all fields of endeavor.

The activities of the Institute, originally confined to the New York area, quickly spread to other cities as the membership increased. Small groups of members in various localities began to hold their own local meetings and elected their own officers for organizing and running these meetings. These groups of members were called "Sections." The first Section, formed in Washington, D. C. in 1914, was promptly followed by the formation of Sections in Boston (1916), Seattle (1916) and San Francisco (1917). It is interesting to note how the development of IRE Sections reflected the growth of the radio field. The first four IRE Sections were formed in large coastal cities due to the predominance of maritime radio just prior to World War I. In 1925 other Sections were formed in Philadelphia, Chicago and Toronto as the broader aspects of radio engineering began to materialize and as the influence of IRE was felt in other countries. The growth of this important "grass roots" activity has continued unabated so that there are today 62 IRE Sections and 15 Subsections in the United States, Canada, Argentina and Hawaii.

The Sections were tied in more closely with the administration of the Institute when in 1946 the Sections were grouped into 8 Regions and each Region was given representation on the IRE Board of Directors. Since then the Regional Plan has provided an effective channel of communication between Board and membership.

The following year the Board authorized the establishment of Student Branches in schools of recognized standing in order that the important segment of student members might be given official and direct support of The Institute. During the elapsed period of time which has elapsed Student Branches have sprung up in 110 colleges.

Perhaps the most basic change in the structure of the Institute which has occurred during its forty years of growth was the establishment in 1948 of the Professional Group system, providing for sub-societies within IRE along lines of technical specialization of the members. This development, although still in its formative stages, has already led to the formation of 18 Professional Groups. Many of these Groups have been active in sponsoring meetings and several have commenced issuing their own technical publications, called "TRANSACTIONS." The successful development of the Professional Group system will do much to counteract the centrifugal tendencies which may be expected to accompany the rapid expansion of the Institute.



Graph indicating tremendous growth of IRE.

The 46 original members of the Institute, located in a 200 mile area, can be justifiably proud of the fact that in forty years their numbers have multiplied to 30,000 engineers and scientists from every corner of the globe. Their trials and tribulations were not in vain. The evolution of Sections, Regions, Student Branches and Professional Groups, the growth in stature and size of the publications of the Institute, the influence and activity of its Technical Committees—all are a tribute to the past, a credit to the present, and hold a promise for the future of the Institute.

The IRE—Cohesion or Dispersion?*

DONALD B. SINCLAIR†, PRESIDENT AND FELLOW, IRE

On its fortieth anniversary, the Institute finds that the character and needs of its membership have undergone a basic change. The future of IRE may depend largely on the solutions of certain major questions now confronting the members and their officers.

There is here presented by the President a clear and thought-provoking analysis of the nature of this fundamental change, the resulting questions that have arisen, and some possible answers. This analysis is recommended to every member as required reading.
—*The Editor.*

When the Institute was formed it comprised a small, cohesive group with a common interest. The interest was the new art of radio, and the group, even at this early date, was aware of this art as an individual technology which could be identified within the broad classification of electrical engineering.

As the years passed by, theory and practice became progressively more highly developed and, by the beginning of World War II, radio engineering had become recognized as a legitimate profession. A highly trained group of engineers had come into existence, creating stratification within the membership, but the basic cohesiveness of the society had not been seriously affected.

This cohesiveness is an interesting phenomenon. Radio seems to possess an element of magic beyond the mere statement of its principles in physical and mathematical terms. No other scientific field has attracted so much interest on the part of nontechnical and semitechnical people, nor participation in such an activity as amateur radio. There seems to be a common bond uniting amateur, technician, and engineer that surmounts differences in technological training. This cohesiveness, in fact, extends to a remarkable degree to the entire industry. It is doubtful if any other technology has developed an equivalent degree of understanding of each other's problems among scientists, engineers, salesmen, and management. Industry and science have developed together, and the Institute has found solid backing by industry whenever it was needed.

At the time of the outbreak of World War II, this common bond still existed, on the whole. The Institute had progressed to a respected position as a professional society, a sizeable nucleus of professionals had emerged, the Proceedings had become an outstanding technical journal, and, at the same time, the membership, as a general rule, could still read it and profit by its contents.

In retrospect, this appears to have been a turning point. Since that time there have been changes of some importance, both in the Institute's field of coverage and in the Institute itself.

One of the important changes has been in the increasing complexity and subtlety of the field which may still be thought of, generically, as radio. Radar and pulse techniques, frequency modulation, microwave plumbing, velocity-modulation tubes, studies of signal-to-noise ratio culminating in the concepts of information theory, all may be considered extensions of radio. And all involve theory so advanced, and techniques so complex and refined, that the gap between professionally trained and lay practitioners has become well-nigh unbridgeable. Following its traditional policy of publishing in the Proceedings the best work being done, the Institute has laid such developments before its members. In the process, the Proceedings has obtained an enviable reputation as an outstanding technical publication. At the same time, it has become well-nigh unintelligible to a large fraction of the membership. An important section of the membership has benefited, but service to another fraction of the membership has deteriorated. Cohesiveness has suffered.

Another important change has been the extension of the Institute's interests into fields arising from radio. These have been many. Computers, industrial heating, electronic controls, nuclear instrumentation, solid-state physics, telemetering, servo-mechanisms, all again involve theory and techniques highly developed for radio. Logically they belong in the sphere of interest of the Institute. And the Institute has taken them under its wing. But again, they are primarily matters of interest to the highly trained professional members whose scope of interest is wide. And, inexorably, there has begun to appear a widening gap between the specialists in these different fields.

At the same time that cohesiveness has deteriorated, the membership has increased. At the end of 1940 the membership was 5,705, and, at the end of 1950, 29,002. This five-fold increase, paralleling the growth in the electronics industry, has occurred in the face of the developing loss of cohesiveness, and is a testimonial to the first-rate job that the Institute has done on whatever it has tackled.

* Decimal classification: R060. Excerpts from an address delivered at the IRE Director's dinner meeting, January 9, 1952.

† General Radio Co., Cambridge, Mass.

During this ten-year period the Institute itself has changed. The expanded effort in the radio field, and the addition of related fields, have resulted in a substantial increase in the burden carried by the editorial department. The demands of increased membership and technical committee activity have taxed the resources of the headquarters staff. The most obvious change has therefore been the expansion of the Institute from a $70,000 operation housed in rented quarters to a $700,000 operation housed in its own building.

Less obvious, perhaps, has been the beginning of a change in the Institute's structure. In recognition of the progressive weakening of common interests among the professional membership, the Professional Group movement has appeared. Through this movement a first step has been made toward restoring cohesiveness among segments of the membership interested in specialized fields. In recognition of the increasing geographic dispersion of the membership, the regional plan has been developed. Through this plan a first step has been made toward assuring improved administration of Institute affairs through on-the-spot consideration of specialized geographic needs of the membership and provision of a means for placing these needs before the Board for direct consideration.

In summary, therefore, the Institute has changed from a small, closely-knit society, covering a relatively narrow field and largely representing the northeast section of the country, to a large society covering diversified fields and representing a membership distributed from coast to coast with specialized problems arising from geographical location. As a result of this change many questions are arising about the Institute's service to its members which must be answered to assure further progress. Recommendations regarding the best way to answer these specific questions are being considered by the Ad Hoc Committee on Institute Projects, and I shall not list my personal recommendations item by item.

I should like, however, to present my own opinions regarding the general course of the Institute's future as foreshadowed by the changes that have already occurred and the pattern of the questions now being asked. These opinions are obviously controversial, and time may easily prove them wrong, but they should at least stimulate constructive long-range thinking at a time when, it seems to me, such thinking is most important.

I believe that the Institute, having taken the first step of organizing Professional Groups, has already begun a process of division that will lead to a grouping of specialized segments rather than a single entity. If this is true, it seems to me that there is a very basic issue of organization that must be decided. Superficially the Institute, as a grouping of specialized societies, might resemble the American Institute of Physics. Any resemblance, other than superficial, however, would seem to me wrong. The Institute has a strong editorial department and a strong administrative system through the Sections. It is a virile, well-managed society. I therefore feel that any grouping of specialized segments should be a closely knit union, with a centralized management that can capitalize on existing strengths, rather than a loose federation of semi-autonomous societies. As a first corollary of this, I think that Group TRANSACTIONS should be developed along the lines now being followed, that is, as satellite publications of the Institute serving the needs of specialists, with the PROCEEDINGS remaining the central official organ. As a second corollary, I think that financial support of local chapters of the national Professional Groups should be administered through the Sections rather than through a new parallel path from the Professional Groups themselves. It seems to me that these policies, if consistently adhered to, will result both in an orderly growth of a well-organized central society, and in a re-establishment, in specialized fields, of a good measure of the cohesiveness that has been lost.

I believe that professional societies in general, and the Institute in particular, have a responsibility to solve the problem of horizontal stratification as well as vertical. Modern society is blessed with devices by the score that have been conceived and designed by engineers, but that are used or kept in order by technicians and service men. This subprofessional group must cope with difficult technical problems without professional training. I think that the Institute should give serious consideration, as an organization of professionals, to the problems of these subprofessionals. I do not think that the caliber of the society or its publications need be lowered to accommodate them, but an extension of the group system might supply a place in the organization and a publication that would provide a means for technical advancement, for the development of standards, and for improvement in relations with engineers and with the public.

Finally, I believe that the Institute has already grown so large and has so expanded its fields of interest, that there is no advantage to be gained by limiting its membership, either by inaction or by design. I therefore favor a campaign to attract additional members. I feel, however, that this campaign should be aimed primarily at professional engineers, for whom the Institute is obviously providing useful services. Any major effort to secure subprofessional members should, it seems to me, wait upon the development of services more specifically devised to meet their needs than those now provided.

I hope that these general thoughts will be of some help, as they bear upon future consideration of Institute projects. It would be helpful if there could be full discussion of these matters in all Sections and Professional Groups at an early date, so that, through organizational channels, the Directors might receive the benefit of our members' thoughts on these important matters.

A Look at the Past Helps to Guess at the Future in Electronics*

WILLIAM C. WHITE†, FELLOW, I.R.E.

A Look at the Beginning

NOT ONLY are we near the mid-point of the twentieth century, but radio, as we know it today, had its beginnings from a group of significant advances that center about the opening of the century. These advances were Maxwell's work prior to 1870, Edison's observation in 1880 of what we now call the "Edison effect," and the experiments of Hertz, published in 1887. Although these occurred prior to 1900, none of them resulted in appreciable commercial applications before the turn of the century. Marconi carried on successful experiments in 1895, but it was not until 1901 that his first trans-Atlantic wireless signal was transmitted.

J. J. Thomson evaluated the charge on the electron in 1900, and deForest added a grid to the Fleming valve in 1906. Unfortunately, research, development, and production or application were not organized and co-ordinated to any great extent before 1910. Maxwell's work in postulating electromagnetic radiation lay unused, as far as our science is concerned, until a number of years after his death when Hertz conducted his experiments, the results of which were published in 1887. Hertz, in turn, was dead before Marconi accomplished actual wireless communication. Thus, progress was discontinuous and largely a matter of circumstance in those early days. Real rapid progress in electronics occurred when two commercial laboratories, those of Western Electric and General Electric, began organized work, in about 1912, on the high-vacuum electron tube along the lines we know it today.

There is a certain momentum to a developmental trend and, therefore, a study of the recent past helps to estimate what is likely to occur in the near future. However, it must be kept in mind that all trends sooner or later come to an end. It is interesting to review some of these trends, although there is always a danger of oversimplification and too broad generalizations.

Some Influences of the Past

For about the past thirty-five years, advances in radio have stemmed largely from the development of new classes or types of tubes. The list substantiating this is long and impressive. There are a number of outstanding examples such as the high-vacuum tube which replaced the original audion by 1917 and made detection and audio, as well as radio-frequency amplification, a practical reality. High-vacuum power tubes, sometimes in combination with thoriated filaments, also permitted tube transmitters to be developed so as to replace spark and arc transmitters. Today it is difficult to realize that the use of tubes for transmitting in place of sparks and arcs was not accomplished without difficulty, delay, and only with much promotional effort. As a matter of fact, some of the early tube transmitters utilized the 500-cycle high voltage normally supplied to the spark gaps for the plate power supply. This was done partially from a desire to utilize as much existing transmitter equipment as possible, but mostly from the fact that existing receivers could receive only damped or interrupted radiation.

The screen-grid tube wholly changed the type of circuits used in receivers as well as transmitters. The ac heater tube wholly changed broadcast-receiver design. The hot-cathode mercury-vapor tube stepped up manyfold the currents that could be handled electronically. Even before thyratrons entered the picture to any great extent, the hot-cathode mercury-vapor rectifier tube made a marked change in the design of radio transmitters.

The development of magnetrons, klystrons, and disk-seal triodes made practical the generation and handling of microwave power, and also made possible microwave radar and other applications of these higher frequencies.

The point I wish to make is that this era, as far as radio is concerned, is probably pretty well over. In other words, new tubes will be much less a controlling factor in radio progress from now on. In fact, this trend may well reverse, which means that, in the future, radio requirements will largely tend to create new tube types. This might be considered a mark of maturity in radio and electronics.

However, the process of new tube types formulating new equipment developments is probably not over in the industrial electronics field. This is a more youthful science than radio, and new developments in it were delayed by World War II with its emphasis on microwaves. It is to be suspected also that the glamour of microwaves still is a powerful influence, and industrial electronics is not yet receiving from laboratories and engineering development groups the attention it deserves.

There is an encouraging note, however, in this industrial tube situation. Scientists and engineers have worked for many years on labor-saving devices. This is now largely a matter of accomplishment. The present trend is what may be termed "routine saving." This is the doing of things electrically that require the use of the human senses, such as touch, hearing, seeing, speaking, and smelling, and also those needing a muscular response, but not requiring the thinking part of the brain. The electron tube, of course, is the main tool in this trend.

This idea is old, as are some of its applications, but more recently the broad field has been dignified by the name of "cybernetics," a new word coined by Professor Wiener of the Massachusetts Institute of Technology, who has written books on the subject. The possible impact of this science on our not-too-far-in-the-future lives literally can be called stupendous. As regards changes in the daily routine of many of us, it may well be more of a factor than the atomic bomb and nucleonics. Some of its implications may be visualized from the following extracts from Professor Wiener's book, "Cybernetics."

"If the seventeenth and early eighteenth centuries are the age of clocks, and the later eighteenth and the nineteenth centuries constitute the age of steam-engines, the present time is the age of communication and control. . . . What distinguishes communication engineering from power engineering is that its main interest is not economy of energy but the accurate reproduction of a signal.

"The organs by which impressions (the signals) are received are the equivalents of the human and animal sense organs. They comprise photoelectric cells and other receptors for light; radar systems, receiving their own short Hertzian waves; hydrogen-ion-potential recorders, which may be said to taste; thermometers, pressure-gauges of various sorts, microphones, and so on. The effectors may be electrical motors, or solenoids, or heating coils, or other instruments of very diverse sorts. Between the receptor, or sense organ, and the effector stands an intermediate set of elements whose function is to recombine the incoming impressions into such forms as to produce a desired type of response in the effectors. Moreover, the information received by the automaton need not be used at once, but may be delayed or stored so as to become available at some future time. This is the analogue of memory.

"Perhaps I may clarify the historical background of the present situation if I say that the first industrial revolution . . . was the devaluation of the human arm by competition of machinery. There is no rate of pay at which a United States pick-and-shovel laborer can live which is low enough to compete with the work of a steam shovel as an excavator. The modern industrial revolution is similarly bound to devalue the human brain at least in its simpler and more routine decisions. Of course, just as the skilled carpenter, the skilled mechanic, the skilled dressmaker have in some degree survived the first industrial revolution, so the skilled scientist and the skilled administrator may survive the second. However, taking the second revolution as accomplished, the average human being of mediocre attainments or less has nothing to sell that is worth anyone's money to buy "[1]

* Decimal classification: R090.1×R010. Original manuscript received by the Institute, September 11, 1951.

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[1] Reprinted by permission, N. Wiener, "Cybernetics," published jointly by the Technology Press, John Wiley and Sons, Inc., New York, N. Y., and Hermann et Cie; 1948.

For those of us who are engineers in the field of electronics, the future appears to be at least reasonably encouraging.

Human skills may thus be applied in the future in an amazing variety of ways. Probably the most important factor for the rapidly increasing applications of cybernetics is the utilization of development in electronics. This, of course, arises from the fact that electron tubes of one form or another are directly responsible to some of the "signals" to which the human senses respond, and also are useful in amplifying the output of other detectors of these "signals" to useful levels.

Getting back to radio, ever since Marconi established radio communication in the early years of the century, there has been a constant trend toward the use of higher and higher frequencies. Here again, tube availability has set the pace. Except in special cases, such as radar, which utilize the magnetron and klystron, radio communication by code and voice has largely utilized the space-charge type of tube, that is, triodes and tetrodes, when suitable ones become available.

In the early years, shortening of leads, and the minimizing of distributed inductances and capacitances in other ways, increased the frequency at which triodes could operate. The acorn tube is a good example of this stage in tube development.

The introduction of the disk-seal principle has made triodes commercially available for general use to over 4,000 mc, but it is probable that this limit may be made to approach 10,000. Probably this frequency is about as high as we will wish to go in radio communication because of interference from snow, hail, and rain, as well as some other associated problems. Therefore, it is entirely possible that the space-charge type of tube will be the mainstay for the communication field for many years to come. Of course, let us hope that some now unforeseen uses will develop for the frequencies above 10,000 mc.

At long last, therefore, it would appear that in radio communication, at least, we may nearly have reached the end of that urge to utilize ever higher frequencies, a trend that has been with us for nearly forty years, and that has exerted a constant pressure on tube engineers. However, the problem of more power at the frequencies already in use will probably be with us for a long, long time.

Some Present Trends

There are a number of other present-day trends in electronics that may be significant.

(1) *Increased Use of Electron Beams*

Many of the new outstanding advances in electronics have taken advantage of our increasing knowledge of electron beams, or what is termed the science of electron optics. Examples of this are the betatron, electron microscope, phasitron, radial beam tubes, cyclophon, traveling-wave tubes, television camera tubes, memory tubes, and the gated beam tube. This trend will probably be accelerated, and it may well be that most of the really new and worth-while electron tube advances from now on will be the result of our increased understanding and application of electron optics. Young engineers studying for, or just entering, the field of electronics, as well as those in laboratory groups, might do well to keep this in mind. The number of engineers and physicists with a thorough knowledge of this branch of science, plus the other factors needed for successful research and development, are entirely inadequate to the possibilities that exist.

(2) *High-Vacuum Techniques*

The production of high vacuua and the many techniques connected with its use are no longer a mysterious occupation limited to a few hundred individuals. It is now a well-established engineering science. There are companies, I think, that would undertake the contract of pumping down an inclosure the size of a classroom to a one-micron pressure, if it could be properly surfaced and also built to withstand the pressures structurally. Modern developments for aircraft flights at very high altitudes and the activity on guided missiles are also helping us to learn how to do many things in a vacuum. There is reason to believe, therefore, that the continuously evacuated or pumped tube may re-enter the picture for certain applications.

(3) *Wide Use of the Chemical Elements*

The manufacture of vacuum tubes has always been characterized by wide use of the lesser-known chemical elements. More than 30 of the elements are commonly employed by the tube engineer, and they span the alphabet from argon to xenon and zirconium. This broad utilization of the list of elements will undoubtedly continue and expand.

(4) *Use of Ceramics*

In many tube applications, glass is almost sure to disappear during the coming years, and will be replaced by ceramic materials. This change allows exhaust and operation at higher temperatures and, in the case of microwave tubes, permits more accurate dimensions and lowers the dielectric loss in the insulating material.

(5) *Developments*

Developments of the past few years in the field of semiconductors have given engineers the crystal diode and the transistor. These devices will undoubtedly replace tubes for many applications that were originally brought into being by the availability of electron tubes. This is no new trend; improved dry type power rectifiers have already replaced many low voltage tubes. Also nonlinear electromagnetic devices and the amplidyne have for some applications superseded thyratron controls. However, electronics has many new areas under exploration, and there is still plenty of frontier territory to provide for continued expansion of our science for a long time to come.

A Present Problem

Probably the most discussed problem in the application of electronics today is that of reliability. This problem has resulted, of course, from the very large number of simultaneously used tubes and circuits which today characterize computers and military applications.

It is becoming evident that no single improvement or accomplishment, such as longer average tube life, will provide the answer to this complicated problem. Probably the general nature of the solution can best be foreseen by recalling the history of the long struggle to obtain safety in railroad operation. During the latter half of the last century, railroad accidents, both to passengers and employees, increased to such an alarming extent that drastic action was necessary. As in the case of tube reliability, there was no one factor which could be improved to obtain the desired result. After some twenty-five years or more of intensive work, tremendous strides were made toward safety. It was accomplished by attacking the problem on many fronts. This, in addition to new developments, involved new rules and regulations, inspection techniques, specifications, and standardization, as well as legislation. One price paid for this marked improvement in safety has been the cautious approach to change which has characterized the railroads and has been a factor in delaying certain improvements.

In a similar way, we must expect that improvement in reliability of electronic equipment will be accompanied by much more rigid and detailed specifications and regulations, more expensive equipment, and other time-consuming requirements and approvals. The trend to smaller and lighter equipment may also for a time be slowed or even reversed.

It will also mean that it will become much more difficult to introduce new ideas and inventions. Only by sticking pretty closely to certain proved features will it be possible to obtain the desired reliability. Radio engineers will chafe at the restrictions and resistance to change which will be one price of increased reliability.

To what extent the coming years in radio and electronics will follow some of the patterns outlined in this article must, for the present, be a matter of opinion; but we can be sure of one thing, they will be busy, challenging, and interesting years for engineers in the field of electronics.

Service Range for Air-to-Ground and Air-to-Air Communications at Frequencies above 50 Mc*

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This paper appears with the approval of the Tutorial Papers Subcommittee of the IRE Committee on Education.—*The Editor.*

Summary—Propagation aspects of air-to-ground and air-to-air communications are analyzed. Contours of constant received signal strength are shown in the form of lobes for various frequencies. It is shown that for systems with equivalent transmitted power, ground-antenna height, and transmitting- and receiving-antenna gain the service range decreases as the frequency is increased. This is due primarily to a decrease in the absorbing area of the receiving antenna and to a larger number of nulls in the lobe structure arising from interference between direct and ground-reflected waves. Ground-station antenna-height diversity and tilted-array ground-antenna systems are discussed as a means of improving coverage as the operating frequency is increased.

I. Introduction

In 1947 the U. S. Air Forces requested the National Bureau of Standards to study propagation aspects of communications systems in the proposed aircraft communications band from 225 to 400 mc. This study was carried out in two parts: air-to-ground communications and air-to-air communications. A considerable amount of theoretical information was furnished to the Department of Defense. In addition, members of the Central Radio Propagation Laboratory participated actively in flight evaluation tests held in 1948 and 1949 and conducted by the U. S. Navy for the determination of propagation characteristics at these frequencies. The results of these tests were made the subject of a report by the U. S. Navy[1] in which the experimental data supported the theoretical treatment to a very high degree. The discussion of the air-to-air problem is based in part on a paper presented at a Naval Electronic Laboratory Symposium.[2]

II. Propagation Aspects of Communications

A. Free-Space Propagation

Consider the transmitter and receiver of a communications circuit to be in free space with no ground or obstructions in any way influencing the circuit. Under these conditions, several well-known fundamental relations exist between transmitter power, antenna gain, field strength, receiver sensitivity, and maximum range of communications.[3,4] First of all, field strength varies inversely as the distance from the source. Furthermore, the field strength at any point in space from an antenna radiating a given amount of power and with a given radiation pattern will be independent of frequency. For instance, for a half-wave dipole antenna radiating one kw, the free-space field strength, E_0, in the direction of maximum radiation at a distance of one mile is 137.6 millivolts per meter for any frequency. The absorbing area of receiving antennas of equivalent directivity decreases with increasing frequency, resulting in less available power at the receiver terminals for given field strengths. It has been shown for these same antennas that in order to deliver a constant voltage across the input of the receiver the field strength must increase in proportion to the frequency. It follows from these considerations that the free-space maximum range for such communications systems is inversely proportional to frequency.

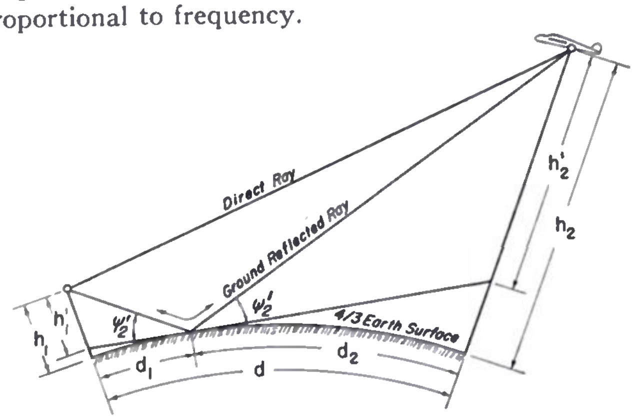

Fig. 1—Geometry for air-to-ground communications.

B. Propagation over the Smooth Earth in the Average Washington, D. C., Atmosphere

Over a smooth spherical earth the space wave at the receiving antenna is composed of a direct wave corresponding to the free-space wave and a ground-reflected wave. Fig. 1 shows the physical representation

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† National Bureau of Standards, Central Radio Propagation Laboratory, Boulder, Colo.

[1] "Ultra High Frequency Propagation Tests over Land and Water," U. S. Naval Air Test Center, Patuxent River, Md., Appendices A, B, and C, Project TED No. PTR EL 929; December, 1949.

[2] K. A. Norton, "Scattering of Radio Waves in Propagation over Irregular Terrain," Presented as a Part of a Symposium on Radio Wave Propagation at the Naval Electronics Laboratory, San Diego, Calif., July 28, 1949.

[3] K. A. Norton, "The calculation of ground-wave field intensity over a finitely conducting spherical earth," Proc. I.R.E., vol. 29, pp. 623–639; December, 1941.

[4] "Propagation of Radio Waves through the Standard Atmosphere," NDRC Committee on Propagation, Summary Technical Report, Washington, D. C., vol. 3, Chap. 1; 1946.

of these waves. The space-wave field strength is the vectorial summation of the direct and ground-reflected wave fields, and for 100-per cent reflection over a plane surface it would vary between zero and twice the free-space field strength. If contours of equal field strength are drawn, they take the form of lobes in vertical cross section. The position of the lobes depends on ground-antenna height, frequency, and, to a lesser extent, on polarization and ground constants. In general, there is a lobe for every half wavelength in height of the ground antenna. Factors are further discussed in appendix.

The magnitude of the ground-reflected wave relative to the free-space wave can be determined from a consideration of the frequency, angle of incidence, polarization, and the ground constants.

The ground-reflected wave is attenuated by a factor which varies between unity and zero, and undergoes a phase transformation as well. As an example, Fig. 2 shows typical values of the reflection coefficient, $|R'|$ and phase angle, c (as defined in the appendix), of the ground-reflection coefficients for horizontal and vertical polarization. In addition to the above changes due to penetration and absorption, the ground-reflected wave energy is diverged upon reflection from a curved surface, such as the spherical earth. For practical purposes this effect is negligible except at low elevation angles, where it should be considered.

The ordinary large-scale atmospheric gradients of refractive index generally cause radio rays to be bent slightly downward. The amount and manner of bending that the ray undergoes at points within the line of sight vary somewhat with time, geographic location, and with altitude. Consequently, for our study of the systematic effects of propagation on the service range we are concerned primarily with the average effect of refraction, and neglect these instantaneous effects. To a very good first approximation the average effect of refraction can be allowed for by assuming that we are dealing with an earth whose radius is four-thirds the actual radius, and that this larger earth has no atmospheric refraction. This will hereafter be called the "four-thirds earth." Second-order corrections can and have been made in this paper. These corrections are quite small and can be neglected completely without risk of appreciable error in the final result. For the reader who is interested in such second-order corrections, their theory and application is indicated in the appendix.

III. Representation of Coverage

The region within which the direct and ground-reflected waves make up the primary mode of propagation is sometimes referred to as the "interference region," and is considered to be that portion of space lying above the radio horizon. The calculations in this paper deal primarily with this region although some consideration is given the diffraction region below the horizon.

In order to show the systematic effects of propagation on the expected coverage, the communications systems are assumed, for the examples in this paper, to have characteristics which are considered typical of an operational air-to-ground system. The transmitter power is assumed to be 6 watts. Service range is shown in terms of the input voltage across the terminals of a receiver with a 50-ohm input impedance. Contours are shown for 3-μv, 6-μv, 12-μv, and 24-μv receiver input voltage. The contour representing maximum coverage which

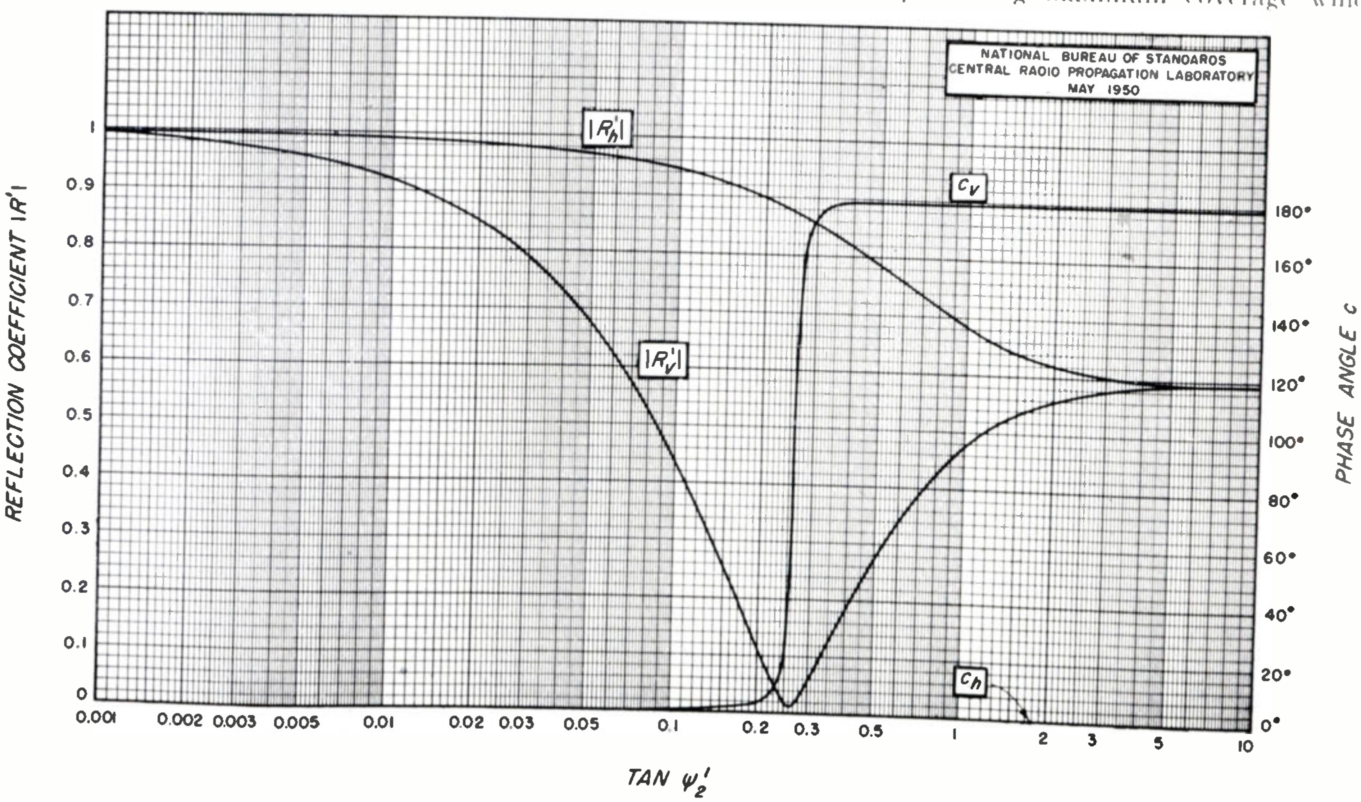

Fig. 2—Magnitude and phase angle of the reflection coefficient for 328 mc. Dielectric constant, =15; conductivity, $=10^{-2}$ mhos/meter; horizontal and vertical polarization.

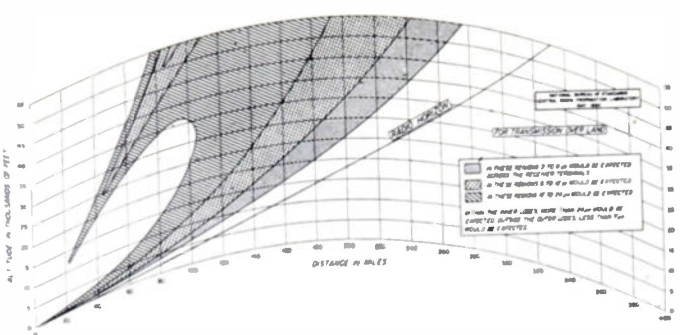

Fig. 3—139-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole; aircraft antenna, half-wave vertical diople; power, 6 watts; height, 35 feet; vertical polarization; assumed communication-system loss, 6 db.

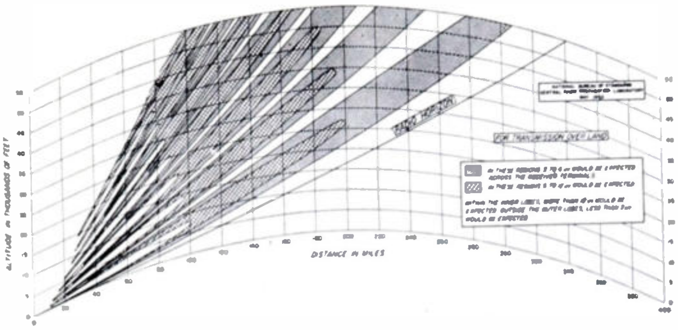

Fig. 6—328-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole; aircraft antenna, half-wave vertical dipole; power, 6 watts; height, 115 feet; vertical polarization; assumed communication-system loss, 6 db.

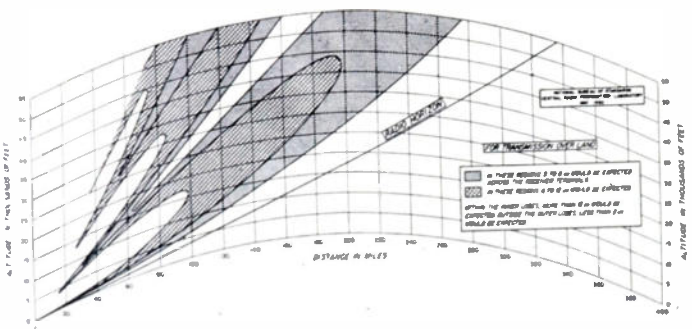

Fig. 4—328-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole; aircraft antenna, half-wave vertical diople; power, 6 watts; height, 35 feet; vertical polarization; assumed communication-system loss, 6 db.

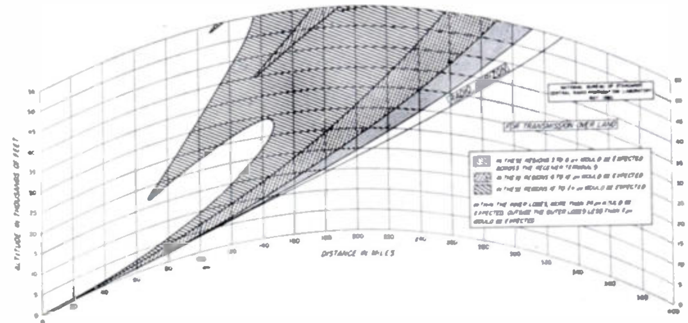

Fig. 7—328-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, eight-element collinear array tilted upward 7.2 degrees; aircraft antenna; half-wave vertical dipole; power, 6 watts; height, 35 feet; vertical polarization; assumed communication-system loss, 6 db.

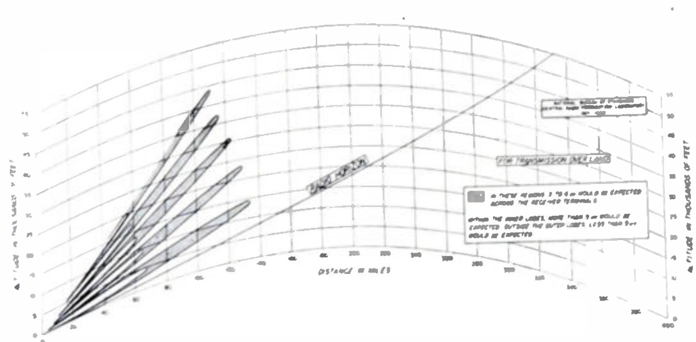

Fig. 5—1,000-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole; aircraft antenna, half-wave vertical dipole; power, 6 watts; height, 35 feet; vertical polarization; assumed communication-system loss, 6 db.

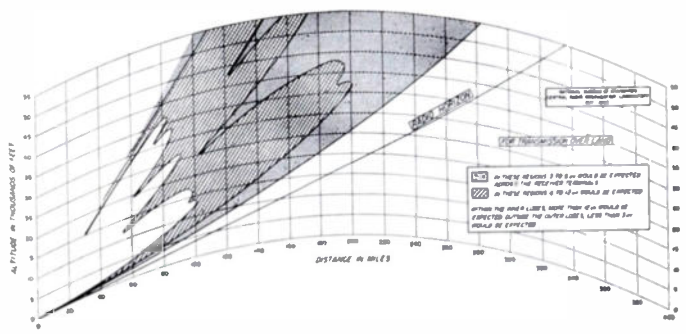

Fig. 8—328-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole (height-diversity reception); aircraft antenna, half-wave vertical dipole; power, 6 watts; height, 35 and 50 feet; vertical polarization; assumed communication-system loss, 6 db.

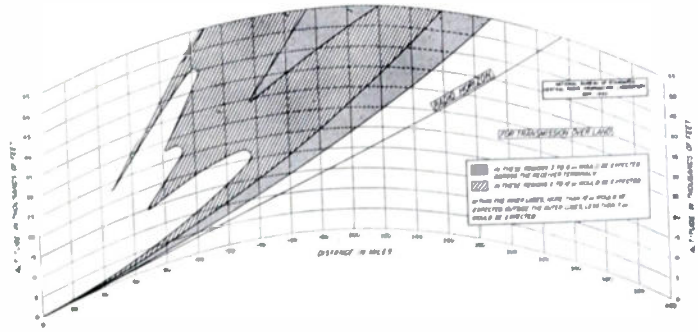

Fig. 9—328-mc radiation pattern for air-to-ground communication over a smooth spherical earth. Ground antenna, half-wave vertical dipole (height-diversity transmission); aircraft antenna, half-wave vertical dipole; power, 6 watts; height, 35 and 50 feet; vertical polarization; assumed communication-system loss, 6 db.

would be applicable to a given communications system would depend to a large extent upon the effective sensitivity of the receivers. From various experiments it has been determined that a value of approximately 3 μv for the minimum usable receiver input signal voltage is representative of operational receivers in good working order. This value is approximately 20 db above kTB noise, assuming 100-kc noise bandwidth. In addition, transmission-line, mismatch, and other circuit losses between transmitter, receiver, and associated antennas are lumped together as a communications system loss. A factor of 6 db is considered typical of such losses, and the voltage contours for the examples in this paper have been prepared assuming this 6-db over-all system loss. Considerable variation from these values can be expected in practice, with a consequent modification of the ranges of communications obtained. These modifications may be considered in terms of the communications system loss; for example, with 12-db system loss our 6-μv contour rather than our 3-μv contour would show the region of communications.

Figs. 3 through 9 show smooth-earth lobe diagrams calculated using the methods outlined in the appendix for various conditions of frequency, polarization, ground-antenna height, and types of ground antennas. Figs. 3 through 5 show the effect on the coverage of increasing the radio frequency. With increasing frequency the number of lobes increases and has a finer structure, while the free-space maximum range decreases. From the above considerations, the free-space maximum range for 3-μv receiver input voltage would be expected to vary under these average conditions inversely with frequency as follows: 139 mc, 509 miles; 328 mc, 214 miles; 1,000 mc, 70 miles. These ranges are characteristic only of the specific conditions assumed, and it should be remembered that receiver and antenna variations can increase them by a factor of 2 or decrease them by a factor of 10, approximately. In the maxima of the lobes the distances to the 3-μv contours are nearly twice these values. Cancellation in the minima is not complete since some of the energy in the ground-reflected wave is absorbed upon reflection. For horizontal polarization, reflection is much more complete so that very deep nulls occur and the phase angle, c, is little different from zero at all angles of incidence. A comparison of Figs. 4 and 6 shows the effect on the lobe structure of increasing the height of the ground antenna. There are more lobes with finer structure for the higher antenna, which indicates, if this is to be avoided, a requirement for an upper limit to the height of the ground antenna.

Antennas encountered in operational installations, in particular those mounted on the aircraft, can be expected to deviate considerably from the pattern of the idealized half-wave dipole. This is largely a matter of directivity, aircraft antenna radiation patterns in general having many lobes. An analysis of the horizontal radiation patterns of three different antennas, operating at frequencies scaled to be the equivalent of from 130 to 175 mc when mounted on a model of a P2V-type aircraft, shows the radiation or reception with reference to an idealized half-wave dipole to be distributed in azimuth as follows: 10 per cent of the directions more than +1.5 db; 50 per cent of the directions more than −3.4 db; 90 per cent of the directions more than −11.0 db. These data are too meager to be considered representative of aircraft antenna directivity variations, but they provide some idea of the allowance which must be made for antenna directivity effects in order to ensure reliable operation. Thus, based on the above data, the communications system loss would be increased to the order of 18 db, and thus the service area would be expected to lie within our 12-μv lobes if operation is to be reliable for 90 per cent of the possible aircraft orientations in level flight.

IV. Operational Aspects

In trying to communicate with the ground, an aircraft flying toward a ground installation at a constant altitude would find, upon entering the lowest lobe, that it would be possible to communicate for some distance until a null is reached, whereupon communications may be impossible until the next lobe is reached. This might occur several times, depending upon the particular manner in which the circuit was set up, i.e., height of aircraft, frequency, and the like. Fig. 10 shows theoretical and experimental curves of receiver input voltage versus distance for an aircraft approaching a ground station at an altitude of 10,000 feet. For this figure a ground-antenna height of 75 feet is used with a frequency of 328 mc and transmission is over water. The three graphs show the effect of polarization on the received signal strength. The experimental curves were obtained simultaneously on three receivers, recording the output of three types of antennas used for picking up the signals emanating from a circularly polarized radiator on the aircraft. For unbiased comparison the voltages for the vertically and horizontally polarized antennas were adjusted upward by 3 db since only half of the power is radiated in either plane component of polarization. The different effects on the lobes are clearly shown. The nulls are very deep with horizontal polarization and communication "drop outs" occur through the higher nulls even at very short ranges. With vertical polarization, the nulls are very evident; but because of more absorption of the ground-reflected energy, they are not nearly so deep. With circular polarization, and at the low angles of elevation where the lobes for each polarization component are nearly superimposed, the field strength in the nulls lies somewhere between that for horizontal polarization and that for vertical polarization. At the higher angles of elevation the two lobe structures are more nearly interposed and some filling in of the nulls occurs under some conditions. It should be noted that the results shown on this figure were obtained with the antennas specially mounted on the aircraft in such a way as to minimize the variations in directivity and the aircraft flew with a constant orientation relative to the ground station. Even under these circumstances there were some departures from the theory, and these are attributed to residual variations in aircraft directivity, instantaneous effects of the atmosphere, and irregularities in the reflecting surface of the water. Similar measurements were made over land with smooth terrain in the vicinity of the ground station, and the results were similar to those shown in Fig. 10. No air-to-ground measurements were made over rough terrain, but some analysis of its expected effects on the lobe structure is given later in this paper.

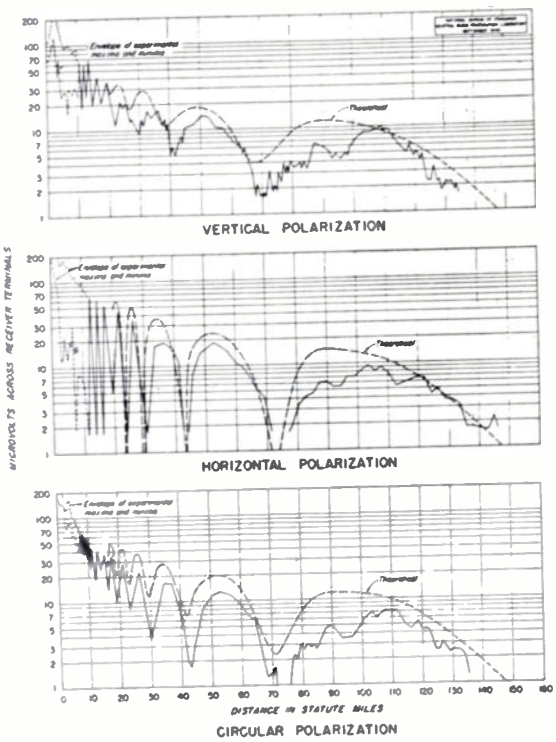

Fig. 10—Observed input voltage variation at ground-station receiver from an aircraft at 10,000 feet transmitting on 328.2 mc. Transmitter power, 6 watts; transmitting and receiving antenna gain, 2.15 db (relative to an isotropic); ground-antenna height, 75 feet; transmission over water; 6 db communication-system loss assumed for theoretical curves.

Because of the serious effect of deep minima upon the reliability of a system for air-to-ground communications, it is felt that the proper way to make coverage comparisons of various systems would involve consideration of combinations of maximum altitudes and distances within which no communication "drop outs" will be expected to occur. These altitudes and distances can be maximized, in general, by decreasing the height of the ground antenna, thus reducing the number of lobes. This, however, decreases the distances at given altitudes to the lower 3-μv contour of the lowest lobe, which is a measure of the low-angle coverage. It is obvious in selecting the height of ground antennas that consideration must necessarily involve a compromise in these factors. It is interesting to note that for a given ground-antenna height, low-angle coverage is relatively independent of frequency. As the frequency increases, the lowest lobe becomes lower, which tends to increase the low-angle coverage; but, at the same time, the free-space maximum range is decreased. In general these factors are of comparable magnitude, and little modification of the low-angle coverage is involved.

V. Methods of Improving Coverage

Recently several methods for improving coverage which do not involve an increase in transmitter power have been proposed and investigated. The most promising methods are those using the following ground-antenna systems:

1. High-gain ground antennas.
2. Tilted-array ground antennas.
3. Height-diversity ground antennas.

The necessity for preserving and improving coverage becomes more pressing as frequencies of operation are increased and as altitude, speed, and range of aircraft are increased. It is interesting to note that lower-frequency coverage patterns can be almost exactly duplicated for higher frequencies by increasing the power in proportion to the square of the ratio of frequencies and using ground-antenna heights of equal numbers of wavelengths. However, the size and weight of higher-powered transmitters limit their use in aircraft.

A. High-Gain Ground Antennas

The use of high-gain directive ground antennas accomplishes essentially the same result as increasing the transmitter power for both the aircraft and the ground station. For transmission from the ground using this type of antenna, the effective radiated power is increased with a corresponding increase in the free-space maximum range. For reception at the ground station, the absorbing area of the antenna is increased and the minimum usable receiver terminal voltage can be obtained from weaker field strengths. This results in an increase in the free-space maximum range corresponding to the transmission case. There are, of course, limits to the feasibility of this method arising from the corresponding reduction in high-angle coverage. However, high-angle coverage usually represents short ranges of communications and the radiated power in these directions need not be as high as for the longer distances.

B. Tilted-Array Ground Antennas

Further improvement with the use of high-gain arrays can be realized by tilting the antenna beam slightly upward. This reduces the intensity of the ground-reflected ray and increases the direct ray, thus substantially filling up the minima with only a small reduction in the maxima. This technique was first proposed by Norton and Omberg[5] to improve radar coverage. Fig. 11 shows the geometry involved in improving coverage with a

[5] K. A. Norton and A. C. Omberg, "The Maximum Range of a Radar Set," Report ORG-9-1, Operation Research Staff, Office of the Chief Signal Officer; February, 1943. Later published, Proc. I.R.E., vol. 35, pp. 4–24; January, 1947.

directive antenna. The angle of tilt here was determined to give maximum improvement in the direction of the null above the first lobe.

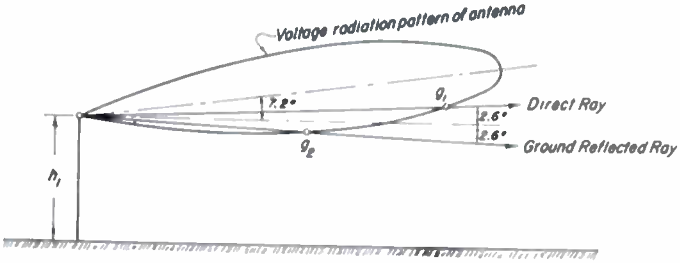

Fig. 11—Suppression of ground-reflected ray obtained by use of high-gain tilted array. Eight-element collinear array phased to tilt pattern upward 7.2 degrees in all directions. Null above first lobe occurs at an elevation of 2.6 degrees with a ground-antenna height of 35 feet and a frequency of 328 mc.
g_1 = antenna voltage-gain factor for direct ray.
g_2 = antenna voltage-gain factor for ground-reflected ray.

C. Height-Diversity Ground Antennas

The use of height diversity in the ground-antenna system makes possible the superpositioning of lobes in such a manner that the minima do not coincide, and filling-in is accomplished. For diversity reception two receivers are used. Reception is first on one receiver and then on the other, depending upon which antenna is in the stronger field; in practice, the two receivers can be combined into one unit having separate RF and IF stages with a common audio section. Fig. 8 shows the expected coverage for a diversity receiving system operating at 328 mc; comparing this with Fig. 4 shows how the minima for the 35-foot antenna are filled in by the lobes from the 50-foot antenna. In order to transmit from these same antennas to the aircraft, a method of transmission is necessary which maintains independence between the waves travelling over the separate paths. Gates[6] outlines one method of doing this which requires the use of two transmitters, one for each of the diversity antennas, transmitting on slightly different frequencies. An example of the expected coverage with this system, using two 6-watt transmitters, is shown in Fig. 9.

In order to determine the heights of the ground antennas for optimum coverage, consideration is given to the diversity-reception problem, which, because of the manner in which the signals are combined, yields somewhat less coverage than the diversity-transmission case. With height diversity, the primary limitation to coverage occurs at the points of lobe crossover. In order to maximize coverage within the 3-μv contours, it is necessary to choose antenna heights such that the 3-μv lobes cross over at the maximum altitude. This is accomplished, in general, when the two lowest points of lobe crossover occur at the same altitude. Over the range of frequencies from 225 to 400 mc, a ratio of heights for the ground antennas of approximately 7 to 10 gives this result.

The use of high-gain antennas with height diversity increases the coverage obtained corresponding to the increase in the free-space maximum ranges obtained but with a consequent reduction of high-angle coverage.

[6] H. P. Gates, "Evaluation of Diversity Systems," USNEL Report No. 80; October 8, 1948.

VI. Polarization Diversity

Circularly polarized transmission may be employed as a form of diversity. Independence between the transmission paths is effected by the polarization. Referring to Fig. 2, it is interesting to note that the phase angle, c, of the reflection coefficient differs for vertically and horizontally polarized waves; hence, the two ground-reflected components cannot arrive simultaneously out of phase with the direct-wave components. For perfect conducting reflecting surfaces, the phase difference of the two ground-reflected components is 180 degrees, and under this condition the maxima and minima for the two polarizations considered separately would be perfectly interposed. However, with ground constants of any practical interest, this effect is very slight at angles of elevation below the Brewster angle. The most pronounced case is that of sea water, as shown in Fig. 10. Here the Brewster angle is fairly low. For poorer conducting surfaces, such as land, the effect is even less noticeable. At angles above the Brewster angle, the maxima and minima are more nearly interposed; but these angles correspond to short ranges of communications, and the effect is of little value here.

VII. Irregular Terrain

In Fig. 1 ground-reflected rays are shown for simplicity to occur at a point. Actually, the entire surface of the earth is illuminated and reradiates elementary waves in all directions. Over smooth earth at any particular receiving location, the resulting intensity of all these waves very nearly equals that of the waves reflected from within a small elliptical area in the neighborhood of the ray reflection point as determined by the laws of geometrical optics. This elliptical area is called a Fresnel zone. The intensity and phase relationships of the remaining waves very nearly cancel each other out.

The length of the ray path to the edge of the entire first Fresnel zone is one-half wavelength longer than the geometrical ray path. It is within this zone that irregularities in terrain would be expected to have the greatest effect upon the reflected signal. In general, the effect of irregularities is to cause filling-in of the lobe minima and less development of the lobe maxima. The dimensions of the major and minor axes of the first Fresnel zone ellipses for the maximum of the first lobe are tabulated for comparison, using a ground-antenna height of 35 feet.

TABLE I

Frequency	Major axis	Minor axis	d_1	d to center of ellipse
139	0.71 mile	100 feet	0.125 mile	0.38 mile
243	1.30 miles	100 "	0.229 "	0.69 "
328	1.76 "	100 "	0.300 "	0.93 "
1,000	5.33 "	100 "	0.944 "	2.83 miles

The centers of the ellipses are displaced from the point of geometric ray-path reflection at d_1 in the direction along the major axis toward the higher antenna. As the grazing angle increases, the ellipses become smaller, thus restricting the area within which irregularities are of greatest importance. Norton and Omberg[5] have shown that the permissible height deviation within the first Fresnel zone for a well-developed k^{th} lobe is $\Delta h = h_1/4(2k-1)$, where h_1 is the ground-antenna height.

When the irregularities in height are large with respect to the above limits, they will be important outside as well as inside the first Fresnel zone; but in this case the average energy of the ground-reflected components is small and the net effect on the received field is correspondingly small.

VIII. Air-to-Air Propagation

As the height of the ground antenna is increased, a very much larger number of lobes appear and larger areas of the earth's surface become of primary importance in considering the effects of reflection. This is the case for air-to-air propagation. The first Fresnel zone areas become quite large and are no longer systematically located in the vicinity of airport areas where the surrounding terrain is likely to be relatively smooth.

Thus a statistical approach is indicated for propagation between aircraft flying over irregular terrain. When the terrain is sufficiently irregular so as to produce ground-reflected path lengths deviating from smooth-earth conditions by as much as $\lambda/8$, the lobe structure begins to break down. When this irregularity is sufficient to produce ground-reflected path-length deviations of, for instance, one wavelength and the irregularities are at random, the phase of the ground-reflected portion of the space wave is assumed to be random and its intensity Rayleigh distributed. This randomness probably occurs largely with respect to variations in space only, but, for a moving aircraft, appears to be a function of time.

An example of the application of this approach is shown in Fig. 12. Here coverage is shown in terms of the percentage of time the receiver input voltage for a moving aircraft exceeds a minimum usable value. As before, the propagation factors assumed are as follows: transmitter power, 6 watts; transmitting and receiver antenna power gain, 1.64 relative to an isotropic antenna; 6-db communications-system loss; vertical polarization; and an operating frequency of 328 mc. In addition, it is assumed that height deviations from smooth-earth conditions are as great as 100 feet, particularly within the area of the first Fresnel zone. Above the shaded area the space wave is considered to be made up of a direct wave plus a Rayleigh-distributed ground-reflected wave. The several curves on the figure show range reliability in terms of the percentage of time the 3-μv receiver input voltage is expected to be exceeded. The region within which the terrain irregularities produce path-length differences from smooth-earth conditions of between one-eighth wavelength and one wavelength is shown shaded with slant lines. This region is considered to be a transition region between the smooth-

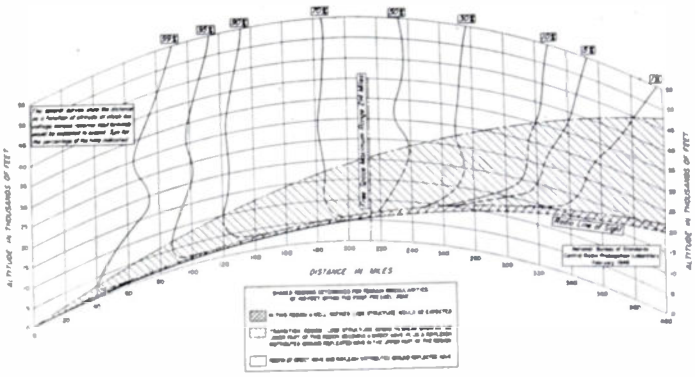

Fig. 12—Received field distribution for air-to-air communication on 328 mc over irregular terrain for two aircraft flying at the same altitude. Transmitting antenna gain, 2.15 db (relative to an isotropic); power, 6 watts; vertical polarization; receiving antenna gain, 2.15 db; assumed communication-system loss, 6 db.

earth ground-reflected wave and the random-phased ground-reflected wave. In this region the range-reliability curves are shown dashed and represent the expected reliability if the ground-reflected wave were Rayleigh-distributed. In the cross-hatched region essentially specular reflection is expected with well-developed lobes.

The power required on other radio frequencies to provide the same grade of service at the ranges indicated on Fig. 12 for 328 mc will be directly proportional to the square of the frequency, i.e., one tenth as much power is required on 104 mc and ten times as much power on 1,040 mc.

Fig. 13 shows two samples of field-strength recordings for 328-mc air-to-air propagation over land. The upper sample corresponds to the case in which the two aircraft

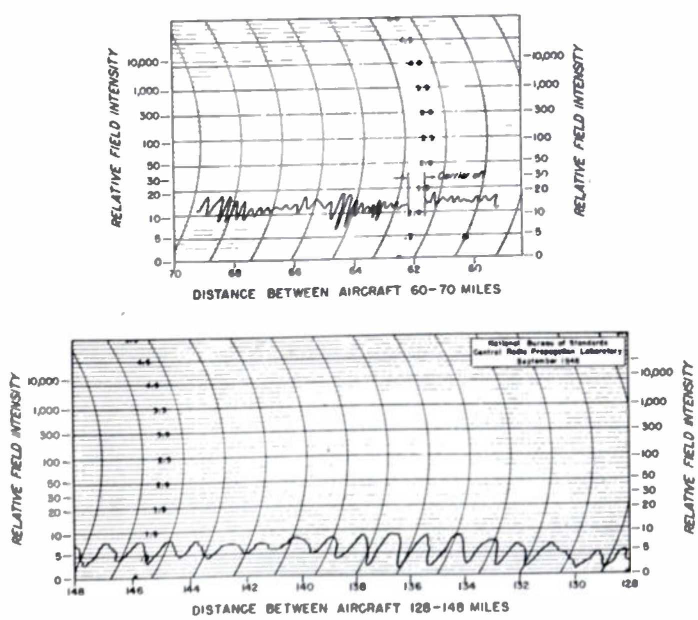

Fig. 13—Field-strength variations observed in air-to-air propagation over irregular terrain. Altitude of both aircraft, 10,000 feet; frequency, 328.2 mc.

are 60 to 70 miles apart, while the lower sample corresponds to the case in which they are 128 to 148 miles apart. Calculations show that the fading minima and maxima are spaced with respect to the distance just about the amount expected for interference between direct and ground-reflected waves at this frequency. Under these conditions, the field-strength maxima should be equal to the sum of the direct-wave amplitude plus the ground-reflected wave amplitude, and the field-strength minima should be their difference. By measuring the relative lobe maxima and minima, it is possible to calculate the value of the ground-reflection coefficient. A separate value of this ground-reflection coefficient can be obtained from each maximum and succeeding minimum.

The distributions of the ground-reflection coefficients obtained in this manner from the data shown in Fig. 13 are shown in Fig. 14. In this figure the quantity, E_R'/E_D', is the rms value of the distribution of E_R/E_D. It should be pointed out that the scales on the graph are arranged in such a manner that Rayleigh-distributed

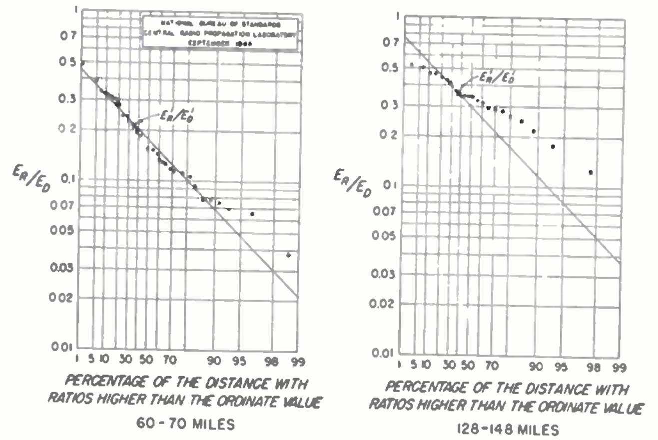

Fig. 14—Distribution of observed effective strength of ground-reflected wave, E_R, relative to the direct wave, E_D, for two aircraft flying over irregular terrain. Frequency, 328 mc; altitude, 10,000 feet. Values of E_R/E_D distributed in accordance with the Rayleigh distribution, $P=100\, e^{-(E_R/E_D)^2/(E_R'/E_D')^2}$, would be on a straight line with slope -1. (From data taken by Collins Radio Co., Cedar Rapids, Iowa).

points fall along a straight line with a slope of minus one. It will be noted that the measurements are Rayleigh distributed at the shorter of the two distance ranges corresponding to a grazing angle at the earth's surface of about 3 degrees. The data at the larger range would not be expected to agree as well with the Rayleigh distribution because of the larger angle of incidence (nearly 89 degrees in this case). Thus it appears that the ground is beginning to appear more nearly smooth to the radio waves at the grazing angle of only one degree.

Meteorological conditions produce pronounced effects upon propagation between aircraft even when they are well within line of sight of each other. The reason for this is evidently the fact that waves between two points, both well above the earth, are incident on the tropospheric layers of variable refractive index at more nearly grazing angles of incidence. In summarizing a large number of experiments and theoretical analyses, Cornell University[7] reports three types of disturbances which cannot be explained in terms of ground reflections and which are evidently associated with the tropospheric propagation medium. These disturbances are (a) intervals of range in which the average signal has decreased markedly (b) intervals of range in which the average signal exceeds the free-space value by as much as 10 or 12 db and (c) intervals of range in which interference-type lobes of large amplitude appear. To date no quantitative information exists indicating a general average of the above conditions. Consequently, no allowance was made for these effects in the theoretical curves of Fig. 12.

Further work on the above-described statistical theory of air-to-air propagation is now in progress at the National Bureau of Standards, and it is expected that further details will soon be available for publication.

IX. Conclusion

The propagation problems involved in the service ranges for air-to-ground and air-to-air communications are primarily the result of lobes caused by interference between the direct and ground-reflected rays as well as a systematic decrease in free-space maximum range with increasing frequency. Coverage diagrams are shown for varying conditions of frequency, polarization, ground-antenna heights, and so on. To form a systematic basis for comparison, the minimum usable receiver input voltage is assumed to be a constant value of 3 μv and antenna radiation patterns are idealized. Wide variations from these conditions can be expected in practice. For simplicity, the effects of the variations can be allowed for by modifying the value assumed for communications-system loss. In general, it is shown that as the frequency of operation is increased, aside from equipment becoming more complex, the propagation characteristics become less suitable for communications. The lobe structure becomes finer with more nulls to contend with and the absorbing area of antennas decreases, thus decreasing the ranges of communications. These restrictions can be overcome within limits by resorting to the use of elaborate ground installations, such as high-gain tilted arrays and height-diversity transmission and reception. Because of the severe restrictions in size and weight of equipment to be carried in the aircraft, little can be done to improve communications at that end. Antennas for aircraft must be relatively nondirective, and thus necessarily be of low gain. If the limitations in spectrum space require the use of higher frequencies, the methods outlined in this paper should prove beneficial in improving coverage.

[7] School of Electrical Engineering, "Summary Report, Air-to-Air and Air-to-Ground Electromagnetic Propagation," Final Report Part II, Prepared under U. S. Air Force Contract AF33(038)-1091, Cornell University, Ithaca, N. Y.; June 1, 1951.

It is shown that air-to-air propagation conditions in this frequency range may advantageously be treated by statistical methods.

ACKNOWLEDGMENT

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APPENDIX

A. Calculation of Coverage Diagrams

A discussion of the procedures used in calculating the coverage diagrams of this paper is presented here to enable the reader to make similar calculations as well as to show the application and limitations to the use of the methods and of the approximations. In the region well above the radio horizon the curves were determined in accordance with the interference theory at large heights, while those well below the radio horizon were determined in accordance with the diffraction theory. That portion of the curves in the immediate vicinity of the radio horizon was obtained by interpolating between these results.

Ground constants representative of the terrain involved are chosen for the determination of reflection coefficients. In this paper, values of the dielectric constant, $\epsilon=15$, and conductivity, $\sigma=10^{-2}$ mhos/meter, were chosen as being representative of overland propagation. For sea-water calculations these values become $\epsilon=81$ and $\sigma=4.64$ mhos/meter. The plane-wave reflection coefficient, $|R'|$, and its phase angle, c, are determined as a function of the ground constants, frequency, and elevation angles by the following equations:

$$R_v' = \frac{n^2 \sin \psi_2' - \sqrt{n^2 - \cos^2 \psi_2'}}{n^2 \sin \psi_2' + \sqrt{n^2 - \cos^2 \psi_2'}} \equiv |R_v'| e^{i(\pi - c_v)} \equiv -|R_v'| e^{-ic_v}$$

(vertical polarization) (1)

and

$$R_h' = \frac{\sin \psi_2' - \sqrt{n^2 - \cos^2 \psi_2'}}{\sin \psi_2' + \sqrt{n^2 - \cos^2 \psi_2'}} \equiv |R_h'| e^{i(\pi - c_h)} \equiv -|R_h'| e^{-ic_h},$$

(horizontal polarization) (2)

where $n^2=\epsilon+ix$, $x=1.79731\cdot 10^4\cdot(\sigma/f_{mc})$, and ψ_2' is the angle the incident ray makes with the reflecting surface. The angle c_v is an angle between 0 and π and equals $\pi/2$ at the pseudo Brewster angle of incidence corresponding to the minimum reflection; c_h is a small negative angle at all angles of incidence. As a typical example, the values of $|R'|$ and c are plotted graphically against tan ψ_2' for both polarizations with $f_{mc}=328$ mc in Fig. 2.

When, as in our case, the reflection takes place over a spherical surface rather than over a plane surface, the rays are diverged, resulting in a further decrease in field strength of the reflected wave. This attenuation is accounted for in a divergence factor, D. To a good approximation, this factor may be expressed as follows:

$$D = \left[1 + \frac{2d_1 d_2}{kad \tan \psi_2'}\right]^{-1/2}. \quad (3)$$

In (3) ka is the effective earth's radius; d, d_1, d_2, and ψ_2' are as shown in Fig. 1. An exact expression for divergence has recently been derived by Riblet and Barker;[8] for our application, however, (3) gives results accurate to four significant figures and has been used in the calculations throughout this paper.

From the interference theory, the smooth-earth space-wave field strength is expressed in terms of a space-wave corresponding to the free-space wave and a direct wave corresponding to the free-space wave and a ground-reflected wave.[3] The resulting wave is the vectorial sum of these two components. Consideration must be given to both the magnitude and the relative phase of these components in order to calculate the space-wave field strength at points in the interference region. If we consider the direct wave to be equivalent to free-space propagation, we have a convenient reference vector of magnitude E_0/d. To this we add the ground-reflected wave of magnitude $(E_0/d)D|R'|$, neglecting the small-order effect of additional distance attenuation due to the longer path length for the ground-reflected ray and neglecting the difference in antenna gains for the different directions of the direct and ground-reflected rays. It can readily be seen that the space-wave field strength varies between $E_0/d(1+D|R'|)$ when the two waves are in phase and $E_0/d(1-D|R'|)$ when the two waves are out of phase.

The relative phase of the two waves at points within the interference region is determined from a consideration of both the phase angle of the reflection coefficient and the geometric path-length difference. The phase angle of the reflection coefficient, $(\pi-c)$, is given by (1) and (2), while the phase lag of the ground-reflected ray due to path length differences is determined from geometrical considerations. Referring to Fig. 1, if we let a denote the actual radius of the earth and ka its effective radius, we may write the following with negligible error:

$$h_1' = h_1 - \frac{d_1^2}{2ka} \quad (4a)$$

$$h_2' = h_2 - \frac{d_2^2}{2ka} \quad (4b)$$

$$\tan \psi_2' = \frac{h_1' + h_2'}{d} = \frac{h_1'}{d_1} = \frac{h_2'}{d_2}. \quad (4c)$$

[8] H. J. Riblet and C. B. Barker, "A general divergence formula," *Jour. Appl. Phys.*, vol. 19, pp. 63–70; January, 1948.

When $k=4/3$, $ka=5{,}280$ miles and (4a) and (4b) become

$$h_1' = h_1 - \frac{d_1^2}{2} \tag{5a}$$

$$h_2' = h_2 - \frac{d_2^2}{2}. \tag{5b}$$

In (5a) and (5b), h and h' are expressed in feet, d_1 and d_2 in miles. The change in phase due to the path-length difference, θ, can be approximately expressed as follows:

$$\theta = \frac{4\pi h_1' h_2'}{\lambda d} \tag{6a}$$

or, more conveniently,

$$\theta = \frac{1.3865 \times 10^{-4}\, h_1' h_2' f_{mc}}{d}. \tag{6b}$$

In (6b) h_1' and h_2' are expressed in feet, d in miles. The relative phase between the direct and ground-reflected wave is $[\theta+(\pi-c)]$. From the rule of cosines the ratio of space-wave field strength to free-spade field strength, $g(\psi_2')$, can be expressed as follows:

$$g(\psi_2') = [1 + (D|R'|)^2 - 2D|R'| \cos(\theta - c)]^{1/2}. \tag{7}$$

In calculating coverage diagrams, it is convenient to determine the variation of $g(\psi_2')$ with height at a fixed distance, for instance, 100 miles. An example of the calculation of this function is presented here for a communication system in which the ground antenna is elevated 35 feet above the surface with a frequency of 328 mc and vertical polarization.

In plotting the function $g(\psi_2')$, as indicated in the tables below, it will be found helpful to proceed as follows. Select some convenient values of d_1 which are not too closely spaced and carry through the computations as indicated. A curve of $(\theta-c)$ as a function of d_1 is next plotted, from which curve the values of d_1 corresponding to $(\theta-c)=n\pi$ will be obtained. These are the critical points in the analysis since it will be noted that values of $(\theta-c)=(2n-1)\pi$ correspond to lobe maxima and values of $(\theta-c)=2n\pi$ correspond to lobe minima. Sufficient other points can then be filled in to get the desired shape of the lobes. It should also be stressed that d_1 is a convenient independent variable upon which to base these calculations because the remaining variables can be computed from it without solving any higher-order equations as would be necessary if h_2 or $(\theta-c)$ had been employed. After calculating a sufficient number of points the entire curve of $g(\psi_2')$ is plotted as a function of altitude, h_2, at 100 miles as shown in Fig. 15.

TABLE II

d_1	$d_1^2/2$	h_1' (5a)	$\tan\psi_2'$ (4c)	d_2 $d-d_1$	h_2' $d_2\tan\psi_2'$	$d_2^2/2$
8.35	35	0	0	91.65	0	4,200
5	12.5	22.5	0.000852	95	427.5	4,512
1	0.5	34.5	0.00653	99	3,416	4,900
0.305	0.046	34.954	0.0217	99.695	11,425	4,970
0.153	0.012	34.988	0.0433	99.847	22,833	4,985
0.1	0.005	34.995	0.0663	99.9	34,961	4,990

TABLE III

h_2 $h_2'+d_2^2/2$	θ (6b)	c Fig. 2	$\theta-c$	$\|R'\|$ Fig. 2	D (3)	$g(\psi_2')$ (7)
4,200	0	0	0	1.0	0	1.0
4,939.5	4.37	0	4.37	0.995	0.567	0.440
8,316	53.59	0.05	53.54	0.945	0.972	0.412
16,395	181.61	0.18	181.43	0.842	0.997	1.839
27,818	363.31	0.35	362.96	0.708	0.999	0.293
39,951	556.39	0.56	555.83	0.582	1.0	1.568

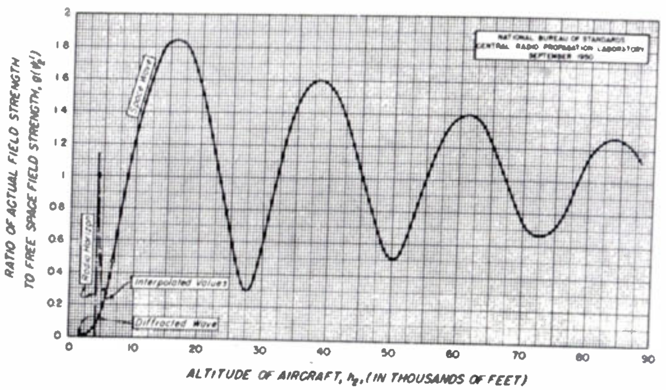

Fig. 15—The variation of field strength with altitude of aircraft at a distance of 100 miles. Frequency, 328 mc; ground-antenna height, 35 feet; transmission over good ground; polarization, vertical.

Near the horizon the application of the interference theory would indicate higher field strengths than would actually be the case, as shown by the dashed portion of the curve on Fig. 15, and in this region other modes of propagation must be considered. The exact computation of the $g(\psi_2')$ function in this region is involved and laborious as shown by van der Pol.[9] However, an excellent graphical interpolation can be made between the results using the interference theory and those using the first term in the diffraction theory.

In considering the diffracted wave, the graphical methods developed by Norton[3] for the solution of the diffracted wave in terms of the surface wave and an appropriate height-gain function are used for the examples in this paper.

The expected coverage for an air-to-ground communications system is expressed in terms of the receiver input voltages. It is convenient first to determine the free-space maximum ranges, d_{fsm}, corresponding to these voltages. Consider first the field strength in free space for a transmitting antenna with power gain, G, relative to an isotropic antenna, in the direction in which we are interested and radiating P_T watts of power. The free-space field intensity at r meters from the transmitting antenna, E_0, is

$$E_0 = \frac{1}{r}\sqrt{30 P_T G_T}\ v/m. \tag{8}$$

[9] B. van der Pol and H. Bremmer, "Further note on the propagation of radio waves over a finitely conducting spherical earth," *Phil. Mag.*, vol. 27, pp. 261–275; March, 1939.

in the equatorial plane of a half-wave dipole antenna radiating 6 watts, $G_T=1.64$ and $E_0=10.68$ Mv/m at one mile. The field strength, E_1, required to deliver the minimum required power, P_R watts, to the receiving antenna is expressed in terms of the gain, G_R, of the receiving antenna and the impedance of free space, $Z_0=120\pi$ ohms, by the following formula:

$$E_1 = \frac{1}{\lambda}\sqrt{\frac{4\pi Z_0 P_R}{G_R}} = \frac{4\pi}{\lambda}\sqrt{\frac{30P_R}{G_R}}\ v/m. \qquad (9)$$

Equating (8) and (9) we obtain the following expression for the free-space maximum range, d_{fsm},

$$d_{fsm} = \frac{\lambda}{4\pi}\sqrt{\frac{P_tG_TG_R}{P_R}}. \qquad (10)$$

Let us apply (10) to the case in which the frequency of operation is 328 mc, the transmitter power is 6 watts, half-wave dipole antennas are employed for both transmission and reception, antenna power gain is 1.64, and 3 μv across the terminals of a receiver with 50-ohms input impedance is assumed to be the minimum detectable signal. Thus, we find the free-space maximum range to be

$$d_{fsm} = \frac{5.68\times10^{-4}}{4\pi}\sqrt{\frac{6\times1.64\times1.64}{1.8\times10^{-13}}}\ \text{miles} = 428\ \text{miles}.$$

Allowing 6 db for communications-system loss, this becomes 214 miles. For other input voltages it can easily be seen that the free-space maximum range is inversely proportional to input voltage. These values are modified by the antenna directivity pattern in directions other than that of maximum radiation. Equation (10) shows that all other propagation factors being equal the free-space maximum range is inversely proportional to frequency.

In plotting the lobe contours on four-thirds earth profile paper several approximations can be made with a high degree of accuracy. When the height of the ground antenna, h_1, is very small compared to the distance to the aircraft, d, points on the coverage diagram having a constant value of $g(\psi_2')$ can be assumed, with negligible error, to lie on a four-thirds earth slant line from the base of the ground antenna then through the height, h_2, calculated for 100 miles to all other distances in question. The error involved here is due to very slight modifications of the reflection coefficient, divergence factor and path-length difference resulting from a slight shift in the point of geometric ray-path reflection. The validity of using this approximation has been determined by calculating values of the $g(\psi_2')$ function, using the methods previously outlined, for distances of 200, 300, and 400 miles and comparing them with those obtained by extrapolating along the slant line through the value calculated for 100 miles. For the particular case in which the ground-antenna height is 35 feet and the frequency of operation is 328 mc with vertically polarized radiation, the errors resulting from this approximation were well within the plotting accuracy of approximately ±100 feet and were not apparent in the comparison. Furthermore, it has been found that the extrapolation works very well in the region at and just above the radio line of sight where other modes of propagation must be taken into account. Values of field strength at distances up to 400 miles in this region show an essentially linear attenuation with distance along the slant lines of extrapolation.

Points on the lobe contours are plotted directly on a sheet of four-thirds earth profile paper, such as that used in Figs. 3 through 9. These charts are designed in such a manner that any straight line in space corrected for the four-thirds earth radius shows up as a straight line on the chart. Points on the lobes are plotted using radial lines from the base of the ground antenna through the altitude at 100 miles to the distance $d=g(\psi_2')\cdot d_{fsm}$ for each input terminal voltage desired as shown in Fig. 16, where $g(\psi_2')$ is the value obtained from Fig. 15. The radio horizon is indicated as an extended line from the

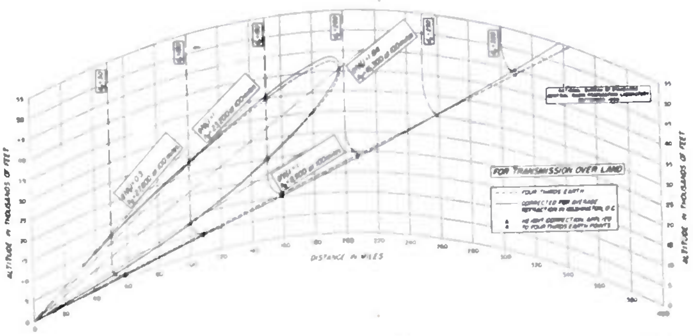

Fig. 16—The graphical development of 328-mc, 6-μv lobe from curve of $g(\psi_2')$ versus height, h_2, at 100 miles (Fig. 15). Ground antenna, half-wave vertical dipole; aircraft antenna, half-wave vertical dipole; power, 6 watts; height, 35 feet; vertical polarization; assumed communication-system loss, 6 db; free-space maximum range, 107 miles.

top of the ground antenna tangent to the ground at distance $d_1=\sqrt{2h_1}$ miles. This line can be plotted very easily using two points on the profile. From (5b), points along the radio horizon occur at heights and distances corresponding to $h_2=d_2^2/2$ and $d=d_2+\sqrt{2h_1}$. For example, when $d_2=200$ miles, $d=200+\sqrt{2h_1}$ miles and $h_2=20{,}000$ feet. Drawing a line through this point tangent to the earth at $d_1=\sqrt{2h_1}$ completes the four-thirds earth radio horizon.

In order to correct more accurately for the average effect of gradients of refractive index, a second-order correction is made to the lobes. This second-order correction, in the examples in this paper, is based on the average refractive conditions in the region of Washington, D. C., as derived by Schulkin, LaBolle, and Herbstreit[10] from an analysis of weather records. The correction to be made to the four-thirds earth ray is shown on Fig.

[10] M. Schulkin, V. LaBolle, J. W. Herbstreit, "Atmospheric refraction corrections for radio field intensity coverage diagrams," in preparation.

17 as a function of distance and altitude for the average refractive conditions in the region of Washington, D. C. It should be noted that this second-order refraction correction is, in any case, quite small, and could be neglected completely without risk of serious error in the

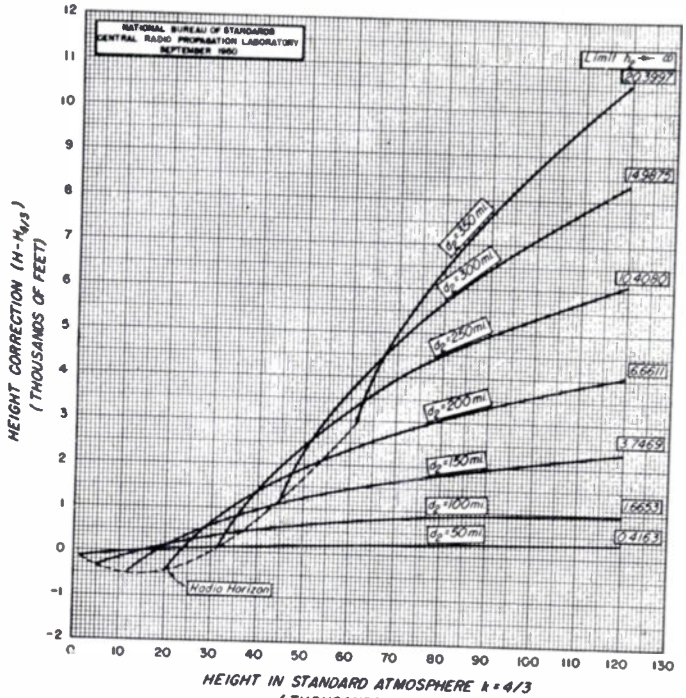

Fig. 17—Approximate height corrections to standard atmosphere field-strength coverage diagrams. Washington, D. C. October mean data.

final results. The several curves on Fig. 17 indicate the height correction to be made to the lobes at distances, d_2, and the four-thirds earth altitude indicated. As an example of the method of applying this correction, it will be noted that on Fig. 16 the four-thirds earth, 6-μv lobe intersects the curve, $d_2 = 100$ miles, at two altitudes, $h_2 = 8{,}800$ feet and 23,500 feet. From Fig. 17 the altitude correction for average refractive conditions is seen to be -170 feet and $+200$ feet, respectively. Similar height corrections are made at all other intersections and the new lobe is drawn in as shown on Fig. 16.

B. *High-Gain Tilted-Array and Height-Diversity Considerations*

The improvement in coverage realized by the use of high-gain tilted arrays and height-diversity ground-antenna systems depends on several factors. Among these are the heights of the antenna above ground, the angle of tilt for tilted arrays, and the spacing for diversity systems. The most serious limitation to communications coverage in most cases lies in the null above the first lobe. For this reason, it is desirable to maximize the improvement in this direction. Fig. 11 shows the geometry involved in tilting an antenna with a narrow beam in vertical cross section for suppression of the ground-reflected energy. For this antenna the gain in the direction of the ground-reflected ray is less than that in the direction of the direct ray.

In order to determine the resulting coverage from such an array, (7) is modified to include the effects of directivity in the ground antenna so that for equivalent distances $g(\psi_2')$ refers the field strength at any given elevation to the free-space field intensity in the direction of maximum radiation,

$$g(\psi_2') = [g_1^2 + (g_2 D \mid R' \mid^2 - 2g_1 g_2 D \mid R' \mid \cos(\theta - c)]^{1/2}. \quad (11)$$

In this equation, g_1 and g_2 are antenna-directivity voltage-gain factors for the direct and ground-reflected rays, respectively, referred to the maximum gain as unity. The $g(\psi_2')$ function at 100 miles can be constructed from the above considerations and the resulting lobe diagram plotted following the methods described before.

To determine the coverage to be expected from height-diversity systems, consideration must be given to the manner in which the signals are to be detected. To receive the signals simultaneously on two antennas, two receivers are used, the signals being combined after detection. Since the receiver which receives weaker signals contributes little or nothing in the way of either signal, or noise, the resulting coverage diagrams are prepared by merely superimposing two lobe diagrams, one for the upper antenna and one for the lower, and by using as contours those representing the stronger signal. The points of lobe crossover now become pseudo-minima and represent the directions of least signal.

Height-diversity, ground-transmission coverage diagrams are calculated in a somewhat different manner than those for reception because of the method in which the signals are transmitted and combined at the receiver. For this type of transmission, two transmitters are employed at the ground installation, one for each antenna. A slight difference in frequency is necessary to maintain independence in transmission. This frequency difference must be small enough so that both signals will pass through the receiver band pass simultaneously, yet it must be sufficiently large so that the beat frequency produced can be eliminated in the receiver audio circuits. The procedure used in determining the coverage to be expected involves adding the power obtained from each transmitter as received at the aircraft. Contours of constant field strength are developed in the manner similar to that outlined before, resulting in a coverage diagram such as Fig. 9.

Interlingua*

An International Auxiliary Language

Scientists and engineers rival exporting and importing industrialists in their urgent need for a simple medium for international communication. Furthermore, with the ever-widening range of communication between distant peoples, a common and simple auxiliary language becomes not only reasonable but essential. In addition, experts in communication are naturally interested in the medium whereby such communication may be effectively conducted.

For these reasons there is here presented a brief description of a basic and constructive project in the field of linguistics—the science and art of international verbal and written communication.—*The Editor.*

"RADIO" comes from Latin; "television," from Latin and Greek. These two languages are drawn upon heavily for the terminologies of science and technology as used in all parts of the modern world. The vast number of internationally identical technical terms constitutes the core of an active international language already in existence in our national tongues.

The International Auxiliary Language Association (IALA) has compiled and standardized a great cross section of this common vocabulary, and has made it available in dictionary form. The name *"Interlingua"* has been given to this living language, which is the natural medium for international communication. The *Interlingua-English Dictionary* and the *Interlingua Grammar*, published by the Storm Publishers, New York, N. Y. in 1951, are the basic manuals for putting the Interlingua into service for scientists. IALA is ready to make translations of abstracts and summaries into the international medium for scientific journals.

The research program of IALA, which has produced the Interlingua, was set up after World War I, at the instance of the International Research Council. The Research Corporation and Rockefeller Foundation have given grants toward IALA's work. The late General J. G. Harbord was Chairman of IALA's Budget Committee. Alfred N. Goldsmith has long been a member of IALA's Board of Directors. The Association is affiliated with Barnard College as its Institute of Interlinguistics.

Members of the Institute are invited to request further information concerning Interlingua from the International Auxiliary Language Association, 420 Lexington Avenue, New York 17, N. Y.

The nature, simplicity of construction, and ease of comprehension of Interlingua can be seen in the text below: a translation from the November, 1951 issue of the PROCEEDINGS OF THE I.R.E., page 1364.

Using Tests to Select Engineers

WARREN G. FINDLEY

Summary—Experience in selecting students for admission to undergraduate engineering colleges provides a clear outline for a program of tests and related procedures that should prove helpful in identifying potential engineering talent early in high school. Qualified students may then be guided toward adequate preparation for engineering training. Such a program would include the following as a basic minimum: the students' average grades, tests of mathematical aptitude, reading comprehension, spatial visualization, and interest inventories.

A recently revised program of examinations are now available for selecting students for graduate study in engineering. Research is being undertaken which gives promise of the development, in the measurable future, of a means of detecting creative talent for scientific research.

When Are Tests Helpful?

The value of standardized tests for selecting engineers must be judged by the extent to which they improve selection over what can be done without tests by using other information routinely available or readily obtainable. That standardized tests will distinguish between superior and inferior applicants for engineering training or employment is not sufficient. Such tests must do the job better or add to what can be done by other methods (i.e., ready-made tests or evaluation of previous school records).

* Decimal classification: 408.9 Original manuscript received by the Institute, November 21, 1951.

Le Uso de Tests pro Seliger Ingerieros

WARREN G. FINDLEY

Summario—Le experientia colligite in le selection de candidatos de admission a polytechnicos inferior (undergraduate engineering colleges) provide nos de clar lineas fundamental de un programma de testes e altere proceduras relate, que se provara utile pro identificar de bon hora in le schola secundari le talento potential de ingeniero. Alora studentes qualificate pote esser guidate verso adequate preparativos pro le instruction professional de ingenieria. Un tal programma includerea le sequente como un minimo basic: le notas median recipite per le studentes, tests de aptitude mathematic, comprehension de lectura, visualisation spatial, e inventarios de interesses.

Un programma recentemente revidite de examinationes es ora disponibile pro seliger studentes pro le studio superior del ingenieria. Recercas es facite que promitte disveloppar, in le futuro measurabile, un medio de deteger creative talentos pro le recerca scientific.

Quando Es Tests Servibile?

Le valor de tests standardisate que se usa pro seliger ingenieros debe esser judicate per le mesura in que illos ameliora le selection ultra lo que pote facer se sin tests per le uso de altere information routinarimente disponibile o facilemente obtenibile. Que tests standardisate pote distinguer candidatos superior e inferior pro instruction o empleo in ingenieria non es sufficiente. Tal tests debe superpassar o supplementar lo que pote facer se per altere methodos (i.e., per tests jam disponibile o per le evaluation de previe notas de schola).

Polaresistivity and Polaristors*

HANS E. HOLLMANN†, ASSOCIATE, IRE

Summary—The induced fibration of magnetic and electric fluids, when acted upon by magnetic or electric fields, manifests itself in two different ways: First, there is a mechanical manifestation in that their viscosity increases, and second an electrical manifestation in that a migratory orientation changes the dielectric constant. If semiconductive particles are suspended in an insulating fluid, the conductivity of the suspension becomes nonlinear as soon as the fibration at a certain threshold field reaches a particular point at which the semiconductive fibers bridge from one electrode to the other.

Particular fluid carriers may be transformed into a solid state, thus making it possible to change the liquid-filled colloidal resistors into dry "polaristors."

The voltage sensitivity is characterized by the "nonlinearity," i.e., the rate at which the differential resistance, under optimum bias, decreases when the bias changes by one volt.

The formation process or the degree of fibration may be checked by means of oscillographic current-voltage characteristics or by means of intermediate-frequency oscillograms similar to modulation trapezoids.

Introduction

SOME TYPES of electromechanical transducers are based on electrostriction and magnetostriction, the deformation of a solid specimen under electric or magnetic stress. In a general sense, electro- and magnetostriction are properties not only of solid substances, but also of certain liquids and suspensions. Aside from the electro-optical effects, electro- and magnetostriction cause the viscosity of such fluids to increase long before a mechanical deformation or shrinkage becomes noticeable. In a broad sense, electrical or magnetic control of viscosity is merely a manifestation of electro- or magnetostriction.

The simplest example is the controllable viscosity of the magnetic fluid,[1–5] a suspension of fine iron powder in oil. When subjected to a magnetic field, the iron particles attract each other and form chains or fibers which predominantly follow the lines of force. Hence, the magnetic fluid thickens and assumes a pastelike consistency; under very strong magnetic fields, the fluid approaches a solid state. One important use to which the magnetic fibration with the associated gel-sol-gel transformation has been put is the magnetic-fluid clutch, in which the thickened fluid transmits a controllable torque from the driving to the driven element.

Long before the development of magnetic fluids, electric prototypes were known. An electric fluid contains polar molecules or molecule complexes or, in a somewhat coarse form, high-dielectric powder suspended in a high insulating fluid, such as oil. When acted upon by an electric field, a similar fibration occurs as in the magnetic case, with the difference that the particles are polarized electrically instead of magnetically.

The classical example of an electric fluid is blood.[6] Wherever the body is subjected to a diathermic field, the shear resistance of the blood increases. Fortunately, the blood circulation is restored by the physiological expansion of the vessels so that the regulative heat flow is not endangered.

Liquids of high polarizability, such as the benzene family, sols, and colloidal suspensions,[7] are other electric fluids. Such suspensions, for example starch, flour, or fine magnesium dioxide suspended in paraffin oil, transformer oil, silicone oil, and the like, exhibit a strong fibration and a well-pronounced change in viscosity and shear resistance.[8] A fibration occurs also in pigments, by means of which the formation of dielectric chains was first shown under the microscope.[9]

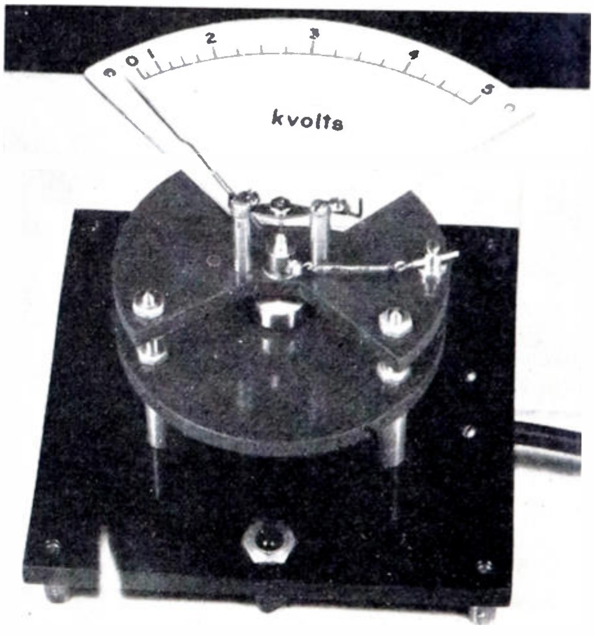

Fig. 1—Electric-fluid viscometer as high-tension voltmeter.

* Decimal classification: R280. Original manuscript received by the Institute January 2, 1951; revised manuscript received December 20, 1951.

† U. S. Naval Air Missile Test Center, Pt. Mugu, Calif.

[1] Jacob Rabinow, "Magnetic fluid clutch," *Tech. Bull. Nat. Bur. Stand.*, vol. 32, pp. 54–60; May, 1948.

[2] "Magnetic fluid uses," *Electronics*, pp. 120–122; September, 1949.

[3] "New uses for magnetic fluids," *Jour. Frank. Inst.*, vol. 248, pp. 155–157; August, 1949.

[4] G. R. Nelson, "Magnetic fluid clutch in servo applications," *Electronics*, pp. 100–103; November, 1949.

[5] E. S. Bettis and E. R. Mann, "Servo employing the magnetic fluid clutch," *Rev. Sci. Inst.*, vol. 20, pp. 97–101; February, 1949.

[6] W. Krasny-Ergen, "Nicht-thermische Wirkungen elektrischer Schwingungen auf Kolloide" (Nonthermal effects of electric oscillations on colloids), *Zeit. fur Hochfrequenz*, vol. 48, pp. 126–133; October, 1936; Report of the International Congress of uhf in Physics, Biology, and Medicine, Vienna, pp. 180–184; 1937.

[7] Y. Björnstahl and K. O. Snellman, "Die Einwirkung eines elektrischen Feldes auf die Viskosität bei reinen Flüssigkeiten und kolloidalen Lösungen" (The influence of an electric field on the viscosity of pure liquids and colloidal suspensions), *Kolloid Zeit.*, vol. 78, pp. 258–272; March, 1937.

[8] W. M. Winslow, "Induced fibration of suspensions," *Jour. Appl. Phys.*, vol. 20, pp. 1137–1140; December, 1949.

[9] J. A. Reising, "Electrostatic responses of pigments in relation to flooding, dispersion, and flocculation," *Indus. Eng. Chem.*, vol. 29, pp. 565–571; May, 1937.

The change of viscosity is studied with a rotational viscometer, which is merely a small electric fluid clutch that actuates a needle or deflects a light beam in counteraction with a restoring spring. The action of such a viscometer may be reversed so that it operates as a high-tension voltmeter. An experimental model is depicted in Fig. 1. A small synchronous motor drives a miniature electric-fluid clutch, the torque of which deflects a needle counteracted by a spring. A dc or ac voltage of the order of kilovolts increases the zero torque, thereby causing the needle to sweep across the scale. The initial portion of the scale has a parabolic calibration, indicating the quadratic character of the relationship between the polarizing field and the shear resistance. The sensitivity naturally depends on the gap but, since a small gap causes an additional phenomenon, namely a compression or jamming of too closely packed particles, the sensitivity is limited.

Electrical Properties of Polarized Colloids

The Dielectric Constant of Colloidal Suspensions

It is obvious that any change in the mechanical properties of the described fluids must be accompanied by a change of their electrical characteristics, primarily their dielectric constant. The problem is very complex because the dielectric constant depends on a multiplicity of parameters. The most important is the mix ratio, usually expressed in terms of the relative volumes v_1 and v_2 of the carrier and the sum of all particles. Other parameters are: shape, size, surface condition, distribution of the particles, and last but not least, the dielectric constants ϵ_1 and ϵ_2 of the components. In order to gain a general understanding, let us consider the liquid-filled condenser shown in Fig. 2. Normally, the particles are in unknown random distribution. It can easily be seen, however, that any distribution (a) must lie between the

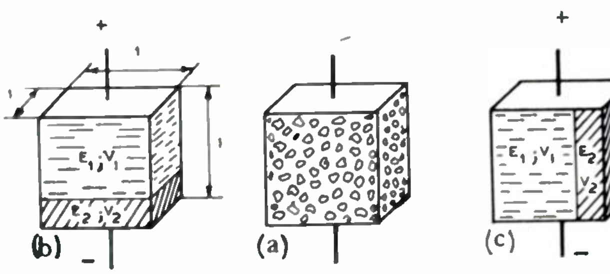

Fig. 2—The particles in a colloidal suspension at random distribution (a) may be visualized as lying between the two theoretical extremes (b) and (c).

two theoretical extremes (b) and (c). In case (b) the particles are packed into a solid layer parallel to the electrodes, i.e., transverse to the field lines. Consequently, the oil and particle layers constitute two series capacitors so that the average dielectric constant ϵ_m of an equivalent homogeneous medium becomes

$$\frac{1}{\epsilon_m'} = \frac{v_1}{\epsilon_1} + \frac{v_2}{\epsilon_2} \tag{1a}$$

or, when the mix ratio $a = v_2/v_1$ is introduced,

$$\epsilon_m' = (1 + a)\frac{\epsilon_1\epsilon_2}{a\epsilon_1 + \epsilon_2}. \tag{1b}$$

Extreme (c) shows the particles compressed into the right portion of the interelectrode space so that the two layers now constitute two paralleled capacitors. Their individual capacities $C_1 = v_1\epsilon_1$ and $C_2 = v_2\epsilon_2$ produce the equivalent dielectric constant

$$\epsilon_m'' = v_1\epsilon_1 + v_2\epsilon_2 = \frac{\epsilon_1 + a\epsilon_2}{1 + a}. \tag{2}$$

The equivalent dielectric constant, under random distribution, cannot be determined accurately. Since it lies between ϵ_m' and ϵ_m'', it must shift toward ϵ_m'' when the particles are aligned. Because of $\epsilon_m'' > \epsilon_m'$, it is apparent that the dielectric constant increases under the influence of the fibration.

The two "mix formulas" may be combined in the exponential form[10]

$$\epsilon_m^p = v_1\epsilon_1^p + v_2\epsilon_2^p, \tag{3}$$

where the exponent $p = \pm 1$ describes the discussed extremes. Since both extremes are only theoretical, the equilibrium of the particles must correspond to a transient condition characterized by a very small value of p; but p increases with increasing fibration and, therefore, with increasing fields. Since it is well known that the forces of electrical attraction are proportional to the second power of the field strength, no great error occurs if p may be set equal to $bE^2 = b(V/d)^2$, where V is the impressed voltage and d the electrode separation. At $p \to 0$, (3) can be written in the form of a series expansion. When higher-order terms are neglected, the relative deviation of the zero dielectric constant,

$$\epsilon_0 = \text{antilog}_e \frac{\log_e \epsilon_1 + a \log_e \epsilon_2}{1 + a}, \tag{4a}$$

with no external field present, then is

$$\frac{\Delta\epsilon}{\epsilon_0} = \frac{b}{2}\frac{E^2}{1 + a}[\log_e^2\epsilon_1 + a \log_e^2\epsilon_2]. \tag{4b}$$

Under practical conditions, the zero value ϵ_0 may be equal to the geometric average of ϵ_m' and ϵ_m''

$$\epsilon_0 = \sqrt{\epsilon_m'\epsilon_m''} = \sqrt{\epsilon_1\epsilon_2\frac{\epsilon_1 + a\epsilon_2}{a\epsilon_1 + \epsilon_2}}.$$

In good approximation, the statement may be made that an almost quadratic relationship exists between ϵ_m an E. This, in fact, has been verified by experiments, even though the influence of all other parameters was not checked.

[10] K. Lichtenecker and X. Rother, "Die Herleitung des logarithmischen Mischungsgesetzes aus allgemeinen Prinzipien der stationären Strömung (derivation of the logarithmic law of mixtures from the general principle of stationary flow)," *Phys. Zeit.*, vol. 32, p. 255–260; June, 1931.

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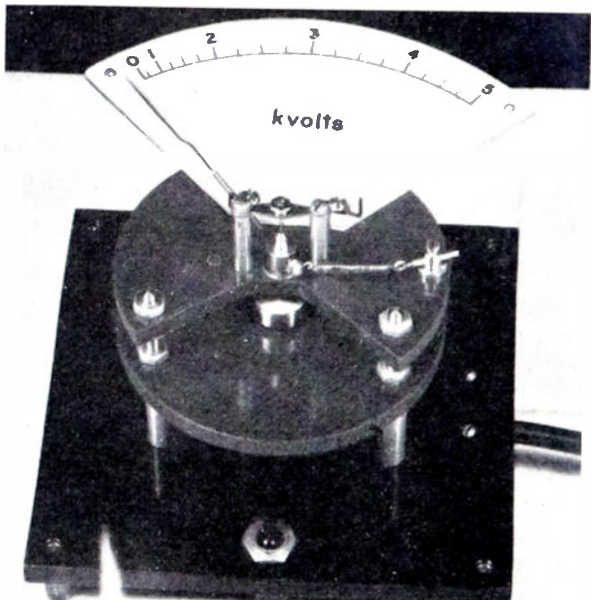

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[4] G. R. Nelson, "Magnetic fluid clutch in servo applications," *Electronics*, pp. 100–103; November, 1949.

[5] E. S. Bettis and E. R. Mann, "Servo employing the magnetic fluid clutch," *Rev. Sci. Inst.*, vol. 20, pp. 97–101; February, 1949.

[6] W. Krasny-Ergen, "Nicht-thermische Wirkungen elektrischer Schwingungen auf Kolloide" (Nonthermal effects of electric oscillations on colloids), *Zeit. fur Hochfrequenz*, vol. 48, pp. 126–133; October, 1936; Report of the International Congress of uhf in Physics, Biology, and Medicine, Vienna, pp. 180–184; 1937.

[7] Y. Björnstahl and K. O. Snellman, "Die Einwirkung eines elektrischen Feldes auf die Viskosität bei reinen Flüssigkeiten und kolloidalen Lösungen" (The influence of an electric field on the viscosity of pure liquids and colloidal suspensions), *Kolloid Zeit.*, vol. 78, pp. 258–272; March, 1937.

[8] W. M. Winslow, "Induced fibration of suspensions," *Jour. Appl. Phys.*, vol. 20, pp. 1137–1140; December, 1949.

[9] J. A. Reising, "Electrostatic responses of pigments in relation to flooding, dispersion, and flocculation," *Indus. Eng. Chem.*, vol. 29, pp. 565–571; May, 1937.

The change of viscosity is studied with a rotational viscometer, which is merely a small electric fluid clutch that actuates a needle or deflects a light beam in counteraction with a restoring spring. The action of such a viscometer may be reversed so that it operates as a high-tension voltmeter. An experimental model is depicted in Fig. 1. A small synchronous motor drives a miniature electric-fluid clutch, the torque of which deflects a needle counteracted by a spring. A dc or ac voltage of the order of kilovolts increases the zero torque, thereby causing the needle to sweep across the scale. The initial portion of the scale has a parabolic calibration, indicating the quadratic character of the relationship between the polarizing field and the shear resistance. The sensitivity naturally depends on the gap but, since a small gap causes an additional phenomenon, namely a compression or jamming of too closely packed particles, the sensitivity is limited.

Electrical Properties of Polarized Colloids

The Dielectric Constant of Colloidal Suspensions

It is obvious that any change in the mechanical properties of the described fluids must be accompanied by a change of their electrical characteristics, primarily their dielectric constant. The problem is very complex because the dielectric constant depends on a multiplicity of parameters. The most important is the mix ratio, usually expressed in terms of the relative volumes v_1 and v_2 of the carrier and the sum of all particles. Other parameters are: shape, size, surface condition, distribution of the particles, and last but not least, the dielectric constants ϵ_1 and ϵ_2 of the components. In order to gain a general understanding, let us consider the liquid-filled condenser shown in Fig. 2. Normally, the particles are in unknown random distribution. It can easily be seen, however, that any distribution (a) must lie between the

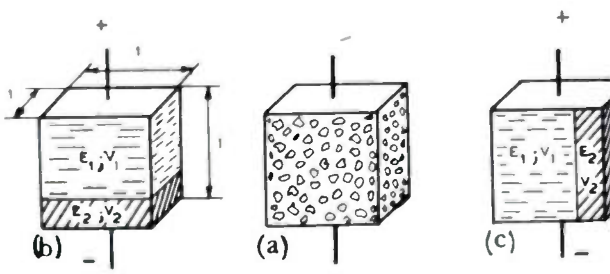

Fig. 2—The particles in a colloidal suspension at random distribution (a) may be visualized as lying between the two theoretical extremes (b) and (c).

two theoretical extremes (b) and (c). In case (b) the particles are packed into a solid layer parallel to the electrodes, i.e., transverse to the field lines. Consequently, the oil and particle layers constitute two series capacitors so that the average dielectric constant ϵ_m of an equivalent homogeneous medium becomes

$$\frac{1}{\epsilon_m'} = \frac{v_1}{\epsilon_1} + \frac{v_2}{\epsilon_2} \tag{1a}$$

or, when the mix ratio $a = v_2/v_1$ is introduced,

$$\epsilon_m' = (1 + a) \frac{\epsilon_1 \epsilon_2}{a\epsilon_1 + \epsilon_2}. \tag{1b}$$

Extreme (c) shows the particles compressed into the right portion of the interelectrode space so that the two layers now constitute two paralleled capacitors. Their individual capacities $C_1 = v_1\epsilon_1$ and $C_2 = v_2\epsilon_2$ produce the equivalent dielectric constant

$$\epsilon_m'' = v_1\epsilon_1 + v_2\epsilon_2 = \frac{\epsilon_1 + a\epsilon_2}{1 + a}. \tag{2}$$

The equivalent dielectric constant, under random distribution, cannot be determined accurately. Since it lies between ϵ_m' and ϵ_m'', it must shift toward ϵ_m'' when the particles are aligned. Because of $\epsilon_m'' > \epsilon_m'$, it is apparent that the dielectric constant increases under the influence of the fibration.

The two "mix formulas" may be combined in the exponential form[10]

$$\epsilon_m^p = v_1\epsilon_1^p + v_2\epsilon_2^p, \tag{3}$$

where the exponent $p = \pm 1$ describes the discussed extremes. Since both extremes are only theoretical, the equilibrium of the particles must correspond to a transient condition characterized by a very small value of p; but p increases with increasing fibration and, therefore, with increasing fields. Since it is well known that the forces of electrical attraction are proportional to the second power of the field strength, no great error occurs if p may be set equal to $bE^2 = b(V/d)^2$, where V is the impressed voltage and d the electrode separation. At $p \to 0$, (3) can be written in the form of a series expansion. When higher-order terms are neglected, the relative deviation of the zero dielectric constant,

$$\epsilon_0 = \text{antilog}_e \frac{\log_e \epsilon_1 + a \log_e \epsilon_2}{1 + a}, \tag{4a}$$

with no external field present, then is

$$\frac{\Delta\epsilon}{\epsilon_0} = \frac{b}{2} \frac{E^2}{1 + a} [\log_e{}^2 \epsilon_1 + a \log_e{}^2 \epsilon_2]. \tag{4b}$$

Under practical conditions, the zero value ϵ_0 may be equal to the geometric average of ϵ_m' and ϵ_m''

$$\epsilon_0 = \sqrt{\epsilon_m' \epsilon_m''} = \sqrt{\epsilon_1 \epsilon_2 \frac{\epsilon_1 + a\epsilon_2}{a\epsilon_1 + \epsilon_2}}.$$

In good approximation, the statement may be made that an almost quadratic relationship exists between ϵ_m an E. This, in fact, has been verified by experiments, even though the influence of all other parameters was not checked.

[10] K. Lichtenecker and X. Rother, "Die Herleitung des logarithmischen Mischungsgesetzes aus allgemeinen Prinzipien der stationären Strömung (derivation of the logarithmic law of mixtures from the general principle of stationary flow)," *Phys. Zeit.*, vol. 32, p. 255–260; June, 1931.

Scientific accuracy requires that an additional phenomenon be mentioned. This occurs at a very low mix ratio and tends to veil the developed analytical result. Let us consider, according to Fig. 3, only a few high-dielectric or even metallic globules under random distribution (a) between two plane electrodes. As long as the

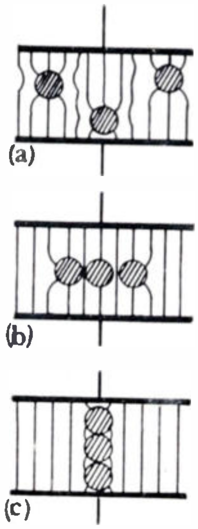

Fig. 3—A high-dielectric mixture under different alignment of the globules.

globules are fairly well separated, the lines of force converge more or less upon each individual particle, which may be substituted by equivalent spheres of larger diameters in a homogeneous field. In other words, the globules, as seen from the electrodes, appear to be magnified and their volumes v_2 seem greater. This magnification decreases when the spheres are aligned either transversely to the electric lines (b) or in the direction of the lines (c). In the latter case, each sphere may be recognized as being in the "electric shadow" of its neighbors so that v_2 assumes almost its true geometric value. This shadow effect counteracts the original increase of the dielectric constant, at least at low mixture ratios, to such a degree that the dielectric constant remains constant, or that even a decreasing capacity may be observed.

In any case, the nonlinear permittivity of a colloidal suspension exhibits a large relaxation time as the result of the sluggish movement of the particles through the viscous medium. Consequently, it is not surprising if the "migratory polarization" builds up slowly when the virgin suspension is polarized by the external field. Inversely, it takes an appreciable time for the particles to revert to their random distribution when the polarizing field is suddenly removed. In this way, a certain analogy with the polarization of polar fluids and substances is quite obvious when an external field tries to orient the dipole molecules in its lines of force. This analogy is made even more striking by the fact that the polarization decreases noticeably under the influence of heat and the associated thermal agitation which interferes with the orientation of the dipoles on the one hand, and the fibration on the other. Furthermore, the virgin condition of the original random distribution is re-established more quickly in a hot suspension, indicating that the lowered viscosity of the carrier plays an important role also. The analogy may even be extended to include a Curie point at which the temperature agitation breaks the alignment of the particles in a way similar to that in which the temperature agitation breaks up the orientation of the polar dipoles.

Since not only high-dielectric but also metallic suspensions exhibit a migratory polarization, it may well be expected that a similar effect also occurs in magnetic fluids. The capacity then may readily be seen to approach an infinitely high value just before the metal fibers produce a short circuit between the electrodes.

Another result of sluggish migratory polarization is hysteresis, as in ferromagnetics and ferroelectrics.

Nonlinear Conductivity of Colloidal Suspensions

After having discussed the mechanical and electrical properties of high-dielectric suspensions and their dependence on external fields, we now turn to semiconductive suspensions, the particles of which exhibit a certain conductivity. The simplest types of such semiconductive oil are suspensions of fine graphite powder or lampblack in silicone oil or commercially available colloidal graphite dispersed or diffused in various oils known under the trade names "Oildag," "Glydag," "Castordag," and so on. The viscosity effect of these oils is almost immeasurably small; the permittivity effect is veiled by the power factor. This, however, is more than compensated for by a controllable and nonlinear conductivity.[11]

Two electrodes, immersed in the semiconductive oil, form a colloidal resistor. The conductivity of the oil in its virgin state, as long as the particles are in a state of random distribution, is practically zero. As soon as the fibration occurs, the semiconductive fibers bridge from one electrode to the other, and thus produce a conductivity. Once the fibers are formed, a certain conductivity remains, except when the fibers are destroyed by vibrations or turbulences. The most important phenomenon is the fact that the conductivity depends to a high degree on the impressed voltage or the passing current. In other words, the conductivity is nonlinear absolutely independently of the relaxation time of the migratory polarization.

The peculiar behavior of a colloidal resistor may best be illustrated through its current versus voltage (I/V) characteristic. A typical example is depicted in Fig. 4. There is no current below a critical threshold voltage V_{th}; but once this voltage is reached, the first fibers bridge the gap from one electrode to the other. As the voltage increases, more and more semiconductive bridges are built up. Simultaneously, the resistance of

[11] H. E. Hollmann, "Semiconductive colloidal suspensions with nonlinear properties," *J. Appl. Phys.*, vol. 21, 402–413; May, 1950.

each individual fiber decreases so that the over-all conductivity increases. The result is a current that rises nonlinearly in accordance with the dotted curve in Fig. 4. Once a certain degree of fibration is attained, i.e., once the suspension is sufficiently polarized, a nonlinear resistance remains and the associated current follows the solid curve. The good frequency response leads to the conclusion that the migratory polarization is responsible only for the threshold field and the state of polarization, whereas the nonlinearity is caused by the mutual attraction between the particles or by some type of electrostriction along the semiconductive chains. This internal compression which causes the viscosity as well as the dielectric constant to increase, strengthens the innumerable transient contacts between all particles as in a microphone.

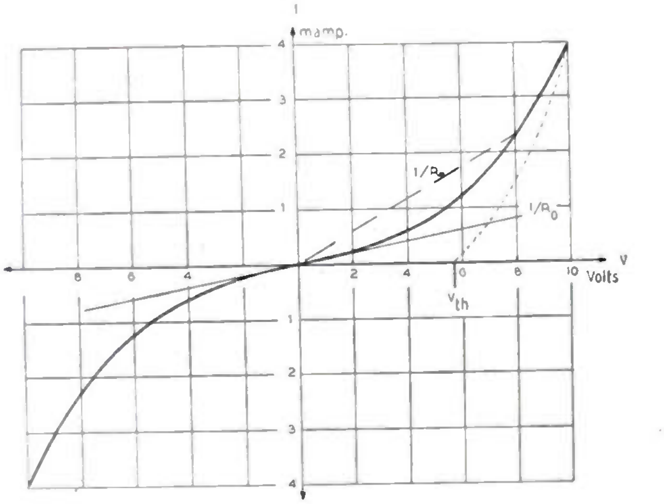

Fig. 4—Current versus voltage characteristic of a colloidal resistor.

As previously mentioned, the viscosity effect is too small to be measured without the use of a supersensitive viscometer. The dielectric constant, however, can be checked as it increases according to the migratory alignment as long as the field strength is below the threshold value.

The colloidal resistors obviously differ from other types of nonlinear resistors, such as vacuum diodes, detectors, and rectifiers, in that they are bipolar. That is, they exhibit a backward-to-forward ratio equal to one. In this respect they are identical with Thyrites.[12,13,14]

Although extensive analytical work has been done to evaluate the conductivity of colloidal suspensions,[15,16] no satisfactory solution has been found even with the assumption of cube-shaped particles because the fibration and the microphonic contact action have not been taken into consideration. Hence we confine ourselves to the experimental result that the quadratic viscosity effect, together with the relationship between mechanical stress and resistivity of the transient contacts, cause I versus V to follow a third power law. This cubical nonlinearity is superimposed upon an initial or zero resistance R_0 as characterized by the tangent through the origin of the I/V curve. As a result, the current is related to the applied voltage by the formula

$$I = \frac{V}{R_0} + \frac{K}{R_0} V^3 = \frac{V}{R_0}(1 + KV^2), \tag{5}$$

where the coefficient K is a characteristic of the colloidal resistor in question. That means K takes into consideration not only the colloidal medium itself, but its state of prepolarization and the electrode separation also.

The nonlinear I/V curve permits definition of two resistances, namely, a dc resistance

$$R_= = \frac{V}{I} = \frac{R_0}{1 + KV^2}, \tag{6a}$$

which is equivalent to the inverse slope of the dashed chord in Fig. 4, and a differential or ac resistance

$$R = \frac{dV}{dI} = \frac{R_0}{1 + 3KV^2}, \tag{6b}$$

which is equivalent to the inverse slope of the curve itself at any arbitrary operating point or bias voltage V_B upon which a low ac voltage is superimposed. In Fig. 5 both resistances are plotted versus V or V_B, respectively. The most important $R_\sim$-curve exhibits a

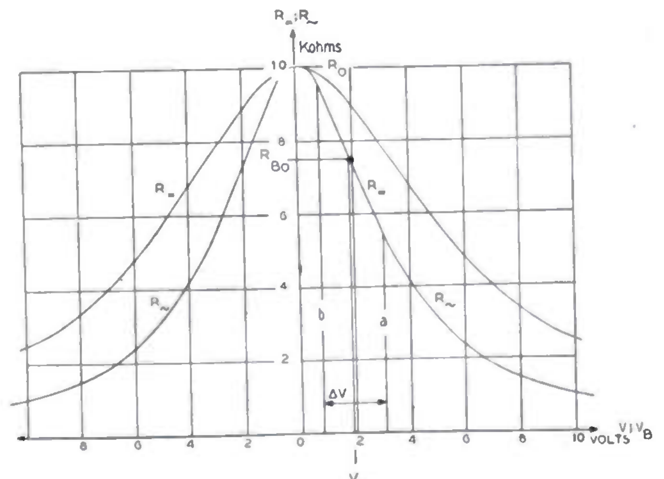

Fig. 5—AC and dc resistances versus voltage according to Fig. 4.

point of inflection and an associated maximum slope at the bias resistance $R_{B0} = 3R_0/4$ and at the optimum bias, $V_{B0} = 1/3\sqrt{K}$. Relative slope at this point then is

$$\frac{\Delta R_\sim}{R_{B0}\Delta V} = 3\sqrt{K}/2 \text{ volt}^{-1}. \tag{7a}$$

[12] K. B. McEachron, "Thyrite," *Jour. A.I.E.*, p. 350; May, 1930.
[13] K. B. McEachron, "Thyrite: a new material for lightning arresters," *G. E. Rev.*, pp. 92–99; February, 1930.
[14] General Electric Bulletin 4/38 A., pp. 1–11.
[15] Guy S. son Frey, "Über die elektrische Leitfähigkeit binärer Aggregate" (On the conductivity of binary aggregates), *Zeit. für Elektrochemie*, vol. 23, pp. 260–274; May, 1932.
[16] F. Wachholtz and A. Franceson, "Dielektrische Messungen an Pigment-Leinöl Suspensionen" (Dielectric measurements on pigment-linseed-oil suspensions, *Kolloid Zeit.*, vol. 92, pp. 75–93, and 155–169; January and February, 1940.

This measure characterizes the voltage sensitivity of any nonlinear resistor far better than the conventional half-crest value, i.e., the voltage at which R_- or $R_\sim = R_0/2$. Hence, the relative slope may be called "ac nonlinearity index" or simply "nonlinearity," whereby it must be kept in mind that the nonlinearity of $R_\sim$ is always meant.

When the $R_\sim$-curve of a certain colloidal resistor is given, the nonlinearity may be evaluated according to the formula

$$NL = \frac{2}{\Delta V}\frac{a-b}{a+b}, \quad (7b)$$

where a and b denote the two ordinates in Fig. 5 which are equidistant from V_{B0} and include ΔV in between. The nonlinearity may readily be seen to indicate the relative change of the differential resistance at optimum bias when the latter is changed by 1 or ± 0.5 volt.

Inversely, the factor K may be evaluated as

$$K = \left[\frac{4}{3\Delta V}\frac{a-b}{a+b}\right]^2. \quad (8)$$

Polaristors

The colloidal resistors in their liquid state exhibit some disadvantages. First, they are very sensitive to mechanical vibrations, i.e., they are microphonic. Even slight turbulences in the electric oil affect the semiconductive fibers and, therefore, the over-all resistance. Second, they are somewhat unstable with respect to their prepolarization, and therefore R_0 may vary. Since the latter is closely related to the state of fibration, it depends primarily on the maximum applied voltage. Consequently, different I/V curves are obtained, depending on whether or not they are measured step-by-step under dc or are presented as oscillograms with the aid of ac. This phenomenon causes R_0 to decrease with increasing ac voltages of audio and radio frequencies, so that a "pseudo rectification" occurs.[11]

In the course of numerous experiments with various types of colloidal suspensions, the question arose as to whether the liquid state is a prerequisite for the described conductivity, or whether it would be possible to form equivalent solid colloids. It was found that the colloidal resistors can be stabilized very successfully by changing from the liquid into the dry state. A solid resistor produced in this manner is called a "polaristor" and the nonlinear resistivity of its preformed medium is called "polaresistivity." The creation of a polaristor is made possible by the fact that the semiconductive fibers retain their form when the originally fluid carrier solidifies after the fibers have been formed. This is particularly true if the polarizing field is maintained during the coagulation.

Wax and paraffin, whose physical states can easily be changed by heat, were utilized as carrier materials during early experiments. A wax polaristor was produced as follows: paraffin was melted and mixed with semiconductive powder, for example, with fine graphite. The hot suspension was then poured between suitable electrodes and immediately polarized. As the paraffin cooled under the steady influence of the polarizing field, the semiconductive fibers and their nonlinear resistance remained while the graphite-paraffin suspension assumed its solid state. The disadvantage is that zero resistance R_0 and nonlinearity NL decrease considerably because of the shrinkage of the paraffin. It may well be understood that this shrinkage is equivalent to an internal mechanical, or more accurately, to a compressional bias.

After it became possible to change the liquid suspensions into dry substances without destroying their electrical performance, a much better carrier material was found, namely, thermo- or cold-setting plastics, known under trade names such as "Castolite," "Plasticast," and so on. Plastic polaristors are produced from colloidal types in that the plastic carrier is hardened by means of the prescribed heat treatment or cold setting after the prescribed plasticizers, hardeners, and promoters have been added.

Various types of electrodes have been utilized for the construction of polaristors. The first type, represented in Fig. 6(a), contains two wires bifilarly wound around an insulator. Assuming a certain polaresistive material, i.e., a certain semiconductive powder, mix ratio, and prepolarization, the nonlinearity depends on the separation of the wires or on the pitch, while the zero resistance depends on the cross section of the conductive path, i.e., on the length of the bifilar helix. Hence, an extensive range of R_0 as well as NL can readily be covered by means of various sizes while the polarization during the process of polymerization permits a certain range of dispersion.

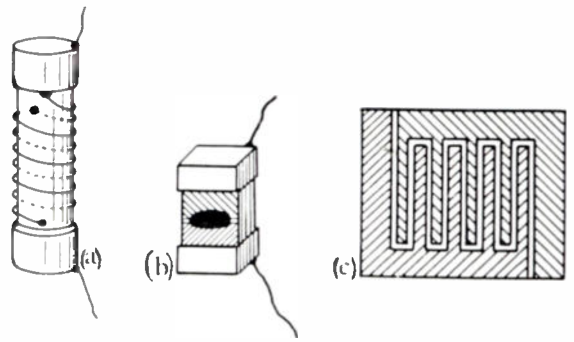

Fig. 6—Various electrode arrangements of polaristors: (a) bifilar wires, (b) mirror electrodes, and (c) microscopically fine grid electrodes.

When a high nonlinearity is required, the second type, the so-called "mirror polaristor" (Fig. 6(b)), is convenient. A minute electrode gap is made by cutting a microscopically fine slot into the silver layer of a mirror. Naturally, this mirror type may be combined with the former cylindrical form by cutting a helical slot into a metal-plated rod or by cutting a meandering line similar to a diffraction grating, as shown in Fig. 6(c). All electrode arrangements are covered with a fine coat of the plastic suspension, which is then hardened by means of the prescribed treatment.

Aside from the electrode separation, the sensitivity obviously depends widely on the colloidal medium itself or, more accurately, on the conductivity and permittivity of the particles, their size, shape, surface condition, and mix ratio. Extensive investigations of various powders, the electric properties of which were graded by means of a special treatment affecting the oxygen content, revealed that titanium dioxide, with a reduced oxygen content, gives satisfactory results. For example, the nonlinearity of a mirror polaristor containing a mixture of $TiO_{1.8}$ and Castolite reaches 16, which is the same order of magnitude as the values of modern crystal diodes and far higher than the nonlinearity of Thyrites.

The polaristors exhibit a surprisingly good frequency response. No drop in nonlinearity could be measured up to the frequencies of many megacycles; a detector action has been measured, even in the range of microwaves.

As it may be expected, the polaristors are very sensitive to temperature because of the difference on the coefficients of thermal expansion of both components. Moreover, the polaristors at a high-current density show an inverse thermistor effect because the energy is dissipated exclusively in the semiconductive chains while the carrier substance warms up only by heat conduction. The thermal expansion of the embedded fibers is equivalent to a mechanical bias, and results in a decreasing R_0 as well as NL. This effect has been shown in semiconductive oil where the expanding fibers even produce a pressure upon the electrodes. Consequently, it may easily be seen that the polaristors are damaged by heavy overloading because the expanding and contracting fibers lose their contact to the electrodes.

Nonlinearity Oscillograms

The measurement of nonlinear characteristics with the aid of dc and an associated step-by-step method is very laborious and time consuming. In the case of polaristors, dc measurements give no reproducible and satisfactory results unless the fibration has reached a final state, i.e., unless the carrier is hardened. It is more convenient to test the polaristors by means of oscillograms under ac which, at the same time, takes care of fibration. In this way, the characteristic values cannot only be observed in the process of production, but also influenced within certain limitations.

The simplest way is an oscillographic representation of the I/V characteristic. A suitable circuit is depicted in Fig. 7(a). The polaristor P under test together with the fixed resistor r forms a nonlinear attenuator which is fed by an ac voltage, e.g., the 60-cycle voltage $V_\sim$. The cathode-ray oscilloscope is fed by the voltage drops across P and r so that the X-deflection occurs proportional to the voltage across the polaristor and the Y-deflection proportional to that across r, which in turn is proportional to the passing current. Fig. 7(b) shows a resulting I/V oscillogram which has the same shape as the static I/V curve shown in Fig. 4.

The nonlinearity becomes more pronounced if the fundamental output voltage across P is eliminated so that only the superimposed harmonics owing to the nonlinearity remain. For this purpose, the polaristor under test forms one arm of the bridge circuit shown in Fig. 8(a). The X-input of the oscilloscope is fed, as before, by the voltage drop across P, but the Y-input is connected across the diagonal branch of the bridge. Provided $r'=r''$, the bridge can be balanced by means of a variable r so that the center portion of the resulting curve coincides with the X-axis as shown in Fig. 8(b). In this case $R_0=r$. The nonlinearity can then be evaluated by means of the slope tan α of any chord connecting two symmetrical points of the oscillographic curve.

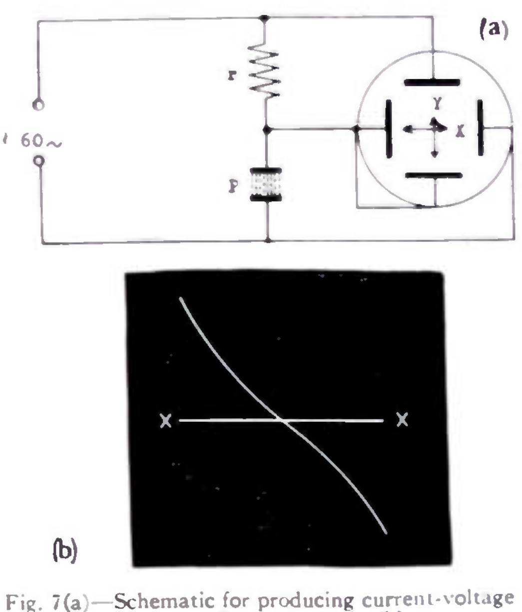

Fig. 7(a)—Schematic for producing current-voltage characteristics as shown in (b).

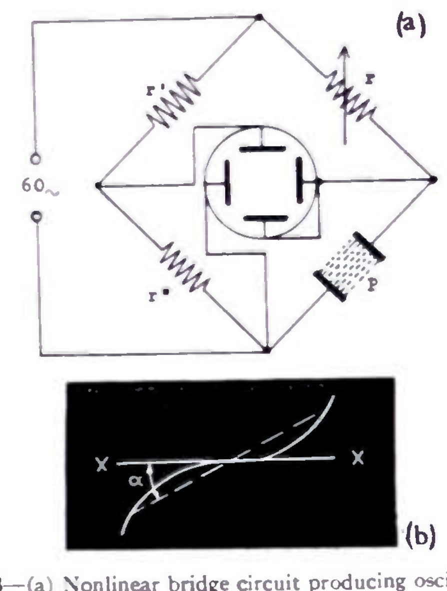

Fig. 8—(a) Nonlinear bridge circuit producing oscillograms as shown in (b).

A more illustrative picture is obtained by a direct representation of the relationship between $R_\sim$ and the applied voltage V_B in the form of an oscillographic $R_\sim$-characteristic. This is achieved by means of the inter-

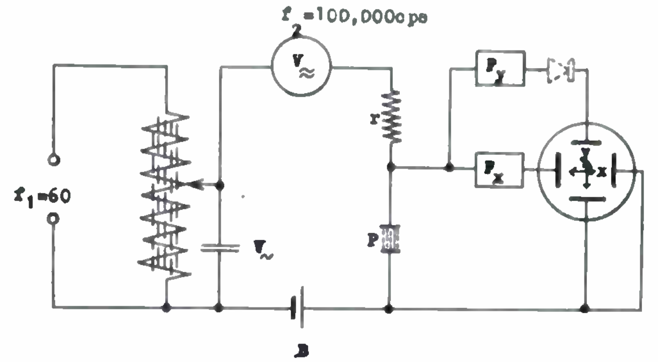

Fig. 9—Schematic circuit of the intermediate-frequency method.

mediate-frequency method according to the schematic shown in Fig. 9. Now, the same nonlinear attenuator as shown in Fig. 7(a) is fed by two voltages with different frequencies, namely, a 60-cycle voltage $V_\sim$ as before, and a superimposed voltage $V_\approx$ of higher frequency, e.g., of 100,000 cps with an amplitude as small as the sensitivity of the oscilloscope will permit. Both oscilloscope inputs are connected across P, but low- and high-pass filters F_x and F_y cause the X-deflection to be produced by the 60-cycle component alone, and the Y-deflection only by the IF. Since the intermediate-frequency component across P is

$$V_{p\approx} = \frac{V_\sim}{1 + r/R_\sim},$$

care must be taken to hold $r/R_\sim \gg 1$ so that

$$V_{p\approx} \doteq V_\approx R_\sim / r,$$

that is, that $V_{p\approx}$ is directly proportional to $R_\sim$. It can easily be seen that the differential resistance is scanned in a rhythm of 60 cycles.

The hexagonal oscillogram shown in Fig. 10(a) is a typical example. The envelopes of the bright area are equivalent to the $R_\sim$-curves shown in Fig. 5.

As soon as the polaristor operates under the influence of a suitable bias which may be produced by the battery B in Fig. 9, the hexagon is unsymmetrically deformed. Provided $V_B = V_{B0}$, the hexagon shrinks to the trapezoid depicted in Fig. 10(b). The length of the X-line indicates a calibration voltage V_{cal}. With the aid of the lengths a, b, c, and d, the nonlinearity can be evaluated according to the formula

$$NL = \frac{2}{V_{cal}} \cdot \frac{d}{c} \cdot \frac{a-b}{a+b}. \qquad (9)$$

The intermediate-frequence method permits a dynamic and instantaneous test, and therefore permits the observation of the fibration during the time the plastic suspension polymerizes. As a result, $V_\sim$ may be varied in order to affect the fibration and the resulting zero resistance R_0 as well as the nonlinearity NL. The two oscillograms shown in Fig. 10(c) explain this process in detail. The outer figure appears in the vicinity of the threshold voltage V_{th}, but shrinks as soon as the driving voltage increases. The inner figure illustrates a certain state of formation characterized by a lower R_0 as well as by a higher nonlinearity. If the test polaristor is permitted to solidify under this state, R_0 and NL do not change to a noticeable degree.

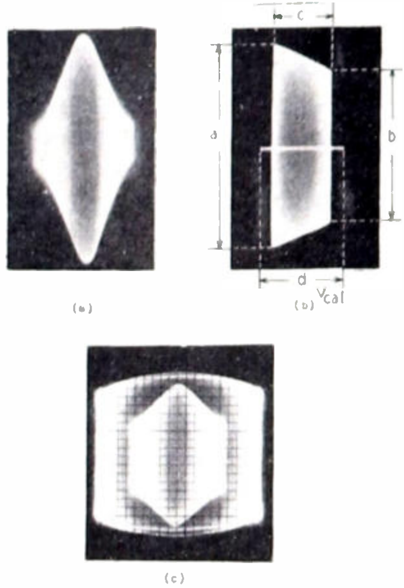

Fig. 10—IF oscillograms; (a) without bias, and (b) at optimum bias. (c) Two oscillograms at various degrees of polarization

In addition, there exists the possibility of inserting a peak-voltage rectifier in series with F_y (Fig. 9) so that the oscilloscope reveals only the upper envelope in the form of a sharp curve.[11] Since such curves, however, do not show the associated X-axis, and since the time-constant of the rectifier device may cause serious distortions and errors, the described IF method without rectification has been found to be more reliable.

Additional Remarks

Beyond the scope of this paper, it may be mentioned that the described IF method permits any nonlinear element to be tested, not only resistors and varistors, but also transistors and reactances such as ferroelectric capacitors and saturable reactors. Experiments with numerous Hi-K condensers revealed the significant fact that some types exhibit a well-pronounced nonlinearity. In addition, some types of carbon potentiometers have been found to be nonlinear, in particular, if they are somewhat worn. Although these nonlinearities are far below the values of polaristors and the like, they may cause noticeable distortions if such a nonlinear potentiometer or capacitor serves as a coupling element in audio amplifiers, and the like, particularly when subjected to a dc bias.

Conclusions

There are many applications, modifications, and possibilities for which the principle of polaresistivity may be very valuable. In many cases, the combination of a high nonlinearity with a high wattage may be in favor of polaristors. They excel all other types of nonlinear resistors not only as far as these two characteristic values are concerned but also in view of their low cost. On the other hand, the polaristors are in their infancy. Various additional problems must be solved before the polaristors become reliable enough for mass production. Such important details as temperature coefficient, frequency response, and noise figure must be studied extensively before the complete picture can be ascertained.

Moreover, it must be kept in mind that the polaresistive colloids are not only electrically nonlinear, but are, at the same time, sensitive to temperature and mechanical strain or stress, so that new sensing elements and electromechanical transducers may be developed. Since we are completely free to intermix a great variety of powders with various properties, such as semiconductive-ferromagnetic particles, photoelectric or photoconductive powders, and the like, the versatility of polaristors may be greatly be enhanced.

An Aural Monitor for Frequency Modulation*

J. L. HATHAWAY†, SENIOR MEMBER, IRE, AND R. E. LAFFERTY†, MEMBER, IRE

This paper is published with the approval of the IRE Professional Group on Instrumentation. —*The Editor.*

Summary—**In television sound and other frequency modulation broadcasting, a practical operating problem exists wherein the usual station monitoring is ineffectual. The requirements of a truly representative aural monitor can be met by proper application and adjustment of a slope detector. Two units embodying this principle are described.**

Introduction

A SITUATION EXISTS at many FM broadcasting transmitters, including those for TV sound, wherein the station aural monitor indicates high-quality sound transmission while receivers, especially in the service area fringes, are sometimes subjected to severe transmitted background noise, such as microphonics, AC hum, or audio squeals. These receivers are not entirely at fault, although the broadcaster believes they are when his monitored sound is from a high-quality station monitor. Paradoxically, this very instrument may block out the background noise heard by distant listeners.

Broadcasters generally utilize accurate equipment for noise measurements as well as for checking frequency deviation and aural quality. But unfortunately, aural-quality monitoring may only be representative of receivers with complete limiting and proper tuning adjustment. The broadcaster has no aural indication of coincidental amplitude modulation because his monitor limits this to below audibility. In the field, especially in the fringe areas, many receivers with incomplete limiting or equivalent function fail to reject coincidental amplitude modulation, giving rise to background noise or distortion, which may be further aggravated by receiver misalignment. Thus, although the broadcaster can measure amplitude modulation to comply with Standards of Good Engineering Practice, he may not have a system for aurally detecting it during actual broadcasting and, therefore, of clearing it at the earliest possible moment. Instead, he may be in a blissful, but unenviable position, until called by distant listeners, or possibly by the FCC.

Requirements

The aural monitor at an FM station should have low distortion and background noise as well as proper frequency response. It should also be responsive to amplitude modulation in a degree at least equivalent to that of receivers in the field with little or no limiting. In fact, for rapid detection of transmission defects, the monitor should be overly sensitive to coincidental AM background modulation products. It should be stable over long periods of time, permit accurate initial adjustment, be easy to operate, and as simple and inexpensive as possible. Two different instruments have been developed with these requirements in view, and have been used successfully in TV sound-transmitter monitoring. The simpler of these is suitable for TV sound on a lower vhf channel, whereas the other is satisfactory for any TV channel.

Theory

One of the earliest known FM demodulators, the "slope detector," is capable of high-quality output, but has not been popular for several reasons, foremost of which is an inherent sensitivity to AM as well as to FM.

* Decimal classification: R148.2×R254.12. Original manuscript received by the Institute, June 4, 1951; revised manuscript received, December 7, 1951.

† National Broadcasting Company, RCA Building, Radio City, New York 20, N. Y.

Since the transmitter aural monitor requires just this characteristic, the slope principle is employed in the present system. FM and AM demodulation are accomplished simultaneously in a detector by detuning for slope detection. Proper slope is required to achieve the desired FM to AM sensitivity ratio, and the range of linearity must bracket the FM deviations. Furthermore, the operating point must be established for minimum FM distortion. This can be accomplished mathematically by equating the second derivative of the universal resonance curve[1] equation to zero, thus locating the inflection point. Fig. 1 shows this, centering the most linear portion of the curve.

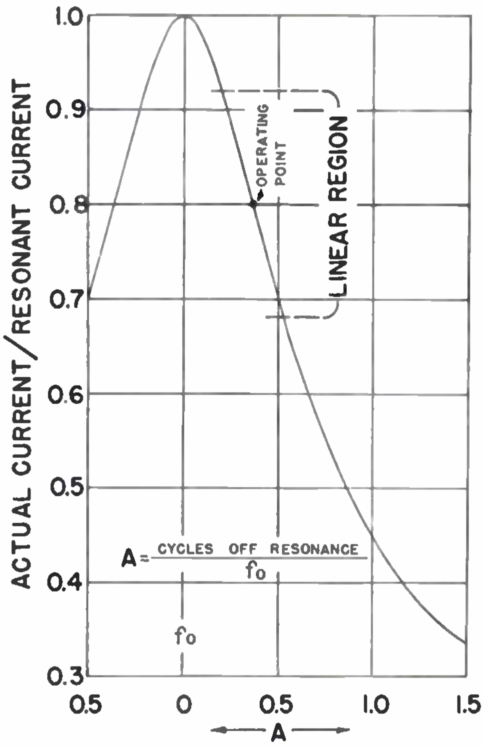

Fig. 1—Universal resonance curve.

The inflection point occurs at approximately $A = 0.37$, and the curve is essentially linear from $A = 0.20$ to 0.54. To allow some margin of safety, this region should be at least equal to 100 kc for use in monitoring a 25-kc deviation transmitter. With this information, an empirical relationship is established for a TV sound monitor, between detector resonance (f_0) and circuit Q:

$$Q = 2.5 f_0 \times 10^{-6}.$$

If Q is excessive, the FM program tends to mask out the AM noise. If the opposite relationship exists, noise below the FCC's "Standards of Good Engineering Practice" limit becomes excessive on the monitored output. Experience has shown that noise from the monitor of an FM transmitter operating just within the FCC requirements, should be about 30 db below program level. This is not annoying to the station personnel under normal conditions; but should the background noise increase above FCC limits, operators immediately become aware of the condition.

For FM transmitters of 75-kc deviation, the preceding formula should be modified by a factor of 3, to achieve the desired sensitivity. That is, the Q should be one-third as high as for 25-kc monitoring. Necessary circuit Q can be easily obtained with sufficient stability for monitoring 75-kc deviation FM in the 100-mc frequency range as well for the TV sound channels 2 to 6. However, circuits of proper Q and controllability are not readily attainable for the TV channels above 6. Here, a heterodyne system may be used to beat down the transmitter frequency. On channel 4, both the direct and heterodyne types operate properly, and have been laboratory as well as operationally tested.

Description

The simple direct-type instrument, shown in schematic and picture in Figs. 2 and 3, is equipped with a capacitive input signal attenuator. With the single circuit tuned for resonance, the meter facilitates adjustment of the input signal to the prescribed value and a

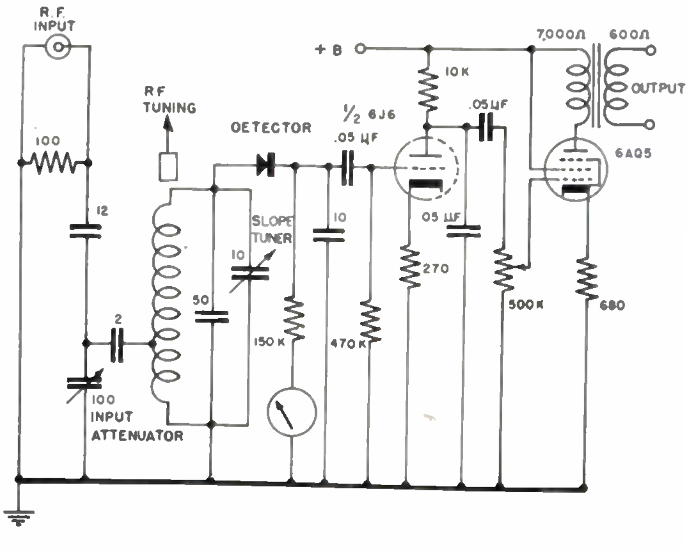

Fig. 2—Direct-type aural monitor.

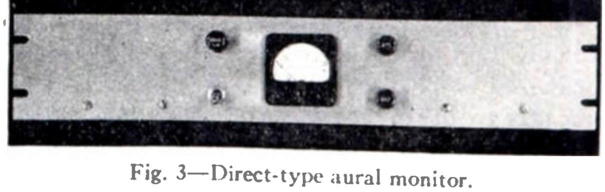

Fig. 3—Direct-type aural monitor.

subsequent detuning for optimum slope detection. Audio output from the detector, incorporating 75-μsec de-emphasis, is fed into a two-stage audio amplifier

[1] F. Terman, "Radio Engineering," Third Edition, p. 42; 1947.

which in turn feeds the low-impedance balanced monitoring line.

The heterodyne-type instrument embodies a few additional circuits and will be described in greater detail. Since ample rf signal is available, no amplifier is utilized and a simple crystal mixer obviates the use of a mixer tube. To meet the stability requirement the local oscillator is crystal controlled, and again for simplicity, the slope detector employes a crystal rectifier.

A capacitive-type attenuator is provided for smooth control of the rf input signal. This, as shown in Fig. 4, couples to a resonance circuit for rejection of spurious input signals, and the FM signal is thence applied to the crystal mixer. A harmonic crystal-controlled local oscillator uses one-half of a double triode with the other half

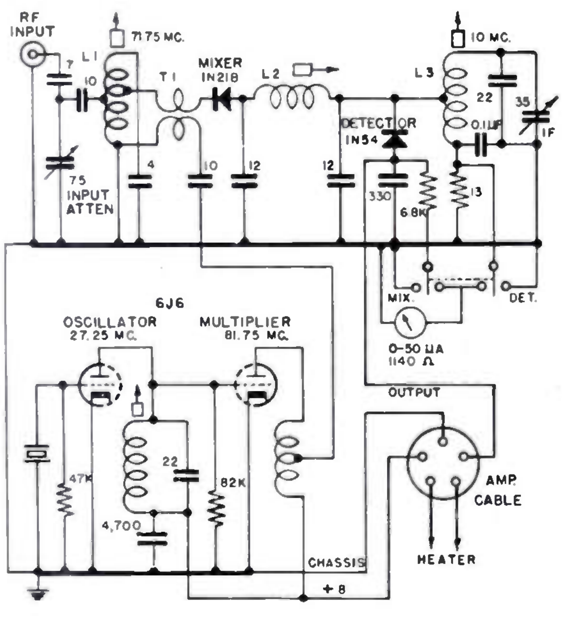

Fig. 4—Heterodyne-type aural monitor.

acting as a frequency tripler for the mixer injection. Mixer output feeds through a Pi-type low-pass filter to eliminate the rf and mixer injection frequencies while delivering the IF to the subsequent tuned circuit. The desired beat signal is thence applied to a single-tuned IF circuit of optimum Q which, with a germanium crystal, serves as the slope detector. Output audio is delivered to a high-quality amplifier having a built-in 75-μsec de-emphasis filter and sufficient audio gain and output power for monitoring. Less than 1.0-per cent rms distortion is introduced in the combined converter-amplifier.

The IF utilized is 10 mc, so that for monitoring TV channel 4 sound (71.75 mc) the local oscillator injection frequency should be either 61.75 or 81.75 mc. The latter is utilized, which, frequency divided by the multiplier ratio of 3, establishes a 27.25-mc harmonic oscillator crystal.

Since detector output is low-level high impedance, the converter is mounted adjacent to the audio unit to avoid stray pickup. A standard rack-type audio amplifier is employed, while a metal box 7 inches×5 inches ×2 inches houses the converter with its meter and controls, as pictured in Fig. 5. Filament and plate power are obtained from the amplifier through a cable and plug which also convey detector output to the amplifier.

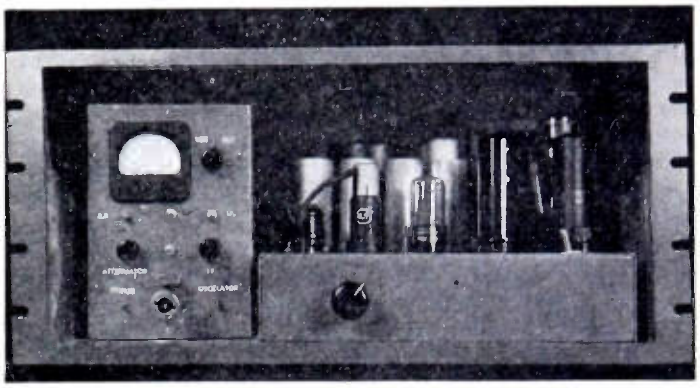

Fig. 5—Heterodyne type aural monitor.

Adjustment

Alignment of the simple direct-type unit is obvious after reference to the heterodyne type. All circuits are first tuned to their proper resonant frequencies. Then with no rf input, the meter is switched to "mixer" and the multiplier tuning inductance adjusted for proper injection level as indicated by a meter reading of 30 (2.6 ma). Next, with the meter in the "detector" position, rf input level is set for a reading of 40 (40 microamperes). This is accomplished roughly with the input attenuator at midposition by properly locating the remote rf pickup loop, and then precisely by the attenuator control. After these adjustments, the IF trimmer capacitor is decreased until the meter indication falls from 40 at resonance down to 32, establishing the correct slope-detection operating point. Should the monitor reveal noise on the transmitter carrier, the IF may be resonated to minimize FM program while further magnifying AM noise for critical examination.

Conclusion

The need for more representative aural monitoring at FM sound transmitters has been recognized for some time. Various methods have been suggested and considered, including for example, the addition of an AM noise component to the standard FM monitor, possibly from the ground-return circuit of the power-amplifier cathodes. This, however, was not too appealing for several obvious reasons. Another possibility was considered, involving a standard type of FM tuner with limiting removed, detuned by a controlled amount to achieve the desired AM sensitivity. Unless the oscillator were crystal controlled, such a monitor would be unreliable, and furthermore, most discriminators would require redesigning to preserve linearity in the off-tuned condition. The present types, on the other hand, are compact and reliable, and have been found to satisfy all presently known requirements.

Measuring Techniques for Broad-Band, Long-Distance Radio Relay Systems*

W. J. ALBERSHEIM†, SENIOR MEMBER, IRE

Summary—Line-up and maintenance of radio relay systems require sensitive yet rapid measurements. These are obtained by scanning the systems response as functions of time, frequency, and amplitude.

Parameters thus scanned include the transient response to step functions; frequency characteristics of gain, phase, impedance and their frequency derivatives; and amplitude characteristics of output nonlinearity and of intermodulation products.

Long-distance radio relay systems may transmit television pictures or multiple telephone conversations over more than a hundred repeaters in addition to the terminals. If their cumulative distortion is to remain within the limits of high-quality transmission, individual repeater or terminal stations must be held to very narrow tolerances by means of sensitive and accurate test equipment. On the other hand, line-up and maintenance of the complex over-all systems makes it desirable to observe the transmission characteristics at one glance instead of relying on point-by-point measurement. This goal is realized by scanning techniques.

In stable systems it is possible to use moderate scanning speeds and to trace the desired characteristic directly on a paper strip by a mechanically guided pen. For instantaneous observation of rapidly changing conditions, a cathode-ray oscilloscope is used with a scanning period shorter than the persistence of vision. Permanent records of the fluorescent traces can be obtained by photography. The system properties may be tested as functions of *time, frequency,* and *amplitude.*

I. Time Functions

The most typical time function of a system is its *transient amplitude response* to impulses or to amplitude steps which for repeated scanning are approximated by square waves. We shall not describe this well-known technique, but refer to one of the earlier papers on the subject published in the Proceedings of the I.R.E.[1] It clearly shows the distortion suffered in transmitting pulses and picture elements, but it does not separate the distortions of gain, phase, and instantaneous amplitude. These data, which are needed to avoid nonlinear distortion and intermodulation, are obtained by "steady-state" frequency and amplitude characteristics.

II. Frequency Functions

Typical parameters dependent on frequency are gain and its frequency derivative, phase, envelope delay distortion, and impedance.

Amplitude scanning devices are well known; one recently described[2] shows their use for wide-band microwave transmission. However, a brief inspection of an RF gain-frequency scanner, such as shown by the block schematic of Fig. 1, will serve to bring out features which most scanning devices have in common.

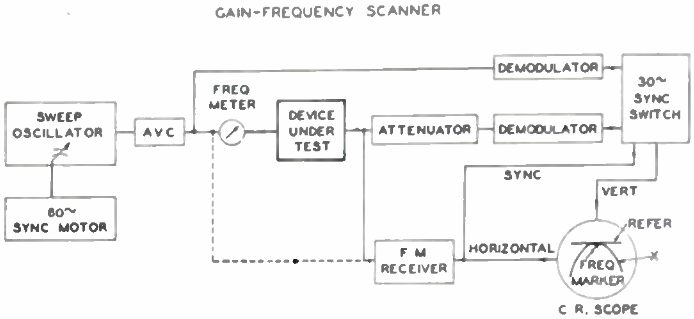

Fig. 1—Gain-frequency scanner.

The system under test must have an input of variable frequency and of constant amplitude. In the circuit shown, this is provided by an LC oscillator swept over the measuring range (from 60 to 80 mc) by a variable condenser driven by a 60-rps synchronous motor and held at constant level by a fast-acting automatic volume-control circuit. The output of the system under test is demodulated by a linear rectifier and impressed on the vertical deflection plates of a cathode-ray scope after suitable amplification. In order to obtain an undistorted scale, the horizontal deflection or, within the linear cathode-ray range, the deflector voltage must be proportional to the instantaneous output frequency.

This proportionality to the independent parameter is a general requirement, whether one scans time, frequency, or amplitude functions. One way to obtain it is to use a linear "sawtooth" time sweep and an equally linear time variation of the parameter. In our case, this would require a shaping of the rotary condenser blades, timing of frequency and cathode-ray sweep, and a brightness control to suppress the return sweep.

All these complications are avoided if the horizontal deflecting voltage is derived from the frequency (which may vary at an arbitrary rate) by a linear frequency detector, such as the FM receiver shown on Fig. 1, connected to either the input or the output side of the tested circuit. The latter connection is preferable when the transmission time of the measured system is an appreciable fraction of the scanning period because it produces a correct cathode-ray picture regardless of the time lag between transmitter and receiver.

* Decimal classification: R200×R480. Original manuscript received May 14, 1951; revised manuscript received, January 24, 1952. Presented at Conference on High-Frequency Measurements in Washington, D. C., on January 12, 1951.

† Bell Telephone Laboratories, 463 West St., New York 14, N. Y.

[1] H. E. Kallmann, "Portable equipment for observing transient response of television apparatus," Proc. I.R.E., vol. 28, p. 351; August, 1940.

[2] F. E. Radcliffe, "Traveling-wave amplifier measurements," *Electronics*, vol. 39, p. 110; August, 1951.

Exact measurement is aided by superimposing the output characteristic upon a standard of comparison, such as a zero line or, in Fig. 1, the input characteristic. This is done by rapid switching between input and output voltage. The switching rate should be faster than the persistence of vision; but it should allow for at least one full scan in each position and, preferably, should be synchronized and phased with a multiple of the scanning period. These requirements are satisfied by a 30-cycle synchronous switch.

As a further aid in calibration, the scanner contains a movable frequency marker. This is obtained by loosely coupling a sharply tuned wavemeter to the test circuit so that it produces a narrow absorption notch in the swept characteristic.

In systems which require very smooth response, the frequency derivative of the amplitude may be more significant than the amplitude itself. The change in equipment from a gain scanner to a gain-slope scanner is simple in principle. Assuming a linear frequency scan, the frequency derivative is proportional to the time derivative. This is obtained by the well-known differentiator circuits, for example, by an amplifier stage with inductive coupling impedance.

The reactive part of the transmission constant, that is, the phase characteristic, is even more important than the amplitude response in FM-microwave radio relays.

A scanning device for the Nyquist diagram, which presents phase as a function of gain, was developed as early as 1934,[3] a phase-frequency scanner in 1941.[4] However, a direct scanning of phase can give accurate results only if the total phase shift of the tested systems is reasonably small. This makes it useless for long-distance radio relays in which the phase shift may run into millions of radians.

It is therefore preferable to measure the frequency derivative of the phase shift, that is, *envelope delay*, and to suppress its constant component. This suppression, too, is imperative because in a transcontinental radio system the total transmission time delay exceeds 15 milliseconds, whereas the envelope delay variations must be measured and controlled to within a few millimicroseconds.

Devices capable of scanning delay distortion to the required precision have recently been described.[5,6] In the scanner developed by the author in co-operation with Hunt, the carrier frequency is subjected to a small amount of 200-kc modulation and swept over the measuring range at a repetition rate of normally 60 scans per second. After passing through the system under test, the 200-kc modulation and the slow sawtooth sweep are recovered by an FM receiver and separated by band-pass filters. The sawtooth voltage is impressed upon the horizontal cathode-ray deflectors. The 200-kc signal contains small phase variations which are proportional to the envelope delay distortion and are transformed into amplitude variations by vector addition to an unmodulated 200-kc wave. This reference wave is derived from the received 200-kc signal by a narrow band filter which strips off the modulation sidebands. Thus it becomes feasible to place the sending and receiving components of the equipment at opposite terminals of a long-distance communication system and to measure the one-way delay distortion.

Impedance is a third parameter related to amplitude and phase. Its control is essential because poor impedance match causes reflections, and in systems containing long transmission lines, objectionable phase and amplitude ripples. Fig. 2 shows one method of impedance mismatch measurement, developed by Bellows, in which the reflection coefficient is separated by a directional coupler and scanned as an amplitude characteristic. A different method based on series or shunt insertion loss was recently published.[7]

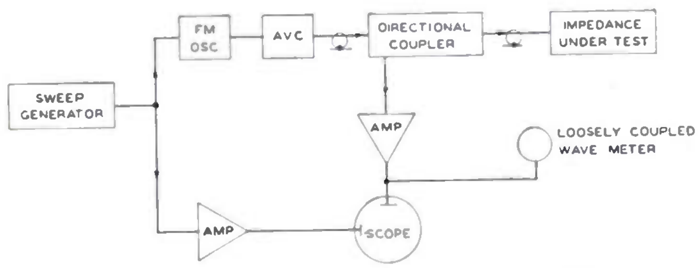

Fig. 2—Scanner for mismatch and reflection coefficient.

III. Amplitude Functions

The amplitude linearity of a relay system for frequency-division-multiplex telephony is vitally important because its lack causes harmonic distortion, intermodulation, and interchannel cross talk.

The most elementary amplitude scan traces the instantaneous output voltage as a function of instantaneous input voltage. In an ideal amplifier this trace is a straight line; in an actual amplifier containing phase shift, it is a Lissajou figure which approximates an ellipse.

Usually a linear phase shift is of no importance, hence the rms output amplitude is shown as a function of input amplitude; this requires linear detection of both input and output signal. In an AM carrier system, a scan of detected output voltage versus instantaneous input voltage indicates the modulation characteristic. Amplitude-modulation scanners have been commercially available for a number of years, and need no further discussion. By the addition of a linear frequency detector, frequency modulation may be scanned in a

[3] E. Peterson, J. G. Kreer, and L. A. Ware, "Regeneration theory and experiment," Proc. I.R.E., vol. 22, p. 1191; October, 1934.
[4] B. D. Loughlin, "A phase-curve tracer for television," Proc. I.R.E., vol. 29, p. 1191; March, 1941.
[5] A. R. Vallerino, "Incremental Delay Indicator," Orally presented, IRE Convention, New York, N. Y.; 1950.
[6] L. E. Hunt and W. J. Albersheim, "A scanner for rapid measurement of envelop delay distortion," Proc. I.R.E., vol. 40, pp. 454–460; April, 1952.

[7] D. A. Alsberg, "A precise sweep-frequency method of vector impedance measurement," Proc. I.R.E., vol. 39, pp. 1393–1400; November, 1951.

like manner. In order to obtain an absolute measure of frequency deviation, it is desirable to superimpose adjustable and calibrated reference deviation lines on the modulation trace by rapid switching methods discussed above.

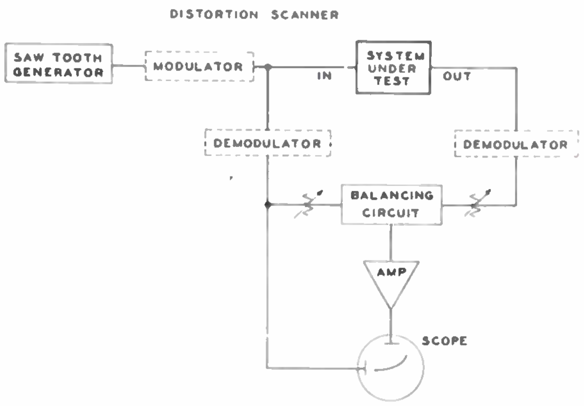

Fig. 3—Distortion scanner.

In a good amplifier system the deviations from linearity are very small and easily obscured by the nonlinearity of the cathode-ray deflection. The precision of indication may be increased a hundredfold by scanning the difference between adjustable fractions of output and input voltages, as in the *distortion scanner* developed by De Lange and shown on Fig. 3. Since most systems are free of distortion at low input levels, the voltage difference is balanced to zero at low levels. The method is applicable to instantaneous voltage, to rms amplitude, and to carrier modulation.

In FM systems, nonlinearity of the transmitter frequency deviation and of the discriminator slope have effects similar to those of amplitude nonlinearity in AM systems. Transmitter and receiver defects can be separately measured if a perfect FM receiver is available for the testing of FM transmitters and a perfect transmitter for receiver tests. Lacking these, it seems impossible to separate transmitter and receiver distortions. However, this separation has recently been accomplished by beat oscillator methods.[5,8]

The *linearity test set* developed at the Bell Telephone Laboratories and schematically shown on Fig. 4(a) and 4(b) subjects the IF transmitter carrier simultaneously to frequency modulation by a small sinusoidal 100-kc voltage and by a large 60-cycle sawtooth voltage. The IF carrier is obtained as the beat frequency of two microwave oscillators. When both the 100-kc and sweep modulations are impressed upon the same microwave FM oscillator, as shown in Fig. 4(a), the scanned 100-kc receiver output amplitude is proportional to the product of transmitter and receiver deviation slopes. Its amplitude variations measure the departure from overall slope linearity.

When the sawtooth sweep of the modulator is counteracted by an opposite sweep of the beat oscillator

[8] J. G. Chaffee and J. B. Maggio, "Frequency modulation terminal equipment for the transcontinental relay system," *Elec. Eng.*, vol. 70, p. 880; October, 1951.

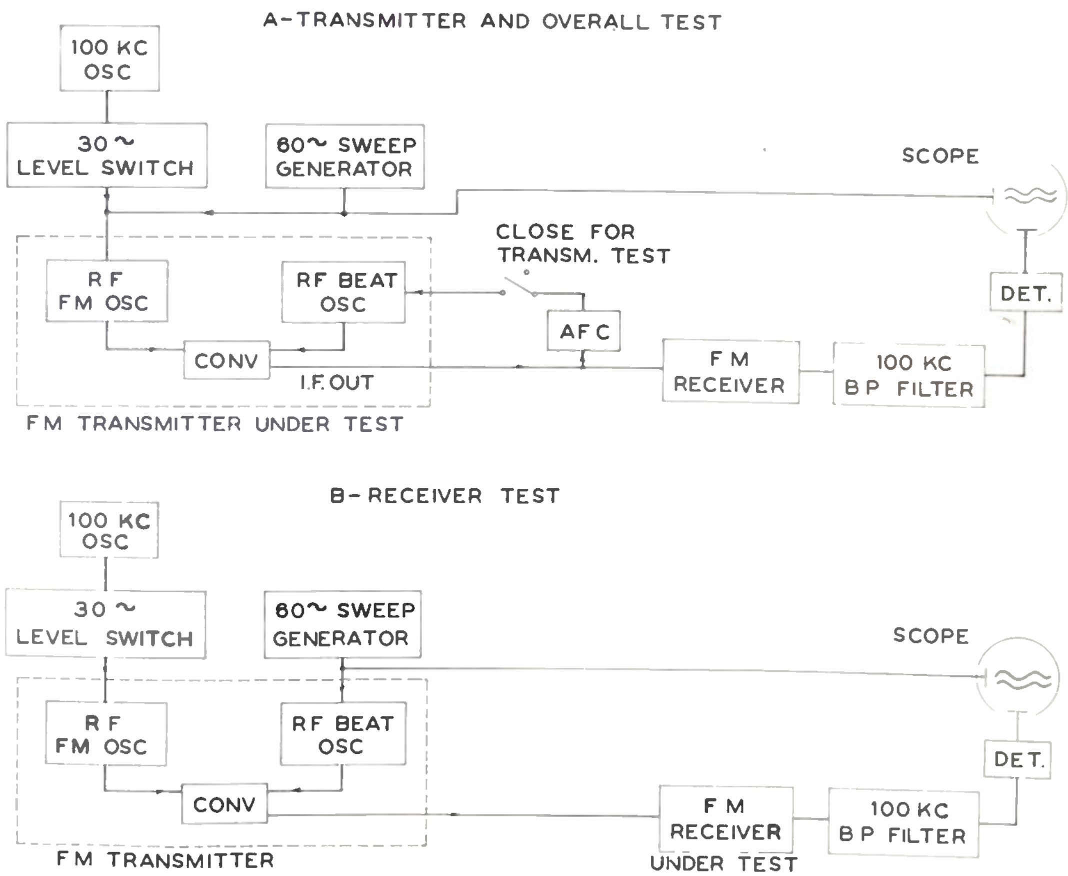

Fig. 4—FM linearity scanner.

(obtained by automatic frequency control), the IF output contains only the 100-kc FM. Detected in a fixed region of the receiver slope, its variations are a measure of transmitter-slope nonlinearity.

When the sawtooth frequency modulation is applied to the beat oscillator only, as shown on Fig. 4(b), the 100-kc modulation occurs in a fixed region of the FM oscillator slope. Hence the variations of the received 100-kc amplitude are a measure of the receiver nonlinearity.[9]

If all characteristics of modulation, transmission, and demodulation were identified and measured, it would become possible to compute the distortion of any given input signal. This, however, is a laborious and indirect procedure. Therefore, various methods have been developed for measuring directly the products of nonlinear distortion. One may, for instance, measure the second and third harmonics of a sinusoidal input signal. However, since in high-quality systems the harmonic content of the input signal may become comparable to that generated in the tested system, it is preferable to use two harmonically unrelated test tones and to measure their second- and third-order intermodulation products. This widely used method still suffers from the defect that the test signals are discrete single frequencies. An irregularity of the transmission characteristic located midway between the two test frequencies does not show up in this distortion test. It was therefore concluded[10] that in a broad-band, radio relay system, and especially in a system used for frequency-division-multiplex telephony, the most significant input signal is a random noise band covering the entire transmission range. This type of test load has the further advantage that the voltage distribution of a large number of telephone channels is approximated by the normal law distribution of random noise.[11] Since it is most important to test those distortion products which fall back into the transmission band, several test-frequency channels are sliced out of the input by narrow-band elimination filters; the filling in of these gaps in the received signal measures the distortion. Usually these test gaps are evenly spread over the transmission band. Their discrete locations hardly impair the generality of test results because each gap receives numerous sum and difference distortions of various orders from all parts of the band.

The distortion amplitude thus measured may be scanned as a function of input amplitude. However, since very large input and output amplitude ratios must be measured, the results are more easily evaluated in the logarithmic (decibel or neper) level scale than in a linear amplitude scale. This can be done by logarithmic compression circuits in combination with linear demodulators, or directly by logarithmic demodulators. A *logarithmic distortion scanner* embodying the above features is shown schematically on Fig. 5.

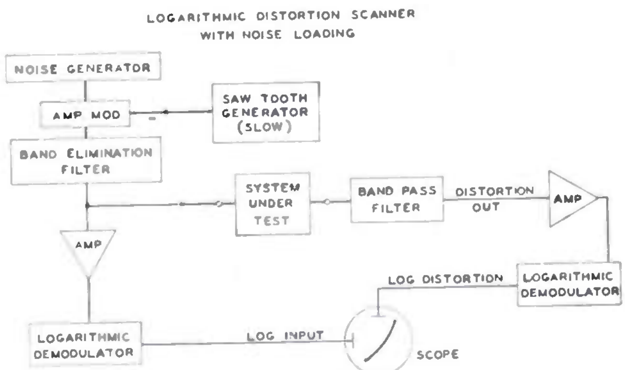

Fig. 5—Logarithmic distortion scanner with noise loading.

Conclusion

Modern measuring techniques tend to produce continuous and preferably instantaneous records of transmission characteristics as functions of time, frequency, and amplitude. The combined use of these techniques has aided in the design, line-up, and maintenance of a broad-band radio relay system which transmits coast-to-coast television pictures or hundreds of telephone conversations per radio channel.

[9] The linearities tested by the above methods refer to slow, quasi-linear modulation. Additional dynamic distortions may occur at high modulating frequencies.

[10] E. Peterson, "Gas tube noise generator for circuit testing," *Bell Lab. Rec.*, vol. 18, p. 81; November, 1939.

[11] It is of course realized that the ultimate test of a multiplex system is its service performance with multichannel speech load, which is usually only available in commercial use.

Correction

A. H. LaGrone, author of the paper, "Volume Integration of Scattered Radio Waves," which appeared on page 54 of the January, 1952 issue of the Proceedings of the I.R.E., has brought the following errors to the attention of the editors:

Line 5 in column 1 should read, " . . . scattered power density per unit area at a . . . "

Line 2 in column 2 should read, " . . . unit area at the receiver point is then P, where . . . "

The third error occurs in the footnote followed by the asterisk. The Naval Research Contract is N5Ori-136, T. O. I. and not M5ori-136, T. O. I. as shown.

Standards on Pulses: Definitions of Terms, Part II, 1952*

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I. INTRODUCTION

THE MEANINGS of commonly used terms in pulse work have often been a matter of disagreement. The IRE Standards Committee, faced with the fact that different technical committees proposed different definitions for the same term, and wishing to try to introduce uniformity where little has existed, set up a special task group with wide representation of special interests to propose standard definitions of terms concerned with pulses. Months of work by the task group and intensive critical review by the Standards Committee has led to results, the second half of which are given below. These are necessarily compromises. The Standards Committee urges IRE members (a) to try to use the terms according to the definitions below so that reasonable uniformity may be achieved, and (b) to take into account that in this particularly controversial region many compromises have been necessary so that favorite meanings and uses may appear not to have been considered, whereas in actuality it is unlikely that the very thorough review of the field has failed to unearth them for the Committee's consideration.

Since many pulse shapes are possible, and a clear concept of the ones under discussion is desirable, it may be helpful to use drawings of pulse shapes, pulse times, magnitudes, durations, and the like to show how these quantities apply.

In these definitions, linear superposition of pulses and possibly other waveforms is understood. Since it is possible to generate pulses whose characteristics may not seem to be adequately covered by the definition, it has been assumed that complex pulses can be analyzed in terms of more fundamental pulses and waveforms somewhat as a complex periodic wave can be considered as the sum of a fundamental and harmonics.

* Reprints of this Standard, 52 IRE 20 S1, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.
Part I of these definitions, constituting the first half, was published in PROC. I.R.E., vol. 39, p. 624; June, 1951, as Standard 51 IRE 20 S1.

II. DEFINITIONS

Note—Terms in italics are defined elsewhere in the Standards on Pulses: Definitions of Terms, Parts I and II.

Bidirectional Pulse. A *pulse* in which the variation from the normally constant value occurs in both directions.

Bidirectional Pulse Train. A *pulse train*, some *pulses* of which rise in one direction and the remainder in the other direction.

Carrier-Frequency Pulse. A carrier, amplitude modulated by a *pulse*. The amplitude of the modulated carrier is zero before and after the *pulse*.

Note—Coherence of the carrier (with itself) is not implied.

Equalizing Pulses (Television). *Pulse trains* in which the *pulse-repetition frequency* is twice the line frequency and which occur just before and just after a vertical synchronizing pulse.

Note—The equalizing pulses minimize the effect of line-frequency pulses on the interlace.

Fruit Pulse (Fruit[1]). A *pulse reply* received as the result of interrogation of a transponder by interrogators not associated with the responsor in question.

Main Bang. Transmitted *pulse*, within a radar system.

Oscillator Starting Time, Pulsed. See *Pulsed Oscillator Starting Time.*

Peak Pulse Power. The power at the maximum of a *pulse* of power, excluding *spikes*.

Peak Pulse Power, Carrier-Frequency. The power averaged over that carrier-frequency cycle which occurs at the maximum of the pulse of power (usually one half the maximum instantaneous power).

Periodic Pulse Train. A *pulse train* made up of identical groups of *pulses*, the groups repeating at regular intervals.

Power, Carrier-Frequency, Peak Pulse. See *Peak Pulse Power, Carrier-Frequency.*

Power, Peak Pulse. See *Peak Pulse Power.*

Pulse, Bidirectional. See *Bidirectional Pulse.*

Pulse, Carrier-Frequency. See *Carrier-Frequency Pulse.*

Pulse Code.

(1) A *pulse train* modulated so as to represent information.

(2) Loosely, a code consisting of *pulses*, such as Morse code, Baudot code, binary code.

Pulse-Code Modulation (PCM). Modulation which involves a *pulse code*.

Note—This is a generic term, and additional specification is required for a specific purpose.

[1] Deprecated.

Pulse Delay, Receiver. See *Transducer Pulse Delay.*

Pulse Delay, Transducer. See *Transducer Pulse Delay.*

Pulse Delay, Transmitter. See *Transducer Pulse Delay.*

Pulse Droop. A distortion of an otherwise essentially flat-topped rectangular *pulse* characterized by a decline of the *pulse* top.

Pulse Frequency Modulation (PFM). A form of *pulse time modulation* in which the *pulse repetition rate* is the characteristic varied.

Note—A more precise term for "pulse frequency modulation" would be "*pulse repetition-rate* modulation."

Pulse Interrogation. The triggering of a transponder by a *pulse* or *pulse mode*.

Note—Interrogations by means of *pulse modes* may be employed to trigger a particular transponder or group of transponders.

Pulse Jitter. A relatively small variation of the *pulse spacing* in a *pulse train*.

Note—The jitter may be random or systematic, depending on its origin, and is generally not coherent with any *pulse modulation* imposed.

Pulse Mode.

(1) A finite sequence of *pulses* in a prearranged pattern used for selecting and isolating a communication channel.

(2) The prearranged pattern.

Pulse-Mode Multiplex. A process or device for selecting channels by means of *pulse modes*.

Note—This process permits two or more channels to use the same carrier frequency.

Pulse Mode, Spurious. See *Spurious Pulse Mode.*

Pulse Modulation.

(1) Modulation of a carrier by a *pulse train*.

Note—In this sense, the term is used to describe the process of generating *carrier-frequency pulses*.

(2) Modulation of one or more characteristics of a *pulse carrier*.

Note—In this sense, the term is used to describe methods of transmitting information on a *pulse carrier*.

Pulse Multiplex. Deprecated. See *Pulse-Mode Multiplex.*

Pulse Phase Modulation (PPM). See *Pulse Position Modulation (PPM).*[2]

Pulse Power, Carrier-Frequency, Peak. See *Peak Pulse Power, Carrier-Frequency.*

Pulse Power, Peak. See *Peak Pulse Power.*

[2] "Standards on Pulses: Definitions of Terms, Part I," vol. 39, pp. 624–626; June, 1951.

Pulse Reply. The transmission of a *pulse* or *pulse mode* by a transponder as the result of an interrogation.

Pulse Separation. The interval between the *trailing-edge pulse-time* of one *pulse* and the *leading-edge pulse-time* of the succeeding *pulse.*

Pulse Shaper. Any transducer used for changing one or more characteristics of a *pulse.*

Note—This term includes *pulse* regenerators.

Pulse Shaping. Intentionally changing the shape of a *pulse.*

Pulse, Single-Polarity. See *Unidirectional Pulse.*[2]

Pulse Spike Amplitude. The *peak pulse amplitude* of the *pulse spike.*

Pulse Tilt. A distortion in an otherwise essentially flat-topped rectangular *pulse* characterized by either a decline or a rise of the *pulse* top.

Pulse Train, Bidirectional. See *Bidirectional Pulse Train.*

Pulse Train, Periodic. See *Periodic Pulse Train.*

Pulse-Train Spectrum (Pulse-Train Frequency-Spectrum). The frequency distribution of the sinusoidal components of the *pulse-train* in amplitude and in phase angle.

Pulse Train, Unidirectional. See *Unidirectional Pulse Train.*

Pulse Valley. The part of the *pulse* between two specified maxima.

Note—Unless otherwise specified, it is to be understood that the maxima are the first and the last.

Pulse Width. See *Pulse Duration.*[2]

Pulsed Oscillator. An oscillator which generates a *carrier-frequency pulse* or a train of *carrier-frequency pulses.*

Note—These *carrier-frequency pulses* may occur as the result of self-generated or externally applied *pulses.*

Pulsed-Oscillator Starting Time. The interval between the *leading-edge pulse-time* of the *pulse* at the oscillator control terminals and the *leading-edge pulse-time* of the related output *pulse.*

Pulses, Equalizing. See *Equalizing Pulses.*

Receiver Pulse Delay. See *Transducer Pulse Delay.*

Single-Polarity Pulse. See *Unidirectional Pulse.*[2]

Spurious Pulse Mode. An unwanted *pulse mode,* formed by the chance combination of two or more *pulse modes,* which is indistinguishable from a *pulse interrogation* or *pulse reply.*

Transducer Pulse Delay. The interval of time between a specified point on the input *pulse* and a specified point on the related output *pulse.*

Note 1—This is a general term which applies to the *pulse delay* in any transducer, such as receiver, transmitter, amplifier, oscillator, and the like.

Note 2—Specifications may require illustrations.

Transmitter Pulse Delay. See *Transducer Pulse Delay.*

Unidirectional Pulse Train. A *pulse train* in which all *pulses* rise in the same direction.

Note—A *unidirectional pulse train* may contain *bidirectional pulses.*

Average Radio-Ray Refraction in the Lower Atmosphere*

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***Summary*—Existing corrections for atmospheric refraction in radio-field intensity computations are reviewed with respect to their application to computation of ray bending. A practical scheme is presented for calculating atmospheric refraction of radio-frequency rays numerically from radiosonde data. Ray-bending computations are made for a range of climatological conditions for rays passing entirely through the atmosphere and arriving or departing tangentially at the earth's surface. Some discussion is included regarding the uncertainty in refractive-index computations from meteorological sounding data.**

I. Introduction

THE PROBLEM of radio-ray tracing is closely connected with the problem of astronomical refraction. The principal difference between optical-

* Decimal classification: R112.42. Original manuscript received by the Institute, May 3, 1951. This paper was presented at the Washington Meeting of the URSI, May 7, 1947, and is a revised version of one originally published by the Central Radio Propagation Laboratory of the National Bureau of Standards as Report CRPL 2-2, August 11, 1947. It is intended to be the first part of a standard-refraction study, the goal of which is the calculation of radio-field intensity as affected by average-refraction conditions when one or both of the terminals is very high, such as a high-flying aircraft or rocket.

The "lower atmosphere" means, arbitrarily, that portion of the atmosphere up to about 18 km. However, an estimate is made in this study of the contribution of the refraction of the atmosphere above this elevation, excluding any ionospheric effects. Measurements of ionospheric effects on radio-ray refraction have been made by R. Payne-Scott and L. L. McCready. "Ionospheric effects noted during dawn observations on solar noise," *Terr. Mag. Atmo. Elec.*, vol. 53, pp. 429–432; December, 1948; and the correction in the *Jour. Geophys. Res.*, vol. 53, p. 98; March, 1949.

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ray and radio-ray refraction is due to the effect of the relatively large contribution of water vapor to the refractive index of air at radio frequencies. The history of atmospheric optical-ray tracing is a long one, and is devoted mostly to the computation of astronomical refraction and somewhat to geodetic surveying. B. Garfinkel[2] has reviewed the past history of astronomical refraction calculations, and has presented a method of obtaining total optical refraction for a ray passing entirely through the earth's atmosphere for angles of arrival less than 10 degrees. In astronomical-refraction investigations, the air density-height distribution function is approximated by plausible physical models. B. Garfinkel also fitted mean sounding data to his model, which featured a temperature lapse up to the tropopause, and then an isothermal stratosphere. His results agree quite well, within error of measurement, with the semi-empirical tables of Radau and Pulkowa.

The tabulated sounding data which Garfinkel used were taken from W. J. Humphreys[3] and were representative of four European cities between the years 1900 and 1912. Since the date of Garfinkel's paper, the United States Weather Bureau has published averages of radiosonde data taken for the most part from July, 1939 to December, 1943, for the North American network of stations from Alaska to the Caribbean Sea.[4] These data are used in the present radio-ray computations.

II. Survey of Radio-Ray Refraction[5] Corrections

The objective of atmospheric radio-wave refraction studies is usually to find a convenient method of correcting the radio-wave field intensity equations for this effect. This final step has not been taken in this paper, but a method of obtaining the true ray bending is presented in Section III. A survey of existing refraction corrections is made herein in terms of ray-bending calculations, instead of the field-intensity computations, for which they were originally designed. It is shown how widely divergent the results of the different methods are from the true refraction. Refraction effects vary with height in the atmosphere and meteorological conditions. The most widely used method of approximation for the effects of atmospheric refraction is the use of an effective earth's radius equal to four-thirds the actual radius. Therefore this approximation is given particular attention.

In the present paper, refraction has been computed from the actual refractive-index structure of the atmosphere as obtained from average meteorological data at locations representative of a range of climatological conditions. The assumption is made, as is usually done, that the atmosphere is horizontally homogeneous. The results have been plotted as the solid curves A, B, and C, in Fig. 1. Curves A, B, and C represent the actual variation of refraction with height, computed by methods presented in this paper, for the years 1939 through 1943, for Fairbanks, Alaska, in April; for Washington, D. C., in October; and for San Juan, Puerto Rico, in July. Broken curves A', B', and C' are the approximations obtained by a method due to Pearcey, see (11), which uses a four-thirds earth's radius factor up to the height at which the refractive index becomes unity and assumes no refraction beyond this point.

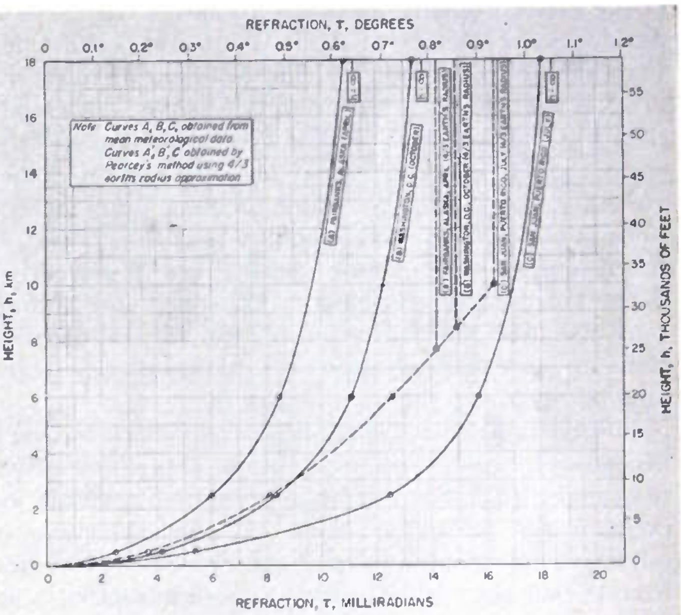

Fig. 1—Bending of a tangential radio ray.

The earliest successful attack on the problem of correcting radio propagation analysis for atmospheric refraction was introduced by Schelleng, Burrows, and Ferrell,[6] when they used the concept of effective earth's radius. They found that a factor of four-thirds was satisfactory for average annual conditions, up to 0.5 km as deduced from Humphreys' weather data. However, the correction thus introduced is only strictly correct for an atmosphere with linear-refractive index gradient of 39×10^{-6} units per km, and thus cannot possibly hold beyond the height (approximately 10 km) at which the refractive index has reached the value of unity. Furthermore, the atmosphere actually has a nonlinear-refractive index gradient, and the true correction will vary with elevation and distance.

Eckart and Plendl[7] showed how to correct field-intensity calculations for atmospheric refraction of nearly horizontal rays by fitting the actual refractive index

[2] B. Garfinkel, "An investigation in the theory of astronomical refraction," *Astronomical Jour.*, vol. 50, p. 169; February, 1944.

[3] W. J. Humphreys, "Physics of the Air," McGraw-Hill Book Co., New York, N. Y.; 1929.

[4] B. Ratner, "Upper Air Average Values of Temperature, Pressure, and Relative Humidity over the United States and Alaska," Climate and Crop Weather Division, U. S. Bureau, Department of Commerce, Washington, D. C.; May, 1945.

[5] Refraction is defined here as the total bending of a ray over a path. The curvature at a point on a curve is defined as the rate of change of tangent direction per unit length of curve at that point, and is measured in terms of change of angle per unit length of path. For a circle, the curvature at any point is constant and equal to the reciprocal of the radius. When the curvature is known along a ray, refraction may be computed as the integral of curvature over the path.

[6] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra short wave propagation," Proc. I.R.E., vol. 21, pp. 427–463; March, 1933.

[7] G. Eckart and H. Plendl, "Die Überwindung der Erdkrummung bei Ultrakurzwellen durch die Strahlenbrechung in der Atmosphare," *Hoch, und Electroak.*, vol. 52, pp. 44–58; August, 1938.

structure by means of two parabolic curves, one from the surface to 5 km, and the other from 5 to 10 km. Above this elevation they considered the refraction negligible. This method furnished correction tables for a mean atmosphere to 10 km. To extend the corrections to greater heights or to different refractive index distributions as encountered in different geographical locations would require additional parabolic fits of data and extensive calculations. There is some question regarding the validity of other approximations which were made by these authors.[8]

Vvedensky and Ponomarev[9] assumed an analytic expression for the variation of dielectric constant with height, $\epsilon(r)=\delta+\gamma a^2/r^2$, where the coefficients δ and γ can be fitted empirically; δ is of the order of 0.8 and γ of the order 0.2 to fit mean conditions near the surface of the earth. For this law of variation, the refractive index decreases to unity at about 10 km and becomes less than one at greater heights. The gradient of refractive index is very closely linear up to this height. This model then has shortcomings similar to those of the effective earth's radius concept treated elsewhere in this paper. However, the authors were able to effect many integrations in closed form, and indicated that they hoped, in the future, to treat a more general law of refractive-index distribution with height. Eckersley and Millington[10] also used this expression for the vertical distribution of dielectric constant in their phase-integral method of solving the wave equation.

In solar-radio emission studies refraction correction is necessary for solar-radio rays. In the case of low-elevation angles (θ_0), Pearcey[11] has presented a simple expression for refraction using the surface value of refractive index and a linear gradient, 39×10^{-6}/km, from the surface to that height at which refractive index becomes unity. Above this height the refraction is considered to have ceased. This is equivalent to using a four-thirds earth's radius factor but, in addition, the final height is specified at which the correction is applicable. This model has the advantage of being simple, and the nature of the approximation for total bending may be examined in the dashed curves of Fig. 1. Pearcey's refraction formula for low ground angles of arrival or departure, $0\leq\theta_0\leq10°$ is given by

$$\tau = \frac{\Delta n_0}{\left(\frac{h_2}{a} - \Delta n_0\right)}\left[\sqrt{2\left(\frac{h_2}{a} - \Delta n_0\right) + \theta_0^2} - \theta_0\right] \quad \text{in radians,} \tag{1}$$

where

τ is the total bending in radians;

Δn_0 is (n_0-1), where n_0 is the value of the refractive index at the earth's surface;

h_2 is the elevation at which Δn becomes zero under a standard refractive index gradient of 39×10^{-6} units/km;

a is the earth's radius, expressed in the same units as h_2.

If the gradient corresponding to any arbitrary effective earth's radius factor k were used, the same formula would hold, the corresponding changes in interpretation of symbols being inserted. Equation (1) may be placed in more convenient form

$$\tau = (k-1)\left[\sqrt{\frac{2h_2}{ka} + \theta_0^2} - \theta_0\right]. \tag{1a}$$

Stickland[12] attempted to correct radio field-intensity calculations for the variation of refraction with height. She found an empirical relation for the average atmospheric refractive index distribution in terms of the ratio m of the ray radius of curvature to that of the earth at any height h in the atmosphere for nearly horizontal rays;

$$m = 2.5 + 1.0h, \tag{2}$$

h being given in km. Meteorological conditions in the lowest kilometer are admittedly quite variable, but the formula was considered valid to 10 km, above which height the refraction was assumed to have ceased. It was then suggested that for computational purposes an average ratio $\bar{m}$ be used between any two points "close together" at heights h_1 and h_2, $\bar{m}=2.5+0.5(h_1+h_2)$. When one point is near the ground, $h_1\cong0$, and then we have

$$\bar{m} = 2.5 + 0.5h. \tag{3}$$

If this average $\bar{m}$ is used for calculating the actual refraction of a tangential ray, incorrect results are obtained. To demonstrate this, the total refraction will be worked out in terms of Stickland's average $\bar{m}$ concept, and it will be shown that these values are in disagreement with the values of refraction as computed by the method outlined later in this paper. Equation (1a) may be used to calculate the refraction for the Stickland approximation up to any height h, by setting $\theta_0=0$ and inserting the value of $\bar{k}$ related to the $\bar{m}$ in (3) by

(earth's curvature) − (ray curvature)
= "effective" earth's curvature

Or;

$$1/a - 1/\bar{m}a = 1/\bar{k}a.$$
$$\bar{k} = \bar{m}/(\bar{m}-1). \tag{4}$$

The bending of a radio ray between the ground and height h, computed by (1a), using the values of $\bar{k}$ ob-

[8] B. A. Vvedensky, and A. G. Arenberg, *Uspekhi Physicheskikh Nauk*, vol. 25, pp. 273–309; 1941.

[9] B. A. Vvedensky, and M. I. Ponomarev, "Application of methods of geometrical optics to the determination of the path of ultra-short waves in a non-homogeneous atmosphere," *Bull. Acad. Sci.* (USSR), no. 9, pp. 1201–1210; 1946 (in Russian).

[10] T. L. Eckersley, and G. Millington, "Application of the phase integral method to the analysis of the diffraction and refraction of wireless waves round the earth," *Phil. Trans. A*, vol. 237, pp. 273–309; June, 1938.

[11] J. L. Pawsey, R. Payne-Scott, and L. L. McCready, "Solar radiation at radio frequencies and its relation to sunspots," *Proc. Roy. Soc. A*, vol. 190, pp. 357–375; August, 1947.

[12] A. C. Stickland, "Refraction in the lower atmosphere and its applications to the propagation of radio waves," *Met. Factors in Radio-wave Propagation*, Phys. Soc. (London), pp. 253–267; 1947.

tained from (3) and (4), has been plotted in Fig. 2 as Curve *A*. Also plotted in this same figure are (a) Curve *B*, the actual refraction for a Stickland atmosphere defined by (2) and computed by the method presented

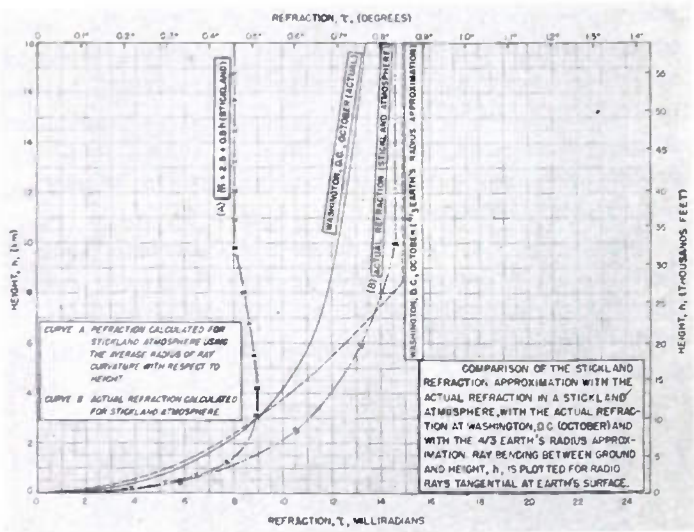

Fig. 2—Comparison of various radio-ray refraction approximations with actual refraction for Washington, D. C., in October.

in this paper; (b) the mean refraction for Washington, D. C., in October; and (c) the four-thirds earth's radius approximation. It may be seen that the Stickland averaging process gives too low values of refraction, and that the simple four-thirds earth's radius approximation is at least as satisfactory as Stickland's refraction correction. The reason for the failure of Stickland's formula in this case is that the method of averaging m has the effect of weighting all height intervals equally, whereas the concept of refraction as the integrated curvature along the ray path allows for the condition that almost horizontal rays have much longer paths in the lower layers where the refraction is greatest.

From Figs. 1 and 2, it may be seen that the refraction approximations in use at the present time are not quite adequate for all seasons, geographical locations, or very high altitudes. However, the effect on field-intensity calculations of the deviations of the refraction approximations from the true refraction has not been determined in this study.

III. Computation of Refraction

A method is now presented for computing the refraction of radio-frequency rays through an atmosphere of known refractive-index distribution, under the usual assumption that surfaces of equal index are spherical and concentric with the earth (horizontal homogeneity). The expression for the refraction $\tau_{1,2}$ of a ray, as illustrated in Fig. 3, between the heights h_1 and h_2 and making an angle θ with the spherical refracting surface at the height h, has been derived, although with different notation, in the literature[13] as

$$\tau_{1,2} = -\int_{n_1}^{n_2} \frac{\cot\theta}{n} dn. \tag{5}$$

Since n varies from at most 1.0004 at the earth's surface to unity at the top of the atmosphere, the factor $1/n$ may be taken as unity with an error of less than 4 parts in 10,000, in the computed refraction. Equation (5) then simplifies to

$$\tau_{1,2} = -\int_{n_1}^{n_2} \cot\theta dn. \tag{6}$$

Another form for (6), is

$$\tau_{1,2} = -\int_{\Delta n_1}^{\Delta n_2} \cot\theta d\Delta n$$
$$= \overline{(\cot\theta)}_{1,2}(\Delta n_1 - \Delta n_2) \text{ in radians,} \tag{6a}$$

where Δn is $(n-1)$ and $\overline{(\cot\theta)}_{1,2}$ is the mean value of the cotangent function over the interval Δn_1, Δn_2. The problem is solved when a proper value of $\overline{(\cot\theta)}_{1,2}$ can be found. Complicating factors are (a) θ is a function of both Δn and h as determined from Snell's law and (b) Δn is only known empirically as a function of h and is given by a table of values computed from radiosonde data. According to Snell's law, the angles θ for a ray path through concentric spherical refracting surfaces are determined from the formula,

$$nr\cos\theta = n_0 r_0 \cos\theta_0 = \text{constant,} \tag{7}$$

where $r = a + h$ is the radius of the spherical refracting surface, at the elevation h, a is the earth's radius, and the subscript zero refers to the ground level.

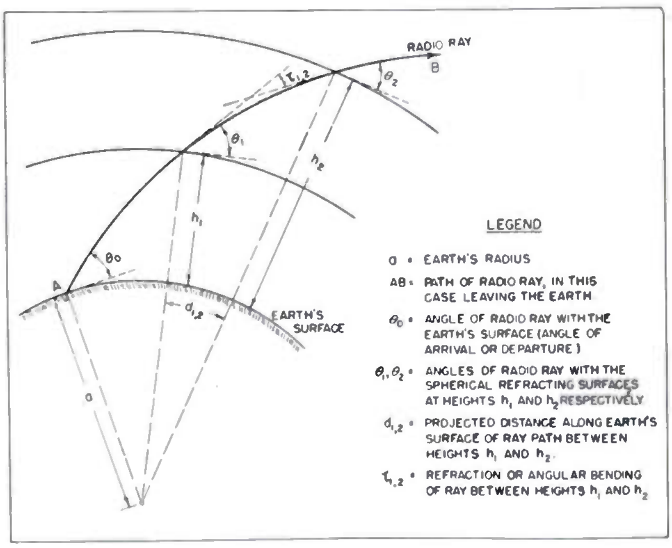

Fig. 3—Refraction of radio ray between heights h_1 and h_2.

At small elevation angles, $\theta_0 \leq 10°$, (7) transforms, neglecting quantities of second and higher orders, to[14]

$$\frac{\theta^2 - \theta_0^2}{2} = \left(\Delta n + \frac{h}{a}\right) - \left(\Delta n_0 + \frac{h_0}{a}\right). \tag{8}$$

[13] W. M. Smart, Textbook on Spherical Astronomy, Cambridge University Press, 2nd ed., p. 64, eq. 18; 1936.

[14] Committee on Propagation, NDRC "Radio wave propagation," Academic Press, Inc., New York, N. Y., p. 165; 1949.

Solving for θ,

$$\theta = \sqrt{\theta_0^2 + 2\left(\Delta n + \frac{h}{a}\right) - 2\left(\Delta n_0 + \frac{h_0}{a}\right)}, \qquad (8a)$$

where h_0 is the station elevation above mean sea level. The angle θ at any elevation may then be calculated for each station from a table of values of Δn against h.

The total refraction must be computed piecewise by using height intervals, over each of which $\overline{\cot \theta}$ may be conveniently determined for application to (6a). The number and size of the required height intervals will depend on the averaging process. The refractive index versus height curve approximated by linear segments is shown in Fig. 4, for Washington, D. C. in October.

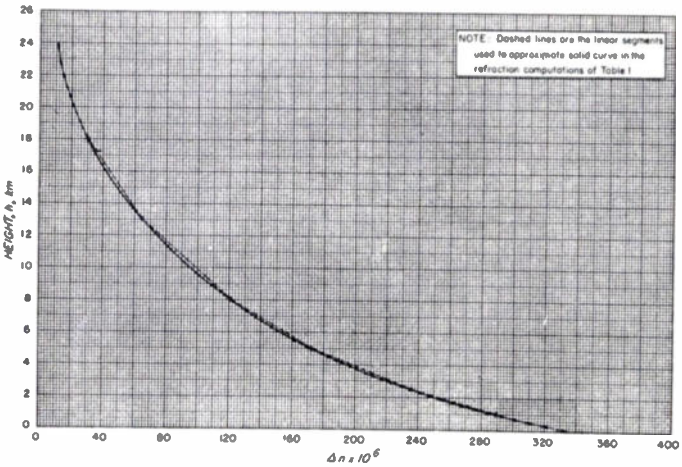

Fig. 4—Average refractive index distribution for Washington, D. C., in October.

It can be shown that $\overline{\cot \theta} \cong 1/\theta_m$,[15] where θ_m is the arithmetical mean of θ on the interval. Thus, for the interval $(\Delta n_1, \Delta n_2)$, $\theta_m = (\theta_1 + \theta_2)/2$. Substituting this approximation into (6a), with θ_m in radians, there results

[15] An infinite integrand occurs when $\theta_0 = 0$ and $n = n_0$. However, this can be evaluated by the standard rules.
F. S. Woods, Advanced Calculus, New Edition, Ginn and Co., Boston, Mass., p. 152; 1934.

$$\tau_{1,2} = \frac{\Delta n_1 - \Delta n_2}{\theta_m} \text{ in radians } (0 \leqq \theta_0 \leqq 10°), \qquad (9)$$

where the angles θ are determined from (8a).

A sample computation of the total refraction of a radio ray arriving tangentially through the mean atmosphere of Washington, D. C., in October is shown in Table 1 (bottom). About six height intervals are required to 18 km. Above 18 km, the contribution to the total refraction is small and may be shown to be approximately $\tau_{>18} = \Delta n_{18} \cot \theta_{18}$, where the subscripts refer to the values of the quantities at 18 km.

For angles of arrival or departure, θ_0, greater than 10 degrees, the total refraction through the entire atmosphere is small, and may be approximated by the first term of the expression obtained by integrating (6), by parts

$$\tau = -\int_{n_1}^{1} \cot \theta dn = (n_0 - 1)\cot \theta_0 - \int_0^{\cot \theta_0} \Delta n d(\cot \theta). \quad (10)$$

The first term is the classical expression for astrononomical refraction at high angles (>25°) (the literature,[13] p. 68), and implies the approximation $\overline{\cot \theta} = \cot \theta_0$. The error made in neglecting the second term may be evaluated from the available radiosonde data by plotting Δn versus $\cot \theta$, as has been done for a 10-degree ray in Fig. 5. In this figure, the high-angle approximation for the total refraction is the area of the box, while the area under the curve is the correction term. The ratio of the two terms is easily obtained with a planimeter, the figure being drawn to a larger scale for the purpose, and applied to (10) to give the true total refraction. The percentage error in total refraction caused by the neglect of the second term decreases with increasing elevation angle. For a 10-degree ray at Washington, D. C., in October, the correction term amounts to 3.47 per cent of the first term. For computations of $\cot \theta$ in such calculations it is suggested that since the increase of θ with elevation is small, that the following formula be used:

TABLE I

Sample Calculation of Refraction—Washington, D. C. Mean October Data $\theta_0 = 0$

(1) h (km)	(2) $h/a \times 10^6$	(3) $\Delta n \times 10^6$	(4) M	(5) $(M - M_0) = \theta_i^2/2$	(6) θ_i^2 (mr)^2	(7) θ_i mr	(8) $\frac{\theta_{i-1}+\theta_i}{2}$ mr	(9) $\Delta n_{i-1} - \Delta n_i \times 10^6$	(10) $\Delta\tau_i$ mr	(11) $\Sigma\Delta\tau$ mr
0.025	4	332	336	0	0	0	0	0	0	0
0.5	79	310	389	53	106	10.3	5.2	22	4.2	4.2
2.5	393	239	632	296	592	24.3	17.3	71	4.1	8.3
5.0	785	172	957	621	1,242	35.2	29.8	67	2.2	10.5
8.0	1,256	120	1,376	1,040	2,080	45.6	40.4	52	1.3	11.8
12.0	1,885	74	1,959	1,623	3,246	57.0	51.3	46	0.9	12.7
18.0	2,827	30	2,857	2,521	5,042	71.0	64.0	44	0.7	13.4
>18.0							71.0	30	0.4	13.8

Where $M = (\Delta n + h/a) \times 10^6$ is the only symbol not previously defined and is the so-called M-value, or *excess modified refractive index*, in millionths.

$$\cot\theta = \cot\theta_0 + (M_0 - M)\cot\theta_0 \csc^2\theta_0, \tag{11}$$

where $M = (\Delta n + h/a)$ is the excess modified refractive index.

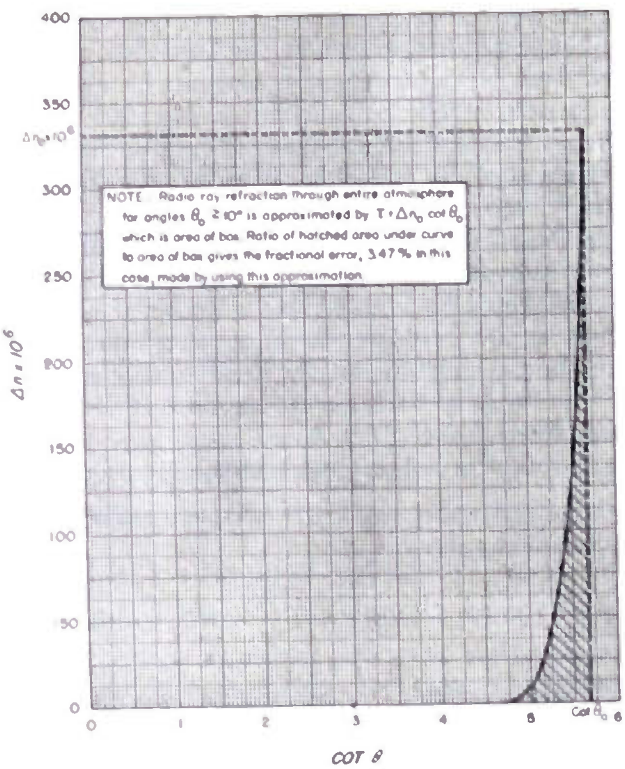

Fig. 5—Computation of refraction for angles of arrival, $\theta_0 \geq 10°$.

The mean-ray bending for radio-frequency rays over the range of angles of arrival or departure has been obtained for San Juan, Puerto Rico (July), and Fairbanks, Alaska (April), as examples of extremes; and for Washington, D. C. (October), as a mean case. These data have been plotted in Figs. 6 and 6(a). In these same figures, computed values of optical refraction have been plotted for Washington, D. C., in October, and compared with the mean-observed astronomical refraction data[16] as adjusted to these same surface-refraction conditions. Optical refraction for Washington, D. C., in October, was computed by the methods of this section, using the optical refractive index distribution obtained by neglecting the explicit water-vapor term in the radio-frequency refractive index computation, see (13a) Section IV. Good agreement is obtained which provides a check of the computational procedure.

It is now possible to determine the ray path as a function of the distance d along the earth's surface as illustrated in Fig. 3. θ has been determined from (8) or (8a), and $\tau_{1,2}$ has been determined from (9) or (10). It may be shown from geometric considerations in Fig. 3 that

$$\tau_{1,2} = \frac{d_{1,2}}{a} + (\theta_1 - \theta_2) \text{ in radians,} \tag{12}$$

where $\tau_{1,2}$, θ_2, and θ_1 are expressed in radians, and d is expressed in the same units as a, the earth radius.

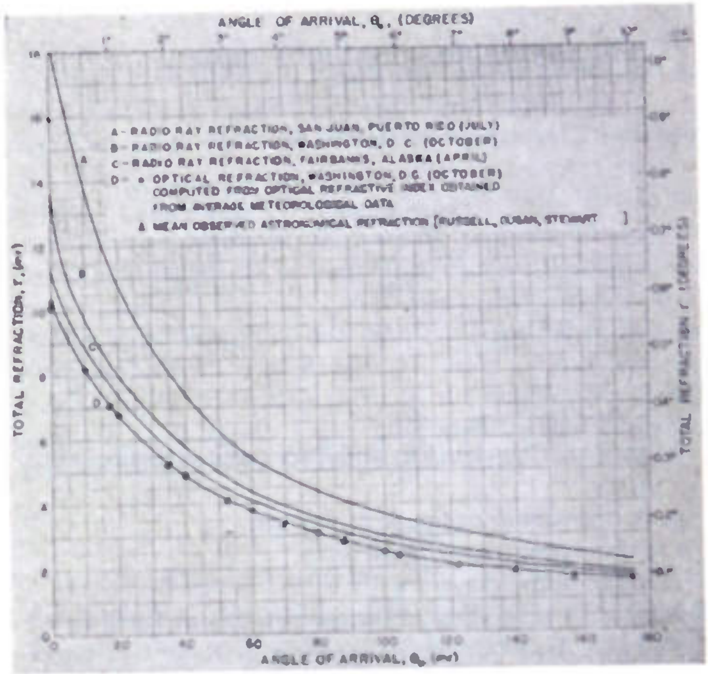

Fig. 6—Total refraction versus angle of arrival, $\theta_0 \leq 10°$.

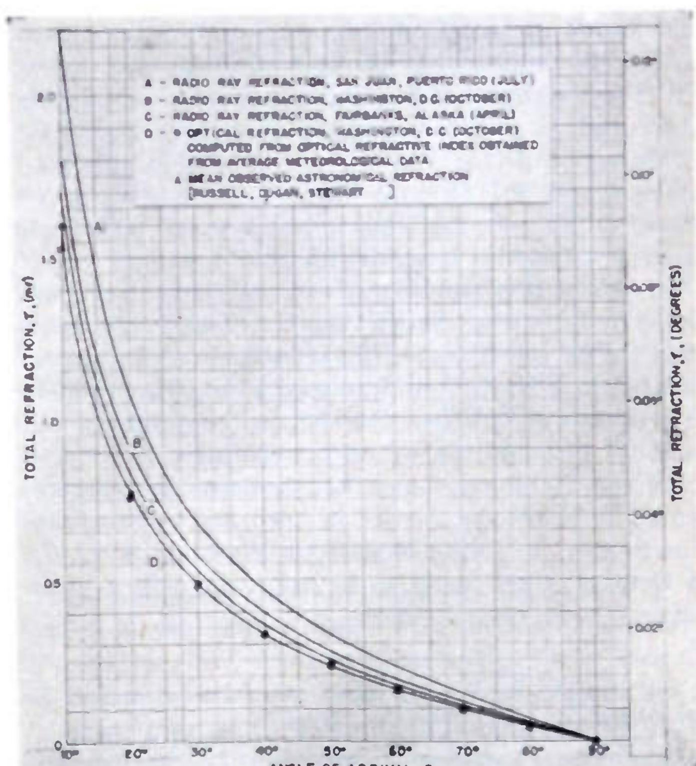

Fig. 6(a)—Total refraction versus angle of arrival, $\theta_0 \geq 10°$.

[16] H. N. Russell, R. S. Dugan, and J. Q. Stewart, "Astronomy," Ginn and Co., Boston, Mass., vol. 1, Appendix, p. vi; 1945.

IV. Refractive-Index Computations from Meteorological Data

The formula for the radio-frequency refractive index of moist air is based on a series of separate laboratory determinations of the dielectric constants of dry air and water vapor. The contributions of each are then combined to yield[17,18]

$$(n - 1) \times 10^6 = 79 \frac{p}{T} (1 + 4800 e/pT). \quad (13)$$

Equation (5) may be written in the form:

	Optical and Radio-Frequency Term	Explicit Water-Vapor Term Required only at Radio Frequencies	
$(n - 1) \times 10^6 =$	$79p/T$	$+ 3.79 \times 10^5 e/T^2$,	(13a)

where

n is the refractive index

p is the total pressure in millibars (mb) and is equal to $p_d + e$.

p_d is the partial pressure of dry air.

e is the water vapor pressure.

T is the temperature in °K.

The formula is applicable for frequencies at least up to 30,000 mc. Although some variation of refractive index might be expected to exist near various absorption lines associated with various elements in the atmosphere, Van Vleck[19] has pointed out that the observed magnitude of the absorption coefficient of oxygen and water vapor in the atmosphere whose absorption lines lie below 30,000 mc would not be expected to be accompanied by appreciable changes in refractive index. Further investigation may show that (13) is applicable at frequencies even considerably in excess of 30,000 mc.

The terms of (13a) have been evaluated in published tables (pages 367–370 and 206–219 of the literature).[14] These tables give the explicit water vapor term as a function of relative humidity and temperature.

In this paper, representative radiosonde data published by the United States Weather Bureau[4] have been used for the calculation of the refractive-index distribution for three North American stations separated widely in geographical location. The data on geographical and seasonal distributions of refractive index have been examined, and it was found that the extremes are represented by Fairbanks, Alaska, in April, and by San Juan, Puerto Rico, in July. The data for Washington, D. C., in October, were found to be approximately a mean between these two extremes.

[17] R. A. Burgoyne, "Nomograms for computation of modified index of refraction," *MIT*, RL #551; April, 1945.

[18] Discussions in the Foreword and following pages of, "Meteorological Factors in Radio-wave Propagation," *Phys. Soc.* (London); 1947.

[19] J. H. Van Vleck, "The relation between absorption and the frequency dependence of refraction," *MIT*, RL #735; May, 1945.

Ordinarily, soundings go up only to 15 or 18 km. However, at these elevations the atmosphere is approximately isothermal (stratosphere), and the refractive-index distribution may easily be extended to great heights, at least to 30 km, which is as far as the isothermal condition prevails.[20] The expression for pressure in an isothermal atmosphere is given by

$$p = p_0 e^{-g(h-h_0)/R'T_s} \quad (14)$$

where

T_s is the constant temperature of the stratosphere in °K, but which may vary with season;

p_0 is the surface pressure in mb which, in this case, refers to the base of the stratosphere, but the actual value is unimportant for the present application, as shown in the following paragraph;

R' is the gas constant for the dry stratosphere;

$h - h_0$ is the height referred to the base of the stratosphere;

g is the acceleration due to gravity.

Assuming no water vapor at these elevations, the refractive index is given by

$$(n - 1) \times 10^6 = 79p/T_s = \frac{79p_0}{T_s} e^{-g(h-h_0)/R'T_s} \quad (15)$$

and hence, $\log (n-1) = k_1 - k_2 h$, where k_1 and k_2 are constants. Thus by plotting the stratosphere data of $\log (n - 1)$ versus h for the location concerned (Fig. 7) a straight line is obtained which allows the extrapolation of refractive index to great heights.

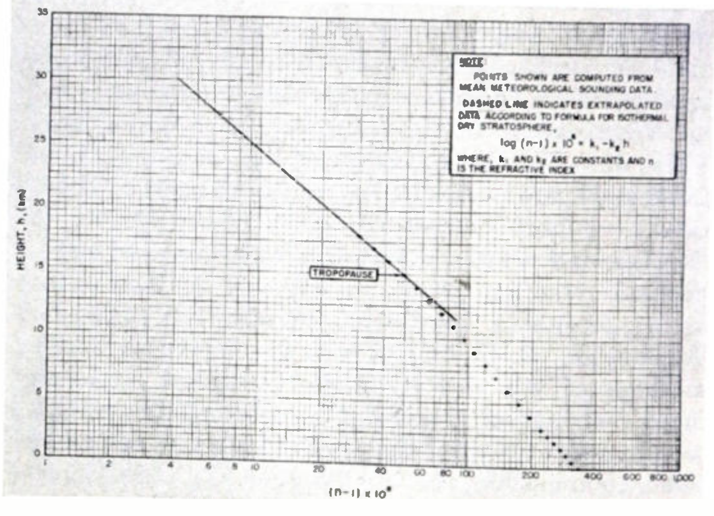

Fig. 7—Refractive index distribution in the troposphere and stratosphere (Washington, D. C., October).

There are limitations on the accuracy of both the refractive-index formula (13) and the meteorological sounding data. These limitations will now be discussed, and it will be shown that there is an uncertainty in the refractive-index calculations of the order of 7 per cent,

[20] C. N. Warfield, "Tentative tables for the properties of the upper atmosphere," *NACA Technical Note* 1200; January, 1947.

about half being due to the uncertainty in the formula coefficients, and half due to the errors of meteorological measurement, the humidity measurements being the least reliable.

Many measurements have been taken on the dielectric constant of dry air, the first being made by Boltzmann in 1872. For sixteen determinations[21] the range of the dry air coefficient was from 72.8 to 80.8, with a mean of 77.9 and a computed probable error of 4 per cent of the mean. The coefficient in (13a) is one per cent larger than the mean value so that 5 per cent has been taken as a reasonable value for the uncertainty in the dry-air contribution. The value of the water vapor refractive index actually used in obtaining (13) is

$$(n_\omega - 1) \times 10^6 = \frac{79e}{T}\left(1 + \frac{4800}{T}\right) \qquad (16)$$

instead of the mean value of the two best determinations over a range of temperatures due to Sanger and Stranathan[17]

$$(n_\omega - 1) \times 10^6 = \frac{79e}{T}\left(0.89 + \frac{4750}{T}\right). \qquad (17)$$

Even at a temperature of 50°C, the values of $(n_\omega - 1) \times 10^6$ obtained from (16) do not differ from those calculated from (17) by more than 2.5 per cent. Thus considering the size of the contribution of each component to the surface value of refractive index at a typical station, such as Washington, D. C., in October, a resultant uncertainty of 4 per cent is obtained.

The uncertainty in the meteorological data to be used in (13) must now be considered. The following error data, used in discussing the dry-air refractive index, was computed from data in the U. S. Weather Bureau memoranda, dated May 15, 1943 and May 19, 1943, and is

TABLE II

PROBABLE ERROR IN COMPUTED PRESSURE AND TEMPERATURE AT FIXED LEVELS FOR A SINGLE SOUNDING FOR U. S. WEATHER STATIONS

Fixed altitude	Pressure mb	Temperature °C.	% error in $(n_d - 1) \times 10^6$
10,000 ft	701.1 ±0.1%	−1.4 ±0.3%	±0.4
20,000 ft	471.8 ±0.28%	−20.6 ±0.48%	±0.8
10 km	270.5 ±0.59%	−44.5 ±0.66%	±1.3
13 km	171.0 ±0.82%	−55 ±0.69%	±1.5
16 km	107.0 ±0.93%	−55 ±0.69%	±1.6

based on Gregg's Standard Atmosphere, which represents the average United States data at 40°N latitude. The error below altitudes of 20,000 feet (≅6 km) in $(n-1) \times 10^6$ for moist air, due to the error in the dry-air refractive index, is then less than about 0.8 per cent. At higher elevations than 10 km, there is a negligible water-vapor contribution, and so the tabulated error then represents the total error. Below these elevations, the effect of the water-vapor refractive index contribution must be considered. For this purpose, the error in relative-humidity measurement is taken as ±10 per cent RH at all values of relative humidity. From this estimate, and an assumed error of one per cent in temperature measurement, the error in $(n_\omega - 1) \times 10^6$ is computed to be about 20 per cent for all elevations up to 6 km for a typical station, Washington, D. C., in October. This results in a 2-per cent error in $(n-1) \times 10^6$ for moist air, just due to the error of measurement of water vapor. Thus the total error in the refractive-index computation below 20,000 feet due to errors of meteorological measurement is 3 per cent, 2 per cent being due to water vapor and one per cent due to the dry air term.

The total uncertainty in the refractive-index calculations is then of the order of 7 per cent, about 4 per cent being due to the uncertainty in the coefficients of the formula, and about 3 per cent due to the errors of measurement of meteorological factors.

V. CONCLUSION

Ray-bending computations are made for a range of climatological conditions for rays passing entirely through the atmosphere and arriving or departing tangentially at the earth's surface. The range of this total angular ray bending extends from 11 milliradians (mr) (0.63°) at Fairbanks, Alaska, in April, to 18 mr (1.01°), at San Juan, Puerto Rico, in July, and is about 14 mr (0.80°), at Washington, D. C., in October. About 90 per cent of the ray bending occurs in the lowest 10 km of the atmosphere. These results are compared with the ray bending computed from the four-thirds effective earth's radius approximation. It has been found that the actual air refraction in the lower atmosphere deviates considerably from that which has been assumed when an effective radius of the earth equal to four-thirds of its value is used as an allowance for this refraction. Although the four-thirds earth's radius assumption provides a reasonably good approximation in the average case, these deviations from the true refraction have been shown to be fairly large both as a function of altitude and of geographical location. The results of this paper, however, have not been applied to field-intensity calculations.

ACKNOWLEDGMENT

This paper was prepared under the direction of T. J. Carroll of the Central Radio Propagation Laboratory. The author also wishes to acknowledge the advice of K. A. Norton, and the assistance of D. L. Randall and V. LaBolle of the Central Radio Propagation Laboratory. The author appreciates the comments and suggestions of L. B. Heilprin, which resulted in the present revised form.

[21] L. G. Hector and D. L. Woernley, "Dielectric constants of eight gases," *Phys. Rev.*, vol. 69, pp. 101–105; February 1–15, 1946.

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Anode Strap (of a Magnetron). A metallic connector between selected anode segments of a multicavity magnetron, principally for the purpose of mode separation.

Cyclotron Frequency. The frequency at which an electron traverses an orbit in a steady, uniform magnetic field and zero electric field. It is given by the product of the electronic charge and the magnetic flux density, divided by 2π times the electron mass.

Cyclotron-Frequency Magnetron Oscillations. Those oscillations whose frequency is substantially the cyclotron frequency.

Critical Field (of a Magnetron). The smallest theoretical value of steady magnetic flux density, at a steady anode voltage, that would prevent an electron emitted from the cathode at zero velocity from reaching the anode.

Critical Voltage (of a Magnetron). The highest theoretical value of steady anode voltage, at a given steady magnetic flux density, at which electrons emitted from the cathode at zero velocity would fail to reach the anode.

Cutoff Field (of a Magnetron) See **Critical Field.**

Cutoff Voltage (of a Magnetron). See **Critical Voltage.**

End Shield (of a Magnetron). A shield for the purpose of confining the space charge to the interaction space.

Interdigital Magnetron. A magnetron having axial anode segments around the cathode, alternate segments being connected together at one end, remaining segments connected together at the opposite end.

Magnetron. An electron tube characterized by the interaction of electrons with the electric field of a circuit element in crossed steady electric and magnetic fields to produce ac power output.

Multicavity Magnetron. A magnetron in which the circuit includes a plurality of cavities.

Multisegment Magnetron. A magnetron with an anode divided into more than two segments, usually by slots parallel to its axis.

Packaged Magnetron. An integral structure comprising a magnetron, its magnetic circuit and output matching device.

π-Mode (of a Magnetron). The mode of operation for which the phases of the fields of successive anode openings facing the interaction space differ by π radians.

Rising-Sun Magnetron. A multicavity magnetron in which resonators of two different resonance frequencies are arranged alternately for the purpose of mode separation.

Split-Anode Magnetron. A magnetron with an anode divided into two segments, usually by slots parallel to its axis.

Traveling-Wave Magnetron Oscillations. Oscillations sustained by the interaction between the space-charge cloud of a magnetron and a traveling electromagnetic field whose phase velocity is approximately the same as the mean velocity of the cloud.

Operational Analysis of Variable-Delay Systems*

L. A. ZADEH†, MEMBER, IRE

Summary—**An operational method of analysis of variable-delay systems such as delay modulators and variable-path communication systems is developed in this article. The output of a variable-delay system is related to its input by a delay operator which has the usual exponential form, but differs from the conventional (time-invariant) delay operators in that the time delay is a function of time. Special consideration is given to systems in which the variation in time delay is due to motion of the receiver or transmitter (source) or both. Such systems—referred to as type-*R*, type-*S* and type-*RS* systems, respectively—are analyzed in general terms. An operational relation is obtained for the correlation function of the output of a type-*R* system, and the result is applied to the determination of the correlation function of a frequency-modulated sound wave.**

Introduction

Various forms of variable-delay systems are frequently encountered by the communication engineer although they are not always recognized as such. A phase modulator, for example, is essentially a type of variable-delay system. A very common form is a communication system in which, as a result of motion of the transmitter or receiver or both, the length of the propagation path is a time-varying quantity. A system of the same type but of a different physical form is a record player in which the angular speed of the turntable is not constant. A different type of system is a vacuum tube in which the electron transit time is a varying quantity. A coder which changes the order of symbols in a sequence is another form of variable-delay system.

A central problem in the analysis of variable-delay systems is that of establishing a relation between the input and output of a specified system. The method of analysis described in this paper is essentially an application of the frequency-analysis technique.[1] More specifically, it is based on the use of delay operators[2] of the form $\exp[\alpha(t)p]$, where $p=d/dt$, and $\alpha(t)$ represents a variable time-delay. The method covers a wide range of special cases and is particularly useful in systems involving a propagation path of varying length.

As a preliminary to the consideration of variable-delay systems, it will be helpful to review briefly the pertinent aspects of the frequency analysis of time-variant systems. This is done in the following section.

Frequency Analysis

A linear time-variant system N may be conveniently characterized by its system function (frequency-response function) $H(j\omega; t)$, which is defined by the relation

$$H(j\omega; t) = \left.\frac{v(t)}{u(t)}\right|_{u(t)=e^{j\omega t}}, \tag{1}$$

where $u(t)$ denotes the input to N and $v(t)$ is the response of N (at rest) to $u(t)$. In other words, $H(j\omega; t)\, e^{j\omega t}$ is the response of N to an input of the form $e^{j\omega t}$.

By superposition, the response of N to an arbitrary input $u(t)$ may be written as

$$v(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} H(j\omega; t) V(j\omega)e^{j\omega t}d\omega, \tag{2}$$

where $U(j\omega)$ is the Fourier transform of $u(t)$. This relation may be written more compactly in an operational form

$$v(t) = H(p; t)u(t), \tag{3}$$

with the understanding that (3) is simply an abbreviated form of (2). When written as $H(p; t)$ the system function constitutes a time-dependent Heaviside operator,[3] and is called the *system operator*.

It should be noted that since the variable t in (2) plays the role of a parameter, it should be treated as such in $H(p; t)$. This means that, in operating on $u(t)$, $H(p; t)$ can be treated as if it were an ordinary Heaviside operator involving t as a parameter.

The *product* of two operators $H_2(p; t)$ and $H_1(p; t)$ is defined as an operator $H_3(p; t)$ such that the result of operating with $H_2(p; t)$ on $H_1(p; t)\, u(t)$ is identical with that of operating with $H_3(p; t)$ on $u(t)$. It can readily be shown that $H_3(p; t)$ is given by the operational relation

$$H_3(s; t) = H_2(p + s; t)H_1(s; t), \tag{4}$$

where s (complex frequency) plays the role of a parameter and $H_1(s; t)$ is the operand. For convenience, (4) may be written in a symbolic form

$$H_3(p; t) = H_2(p; t) * H_1(p; t) \tag{5}$$

or more simply

$$H_3 = H_2 * H_1, \tag{6}$$

where the symbol $*$ has the usual properties of the algebraic product except for commutativity.

The *inverse* of an operator $H(p; t)$ is denoted by $H^{-1}(p; t)$ or H^{-1}, and is defined by

$$H * H^{-1} = H^{-1} * H = 1. \tag{7}$$

H^{-1} is essentially the system operator of the inverse of N, that is, of a system N^{-1} such that a tandem combina-

* Decimal classification: R140. Original manuscript received by the Institute, June 1, 1951.

† Department of Electrical Engineering, Columbia University, New York, N. Y.

[1] L. A. Zadeh, "Frequency analysis of variable networks," Proc. I.R.E., vol. 38, pp. 291–299; March, 1950.

[2] L. A. Zadeh, "Correlation functions and spectra of phase- and delay-modulated signals," Proc. I.R.E., vol. 39, pp. 425–428; April, 1951.

[3] L. A. Zadeh, "Time-dependent Heaviside operators," *Jour. Math. Phys.*, vol. 30, pp. 73–78; July, 1951.

tion of N and N^{-1} is equivalent to a system whose output is identical with the input.

Delay Operators

A *delay operator* is a time-dependent Heaviside operator of the form

$$H(p; t) = e^{-\alpha(t)p}, \tag{8}$$

where $\alpha(t)$ is a function of time. The result of operating with $e^{-\alpha(t)p}$ on a signal $u(t)$ is

$$v(t) = e^{-\alpha(t)p}u(t) = u[t - \alpha(t)]. \tag{9}$$

This transformation may be regarded as a change of the time scale t in accordance with the relation $t' = t' - \alpha(t')$.

From (9) it follows that the result of operating with $e^{-\alpha(t)p}$ on a unit impulse $\delta(t - t_0)$ is

$$v(t) = \delta[t - t_0 - \alpha(t)]. \tag{10}$$

This represents an impulse (but not a *unit* impulse) occurring at the instant $t = t_1$, where t_1 is the solution of

$$t_1 - t_0 - \alpha(t_1) = 0. \tag{11}$$

The "amplitude" of this impulse is the reciprocal of $1 - \dot{\alpha}(t_1)$, where the dot indicates differentiation with respect to time. A graphical construction for finding the time delay $t_1 - t_0$ is illustrated in Fig. 1.

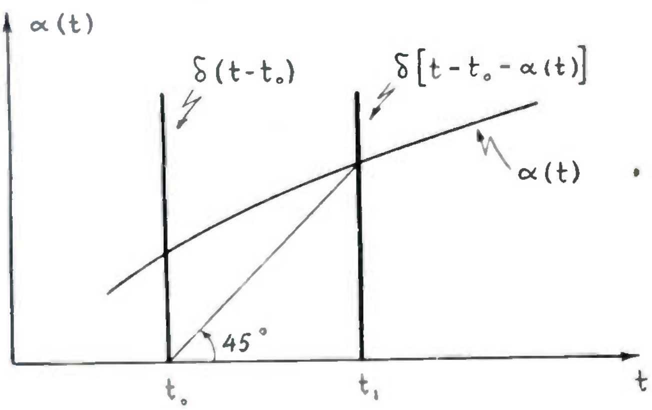

Fig. 1—Graphical determination of the impulsive response.

The product of two delay operators $e^{-\beta(t)p}$ and $e^{-\alpha(t)p}$ is another delay operator $e^{-\gamma(t)p}$,

$$e^{-\gamma(t)p} = e^{-\beta(t)p} \cdot e^{-\alpha(t)p}. \tag{12}$$

The expression for $\gamma(t)$ in terms of $\alpha(t)$ and $\beta(t)$ can readily be obtained by applying (4)

$$e^{-\gamma(t)s} = e^{-\beta(t)(p+s)}e^{-\alpha(t)s}$$
$$= e^{-\beta(t)s - \alpha[t-\beta(t)]s}. \tag{13}$$

Since (13) should hold for all s, it follows that

$$\gamma(t) = \beta(t) + \alpha[t - \beta(t)], \tag{14}$$

which is the desired expression. A graphical procedure leading to $\gamma(t)$ is shown in Fig. 2.

The inverse of a delay operator $e^{-\alpha(t)p}$ is likewise a delay operator. Writing it as $e^{\beta(t)p}$ and making use of (14) [by setting $\gamma(t) = 0$], one obtains the following relation between $\alpha(t)$ and $\beta(t)$:

$$\alpha(t) = \beta[t - \alpha(t)]. \tag{15}$$

An explicit expression for $\beta(t)$ may be formulated by setting $t' = t - \alpha(t)$ and solving for t in terms of t'. This yields

$$\beta(t') = \alpha(t). \tag{16}$$

It will be noted that in order that the inverse exist, it is necessary that t be a single-valued function of t'. This in turn requires that $t - \alpha(t)$ be nondecreasing, i.e.,

$$\frac{d\alpha(t)}{dt} \leq 1. \tag{17}$$

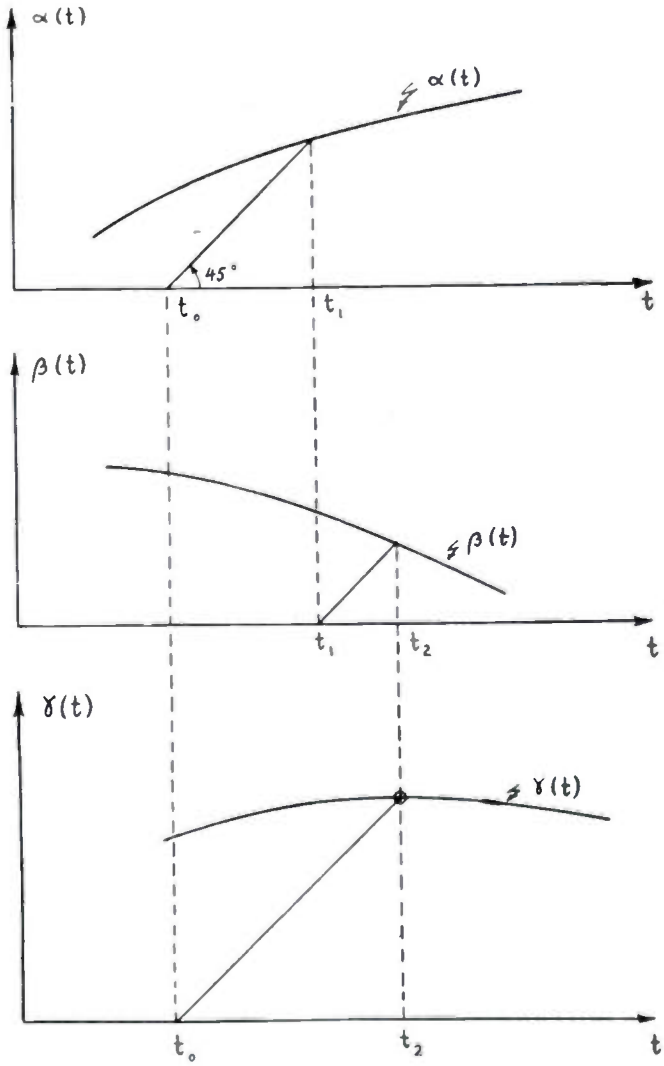

Fig. 2—Graphical determination of the product of two delay operators.

To summarize, the inverse of a delay operator $e^{-\alpha(t)p}$ in which $\alpha(t)$ satisfies (17) is a delay operator $e^{\beta(t)p}$ in which $\beta(t)$ is given by (16). [The term "delay" used in connection with $e^{\beta(t)p}$ should not be taken literally since $\beta(t)$ might be a positive quantity.] A graphical construction for determining the inverse of a delay operator is illustrated in Fig. 3. Referring to this figure it will be noted that in order that A be unique, the slope of the $\alpha(t)$ curve should be less than or equal to unity. This verifies the condition expressed by (17).

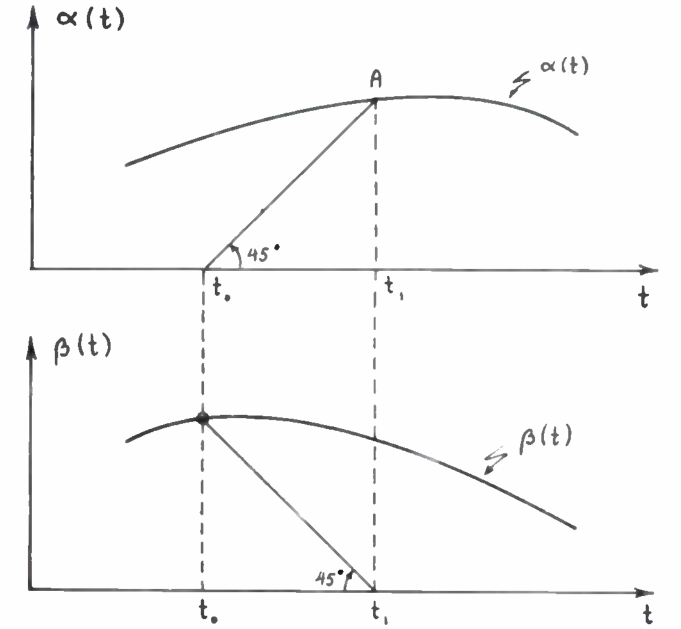

Fig. 3—Graphical determination of the inverse of a delay operator.

Variable-Delay Systems

The differences between various types of variable-delay systems lie chiefly in the causes of variation in the time delay. The most common type of system is one in which the variation in time delay is due to a variation in the distance between the source (transmitter) and receiver. In what follows, the analysis will be restricted to systems of this type, although the same or similar techniques may be applied as well to other types of variable-delay systems.

Consider first a variable-delay system in which the variation in time delay is brought about by a variation in the position of the receiver. Such a system may be represented schematically as in Fig. 4 and may be thought of as a transmission line in which the input end S (source) is fixed, while the position of the output end R (receiver) varies with time. The origin O is chosen to coincide with S. The distance OR is denoted by $\rho(t)$.

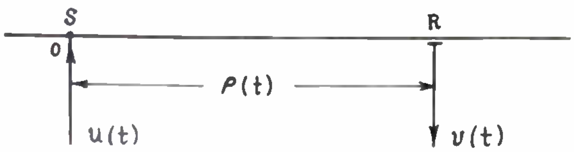

Fig. 4—Schematic representation of a type-R system.

The input to the system is applied at S and is denoted by $u(t)$. The output is obtained at R and is denoted by $v(t)$. The line is assumed to be lossless and, for convenience, the velocity of propagation along the line is assumed to be equal to unity.

A system of the type described above will be referred to as a type-R system. Referring to Fig. 4 it is seen that the output at the instant t is equal to the input at the instant $t-\rho(t)$. Thus

$$v(t) = u[t - \rho(t)] \qquad (18)$$

or equivalently

$$v(t) = e^{-\rho(t)p}u(t). \qquad (19)$$

From this it follows that the system operator of a type-R system is given by

$$H_R = e^{-\rho(t)p},$$

where $\rho(t)$ is, in effect, the normalized distance [ρ = distance/velocity] between the source and receiver.

Next consider a variable-delay system in which the variation in time delay is the result of a variation in the position of the source. A system of this type will be referred to as a type-S system; it may be represented as in Fig. 5. In this case, the origin O is chosen to coincide

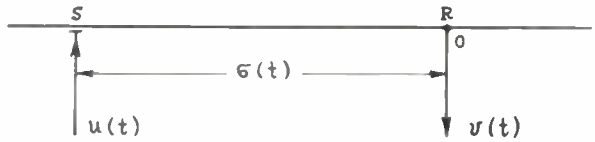

Fig. 5—Schematic representation of a type-S system.

with the receiver R; the distance SO is denoted by $\sigma(t)$; and, as before, the velocity of propagation is taken to be equal to unity.

By using the same reasoning as in the case of a type-R system, the relation between $u(t)$ and $v(t)$ is found to be

$$v[t + \sigma(t)] = u(t), \qquad (21)$$

which in operational form reads

$$e^{\sigma(t)p}v(t) = u(t), \qquad (22)$$

or equivalently

$$v(t) = [e^{\sigma(t)p}]^{-1}u(t), \qquad (23)$$

where $[e^{\sigma(t)p}]^{-1}$ is the inverse of $e^{\sigma(t)p}$. Consequently, the system operator of a type-S system is given by

$$H_S = [e^{\sigma(t)p}]^{-1}, \qquad (24)$$

where $\sigma(t)$ is, in effect, the normalized distance [σ = distance/velocity] between the source and receiver.

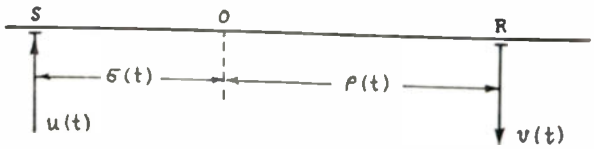

Fig. 6—Schematic representation of a type-RS system.

Frequently, a variable-delay system which is not of R or S type may be regarded as a combination of two or more systems of this type. An example of such a system—one in which both R and S are variable—is illustrated in Fig. 6. Let O be a fixed reference point and let $OR=\rho(t)$ and $SO=\sigma(t)$. It is evident that the system in question may be regarded as a tandem combination of a type-S system (SO) and a type-R system (OR). Consequently, the system operator H_{RS} of the composite system is the operational product of $H_R = e^{-\rho(t)p}$ and $H_S = [e^{\sigma(t)p}]^{-1}$,

$$H_{RS} = e^{-\rho(t)p} * [e^{\sigma(t)p}]^{-1}. \tag{25}$$

This operator may be expressed as a single delay-operator of the form $e^{-\gamma(t)p}$ by using the procedures described in Delay Operators.

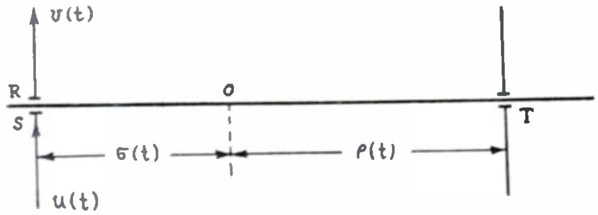

Fig. 7—A variable-delay system involving a reflecting target.

A somewhat more complex variable-delay system is illustrated in Fig. 7. In this case S represents both the source and receiver while T represents a moving reflecting target. Neglecting attenuation, the system in question may be regarded as a tandem combination of two variable-delay systems of the type considered in the preceding example. Thus, the over-all system operator is given by

$$H = H_{RT} * H_{TS}, \tag{26}$$

where H_{RT} and H_{TS} are of the same form as H_{RS} [(25)]. Consequently, the system operator of the variable-delay system under consideration reads

$$H = e^{-\sigma(t)p} * [e^{\rho(t)p}]^{-1} * e^{-\rho(t)p} * [e^{\sigma(t)p}]^{-1}. \tag{27}$$

For specified $\rho(t)$ and $\sigma(t)$, this expression may be reduced to a single delay operator of the form $e^{-\gamma(t)p}$ through the application of the procedures described in Delay Operators.

A basic problem which frequently arises in connection with variable-delay systems is that of determining the spectral density—or, equivalently, the correlation function[4]—of the response of the system to a random input. As is shown in footnote reference 2, if $v(t)$ and $u(t)$ are related by a delay operator $e^{-\alpha(t)p}$,

$$v(t) = e^{-\alpha(t)p}u(t); \tag{28}$$

and if $\alpha(t)$ and $u(t)$ are statistically independent, then the correlation function of $v(t)$, $\psi_v(\tau)$ is related to that of $u(t)$, $\psi_u(\tau)$ by the operational relation

$$\psi_v(\tau) = \psi_H(p;\tau)\psi_u(\tau); \tag{29}$$

$\psi_u(\tau)$ plays the role of the operand and $\psi_H(p;\tau)$, the correlation function of $H = e^{-\alpha(t)p}$, is the time-average of the quantity $\exp\{[\alpha(t+\tau) - \alpha(t)]p\}$,

$$\psi_H(p;\tau) = \lim_{T\to\infty} \frac{1}{2T}\int_{-T}^{T} \exp\{[\alpha(t+\tau) - \alpha(t)]p\}\,dt. \tag{30}$$

[4] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

Once $\psi_H(p;\tau)$ has been determined, the evaluation of $\psi_b(\tau)$ is reduced to the straightforward process of operating with $\psi_H(p;\tau)$ on $\psi_u(\tau)$ in accordance with (29) and the defining (2).

A practical application for these results is furnished by the case of a record player in which the angular velocity of the turntable oscillates around a constant value—resulting in the frequency-modulation of the recorded sound. Considering only the effect of variable angular velocity, the system in question may be regarded as a variable-delay system of type R, whose input and output are the recorded and reproduced sounds, respectively. The system operator is of the form $e^{-\alpha(t)p}$, with $\alpha(t)$ of the form

$$\alpha(t) = \alpha_0 \sin \omega_0 t, \tag{31}$$

where α_0 and ω_0 are constants.

In this case, the expression for $\psi_H(p;\tau)$ becomes

$$\psi_H(p;\tau) = \frac{1}{2\pi}\int_0^{2\pi} \exp\{[\sin \omega_0(t+\tau) - \sin \omega_0 t]\alpha_0 p\}\,d(\omega_0 t). \tag{32}$$

The integral involved in (32) is readily evaluated, yielding the following expression for $\psi_H(p;\tau)$:

$$\psi_H(p;\tau) = I_0\left(2\alpha_0 p \sin\frac{\omega_0\tau}{2}\right), \tag{33}$$

where I_0 is a modified Bessel function of the first kind and zero order.

In consequence of (29), the correlation function of the reproduced sound is given by the operational relation

$$\psi_v(\tau) = I_0\left(2\alpha_0 p \sin\frac{\omega_0\tau}{2}\right)\psi_u(\tau), \tag{34}$$

where $\psi_u(\tau)$ is the correlation function of the recorded sound.

In many practical cases $\psi_u(\tau)$ is either of the form

$$\psi_u(\tau) = e^{-a|\tau|}\cos b\tau, \tag{35}$$

where a and b are constants, or is a sum of terms of this form. It can readily be verified that, when $\psi_u(\tau)$ is expressed by (35), the expression for $\psi_v(\tau)$ becomes

$$\psi_v(\tau) = \frac{1}{\pi}\int_{-\tau-2\alpha_0\sin\omega_0\tau/2}^{-\tau+2\alpha_0\sin\omega_0\tau/2} [4\alpha_0^2\sin^2(\omega_0\tau/2) - (\tau+x)^2]^{-1/2} e^{-a|x|}\cos bx\,dx,\quad \tau > 0, \tag{36}$$

where x is the variable of integration. This integral can be expressed in the form of an infinite series, but for practical purposes it is more conveniently evaluated by the use of numerical integration.

Concluding Remarks

The operational method described in this paper is based on the use of time-dependent Heaviside operators of the form $e^{-\alpha(t)p}$ (delay operators), and involves the operational multiplication and inversion of such operators. The method yields the system operator $H(p; t)$, generally in the form of a product of two or more delay operators each of which represents the system operator of a part of the over-all system. Once $H(p; t)$ has been determined, the response of the system to a specified input $u(t)$ may be found from the operational relation $v(t) = H(p; t)\, u(t)$, in which $H(p; t)$ should be treated as if it were an ordinary Heaviside operator involving t as a parameter.

An important class of variable-delay systems which is not considered explicitly in this paper is one in which the variation in time delay is due to time-varying velocity of propagation through the medium between the source and receiver. Although the system operator of a system of this type has the usual form $H(p; t) = e^{-\alpha(t)p}$, the determination of $\alpha(t)$ is, in general, more difficult than in the case of systems in which the velocity of propagation is constant.

A Sampling Analogue Computer*

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Summary—The sampling analogue computer described may be used for various algebraic operations. Using standard electronic components, and with electrical inputs, only fifteen tubes are required to produce an output accuracy better than one per cent of full scale.

I. Introduction

THE SAMPLING ANALOGUE COMPUTER is an all-electronic device capable of performing various functions, such as multiplying, dividing, adding, subtracting, raising to various powers, and extracting roots. The computer consists of two main parts, one called the "algebraic unit" and the other the "pulse converter." Using all standard electronic components and a 15-tube complement, the average accuracy of the computer is better than one per cent of full scale.

The algebraic unit receives the inputs for multiplying, dividing, powers, and roots; it performs the indicated operation, and produces a narrow pulse the amplitude of which is equal to the computed result. The basic idea of this unit was developed by Hirsch of the Hazeltine Corporation, and by Felker of Bell Telephone Laboratories, and has been described in their reports.[1]

The output of the algebraic unit forms the input to the pulse converter, the main function of which is to convert the pulse from a high-impedance source into a steady voltage at a low-impedance level. Two collateral functions of the pulse converter are addition-subtraction, and multiplication by a constant.

Where much input data must be handled, computer economy may be achieved by multiplexing the input signals. The computer settling time is determined by the pulse converter, and this has been designed to accept the abrupt changes that come from the sudden switching from one set of inputs to another.

II. Theory of Operation

A. Algebraic Unit

Referring to Fig. 1, X, Y, and Z can be considered to be constant (in general, the sampling rate is rapid enough so that X, Y, and Z change very little over a few sampling periods). The Z and X switches are closed for only the very short time (compared to the sampling period) necessary to charge the time-constant capacitor through the switch resistance. In practice, this can be made less than 5 μ sec.

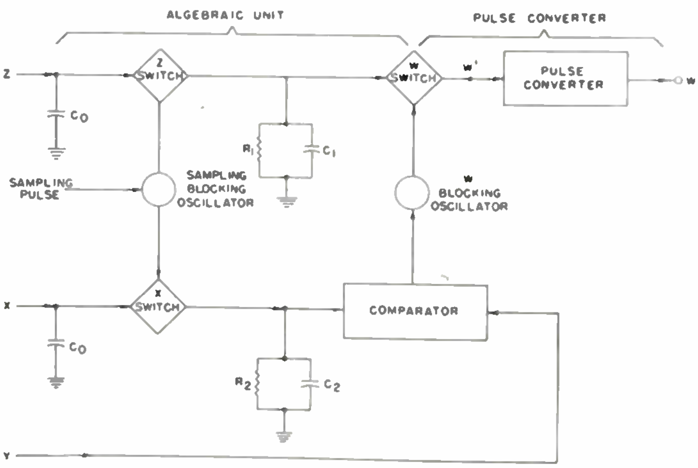

Fig. 1—Block diagram of sampling analogue computer.

The X and Z switches open simultaneously and the voltages across C_1 and C_2 decay exponentially. When the voltage across C_2 is equal to the Y voltage, the comparator produces a trigger pulse to trigger the W blocking oscillator, which briefly closes the W switch and produces an output answer pulse by sampling the voltage across C_1 at that instant.

Mathematically,

$$W' = Z\epsilon^{-t/R_1C_1} \quad \text{and} \quad Y = X\epsilon^{-t/nR_1C_1}, \qquad (1)$$

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[1] C. J. Hirsch, Hazeltine Report #2068W; January 29, 1948. J. H. Felker, Bell Lab. Report Case #25653-3; April 22, 1948.

where

$$n = R_2C_2/R_1C_1.$$

In practice, there is a time delay between the comparator output pulse and the closing of the W switch. Thus the voltage in the Z time constant discharges to a slightly lower value than the Y voltage. Rewriting the W' equation and solving it,

$$W' = Z\epsilon^{-(t+\tau)/R_1C_1} = Z\left(\frac{Y}{X}\right)^n \epsilon^{-\tau/R_1C_1} = KZ\left(\frac{Y}{X}\right)^n, \qquad (2)$$

where K is less than one.

For the application to be considered later, $n=1$, and

$$W' = KYZ/X. \qquad (3)$$

The W' pulse is then fed into the pulse converter, and the output direct voltage is

$$W = GKZ\frac{Y}{X}, \qquad (4)$$

where G is the voltage gain of the pulse converter.

It is interesting to note that by feeding the output of the pulse converter into the X switch, the sampling computer computes the square root of the product of the quantities. For this case, $W = C\sqrt{ZY}$, where $C = \sqrt{GK}$.

B. Pulse Converter

A block diagram of the pulse converter is shown in Fig. 2.

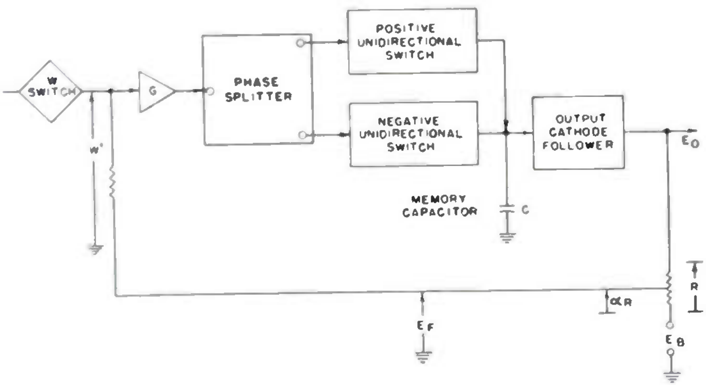

Fig. 2—Block diagram of pulse converter.

The output voltage E_0 is compared at the W switch with the correct answer. If there is a difference, an error pulse is produced, the polarity of which represents the sign of the error. This error pulse is amplified and then, depending on the polarity, either increases or decreases the voltage across the memory capacitor. The converter as shown in Fig. 2, with $E_B = 0$, gives an output voltage very nearly equal in amplitude to the correct answer.

By connecting the output resistor R to a bias voltage, E_B, a constant is either added or subtracted from the output voltage. Thus, addition or subtraction can be performed.

A constant gain can be realized from the pulse converter if only a fraction α of the output voltage is fed back. Thus, multiplication of the answer by a constant can be performed.

The following analysis illustrates the above discussion. Assuming no error in the pulse converter, the action of the circuit is to cause the feedback voltage, E_F, to equal the correct answer, W'.

$$W' = E_F - E_0 - (1 - \alpha)E_B, \qquad (5)$$

or

$$E_0 = GW' - (G - 1)E_B \text{ where } G = 1/\alpha. \qquad (6)$$

The output voltage consists of the answer voltage W' multiplied by a gain G, with a quantity $(G-1)E_B$ subtracted. Both the multiplication and the subtraction characteristics are used in the applications discussed later.

The maximum value of the gain G is limited by rf leakage through the W switch, contact and thermal potential in the W switch, and the finite time constant associated with the memory capacitor.

III. Discussion of a Particular Computer and its Practical Limitations

The particular computer to be discussed (see Fig. 3, following page) is designed to sample at a 400 cps rate. The algebraic unit and pulse converter are discussed separately.

A. Algebraic Unit

Blocking Oscillator: Five-μsec selector pulses to close the switches are supplied by blocking oscillators.

Four-Diode Switch: The particular form of four-diode switch used in this application is shown in Fig. 3. The gate is self-biased because of the time constant R_gC_g. Operation is as follows:

The selector pulse is supplied by the blocking oscillator. For most of the duration of the flat-topped 5-μsec pulse, the switch is closed. During this time, the capacitor C_g is charged, through the diode and reflected blocking oscillator source impedance, to a value almost equal to the peak value of the pulse. The value of bias capacitor must be chosen small enough so that the bias voltage equals approximately the pulse amplitude, and yet large enough so that the switch remains closed throughout the selector-pulse duration. The optimum value of C_g depends on so many factors, including pulse shape, that for any given application it is best found by trial and error.

In the time interval between selector pulses, it is necessary for the capacitor C_g to discharge slightly in order to permit the next selector pulse to close the switch. This slight discharge is obtained by shunting C_g with an appropriate bias resistor, R_g.

The switch remains open in the interval between selector pulses because of the self-bias maintained across the bias time constant R_gC_g if the signal input level is kept below the maximum self-bias voltage. The value of the self-bias resistor, R_g, is chosen such that the bias

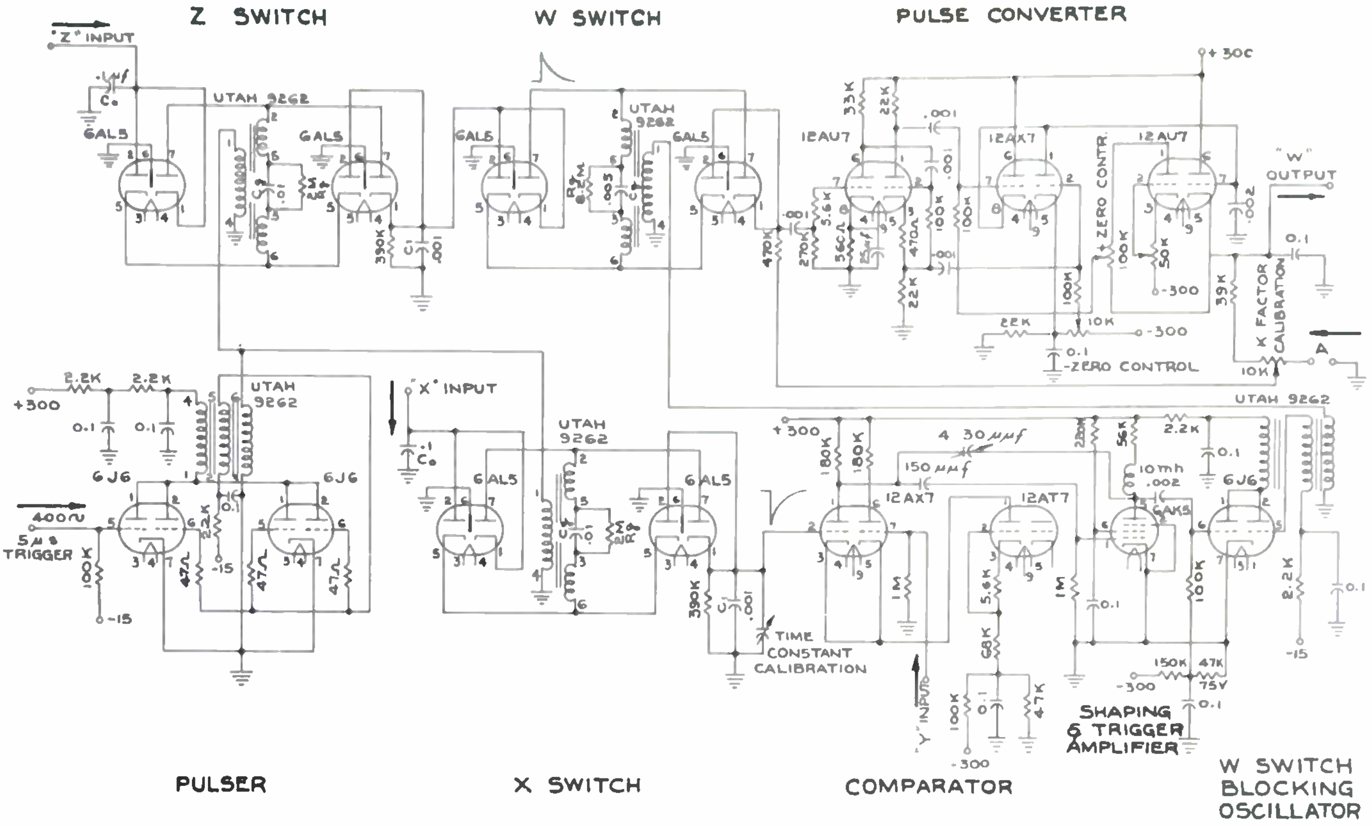

Fig. 3—Circuit diagram of a sampling analogue computer.

time constant is long compared to the sampling period. A time constant longer than six sampling periods is satisfactory. If the time constant is made too long, the switch will close only at the tip of the pulse. A value for this time constant is best found by trial and error.

The switch resistance is of the order of 200 ohms using 6AL5 diodes. The diodes used are carefully chosen to have equal residual potential (contact plus thermal potential).

Comparator and Shaping Circuits: A double-triode discriminator with constant-current cathode load is used as a comparator. Depending on the polarity desired, output can be obtained from either plate in the form of a pulse. The comparator output pulse is shaped and amplified, and then used to trigger the *W* blocking oscillator. The comparator output pulse is first differentiated, and the positive overshoot clipped by grid conduction. The differentiated pulse is strongly amplified and clipped again. The resultant waveform has a steep leading edge, and provides an accurate trigger pulse for the *W* blocking oscillator.

B. Pulse Converter

In the discussion to follow, the *Z* input voltages are positive and are fed from the *Z* time constant, through the output *W* switch, directly into the pulse converter. The converter is designed to accept positive inputs from 0 to 150 volts, which is greater than the range of any *Z* input. The pulse converter performs the following duties: (a) Converts the answer from a 5-μsec voltage pulse into a steady voltage equal to the peak value of the pulse. (b) Converts the answer from a high-impedance level to a low-impedance level. This permits one computer to feed another computer, or meter, without appreciable error. (c) A very accurate and stable variable gain can be obtained. This permits convenient scaling-factor adjustment. (d) Addition and subtraction can be performed conveniently by applying a bias, E_B, to the output resistor, as shown in Fig. 2.

For purposes of analysis and discussion the pulse converter can be separated as follows:

Amplifier and Phase Splitter: As mentioned earlier in this paper, an error pulse is obtained because of the difference between the *W* switch output and the feedback voltage. The polarity of the error pulse indicates the sign of the error between the output voltage, E_0, and the correct answer, *W*.

The error pulse is amplified in a single-stage pulse amplifier, and then fed into a conventional phase splitter. The two outputs of the phase splitter are necessary to drive the two unidirectional switches.

Because of over-all feedback considerations, it is desirable to design the amplifier and phase splitter to have a band-pass characteristic just wide enough to pass the 5-μ sec pulse properly.

Unidirectional Switches and Output Cathode Follower: The outputs from the phase-splitter are fed into two unidirectional switches. These unidirectional switches pass only positive pulses. If the polarities are traced from the input of the pulse amplifier, it is seen that a positive-error pulse will close the switch connected to the plate circuit of the phase splitter for the duration of the pulse. This switch is called the positive unidirectional switch.

The memory capacitor is given a positive increment of voltage when the positive unidirectional switch closes, and a negative increment when the negative unidirectional switch closes. In between operation of the switches, the capacitor voltage remains constant (hence the name, "memory capacitor").

The output cathode follower is used to transfer the voltage across the memory capacitor to the output load without loading the memory circuit. The incremental output impedance, including the effect of feedback, is of the order of 1 to 2 ohms over its working range.

The constant current tube in the cathode circuit of the output cathode follower is used to provide bias for the positive unidirectional switch. With the bias provided in this fashion, it can be shown that the grid-cathode voltage on each switch tube is roughly independent of the voltage level across the memory capacitor. This permits balancing the leakage current through the two switches over a wide range of output voltages. The grid-to-cathode bias on the switch tubes is adjusted to approximately cutoff when no error signal is applied. By slight adjustment, the leakage current through each tube can be made equal to that in the other, and will not charge or discharge the memory capacitor in between error signals.

IV. Discussion of Requirements Due to Time Sharing

A. Effective Input Impedance of Z and X Switch

The effective input impedance of the Z and X switches is the same if the value of resistance and capacitance loading the ouput of each switch is the same. This is true for this application.

By choosing the value of C_0 (see Fig. 3) roughly 500 times the value of C_1, the value of the input Z drops one-half per cent during the short time that the switch is closed (assuming that the source time constant is long compared to the time that the switch is closed and, hence, cannot regulate the voltage Z). If we further require that the time constant formed by the resistance of the Z voltage source and input capacitor, C_0, be less than or equal to one-half the sampling rate, T_S, we are assured that the value of the Z input voltage across C_0 just prior to the opening of the Z switch is within 0.01 per cent of its value in absence of any loading on C_0. The required resistance, R_S, of the source is given by

$$R_s \leq \frac{T_S}{2C_0}. \tag{7}$$

B. Settling Time of Pulse Converter

From the detailed description of the pulse converter given in Section II, it is clear that the output voltage changes by an increment each time an error pulse occurs. In between error pulses, the output voltage remains sensibly constant because of the memory capacitor and associated circuits. Because the output voltage changes by incremental steps, a number of sampling periods is required to change the output voltage by a given amount. Increasing the sampling rate reduces the settling time required. A simple analysis indicates that the maximum increment, $\Delta V_{max.}$, in the voltage is given by

$$\Delta V_{max} = \frac{I_{max}\,\tau}{C}, \tag{8}$$

where

I_{max} = maximum cathode current in either unidirectional switch tube

τ = width of error pulse

C = value of memory capacitor.

It is desirable to make the voltage increment, ΔV_{max}, as large as possible by maximizing (8). The error pulse width, τ, is determined by computer accuracy consideration and cannot be varied. The minimum size of the memory capacitor is determined by how well the leakage currents in the switch tubes can be balanced over the working range of voltage. Too small a value of capacitor results in an appreciable exponential decay or rise of output voltage between error pulses. An optimum value of capacity is best found experimentally and depends on the amount of variation that can be tolerated in the voltage output. The remaining term in (8) is the maximum value of the cathode current flowing in either undirectional switch. This can be controlled over a wide range of values by proper choice of tubes.

In the present system the settling time of a pulse converter is approximately 30 milliseconds. This settling time can probably be reduced considerably by the methods mentioned above. If some reduction in computer accuracy can be tolerated, the settling time can be reduced proportionately as the sampling rate and/or the error pulse width is increased.

V. Accuracy and Limitations

The errors in the algebraic unit consist of two types: First, computing error—that is, error in each individual computation assuming an infinite sampling rate. Secondly, sampling error—that is, error due to approximating a continuous function by samples of finite width obtained at finite time intervals.

Sources of error are listed below in the order of their importance at the present stage of development of the computer. The use of a pulse converter is assumed.

A. Computing Error

Comparator: A double-triode discriminator[2] with a constant-current cathode load has been found to give satisfactory results when computer accuracies not much better than one per cent are required. For higher accuracies, more complicated comparators are available.

Switches: Switches must be closed very quickly and for only a short period of time. For most applications, the speeds required can be obtained only by using

[2] Waveforms-Radiation Lab. Series, vol. 19, pp. 335–338; McGraw-Hill Book Co., Inc., New York, N. Y.; 1949.

electronic switches. Of the many types available, the four-diode switch[3] was found to give the best results for this application. Because of the high accuracy required of the X and Z time constants, germanium crystals could not be used.

The sources of error using thermionic diodes may be divided as follows:

For low signal input, contact and thermal potentials cause error. By carefully choosing tubes and reducing heater voltage slightly, this can be minimized.

Because of the shunt and stray capacitance across the diode, rf current from the selector pulse can "leak" through the switch unless the switch is carefully balanced. Neutralizing capacitors can be used to offset any capacitance unbalance due to tubes, writing, and the like. The rf leakage is most serious at the output of the W switch because of the high-output rf impedance, and it is desirable to remove any overshoot on the selector pulse.

In general, the rf leakage can be made small, but the contact and thermal potential unbalance places a lower limit on the magnitude of the input signal that can be used with diode switches.

B. Sampling Error

The sampling error, which can usually be made much smaller than the individual computation error, is due chiefly to two sources:

First, sampling the continuously varying input voltages at finite intervals. This error can be minimized by making the sampling rate much larger than the highest frequency component in the input voltages.

Second, finite time necessary to operate the switches properly. Because of the finite switch resistance, it is necessary to close the switch for a definite time to charge the capacitor on the output side of the switch completely. As the input voltage is varying during this time, the output voltage is some average of the input voltage over the interval. Switch resistance and the interval during which the switch is closed should be as short as possible.

In the application to the discussed below, the sampling error is negligible compared to the computational error.

A further error arises in the conversion of the answer pulse (i.e., output pulse of algebraic unit) into a direct voltage. The following factors cause errors in the output of the pulse converter:

(a) The pulse converter requires a finite error pulse for operation. The rf leakage through the W switch produces extraneous error pulses in the pulse converter. A lower limit on the size of the error pulse is caused by these extraneous pulses and the requirement of stability in the feedback loop.

(b) Finite memory—ideally the time constant associated with the memory should be infinite. For reasons discussed under Section IV, it is desirable to keep the memory capacitor as small as possible. In practice, this time constant is very long, but finite. It further varies slightly with voltage. This finite memory results from the unbalance between tubes used as unipolarity switches, which can be adjusted for balance at only one point. Because of finite memory, the converter has a tendency to "hunt" around the correct value.

(c) Settling time—the converter has a finite settling time depending on the sampling rate of the parametric computer.

The computer shown in Fig. 3 has an rms error of 1.0 per cent of full scale, with the pulse converter set for unity gain and the inputs variable as follows:

$$X \rightarrow 5\text{–}90, \quad Y \rightarrow 2\text{–}85, \quad Z \rightarrow 0\text{–}120 \text{ volts.}$$

If the value of Y is fixed at a single value and the cathode bias on the comparator optimized, the rms error is 0.7 per cent of full scale, with the pulse converter set for unity gain.

Table I shows a typical set of readings for a pulse converter with a gain of one. The maximum error is one per cent of full scale, and occurs at the maximum value. With higher gains the percentage error increases. At a gain of 15 the maximum error is of the order of 2 per cent of full scale from 0 to 150 volts. The error increases very rapidly for gains greater than 20.

TABLE I

PULSE CONVERTER RESPONSE (GAIN = 1)

E_{in} (to W switch)	E_{out} (Pulse converter output)
1 volt	0.45 volt
2	1.6
3	2.56
4	3.62
5.0	4.6
10.0	9.75
20.0	20.0
30.0	30.0
40.0	40.25
50.0	50.4
60.0	60.4
70.0	70.7
80.0	80.8
90.0	90.9
100.0	101.0

VI. CONCLUSION

Major sources of error in the over-all computer are due to: pulse converter; rf leakage in four-diode switch, time-constant unbalance, and the like and comparator.

With reasonable effort, the present accuracy of one per cent could be improved by a factor of 5 or 10. Further improvement will probably require considerable effort.

VII. ACKNOWLEDGMENT

The authors wish to thank C. J. Hirsch and J. H. Felker for copies of their original reports and helpful discussions.

The development work described in this paper was done under contract between the University of Pennsylvania and the U. S. Army Signal Corps Engineering Laboratories.

[3] Waveforms-Radiation Lab. Series, vol. 19, pp. 374; McGraw-Hill Book Co., Inc., New York, N. Y.; 1949.

Development of VHF Field-Intensity Standards*

F. M. GREENE†, MEMBER, IRE, AND M. SOLOW†

THIS ABSTRACT briefly describes the development and design of field-intensity standards now in use at the National Bureau of Standards, for the calibration of commercial field-intensity meters in the frequency range 30 to 300 mc. The two basic methods of measuring absolute field intensity have been used. These are: (a) the "standard-antenna" method in which the electric component of the field is evaluated in terms of the induced voltage and the effective length of a horizontal half-wave-length receiving dipole from the relationship[1]

$$|E| = V_{oc}/l_h; \quad (1)$$

and (b) the "standard-field" method in which the field is evaluated in terms of the current I, the effective-length of a horizontal half-wavelength transmitting dipole, the path lengths R_1 and R_2, respectively, of the direct and ground-reflected rays, and Γ, the complex reflection coefficient of the ground for horizontal polarization, the relationship being

$$|E| \simeq \frac{60\pi l_h I}{\lambda}\left|\frac{1}{R_1}+\frac{\Gamma}{R_2}\epsilon^{-jk(R_2-R_1)}\right|, \quad (2)$$

where $k=2\pi/\lambda$, and $\lambda=$ the operating wavelength.

In using the standard-antenna method, the induced voltage was measured directly by means of a relatively high-impedance silicon crystal diode mounted coaxially in the gap at the center of the receiving antenna. By using a crystal having an rf resistance of several thousand ohms, the need for knowing the antenna impedance was eliminated. The filtered dc output of the calibrated crystal diode was measured on a slidewire potentiometer. The resulting sensitivity permitted establishing a field-intensity reference in the range 0.02 to 1.0 volt per meter at 100 mc.

In using the standard-field method the current at the center of the horizontal transmitting dipole was measured by means of a calibrated vacuum thermocouple. A half-wavelength balanced transmission line was terminated at opposite ends in the antenna and thermocouple, and a balanced line from the rf generator was tapped in at the mid-point. Thus the current in the antenna and in the thermocouple was essentially the same, regardless of their respective impedances.

The effective length of the receiving and transmitting dipoles was determined analytically from Schelkunoff's analysis of the cylindrical dipole by integrating the relative current distribution obtained for various length-to-diameter ratios. This was also determined experimentally by evaluating, by means of (1) and (2), the field as a function of height directly beneath the horizontal transmitting dipole mounted at a fixed height, h_1, of 9.27 meters. Results of a typical run ($f=100$ mc) are shown in Fig. 1. The

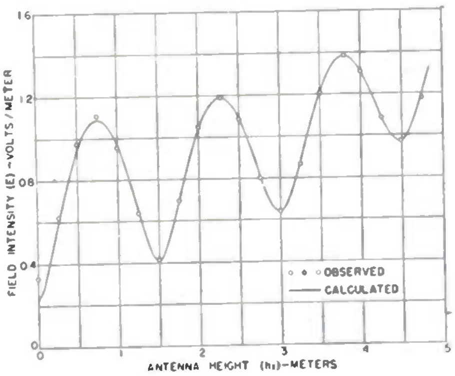

Fig. 1—Vertical-incidence measurements.

reflection coefficient of the ground at the time was accurately determined for vertical incidence by measuring the ratio and position in space of the maxima and minima of the resulting standing wave. Solving (1) and (2) for the effective length then yielded a number of separate evaluations, the average being $l_h=0.96$ meter for the antennas used. The results of the analytical process gave $l_h=0.97$ meter, while the classical value, assuming sinusoidal current distribution, is $l_h=\lambda/\pi=0.955$ meter.

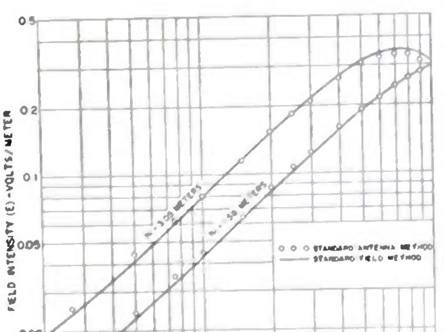

Fig. 2—Variable-height measurements.

The probable accuracy of the two standards was determined by inter-comparing, over a wide range of the variables involved, numerous values of absolute field intensity measured simultaneously by each. One set of results is shown in Fig. 2, where the field intensity was determined as a function of the height of the receiving dipole for each of two heights of the transmitting dipole and a fixed horizontal distance of separation $d=30.5$ meters. Another set of results is shown in Fig. 3, where the field intensity was determined as a function of the horizontal distance of separation between transmitting and receiving dipoles for each of two heights of the transmitting dipole and a fixed height of the receiving dipole, $h_2=3.05$ meters. In both cases the ground constants ϵ_r and σ, as determined from vertical-incidence measurements,[2] were $\epsilon_r \simeq 15$, $\epsilon_r \gg 60\lambda\sigma$, $f=100$ mc.

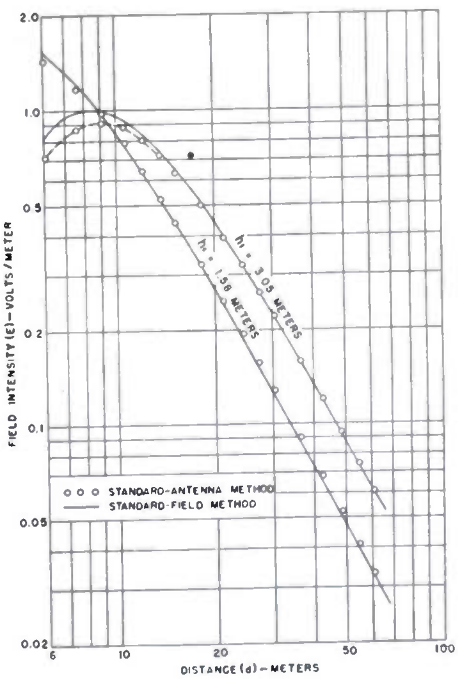

Fig. 3—Variable-distance measurements.

The results seem to indicate that if sufficient precautions are taken with the instrumentation and techniques used, it is possible, under a limited range of conditions, to establish reference values of horizontally-polarized electric field intensity for calibration purposes which will have an accuracy of the order of 5 per cent in the frequency range up to possibly 150 mc and somewhat larger up to 300 mc. However, additional errors would be involved in the use of these standards for the calibration of commercial instruments which would perhaps double the above figure for a certified calibration.

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This is an abstract of the original paper presented at the URSI-IRE Meeting, May 2, 1949, Washington, D. C., and which appeared in the Jour. Res. Nat. Bur. Stand., vol. 44, no. 5, pp. 527–547; May, 1950, RP 2100. Copies for sale by the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. Price 15 cents.
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[1] Rationalized mks units are employed.

[2] Details of the antenna construction, the calibration procedures, the method used to determine the reflection coefficient of the ground, and the results of the various field tests are given in the original paper.

Isotropic Artificial Dielectric*

R. W. CORKUM†

Summary—**Isotropic artificial dielectric media composed of a three-dimensional cubic array of metal or dielectric spheres have been investigated.**

Theoretical expressions using the Clausius-Mossotti relation have been derived for the index of refraction, dielectric constant, and magnetic permeability of this type of dielectric. These quantities are independent of frequency so long as the size of the spheres and the spacing between spheres are small compared to the wavelength within the resulting dielectric media.

Samples using steel and fused-quartz spheres have been fabricated, and the dielectric properties measured in rectangular waveguide at a frequency of 5,000 megacycles per second. Standard waveguide techniques are readily adaptable to this type of dielectric.

Experimentally determined values of the dielectric properties are in good agreement with theoretical values, and the theoretical expressions are assumed to be valid.

I. Introduction

ISOTROPIC artificial dielectric media composed of a three-dimensional cubic array of metal spheres and dielectric spheres have been investigated. Such an investigation was deemed necessary to determine first, proper theoretical expressions for predicting the dielectric properties of this type of artificial dielectric, and secondly, appropriate experimental techniques for measuring these properties.

The theoretical expressions derived in this paper use the Clausius-Mossotti relation. Kock[1] suggested using this relation to derive the dielectric constant of a metal-sphere array. However, its use has been extended to include the magnetic permeability of such an array, as well as the dielectric constant of a dielectric-sphere array. The metal-sphere array had previously been investigated theoretically by Pasternack,[2] using a modification of the Wigner-Seitz-Slater cellular method, and by Lewin,[3] using the Mie solution for scattering an incident plane wave by a conducting sphere.

Measurements on the samples constructed were taken in rectangular waveguide, using standard waveguide techniques, and the major problem involved was that of determining whether these techniques are applicable to this type of artificial dielectric.

The basic theory of this type of artificial dielectric will be given, the apparatus described, and data presented and evaluated.

* Decimal classification: R281. Original manuscript received by the Institute, April 6, 1951; revised manuscript received, October 22, 1951.

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[1] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, vol. 27, pp. 77–80; January, 1948.

[2] S. Pasternack, "Microwave Lenses," University of Pennsylvania, Final Report, AFCRL Contract No. W28-099-ac-112; July, 1947.

[3] L. Lewin, "The electrical constants of a material loaded with spherical particles," *Jour. I.E.E.*, vol. 94, pp. 65–68, pt. 3; January, 1947.

II. Theory and Calculations

A. Dielectric Constant of Metal-Sphere Arrays

If perfectly conducting spheres, arranged in a cubic array, are placed in a uniform static electric field or in an alternating electric field such that the wavelength within the resulting dielectric medium is much larger than the diameter of the spheres, the free charges upon these spheres are displaced by the applied field, and the spheres may be replaced by electric dipoles. The polarizability of the sphere is related to the artificial dielectric coefficient of the medium by the Clausius-Mossotti equation given in Appendix I as follows:

$$\frac{\alpha_e N}{3V} = \frac{K_e - 1}{K_e + 2}, \tag{1}$$

where α_e is the electric polarizability of the sphere, N/V the number of spheres per unit volume, and K_e the relative dielectric coefficient of the medium. Equation (1) may be inverted to yield

$$K_e = \frac{1 + (8/3)\pi a^3 N/V}{1 - (4/3)\pi a^3 N/V}, \tag{2}$$

since $\alpha_e = 4\pi a^3$ for a metal sphere of radius a.

B. Calculation of Permeability of Artificial Dielectric

Similarly, from the Clausius-Mossotti equation, we find for the metal-sphere array

$$K_m = \frac{1 - (4/3)\pi a^3 N/V}{1 + (2/3)\pi a^3 N/V}, \tag{3}$$

since $\alpha_m = -2\pi a^3$, the magnetic polarizability of the sphere, with K_m the relative magnetic permeability of the medium.

In the above formulas we have assumed the spheres to be in free space, or the effect of the structural supporting material to be negligible. If, however, the spheres are imbedded in a material whose permittivity and permeability are k_1 and μ_1, respectively, and the above formulas are expressed in terms of f, the fractional volume filled by the metal spheres, we have

$$K_e = k_1\left(\frac{1 + 2f}{1 - f}\right), \qquad K_m = \mu_1\left(\frac{1 - f}{1 + f/2}\right), \tag{4}$$

where

$$f = \frac{4}{3}\pi a^3 \frac{N}{V} = \frac{4}{3}\pi a^3 \frac{1}{S^3}$$

for the cubic array. (S is the distance between the centers of adjacent spheres.)

C. Dielectric Constant of Dielectric-Sphere Arrays

Similarly, approximate theoretical expressions for the dielectric quantities of a cubic array of dielectric spheres are obtained by substituting the polarizability for dielectric spheres in (1). The resulting dielectric quantities will depend on the material of the sphere as well as sphere spacing and size. If k_e is the relative dielectric constant of the sphere material, then, in terms of the preceding notation, we have the following relation:

$$K_e = k_1\left(\frac{1 + 2fC}{1 - fC}\right), \tag{5}$$

where

$$C = \frac{k_e - 1}{k_e + 2}.$$

If the supporting material is dielectric, the permeability of the resulting medium will be unity. If it is not, $K_m = \mu_1$.

D. Index of Refraction of Metal-Sphere and Dielectric-Sphere Arrays

These arrays of spheres will retard electromagnetic waves just as an ordinary dielectric does. The effective index of refraction is given by

$$n = \sqrt{K_m K_e}. \tag{6}$$

Consequently, the index of refraction for the metal-sphere array is

$$n = n_1\sqrt{\left(\frac{1 + 2f}{1 + f/2}\right)} \tag{7}$$

and the index of refraction for the dielectric-sphere array is

$$n = n_1\sqrt{\left(\frac{1 + 2fC}{1 - fC}\right)}, \tag{8}$$

where n_1 is the index of refraction of the supporting medium.

The theoretical limits of the obtainable dielectric quantities depend on the packing efficiency of the spheres. In a cubical lattice, the maximum fraction of the volume that can be filled with spherical particles is $f = \pi/6 = 52.4$ per cent for equal-sized particles.[4]

E. Losses in Artificial Dielectric

Formulas for the losses within the metal-sphere array have been derived by Pasternack and are included in his paper. No attempt has been made to verify these formulas. However, we may say, qualitatively, that the losses within the metal-sphere array are small inasmuch as the fields penetrate only a small distance into the metal surface (as is shown by the skin depth). The losses will be due primarily to eddy currents induced on the surface of the spheres, rather than to hysteresis or other ferromagnetic effects.

In the case of the dielectric spheres, losses will be caused by the material of the spheres. The dielectric spheres used were of fused quartz, which has a very low loss factor.

When the loss factor is very small (as in the two types of sphere discussed above) its accurate measurement is difficult because of the limited sensitivity of the detector and the presence of other losses (e.g., resistance losses in the junction of the cavity, metal losses within the cavity itself, and so on). These other losses, though small, are comparable in magnitude with the losses within the dielectric itself.

III. Description of Apparatus

A. General

The use of waveguide techniques for the measurement of dielectric properties seemed more desirable than free space or other resonant cavity methods of measurement because of the small-sized sample required, the simple geometry of the sample, and the single-mode propagation within the waveguide. Furthermore, the necessary components for such a waveguide device were readily available (see Fig. 1).

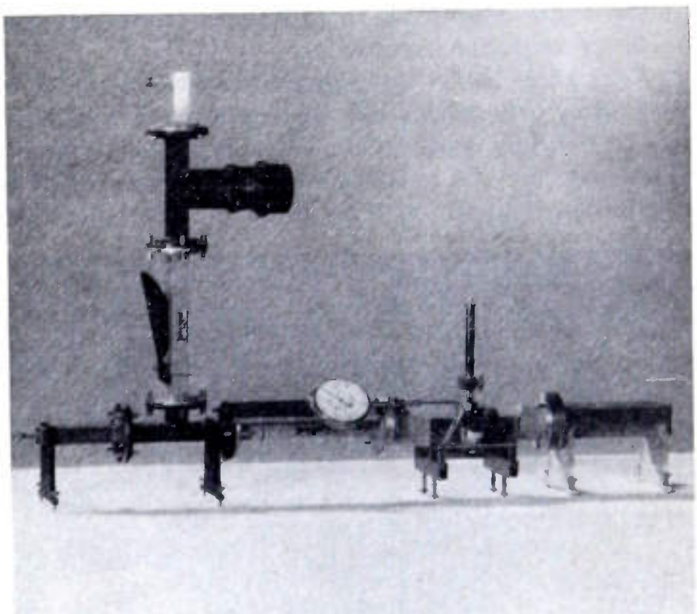

Fig. 1—Waveguide unit.

Since it was the goal of this research to obtain reasonably accurate data rather than to develop extremely accurate equipment, the microwave unit was built and put to use without making extensive improvements or correcting minor difficulties. The device was immediately adaptable to use in measuring ordinary dielectric

[4] This means that in the case of a metal-sphere array in air, the maximum value of n is 1.273, whereas for the dielectric-sphere array, the maximum value of n will depend on the sphere material.

materials as well as the artificial dielectric. A block diagram is shown in Fig. 2.

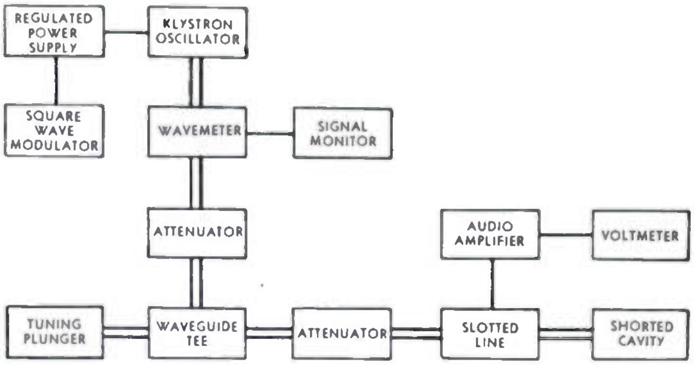

Fig. 2—Block diagram of a waveguide system.

B. Shorted Cavity

The shorted cavity that receives the sample to be measured is a unit of cross section 1.874×0.937 inches ID at one end, matching into RG/49-U guide (ID 1.874×0.874 inches). A tapered section 1.522 inches long leads smoothly to a cross section of 1.874×0.937 inches ID. The propagation properties of the resulting waveguide have not been altered since the a dimension is not changed. The gradual tapering of the b dimension from 0.874 to 0.937 inch insures very small reflections from points of discontinuity. The b dimension was made equal to one-half the a dimension. With this relationship between a and b, the choice of a cubic cell size such that an even integral number of cells will fit in the a dimension leads to an integral number of cells in the cross section as a whole.

The sample holder is terminated by a flat metal plate accurately perpendicular to the axis of the guide and recessed to insure good mechanical and electrical contact. This plate may be removed to insert the sample. It is found that the position of the mechanical and electrical shorts may not be identical because of bad contact between the shorting plate and the walls of the cavity. To insure a constant position of the electrical short for successive measurements, the shorting plate was secured firmly by two screws (see Fig. 3).

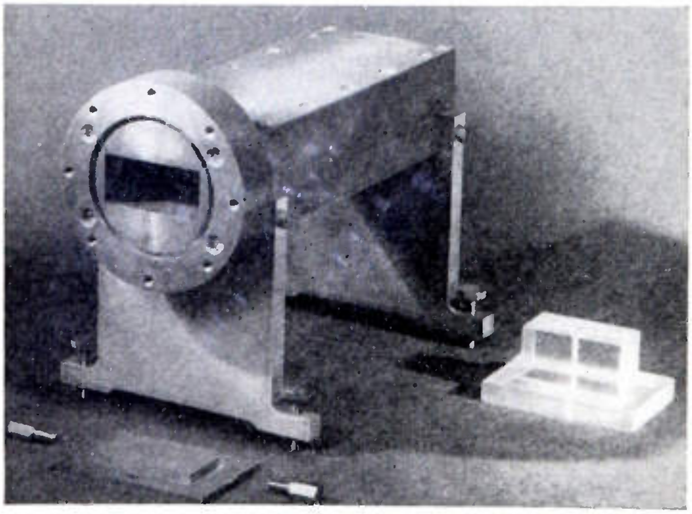

Fig. 3—Shorted cavity.

A choke-flange coupling was used to join the cavity to the slotted line.

IV. Description of Samples

The sample of artificial dielectric was composed of a cubic lattice of $\frac{1}{4}$-inch diameter spheres held in place by layers of Styrofoam (Dow Chemical Company) in which hemispherical recesses were milled to receive the spheres (see Fig. 4).

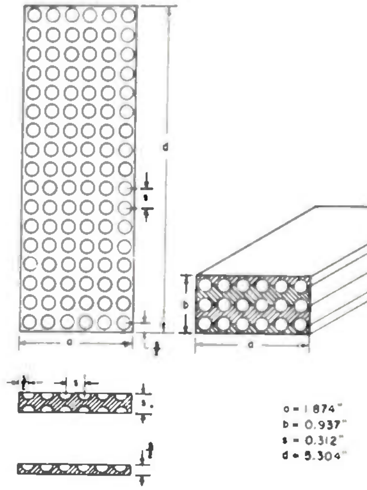

Fig. 4—Styrofoam sphere spacing structure.

The spheres formed a cubic lattice similar to a natural crystalline lattice. The distance between centers of adjacent spheres was 0.312 inch, allowing an integral number of cubic cells 0.312 inch on the side within the cross-sectional opening. The distance between the center of the outer sphere and the wall of the waveguide was one-half the spacing between the centers of the spheres.

The metal spheres used were commercially ground steel balls with tolerances held within 0.001 inch on the diameter. The dielectric spheres used were made of fused quartz ($\epsilon = 3.78$), with like tolerances. The spacing between the spheres was held within ± 0.003 inch.

The $\frac{1}{4}$-inch diameter spheres were chosen because, with the convenient spacing between centers of adjacent spheres of 0.312 inch offered by this size waveguide, these spheres gave a fractional volume of about 27 per cent, approximately equal to one-half the maximum value of f for this type of lattice. The sample (Fig. 4) was 13.472 cm long, comprising 17 transverse rows each containing 18 individual spheres. The wavelength within the samples is, by coincidence, approximately equal for both the metal and the dielectric array. The diameter of the spheres was approximately $\lambda/8$ and the spacing between edges of the spheres about $\lambda/32$, where λ is the wavelength within the material. Only one length of the structure supporting the metal or dielectric spheres was required. For successive measurements the sample was shortened by removing one transverse row of spheres at a time, correction being made for the residual Styrofoam.

V. Method of Measuring Samples

Measurements were made in rectangular waveguide at a frequency of 5,000 mc (guide wavelength 7.724 cm). Von Hippel's shorted line method[5] was used to measure the dielectric properties. The same general procedure was used for the metal-sphere array as for an ordinary dielectric material. However, in order to determine the magnetic permeability and the magnetic loss factor in addition to the dielectric constant and the dielectric loss, it was necessary to obtain data with the length of sample at two positions in the guide—against the shorting plate and at a distance of $\lambda_g/4$ from the shorting plate, where λ_g is the guide wavelength.

The experimentally measured quantities are the standing-wave ratio and the distance of the first minimum of the standing-wave pattern from the air-sample interface. These two quantities for the two positions of sample suffice to determine K_e, K_m, tan δ_e, and tan δ_m.

The measured values of a dielectric loss factor, tan δ_e, and the magnetic loss factor, tan δ_m, include the losses due to the guide and the imbedder material, as well as to the array of spheres. No attempt has been made to separate component values of individual loss.

The standing-wave ratio being quite large, it was measured by the double-power-point method. This requires a determination of the probe position and of the intensity of the field. The value of the field intensity at the node is determined; the points, one on either side of the node, for which the value of the field intensity is twice that at the node, are then found. The distance Δx between the double-power points determines the SWR as follows:

$$SWR = \frac{P_{\min}}{P_{\max}} = \frac{\pi \Delta x}{\lambda_g}.$$

We thus obtain simultaneously the SWR and the position of the first minimum from the interface.

The calculation procedure to determine K_e, K_m, tan δ_e, and tan δ_m from the experimental quantities is outlined in Appendix IV. The necessary relations and equations are those used with ordinary dielectric materials, and are readily available.[6]

VI. Estimated Accuracy of Measurement

To estimate the accuracy of measurements themselves—that is, experimental procedure rather than application of theory—a sample block of polystyrene was measured with this equipment. From the average values obtained for the constants, the estimated maximum error in the measurement of K_e and K_m is of the order of 3 per cent.

[5] C. G. Montgomery, "Techniques of Microwave Measurements," Radiation Laboratory Series, vol. 11, McGraw-Hill Book Co., New York, vol. 11, p. 625; 1947.

[6] W. B. Westphal, "Techniques and Calculations Used in Dielectric Measurements on Shorted Lines," NDDC Report No. 490; August, 1945.

VII. Experimental Data

A. Table of Constants of Metal-Sphere Array

Values of the constants of the metal-sphere artificial dielectric calculated from experimental data are shown in Table I.

TABLE I

Transverse Rows	Length of Sample (cm)	n	K_e	K_m
1	0.792	1.284	2.391	0.690
2	1.585	1.245	1.999	0.776
3	2.377	1.231	2.664	0.569
4	3.170	1.222	1.982	0.753
5	3.962	1.220	2.242	0.663
6	4.755	1.201	2.703	0.534
7	5.547	1.205	2.202	0.660
8	6.340	1.206	2.202	0.660
9	7.132	1.198	2.015	0.712
10	7.925	1.197	2.515	0.569
11	8.717	1.202	2.280	0.634
12	9.510	1.199	2.289	0.628
13	10.302	1.118	1.487	0.840
14	11.095	1.202	2.374	0.608
15	11.887	1.201	2.287	0.631
16	12.680	1.200	2.364	0.609
17	13.472	1.202	2.106	0.686
Theoretical		1.182	2.169	0 644

B. Table of Constants of Dielectric-Sphere Array

Values of the constants of the dielectric-sphere artificial dielectric calculated from experimental data are given in Table II.

TABLE II

Transverse Rows	Length of Sample (cm)	n	K_e	K_m
1	0.792	1.311	1.560	1.101
2	1.585	1.263	1.470	1.085
3	2.377	1.249	1.587	0.983
4	3.170	1.242	1.423	1.084
5	3.962	1.237	1.540	0.993
6	4.755	1.234	1.477	1.030
7	5.547	1.232	1.570	0.966
8	6.340	1.231	1.514	1.001
9	7.132	1.056	0.644	1.731
10	7.925	1.228	1.470	1.026
11	8.717	1.227	1.529	0.986
12	9.510	1.227	1.533	0.982
13	10.302	1.226	1.442	1.043
14	11.095	1.225	1.486	1.010
15	11.887	1.224	1.506	0.994
16	12.680	1.224	1.434	1.044
17	13.472	1.223	1.503	0.996
Theoretical		1.221	1.490	1.000

C. Explanation of Graphs

Fig. 5 is a plot of the index of refraction versus the length of sample for the metal-sphere and dielectric-sphere arrays. The length of sample is expressed in terms of the number of transverse rows of spheres. One transverse row of spheres is 0.792 cm long.

It may be seen that, as the length of the sample increases, the index of refraction approaches a definite value of $n = 1.20$. The index, of course, should be constant for all lengths. With this technique of measurement, however, data for the shorter lengths of sample are less reliable. For the metal-sphere array, the percentage error between the theoretical value and the experimental value of n is nowhere greater than 9 per cent, and for samples of 3 or more rows, the error is less than 5 per cent.

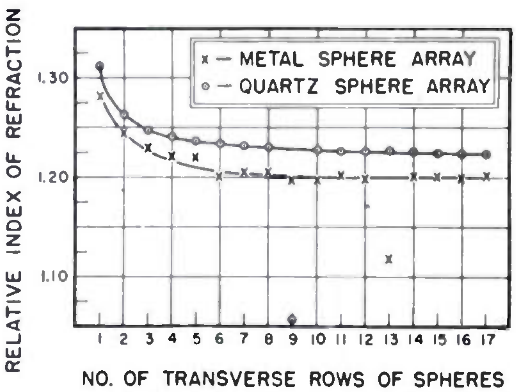

Fig. 5—Relative index of refraction versus sample length.

The plot of the index of refraction for the dielectric-sphere array produces a much smoother curve and approaches the value of $n = 1.233$. The percentage error between the experimental and the theoretical value for n is much less than that for the metal-sphere array. For lengths of sample longer than 3 rows, the percentage error is less than 2 per cent, except for the 9-row length of sample. The graph shows that for both arrays the value of n is greater than the theoretical value except for one point on each of the curves.

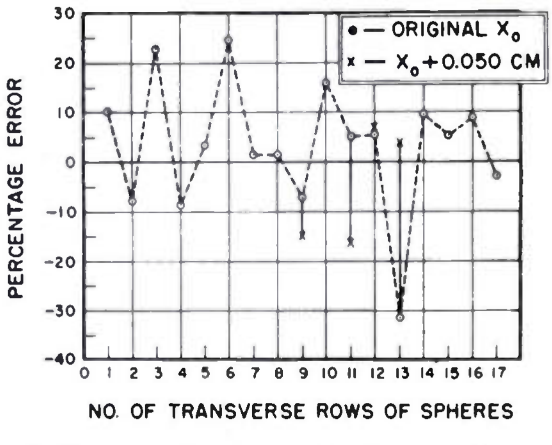

Fig. 6—Percentage error of K_e versus sample length—metal spheres.

Fig. 6 is a plot of the percentage error between the experimentally determined and the theoretical value of K_e for the metal-sphere array. Unfortunately, for four of the sample lengths measured, a minimum occurred at the air-dielectric interface, or very close to it. The measurement of X_0—the distance of the first minimum of the standing-wave pattern from the air-dielectric interface—is extremely critical in this region because tangent functions are used in the calculations. Because the sample length must be in terms of integral numbers of transverse rows of spheres, it is impossible to investigate lengths of sample differing by small increments from these critical lengths, as is customary when measuring ordinary dielectrics. Therefore, for a more accurate estimate of K_e, and also K_m, a large number of sample lengths should be measured, and the critical lengths ignored.

The graph indicates that the value of the percentage error of K_e fluctuates about zero. To show the criticalness of X_0, 0.050 cm was added to the original value of X_0, and K_e recalculated from this value of X_0. A very large change occurs in the sample with 13 transverse rows, whereas, for the same change in X_0, a very small change occurs in the sample with 12 rows. This change in X_0, however, does not cause the value of the K_e error to vary in the same direction for different lengths of sample. Possible error in X_0 can be attributed to two causes: First, the actual length of sample is not known definitely. (This is the interface problem which will be discussed presently.) Second, the value of λ_g is critical in the calculation of X_0.

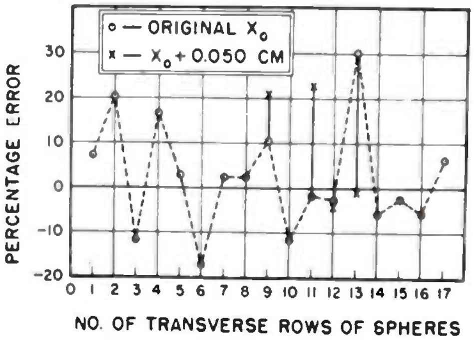

Fig. 7—Percentage error of K_m versus sample length—metal spheres.

The standing-wave pattern is investigated at a distance of two or three guide wavelengths from the sample. Therefore, any error in λ_g is multiplied two or three times in the determination of X_0. It was found that, because of the instability of the klystron oscillator, the wavelength varied by 0.2 per cent in the course of measurements, giving a maximum error of 0.018 cm in the guide wavelength.

Fig. 7 is a plot of the percentage error of K_m versus the length of sample for the metal-sphere array, together with a change in X_0 of +0.050 cm. Again the percentage error fluctuates about zero.

Fig. 8 is a similar plot of the percentage error of K_e versus the length of sample for the dielectric spheres. The percentage errors are far less, and again fluctuate about zero. The percentage error in K_e resulting from the addition of 0.100 cm to X_0 is also indicated.

The following circumstance may seem strange at first sight: Despite the fact that K_e, K_m, and n are all obtained from the same experimental data for both arrays, at some bad points the percentage error in K_e and K_m may be as much as 30 per cent, whereas, except for one bad value in the quartz-sphere array, the error in n is never greater than 5 per cent throughout the region from 3 to 17 rows. This discrepancy in magnitude of error requires explanation.

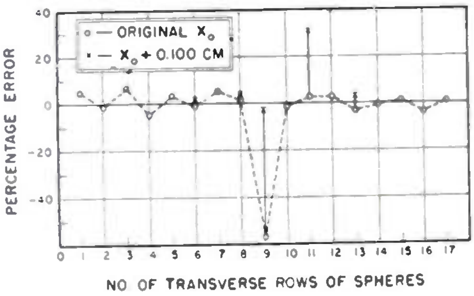

Fig. 8—Percentage error of K_e versus sample length—dielectric spheres.

The answer rests on a relation between the percentage errors in K_e and K_m, which could no doubt also be derived from the Westphal equations. For samples from 3 to 17 rows long, it is found that the algebraic sum of the percentage errors in K_e and K_m is ten or less. In other words, no matter how large their individual values, $\Delta K_e/K_e$ and $\Delta K_m/K_m$ are approximately equal in magnitude and opposite in sign for any given sample length. It can be easily shown that a 5 per cent error in n is not incompatible with relatively large percentage errors in K_e and K_m, subject only to the condition that

$$\frac{\Delta K_e}{K_e} + \frac{\Delta K_m}{K_m} \leqq 0.10.$$

Since

$$n = \sqrt{K_e K_m}$$

$$\frac{\Delta n}{n} = \frac{1}{n}\left\{\frac{\sqrt{K_m}}{2}\frac{\Delta K_e}{\sqrt{K_e}} + \frac{\sqrt{K_e}}{2}\frac{\Delta K_m}{\sqrt{K_m}}\right\}$$

$$= \frac{1}{2}\left\{\frac{\Delta K_e}{K_e} + \frac{\Delta K_m}{K_m}\right\} = 0.05.$$

That is, the percentage error in n is within 5 per cent, and this value checks exactly with that given by Fig. 5.

D. Interface Problem

The relative index of refraction at the boundary increases from unity to the true or bulk index of the dielectric through a transition region. The extent of this region is probably of the order of one cell length, which in this case would be small compared to a wavelength. The problem, then, of correcting measurements made on such a sample is one of finding the electrical equivalent of its length. Such an equivalent length is thus defined as having a constant bulk index throughout its extent with no transition region. In general, the equivalent length will not be the same as the physical length of the sample measured. The transition region and equivalent length may be represented as in Fig. 9.

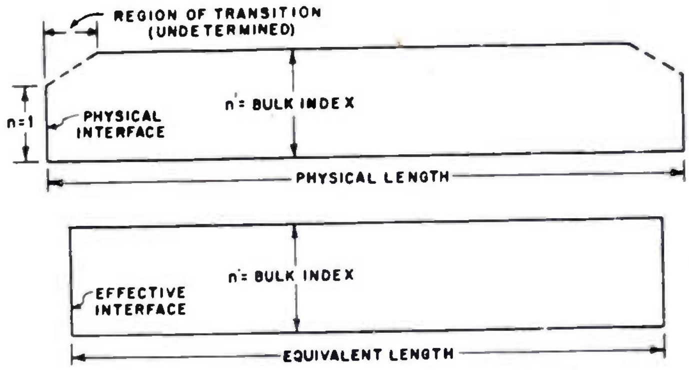

Fig. 9—Transition region and equivalent length.

In this work it has been assumed that the effective interface coincides exactly with the physical boundaries. We now consider the effect of such an approximation.

First, if the effective interface does not coincide exactly with the physical boundary, an error will result because the measured X_0, the distance of the first minimum from the interface, is not the proper one to use in the conventional formulas from which the dielectric quantities are calculated.

Second, an additional error will be made by assuming that a short circuit exists when the sample is placed against the terminating metal plate, and an open line when it is placed at an assumed distance of $\lambda_g/4$ from the terminating plate. Nevertheless, as measurements show, the approximation is satisfactory in that it does give consistent values for the measured dielectric quantities for any sample longer than a few cells. For the latter, the effects discussed above are undoubtedly more important. The interface problem is usually side-stepped in the case of ordinary dielectrics, such as crystals and homogeneous media, because of the microscopic dimensions in the atomic "array" elements.

VIII. Conclusion

Since the theoretical expressions derived from the Clausius-Mossotti equation and the measurements seem to support each other, it follows that the general application of this equation to the cubic lattice is reasonable. It allows one to calculate within a few per cent what to expect for the resulting permeability, dielectric constant, and index of refraction for an artificial dielectric with this geometry.

A. Practicality

It is seen that the cubic array of metal spheres in air is not capable of producing an index of refraction greater than 1.273, and the practical value obtainable is much less. An additive effect, however, may be obtained by imbedding the metal spheres in a supporting medium of higher dielectric constant. To produce a relatively large value for n with this type of array, the fractional volume filled must be quite large. Hence the weight as well as the mechanical difficulties is a factor to be considered. The metal spheres, however, can be replaced by metal-plated spheres of other materials, or the spherical recesses themselves may be metal plated.

The array of dielectric spheres is capable of producing higher values of n, depending upon the material of the sphere and the fractional volume filled. Accordingly, there are three parameters which may be varied to produce a prescribed index of refraction. Although large values of n are obtainable with proper choice of sphere material, losses may become prohibitive. Use of spheres of the low-loss, high dielectric constant, titanium compounds, or some similar material, would give large resultant values of n combined with low loss. Titanium compound spheres are being fabricated for use in this type of array. The nonferroelectric compounds will be used to produce an artificial dielectric medium with a high index of refraction, and to produce a highly refractive medium whose index can be varied by applying a static field across the medium.

B. Measurement Techniques

The waves considered in this series of experiments can all be regarded as a superimposition of plane waves; the theory based on a plane-wave analysis is therefore applicable here, and in fact to all wave types which can be synthesized from plane waves. The type of field used merely influences the method by which experimental observations are reduced to the final values of dielectric constant and magnetic permeability. Within the scope of the analysis presented, the electric and magnetic susceptibilities are independent of the fields.

It is felt that one of the contributions of this paper is that it indicates that the standard waveguide procedures, such as using the shorted-line method of measuring dielectric properties in rectangular waveguide, are readily adaptable to artificial dielectrics composed of arrays of metal or dielectric spheres. However, relatively large experimental errors may be easily made when a minimum or maximum occurs at or near the air-dielectric interface. A minimum will occur at this interface both when the sample is $n\lambda_{m/2}$ long and placed against the shorting plate, and when the sample is $(2n+1)\lambda_{m/4}$ long and placed $\lambda_{g/4}$ from the shorting plate. A maximum will occur at the interface when the sample is $(2n+1)\lambda_{m/4}$ long and placed against the shorting plate, and when the sample is $n\lambda_{m/2}$ long and placed $\lambda_{g/4}$ from the shorting plate. (λ_m is the wavelength within the material in the guide.) Therefore, the optimum length of sample to measure is an odd multiple of eight wavelengths.

C. Possible Extensions in This Field

Since the Clausius-Mossotti equation appears to be satisfactory for these two particular arrays, it might be worthwhile to consider how the polarizability equations must be modified to deal with spheres whose diameter approaches a wavelength.

In addition, the interface problem should be investigated both theoretically and experimentally to determine the equivalent length of sample appropriate to the shorted-line method of measuring the dielectric quantities.

IX. Acknowledgments

The author wishes to acknowledge the substantial contributions of Dr. M. W. P. Strandberg, who acted as consultant on this project. His comments on the theoretical aspects of the subject and his preparation of Appendices I–III aided materially in determining the course of research. Gratitude is expressed to R. E. Hiatt, who supervised and directed the research program, and to F. J. Zucker for valuable suggestions and criticisms in preparing this report.

Appendix I

Electric Polarizability

It may be worthwhile to review our picture of isotropic dielectric materials to see more readily how artificial dielectrics are possible.

The fundamental and primary field vector is the vector field strength $\vec{E}$, and it is defined in terms of the force $\vec{F}$ on a charge q; thus $\vec{F}=q\vec{E}$. Now $\vec{E}$ may be due to Coulomb forces resulting from free or induced charges, or both. Since it is generally desirable to calculate in terms of free charges (at our disposal), it is convenient to introduce another vector $\vec{D}$ related only to the free charges in a given physical situation. Since $\vec{D}$ is independent of the charges induced on the surface of a dielectric by the electric field, we may say that the component of $\vec{D}$ normal to the surface of the dielectric is continuous. This is our first *boundary condition*, and it arises, of course, from our qualitative definition of $\vec{D}$.

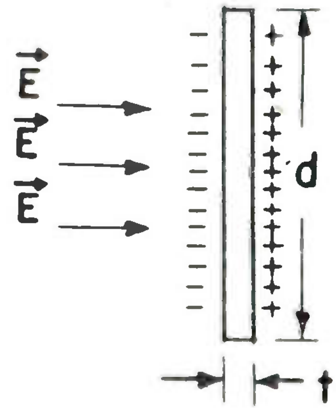

Fig. 10—Bound charges induced on dielectric disc.

We may conveniently relate $\vec{D}$ to $\vec{E}$ in the following way. We consider a thin disk of dielectric with a diameter d much greater than the thickness t. This disk is so placed in an electric field $\vec{E}$, that $\vec{E}$ is normal to the faces of the disk. Because the electrostatic force attracts the negative charges and repels the positive, bound charges will be induced on the faces of the disk by the field $\vec{E}$, as shown in Fig. 10. Considering the section in the center of the disk, we may say that $\vec{E}_D$, the field inside the dielectric, is equal to the imposed field $\vec{E}$ plus the (vector) field due to the induced surface charges.

$$\vec{E}_D = \vec{E} + \vec{E}_i, \tag{9}$$

where $\vec{E}_D$ is the field in the dielectric; $\vec{E}$, the field in free space, and $\vec{E}_i$, the field due to induced charges. On the other hand $\vec{D}$, resulting from the free charges which create $\vec{E}$, is in the direction of $\vec{E}$ and is normal to the dielectric; hence it is continuous. Therefore,

$$\vec{D}_D = \vec{D}, \tag{10}$$

where $\vec{D}$ is the displacement in free space and $\vec{D}_D$, the displacement in the dielectric. We may arbitrarily set

$$\vec{D} = \epsilon_0 \vec{E} \tag{11}$$

for free space, since $\vec{D}$ has not yet been quantitatively defined. Combining (9) and (11), we have

$$\epsilon_0 \vec{E}_D = \vec{D} + \epsilon_0 \vec{E}_i. \tag{12}$$

Now we may readily show from Coulomb's law ($|\vec{E}| = q/4\pi\epsilon_0 r^2$) that the field between two plane charged surfaces with surface-charge density σ is σ/ϵ_0. This condition is met in our disk since we have said that $d \gg t$. We note also that a small section of the disk of cross-sectional area A (see Fig. 11) forms a dipole moment of magnitude $\sigma_i A t$ and, since the volume is At, the dipole moment per unit volume is $\sigma = \vec{P}$. The dipole moment is defined to have a direction from the negative (−) to the positive (+) charge, and this is opposite to the direction of the field $\vec{E}_i$.

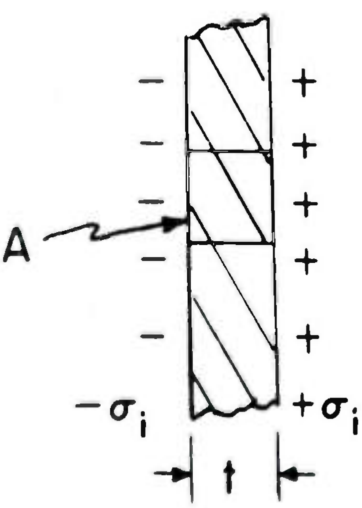

Fig. 11—Two-plane charged surfaces with surface charge density σ.

Since $|\vec{E}_i| = \sigma_i/\epsilon_0$ and $|\vec{P}| = \sigma_i$, we may write $\epsilon_0 \vec{E}_i = -\vec{P}$. We may then write (12) as

$$\epsilon_0 \vec{E}_D = \vec{D} - \vec{P} \tag{13}$$

or

$$\vec{D} = \vec{D}_D = \epsilon_0 \vec{E}_D + \vec{P}.$$

In general $\vec{P}$ is proportional to $\vec{E}_D$ so that at this point we introduce Maxwell's assumption

$$\vec{P} = \epsilon_0 \chi \vec{E}_D,$$

obtaining

$$\vec{D}_D = \epsilon_0 \vec{E}_D (1 + \chi),$$

where χ is the electric susceptibility, or

$$\vec{D}_D = K_e \epsilon_0 \vec{E}_D = \epsilon \vec{E}_D, \tag{14}$$

where K_e is the relative dielectric constant. The existence of a dielectric constant greater than one, then, is due to the existence of a finite dipole moment per unit volume. For our purposes (13) will be the most convenient equation from which to calculate K_e. The arguments used above may be extended to a body of any shape, for any body may be thought of as being composed of our thin disks; and if t and d are taken sufficiently small, $\vec{E}$ likewise will be uniform to any desired approximation.

Our problem, then, in calculating the dielectric constant of artificial dielectrics, is to calculate $\vec{P}$, the dipole moment per unit volume.

If we consider a perfectly conducting sphere in a uniform field $\vec{E}$, we know that charges will be induced on the surface so that no tangential component of $\vec{E}$ exists at the surface of the sphere (see Fig. 12). A tangential component of $\vec{E}$ would create large surface currents; hence it cannot be physically allowed. This solution is

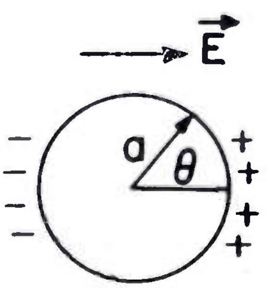

Fig. 12—Perfectly conducting sphere in a uniform field.

well known, and we say that the charges distribute themselves so that the sphere has a dipole moment $\vec{p}$. The tangential component of the dipole $\vec{E}$ field is its θ component,

$$\vec{E}_{\theta i} = \frac{\vec{p} \sin\theta}{4\pi\epsilon_0 r^3}.$$

The θ component of the imposed field is $\vec{E} = -\vec{E} \sin\theta$. At the surface ($r=a$) the resultant of these fields must be zero; this relation determines $\vec{p}$, the effective dipole moment of the sphere.

$$-\vec{E}\sin\theta + \frac{\vec{p}\sin\theta}{4\pi\epsilon_0 a^3} = 0 \tag{15}$$

$$\vec{p}_- = 4\pi\epsilon_0 a^3 \vec{E} = \alpha_e \epsilon_0 \vec{E},$$

where α_e is the electric polarizability of the sphere. With N such spheres per unit volume, the dielectric constant, from (13), (14), and (15) proves to be

$$\vec{P} = \vec{p}_- N/V$$
$$K_e = (1 + 4\pi a^3 N/V) = (1 + \alpha_e N/V). \tag{16}$$

It is well to examine at this point what restrictions our derivation has placed on the use of relation (15).

First, we have said that $\vec{E}$ is uniform. If we have a lattice of spheres, $\vec{E}$ will be, in fact, our imposed field plus the field due to induced charges on the nearest spheres. The field from the induced charges, being a dipole field, falls off as $1/r^3$; thus, if the spheres are $6a$ on centers, the field due to the dipole moment of one sphere will, at the adjacent sphere, be reduced by a factor of $1/(5)^3$ or $1/125$ from its value at the charge-bearing sphere. Since the field due to the induced charges at the sphere itself can be no greater than the imposed field, the field due to these induced charges at the adjacent sphere must be less than $\vec{E}/125$—a negligible quantity. Equation (8) then is applicable when the center-to-center spacing of the spheres is greater than six times the radius (see Fig. 13).

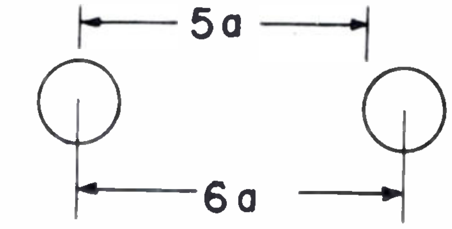

Fig. 13—Conducting spheres spaced at least six times the radius apart.

When we have radiating (alternating) fields, there is a spatial variation of $\vec{E}$ which takes place in distances comparable with the wavelength of the radiation. Thus, in order to meet the criterion for a uniform field, we say that a must be much less than one wavelength, λ. The boundary conditions for electromagnetic radiation and static fields are the same; thus, our derivation of (16) is satisfactory so long as $a \ll \lambda$ and the spheres are widely spaced.

When the spheres are closely spaced, the fields due to the induced charges on adjacent spheres may be taken into account by the ingenious method of Clausius and Mossotti. To calculate the electric field acting on the sphere, we remove it and ask what fields exist in the absence of one sphere or, in other words, in the absence of an element of our dielectric (see Fig. 14). The field existing in the empty hole will be that due to the average field in the dielectric minus that due to the induced dipole moment of the sphere which we have removed. As long as the array of spheres has three-dimensional symmetry, it can be shown that the material is a continuous isotropic medium.[7] We proceed to calculate the field in the cavity carved out by the removal of one sphere. We remember that $\vec{D}$ normal to the dielectric surface must be continuous and that, as a consequence, the magnitude of $\vec{P}$ normal to the cavity surface equals the negative of the induced surface charge density σ_i. Thus

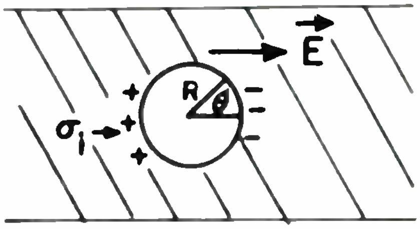

Fig. 14—Spherical cavity in dielectric.

we may write $\sigma_i = -|\vec{P}| \cos\theta$. We may calculate the horizontal component of the field due to σ_i at the *center* of the cavity readily from Coulomb's law

$$|(\vec{E}_i)|_{\text{Horizontal}} = \int_s \frac{\sigma_i ds}{4\pi\epsilon_0 R^2} \cos\theta$$
$$= \int_0^\pi \frac{|\vec{P}|\cos\theta}{4\pi\epsilon_0 R^2} \cos\theta \cdot 2\pi R \sin\theta d\theta$$
$$= \frac{|\vec{P}|}{2\epsilon_0} \int_0^\pi \cos^2\theta \sin\theta d\theta = \frac{|\vec{P}|}{3\epsilon_0}. \tag{17}$$

We have calculated this at the center of the cavity, but we may show generally that the field, due to the induced surface charge (Fig. 15), is everywhere in the cavity horizontal and equal to $|\vec{P}/3\epsilon_0|$. First, we calculate the dipole moment of the sphere due to the induced charges

$$d|\vec{p}_-| = \sigma_i(2\pi R \sin\theta) \times R d\theta \times 2R\cos\theta = dq \cdot l$$

$$\therefore \quad |\vec{p}_-| = \int_s |dp| = -4\pi R^3 |\vec{P}| \int_0^{\pi/2} \cos^2\theta \sin\theta d\theta$$

$$\vec{p}_- = -\frac{4\pi R^3}{3} \vec{p}. \tag{18}$$

This we could have deduced without the computation, for in forming the cavity, were moved material of volume

[7] Intuitively, we may see that, although the array itself has three orthogonal axes and is, therefore, nonisotropic, the electromagnetic waves are in no way confined to travel along those axes. They may, indeed, spread freely in all directions and thus behave as if in an effectively isotropic medium.

$4\pi R^3/3$, and the dipole moment per unit volume was $\vec{P}$. This is the same as inserting a dipole of moment $-(4\pi R^3/3)(\vec{P})$. Now we apply our boundary condition that $\vec{D}$ normal to the surface is continuous. We write normal components with a subscript n while the primed quantities refer to vectors inside the cavity.

$$\vec{D}_n = \vec{D}_n'.$$

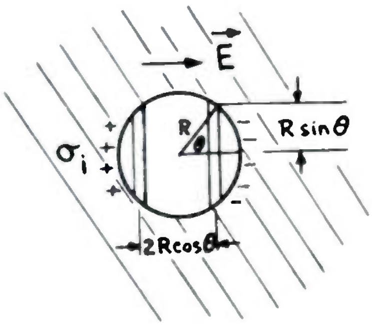

Fig. 15—Induced surface charge in the cavity.

We assume $\vec{E}'$ is horizontal and find that this leads to no inconsistencies. Now $\vec{D}_n$ is composed of the normal component due to the dipole and due to $\vec{E}$; thus,

$$\vec{D}_n = \epsilon\left[\vec{E}\cos\theta - \frac{2\vec{p}_-\cos\theta}{4\pi\epsilon R^3}\right] = \vec{D}_n' = \epsilon_0\vec{E}'\cos\theta \qquad (19)$$

when we remember the expression for the vertical field due to a dipole moment $\vec{p}$. Inserting the value for $\vec{p}_-$ in terms of $\vec{P}$, we find

$$\epsilon_0\vec{E}' = \epsilon\vec{E} - 2\vec{P}/3.$$

But $\epsilon\vec{E}$ is given as

$$\epsilon\vec{E} = \epsilon_0\vec{E} + \vec{P}.$$

Therefore, we have

$$\vec{E}' = \vec{E} + \vec{P}/3\epsilon_0, \qquad (20)$$

which is identical to the value calculated above for the field at the center of the cavity. The term $\vec{P}/3\epsilon_0$ is the additional field due to the surrounding polarized spheres. The dipole moment for a sphere is thus

$$\vec{p}_- = \alpha_e\epsilon_0\vec{E}' = \alpha_e\epsilon_0(\vec{E} + \vec{P}/3\epsilon_0)$$

so that

$$\vec{P} = \vec{p}_-N/V = (N/V)\alpha_e\epsilon_0(\vec{E} + \vec{P}/3\epsilon_0),$$

$$\frac{N\alpha_e}{V} = \frac{3\vec{P}}{3\epsilon_0\vec{E} + \vec{P}}.$$

Remembering that, according to the usual notation,

$$\epsilon\vec{E} = \epsilon_0\vec{E} + \vec{P}$$

or

$$\vec{P} = \vec{E}(\epsilon - \epsilon_0),$$

we have the usual Clausius-Mossotti relation

$$\frac{\alpha_e N}{3V} = \frac{\epsilon - \epsilon_0}{2\epsilon_0 + \epsilon} = \frac{K_e - 1}{K_e + 2}. \qquad (21)$$

We may note that (21) was derived under the assumption of an isotropic medium; thus it is applicable, in general, only in that case (i.e., in the case of spherical balls in a cubic lattice).

Since this expression, however, relates the electric polarizability of the sphere α_e to the effective dielectric constant K_e, we may apply it either to conducting or dielectric spheres. We have seen that, for a conducting sphere,

$$\alpha_e = 4\pi a^3,$$

and we may similarly calculate a value of α_e for a dielectric sphere.

We proceed as in the above calculations. The volume of the sphere is $4\pi a^3/3$ so that the dipole moment of the sphere is

$$\vec{p}_- = \vec{P}4\pi a^3/3. \qquad (22)$$

Remembering that $\vec{D}_n$ is continuous, and assuming that $\vec{E}'$ in the sphere is uniform (we will check this, as before, by noting that no inconsistency arises), we have

$$\vec{D}_n = \epsilon_0\left[\frac{2\vec{P}\cos\theta}{3\epsilon_0} + \vec{E}\cos\theta\right] = \vec{D}_n' = (\epsilon_0\vec{E}' + \vec{P})\cos\theta.$$

Since the $\cos\theta$ term cancels out, our assumption that E' is uniform is upheld, and we have

$$\epsilon_0\vec{E} = \epsilon_0\vec{E}' + \vec{P}/3. \qquad (23)$$

But, using our notation of (14), this may be rewritten as

$$\epsilon\vec{E}' = \epsilon_0\vec{E}' + \vec{P},$$

or

$$\vec{E}' = \frac{\vec{P}}{\epsilon - \epsilon_0}. \qquad (24)$$

So by combining (23) and (24), we have

$$\vec{P} = \frac{3\epsilon_0(\epsilon - \epsilon_0)}{2\epsilon_0 + \epsilon}\vec{E},$$

which means that the dipole moment of the dielectric sphere in air is

$$\vec{p}_- = \frac{4}{3}\pi a^3\vec{P} = 4\pi a^3\epsilon_0\left(\frac{\epsilon - \epsilon_0}{2\epsilon_0 + \epsilon}\right)\vec{E} = \epsilon_0\alpha_e\vec{E},$$

or

$$\alpha_e = 4\pi a^3\left(\frac{\epsilon - \epsilon_0}{2\epsilon_0 + \epsilon}\right) = 4\pi a^3\left(\frac{K_e - 1}{K_e + 2}\right), \qquad (25)$$

where K_e is the dielectric coefficient of the sphere material. We may, of course, use this polarizability, α_e, in the usual way in the Clausius-Mossotti relation (21).

APPENDIX II

RADIAL FIELD OF A DIPOLE

Consider two charges, $+q$ and $-q$, separated by a small distance l, as shown in Fig. 16. The electric field due to the $+q$ is

$$|(\vec{E}_+)_r| = \frac{q}{4\pi\epsilon_0(r + (l\cos\theta)/2)^2} = \frac{q}{4\pi\epsilon_0 r^2}\cdot\frac{1}{(1 + (l\cos\theta)/2r)^2},$$

and that due to $-q$ is

$$|(\vec{E}_-)_r| = \frac{-q}{4\pi\epsilon_0 r^2}\cdot\frac{1}{(1 - (l\cos\theta)/2r)^2}.$$

If l is much smaller than r, we may expand these as

$$|(\vec{E}_+)_r| \cong \frac{q}{4\pi\epsilon_0 r^2}\left(\frac{1 - l\cos\theta}{r}\right);$$

$$|(\vec{E}_-)_r| \cong \frac{-q}{4\pi\epsilon_0 r^2}\left(\frac{1 + l\cos\theta}{r}\right).$$

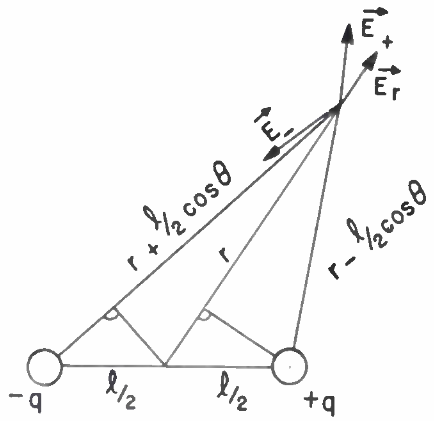

Fig. 16—The electric field due to two separated opposite charges.

Now since $\vec{E}-$ and $\vec{E}+$ are nearly along r, we may say that the magnitude of $(\vec{E})_r$ is equal to the sum of $E+$ and $E-$, or

$$|(\vec{E})_r| = \frac{q}{4\pi\epsilon_0 r^2}\left(\frac{2l\cos\theta}{r}\right) = \frac{2(ql)\cos\theta}{4\pi\epsilon_0 r^3}.$$

The quantity (ql) is called the dipole moment $\vec{p}_-$, and the approximation sign becomes an equality as $l \to 0$; thus,

$$|\vec{E}_r| = \frac{2|\vec{p}_-|\cos\theta}{4\pi\epsilon_0 r^3}.$$

The projections of $\vec{E}+$ and $\vec{E}-$ in the θ direction are simply, from geometry,

$$|(\vec{E}_+)| = \frac{l\sin\theta}{2r}, \qquad |(\vec{E}_-)| = \frac{l\sin\theta}{2r}.$$

Thus, E_θ, being the sum of these terms, is

$$E_\theta = |\vec{E}_\theta| = \frac{2q}{4\pi\epsilon_0 r^2}\,\frac{l\sin\theta}{2r} = \frac{|\vec{p}_-|\sin\theta}{4\pi\epsilon_0 r^3}.$$

APPENDIX III

MAGNETIC POLARIZABILITY

In the usual case, artificial dielectrics are in electromagnetic fields so that we must also consider the behavior of conducting spheres in magnetic fields.

Magnetic fields are treated by a method quite analogous to that used for electric fields. For this reason our treatment here will be concise and will be such as to yield rapidly an evaluation of the effect of a conducting sphere in an alternating magnetic field. This is the only case where such a treatment has relevance. (See Fig. 17.)

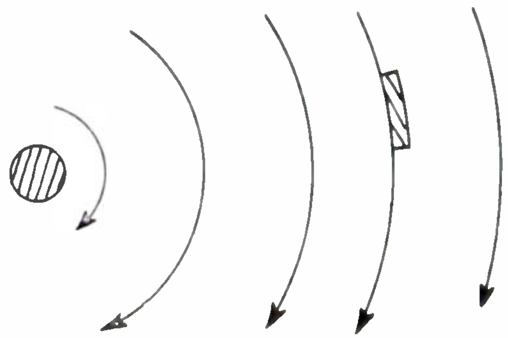

Fig. 17—Magnetic field around a wire conducting a current.

We have a vector $\vec{B}$ related to the force on a magnetic pole or on a conducting wire. This force depends upon the properties of the medium (just as $\vec{E}$ did above); so for ease in calculation we choose a second vector $\vec{H}$ which, in the absence of fixed magnetic poles, depends only upon the real magnetizing current at our disposal. The proper boundary condition for $\vec{H}$ is that the tangential component of $\vec{H}$ be continuous. This may be seen as follows: Consider the field around a wire conducting a current. If we put a "pencil" of magnetic material in this field so that $\vec{H}$ is tangential to the axis of the pencil, we require that $\vec{H}$ be independent of the magnetic field due to induced poles on the ends of the rod, or that $\vec{H}$ be continuous across the surface. This is not sufficient to define $\vec{H}$ since, if the wire were a thin-walled tube, then $\vec{H}$ would be finite outside the tube and zero inside. To cope with this case we may say that the discontinuity in the tangential component of $\vec{H}$ is equal to the real surface current density. This is Ampère's law.

To return to our magnetic rod, we know that $\vec{B}$ is not equal inside and outside the pencil because of the induced N and S poles. This effect was accounted for by Ampère with induced "surface currents," which play the same role as the induced surface charges in our electrostatic picture. Or, stated otherwise, the induced poles on the ends of the rod could be replaced, as far as forces are concerned, by a properly wound and excited solenoid. (See Fig. 18.)

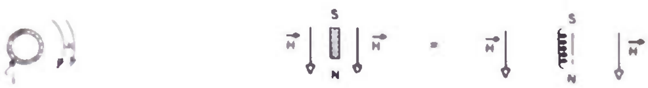

Fig. 18—"Surface currents," which play the same role as the induced surface charges in our electrostatic picture. Or, stated otherwise, the induced poles on the ends of the rod could be replaced, as far as forces are concerned, by a properly wound and excited solenoid.

We define the magnetic moment of a loop of area A carrying a current I as IA with a direction normal to A (i.e., $\vec{m}=IA\vec{n}$). (See Fig. 19.) Now consider a real solenoid with current I and length L, containing N turns. H outside is zero and, therefore, since the current density is (NI/L) ampere turns/m. $\vec{H}$ inside is $\vec{n}NI/L$. In free space we will take $\vec{B}=\mu_0 H$, since both arise from the effect of the same current. Thus, in the case where we have replaced our magnetic rod by an equivalent solenoid, we may say that $\vec{B}$ inside $=\vec{B}$ outside $+\mu_0\vec{n}NI/L$. Now we note that the magnetic moment of a section length l of our hypothetical solenoid is $\vec{m}=\vec{n}\,lA\,NI/L$, or the magnetic moment per unit volume is $\vec{M}=IN\vec{n}/L$. Since A has normals on opposite sides, we choose that normal which makes the following notation hold:

$$\vec{B}\text{ inside} = \vec{B}\text{ outside} + \mu_0\vec{M}. \tag{26}$$

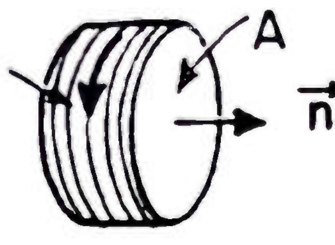

Fig. 19—Equivalent solenoid.

Or, if $\vec{M}$ is proportional to $\vec{H}$, we may write

$$\vec{B} = \mu_0\vec{H} + \mu_0\chi\vec{H} = \mu\vec{H} = K_m\mu_0\vec{H}, \tag{27}$$

where χ is the magnetic susceptibility and K_m the magnetic permeability. It is well to note the similarities and dissimilarities in the corresponding electric field equations.

$$\text{always true}\left\{\begin{aligned}\vec{D} &= \epsilon_0\vec{B} + \vec{P}\\ \vec{H} &= \frac{\vec{B}}{\mu_0} - \vec{M}\end{aligned}\right\}\left\{\begin{aligned}&= \epsilon\vec{E} \quad \vec{P} = \epsilon_0\chi_e\vec{E}\\ &\qquad\text{But;}\\ &= \frac{\vec{B}}{\mu} \quad \vec{M} = \chi_m\vec{H}\end{aligned}\right\}\text{often true.} \tag{28}$$

We now find the dipole moment of a conducting sphere in order to evaluate M, the magnetic moment per unit volume. In alternating electric and magnetic fields, the fields inside a conductor are vanishingly small because of the "skin effect." (See Fig. 20.) Thus, with a tangential $\vec{H}$ outside a sphere and zero $\vec{H}$ inside (for a perfect conductor), the surface current density is simply equal to the tangential $\vec{H}$. We must be careful, however, to remember that the tangential $\vec{H}$ is calculated using

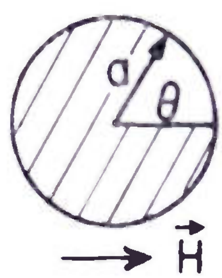

Fig 20 -Conducting sphere in magnetic field.

the total field resulting from both the imposed field $\vec{H}$ and the field due to the current on the surface of the conducting sphere. The field due to the current on the sphere must be that due to a magnetic dipole, since only this field has the axial symmetry demanded by the uniform imposed $\vec{H}$. So we may write

$$\vec{H}_{\text{tangential}} = -\vec{H}\sin\theta + \frac{\vec{m}\sin\theta}{4\pi a^3} = \vec{J}, \tag{29}$$

where J is the surface current density.

Now, we inquire, what is the density of the dipole moment for such a current distribution? The moment for a thin shell (Fig. 21) is

$$\begin{aligned} d\vec{m} &= \pi(a\sin\theta)^2\cdot\vec{J}ds\\ \vec{m} &= \pi a^3\int_0^\pi \vec{J}\sin^2\theta d\theta. \end{aligned} \tag{30}$$

Using the value of $\vec{J}$ from (29), we have

$$\begin{aligned} \vec{m} &= \pi a^3\left[-\vec{H} + \frac{\vec{m}}{4\pi a^3}\right]\int_0^\pi \sin^3\theta d\theta\\ &= \frac{4}{3}\pi a^3\left[-\vec{H} + \frac{\vec{m}}{4\pi a^3}\right]. \end{aligned} \tag{31}$$

Solving for $\vec{m}$, we find

$$\vec{m} = \frac{-4/3\pi a^3\vec{H}}{1-1/3} = -2\pi a^3\vec{H}. \tag{32}$$

The dipole moment per unit volume is then

$$\vec{M} = \frac{N}{V}\vec{m} = -2\pi a^3 \frac{N}{V}\vec{H} = \frac{N}{V}\alpha_m \vec{H} \qquad (33)$$

so that

$$\alpha_m = -2\pi a^3$$

$$\mu_0(\vec{H} + \vec{M}) = \mu\vec{H}$$

$$\mu = \mu_0(1 + \alpha_m).$$

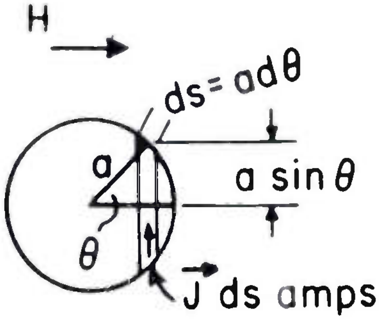

Fig. 21—Thin conducting shell in magnetic field.

As in the dielectric case, (33) is valid only for widely spaced spheres.

To take care of the case for closely spaced spheres, we must make a calculation analogous to that done by Clausius and Mossotti. We assume, as in the dielectric case, that the medium is isotropic (three-dimensional symmetry) and remove one sphere from the medium to evaluate the average $\vec{H}$ in the cavity. We know the tangential $\vec{H}$ is continuous at the cavity surface (Fig. 22). Remembering that $\vec{M}$ is the dipole moment per unit volume, we see that the cavity acts as if we had put into a uniform medium a dipole of moment $\vec{m} = -4\pi R^3\vec{M}/3$.

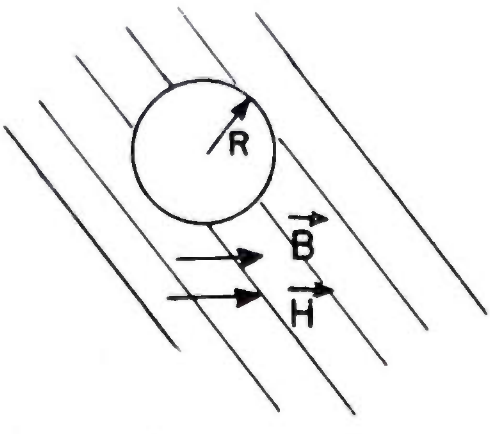

Fig. 22—Spherical cavity in dielectric.

Equating the tangential $\vec{H}$ inside and outside, we find (assuming $\vec{H}$ inside is uniform as before)

$$-\vec{H}\sin\theta + (\vec{m}\sin\theta)/4\pi R^2 = -\vec{H'}\sin\theta. \qquad (34)$$

The sin θ terms cancel, so $\vec{H'}$ is uniform.

$$\vec{H'} = \vec{H} + \vec{M}/3, \qquad (35)$$

and we have from (33)

$$\vec{M} = (N/V)\alpha_m(\vec{H} + \vec{M}/3), \qquad (36)$$

which using the definition

$$\mu\vec{H} = \mu_0\vec{H} + \mu_0\vec{M} \qquad (37)$$

or

$$\vec{H}(\mu - \mu_0) = \mu_0\vec{M}$$

yields

$$\frac{N\alpha_m}{V3} = \frac{\mu - \mu_0}{2\mu_0 + \mu} = \frac{K_m - 1}{K_m + 2}. \qquad (38)$$

Equation (38), then, with α_m from (33), may be used for any artificial dielectric lattice where the spacing of spheres is uniform on the three-space axes.

APPENDIX IV

DIAMAGNETIC DIELECTRIC MEASUREMENTS

Sample: Metal-Sphere Array	λ_g = 7.724 cm	K_e = 2.2018
	Date = March 2, 1950	K_m = 0.6599
Thickness: = 5.547 cm	Temp. = Room	$\tan\delta_e$ = 0.00742
		$\tan\delta_m$ = 0.00072

I. Sample on terminal plate

① $X_0 = 0.195$

② $\Delta x = 0.046$

③ $E = 0.0187$

④ $\tan(360\ X_0/\lambda_g) = 0.1602$

⑤ $[③^2 + ④^2]^{1/2} = 0.1613$

⑥ $\tan^{-1}\frac{④}{③} = 83.342°$

⑦ $\tan^{-1}③ \times ④ = 0.172°$

⑧ $[1 + ③^2 \times ④^2]^{1/2} = 1.0000$

⑨ $\tan\rho_l = \frac{⑤}{⑧} e^{j[-⑥+⑦]} = 0.1613 e^{j(-83.170°)}$

II. Sample $\lambda_g/4$ from terminal plate

$\textcircled{1}$ $X_0 = 2.641$

$\textcircled{2}$ $\Delta x = 0.123$

$\textcircled{3}$ $E = 0.0500$

$\textcircled{4}$ $\tan(360\, X_0/\lambda_g) = -1.5457$

$\textcircled{5}$ $[\textcircled{3}^2 + \textcircled{4}^2]^{1/2} = 1.5465$

$\textcircled{6}$ $\tan^{-1} \frac{\textcircled{4}}{\textcircled{3}} = -88.147°$

$\textcircled{7}$ $\tan^{-1} \textcircled{3} \times \textcircled{4} = -4.419°$

$\textcircled{8}$ $[1 + \textcircled{3}^2 \times \textcircled{4}^2]^{1/2} = 1.0630$

$\textcircled{9}$ $\tan \rho_{II} = \frac{\textcircled{5}}{\textcircled{8}} e^{j[-\textcircled{6}+\textcircled{7}]} = 1.5419 e^{j(83.728°)}$

$\textcircled{10}$ $$Ce^{j\phi} = \frac{1}{\tanh \rho_I \tanh \rho_{II}} = 4.0208 e^{j(-0.558°)}$$

$\textcircled{11}$ $$\tanh \gamma_2 d = \left[\frac{\tanh \rho_I}{\tanh \rho_{II}}\right]^{1/2} = 0.3234 e^{j(-83.449°)}$$

$\textcircled{12}$ $A + jB = 0.0334 + j5.972$

$\textcircled{13}$ $1/\lambda_c^2 = 0.0110$

$\textcircled{14}$ $1/\lambda_g^2 = 0.0168$

$\textcircled{15}$ $$\tan^{-1} F = \tan^{-1}\left[\frac{2AB}{B^2 - A^2 + (2\pi d)^2\, \textcircled{13}}\right] = 0.466°$$

$\textcircled{16}$ $$\delta_e = 1/2\left(2\,\textcircled{15} - \phi - \tan^{-1}\frac{2AB}{B^2 - A^2}\right) = 0.425°$$

$\textcircled{17}$ $\delta_m = \textcircled{15} - \textcircled{16} = 0.041°$

$\textcircled{18}$ $\tan \delta_e = 0.00742$

$\textcircled{19}$ $\tan \delta_m = 0.00072$

$\textcircled{20}$ $$G = \frac{\textcircled{13} + (B^2 - A^2)/(2\pi d)^2}{[\textcircled{13} + \textcircled{14}][1 - \textcircled{18}\,\textcircled{19}]} = 1.4529$$

$\textcircled{21}$ $$C' = \frac{C(1 + \textcircled{19}^2)}{\left[\left\{(1 - \textcircled{18}\,\textcircled{19})\left(1 + \frac{\textcircled{13}}{\textcircled{14}}\right) - \frac{\textcircled{13}}{\textcircled{14}\,\textcircled{20}}\right\}^2 + (\textcircled{18} + \textcircled{19})^2\left(1 + \frac{\textcircled{13}}{\textcircled{14}}\right)^2\right\}\Bigg]^{1/2}} = 3.3368$$

$\textcircled{22}$ $K_e = [\textcircled{20} \times \textcircled{21}]^{1/2} = 2.2018$

$\textcircled{23}$ $$K_m = \left[\frac{\textcircled{20}}{\textcircled{21}}\right]^{1/2} = 0.6599.$$

Loss of Thermionic Emission in Oxide-Coated Cathode Tubes Due to Mechanical Shock*

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Summary—From the data presented here, it will be shown that large variations in the ability of a commercial pentode to maintain emission levels in the presence of shock are observed in groups of the same type from different production lots. Also, marked differences will be observed in a triode type having the same production date but exhausted on different days. It will be noted, too, that a standard A.S.T.M. diode displays only a very minor drop in activity after shock but when the diode has the cathode coating lengthened or a grid is inserted into the structure, the decrease in cathode activity with shock increases. Gas is observed momentarily immediately after shock in the "diodes" in which the grids are inserted. Replacing bumper type micas with smaller ones which do not contact the bulb result in less decrease in emission with shock, and increasing the number of micas in the standard diode has the effect of increasing the loss in activity after shock. In addition, it will be shown that the use of micas of different grades has, we believe, significant differences in the effect of shock on cathode activity.

Introduction

During the past few years a number of rugged receiving tubes have been developed. Part of the requirements for these tubes is that they be able to withstand repeated shock accelerations of five hundred to one thousand times gravity without significant impairment of performance. To simulate service conditions, the U. S. Navy has developed special shock equipment for testing these tubes. It is the purpose of this paper to discuss various experimental results in connection with a little-understood type of failure which occurs quite generally among rugged tubes, that of failure or decrease of thermionic emission from the cathode as a result of shock. Means have been found for reducing the effect, and a theory which can possibly explain the cause is discussed.

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† Receiving Tube Division, Raytheon Manufacturing Co., Newton 58, Mass.

Description of Shock Machine

The essential parts of the shock machine used for all of our tests are shown in Fig. 1. The table on which the tubes are mounted is free to move horizontally on rollers. After the hammer blow, the table is decelerated by hydraulic shock absorbers. The pendulum hammer swings on a two-foot arm, and delivers a horizontal blow to the table at the bottom of its swing. The anvil which the hammer strikes is a flat rectangular-shaped plate at the extreme right end of the table. The hammer head is the massive rectangular steel part with a cylindrical insert, visible immediately to the right of the anvil and below the large sheet metal shield. The intensity of the shock is controlled by presetting the angle to which the hammer is raised for the blow. The table weighs ninety pounds, and the hammer weighs one hundred and ten pounds. The tubes are shown mounted in the four possible orientations, the external shield can having been removed from one of them while taking the photograph. In a regular shock test, tubes are subjected to five blows in each of the four positions.

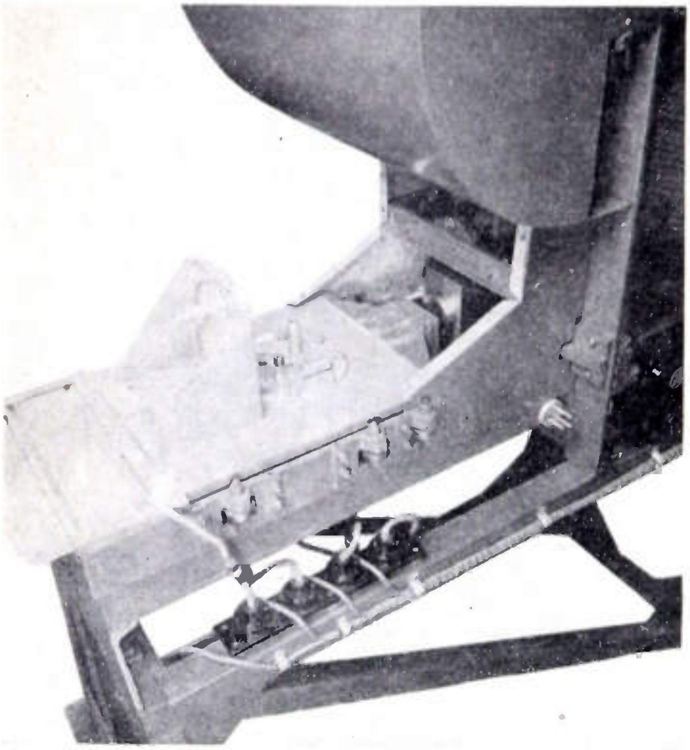

Fig. 1—Shock machine.

Sample Results on Some Commercial Tubes

Many tubes which withstand the higher shock intensities sufficiently well to be tested electrically, show a substantial impairment of emission and associated electrical characteristics as a result of the shock. The cathode coating of the shocked tubes retains its normal appearance, and the amount of distortion of grids or other elements is too small to account for the observed changes. Also, the electrical characteristics changes are too large to be explained by shifting of contact potential. Fig. 2 gives sample data on a commercial pentode and triode. These data are not to be considered as typical of individual lots, but rather as representative of the large range of variation from lot to lot for which no assignable cause has been found. This figure shows mutual conductance, G_m, at various shock levels for two receiving tube types, a triode and a pentode both having indirectly heated cathodes. The shock levels are represented by the angle which the hammer makes with the horizontal. To convert to g's approximately, the number of degrees is multiplied by 15. Curve 1 shows one production date of the pentode, and it had little decrease in G_m after the various shock levels. Another production date of the same tube type shown by curve 2 was extremely poor for G_m after each of the shock levels. All the data in this paper represent averages from at least six tubes.

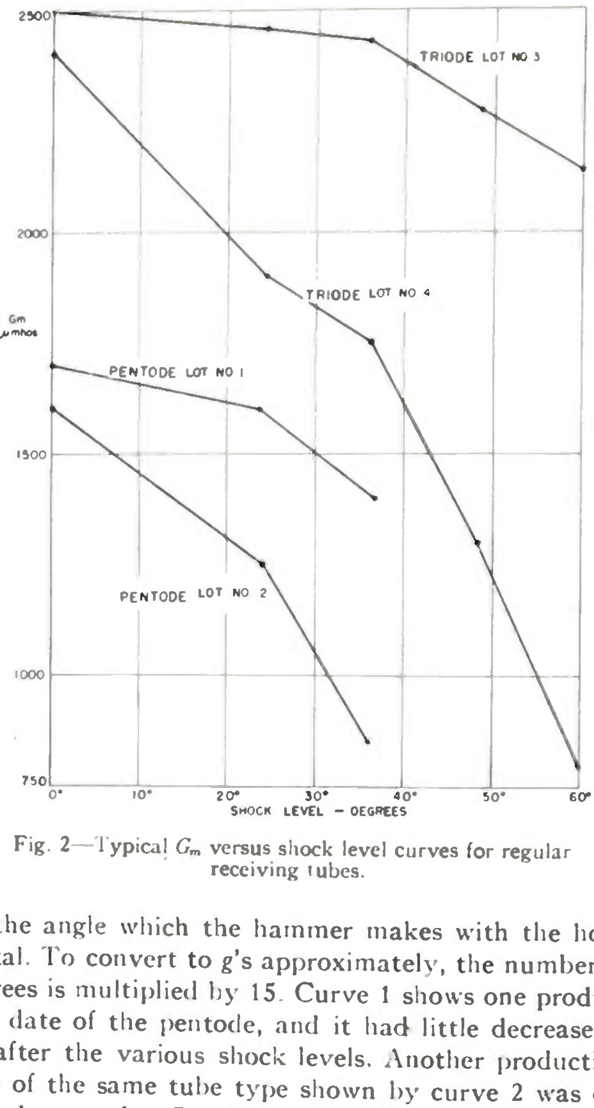

Fig. 2—Typical G_m versus shock level curves for regular receiving tubes.

Curves 3 and 4 show results after shock on two lots of triodes from the same mount production date, but exhausted on different days. Here a remarkably wide difference was obtained between these lots. This would indicate that exhaust processing was an important factor to consider, but further tests covering wide variations in exhaust conditions showed that this variable had little effect on this loss of emission phenomenon. All we could conclude was that here again we had a wide difference for which we could find no assignable cause. These types consistent with usual practice were shocked with the heaters on at rated voltage, but without voltages on the other electrodes. Impairment of emission is especially serious under these conditions. Tubes which are shocked with the heaters cold are generally much better than tubes shocked with the heaters hot, and, in most cases, tubes shocked with full operating voltages are better than those shocked with heater voltage only. In a comparison test, a group of triodes shocked at 48° level with full operating voltages suffered a 25 per cent drop in emission, while a group of triodes from the same lot shocked with heaters cold did not drop significantly. If tubes which have suffered loss of emission after shock are operated at regular life conditions for a long period of time, the emission loss is partially restored. When some of the tubes were operated with heater voltage much higher than normal, the emission loss was restored to even a greater degree and in less time. The data showing these phenomena are tabulated below:

TABLE 1
SPECIAL MULTI-SPACER TUBES[1]

	Per Cent Drop of Original Readings After 48° Shock Level	Per Cent Drop After Being Operated at Regular Life Conditions—24 hours	Per Cent Drop After Being Operated at 100 Per Cent Ef Overload for 1 Minute, then Regular Life at 30 Per Cent Overload for 1 Hour
Lot 1	49	29 per cent	
Lot 2	48		14 per cent

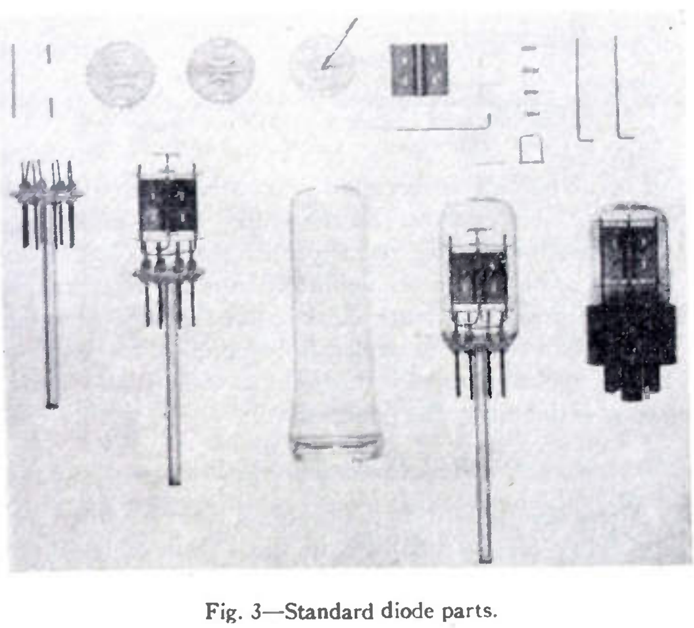

Fig. 3—Standard diode parts.

RESULTS USING STANDARD DIODE

To minimize production variables, most of the tests were conducted on a special diode. This tube is the so-called industry type standard diode as specified for testing tube materials by Section "A" of the A.S.T.M.[2]

[1] Same tubes as those described in Fig. 5.
[2] R. L. McCormack, "A standard diode for electron tube oxide coated cathode core material approval," PROC. I.R.E., vol. 37, pp. 683–687; June, 1949.

Fig. 3 shows the various parts and completed assembly of the standard diode structure which was used. It was felt that more precise results could be obtained from this diode than from the running of tests on commercial tubes through the factory. With this standard diode, laboratory and engineering control could be exercised over the processing and introduction of the specific variables we were interested in evaluating. The button stem replaces the regular flat-press or in-line stem for greater mechanical ruggedness.

Fig. 4, the top curve, shows the emission characteris-

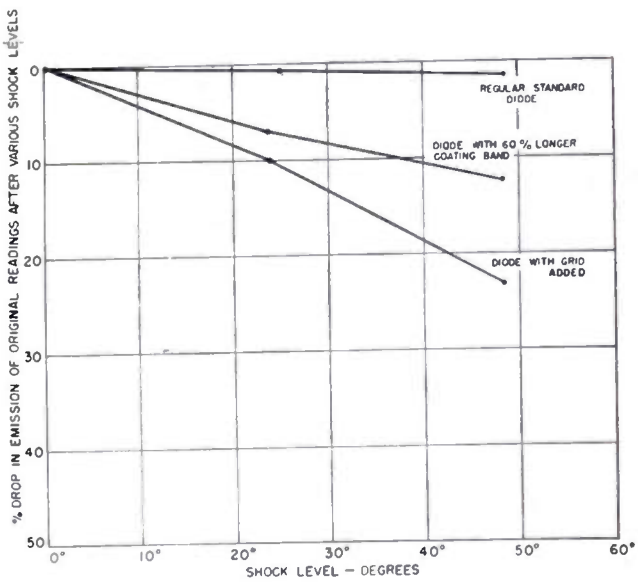

FIG. 4—Per cent drop in emission versus shock level for regular diodes, diodes with long cathode coating band, and diode with addition of a grid.

tics after shock on our standard diode. The coating covers only the middle 50 per cent of the part of the cathode between the spacers. Below this is shown the curve of the standard diode with a 60 per cent longer cathode coating band. This is more nearly the length of the coated band in a commercial tube. The very bottom curve shows the after shock emission characteristics of the standard diode with a grid added between the plate and the cathode. It is seen that the regular standard diode emission curve is relatively flat through the various degrees of shock level, while the other two curves have a significant drop.

This characteristic of the standard diode to resist loss in emission after impact was shown in more than a dozen tests, and we felt it to be highly significant. Any attempt to complicate the original simple diode structure resulted in considerably more loss of emission after impact. One important feature was noted in the laboratory triodes which had been constructed. After shock these tubes were tested on a standard gas test. When the tube was first inserted, the gas meter showed a sudden rush of current of the order of 5 to 8 microamperes, which immediately subsided to a negligible reading of less than 1 microampere. This would seem to indicate the presence of some surge gas, which was quickly cleaned up by being absorbed by the cathode or getter, or some other tube element.

Of course, on this structure there were four extra chances for abrasion of metal against mica during shock, because of the grid side rods. It was reasoned that this abrasion released nascent gas which consequently would poison the emission.

Another phenomenon which had frequently been noticed after shock was a fraying of the resilient tabs and edges of the mica spacers, where they had struck against the glass bulb. It has been felt that during this impact considerable gas is liberated from the mica spacers. On one commercial type triode a change was made from a large bumper point mica spacer to a small spacer which could not touch the bulb during shock. This made a considerable improvement in the emission characteristics after shock, thus lending more credence to the gas-from-mica theory.

To investigate this further, a comprehensive study was made of the effect of the quantity of mica spacers on this loss of emission phenomenon. If this were a gas effect, then one would anticipate more trouble as the number of micas increased.

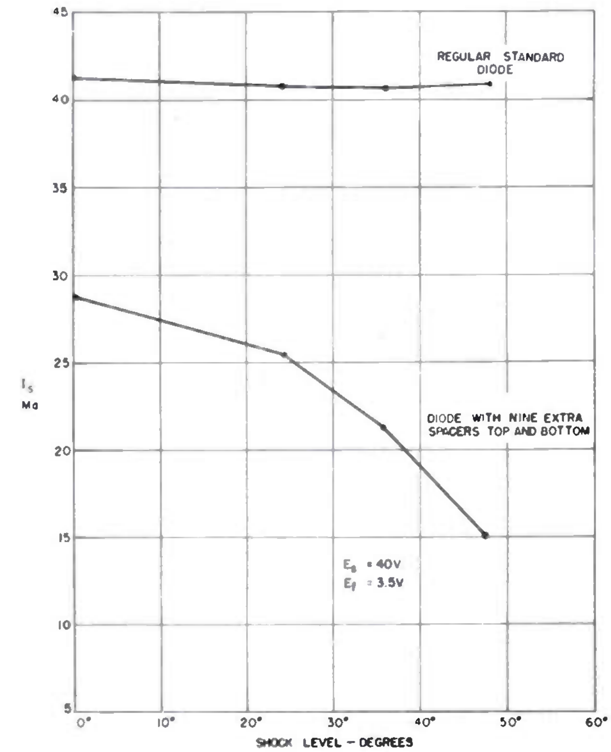

FIG. 5—Emission versus shock level for single spacer diodes and diodes containing nine extra spacers top and bottom.

Fig. 5 shows plots of two lots of diodes. The emission after shock characteristics on the top curve shows the control standard diode, while the curve below this shows the results on a diode made with nine extra mica spacers top and botton, everything else in the two tubes being similar. These additional spacers had the cathode holes reamed out so that they did not touch the cathode.

These results are remarkably significant. The extra spacer tubes, in addition to being considerably poorer initially, resulted in 50 per cent drop in emission after 48° shock level, while the regular diode with single spacers showed virtually no drop. These results obtained on the effect of mica quantity led us to investigate micas from various sources. Fig. 6 shows the results obtained

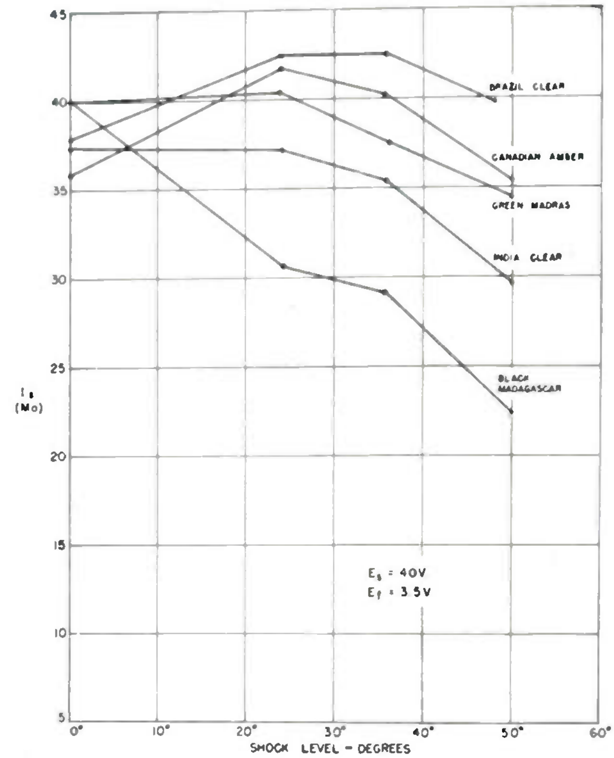

FIG. 6—Emission versus shock level for various lots of diodes having mica spacers from different sources.

on regular single spacer diodes made with five different types of mica as shown on the graph. These and subsequent results indicate that the mica quality is important in that some grades are definitely inferior. However, in regular production runs, clear stained grades were not better than the grade currently being used. In this test, Brazilian clear stained is the best lot, while Black Madagascar is definitely the poorest.

CONCLUSION

These results lead to the conclusion that, while there are many factors involved in the effects of shock on cathode activity, mica is perhaps the most important variable. One possible explanation for this is that during shock, considerable gas is evolved from the mica spacers, causing impairment of emission. Gas has been observed which substantiates the theory. While this suggests means for minimizing this loss of emission effect, there is still much work to be done to solve the problem completely.

ACKNOWLEDGMENT

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Use of Thoriated-Tungsten Filaments in High-Power Transmitting Tubes*

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Summary—**Experience gained in field and laboratory tests over the past ten years has shown that thoriated-tungsten filaments can be used in high-power transmitting tubes with life and reliability equal to that of the low-power types. This achievement can be attributed to improved materials such as oxygen-free high-conductivity copper and platinum-clad molybdenum, improved processing, and better exhaust facilities. Better appreciation of the mechanical problems and of the relationship between the carbide content, operating conditions, and life have all been important factors.**

I. Introduction

THIS paper describes the development and use of thoriated-tungsten filaments in high-power transmitting tubes, such as the RCA-5671, RCA-5770, and RCA-5771. Tubes operating with a dc plate voltage above 5,000 volts are considered to be in the high-power class.

Much has been written on the theoretical aspects of electron emission from thoriated-tungsten filaments since Langmuir[1] discovered that the emission from a tungsten filament containing 1 to 2 per cent of thorium oxide was much higher than that obtained from pure tungsten. Theoretically, the emission efficiency of a thoriated-tungsten filament is 70 to 100 milliamperes per watt of heating power, as compared with 4.25 to 8.5 milliamperes per watt for pure tungsten in the operating temperature ranges of 1,950 to 2,000 degrees K and 2,500 to 2,600 degrees K, respectively. Use has been made of this feature in the design of low-power transmitting tubes, and a long history of reliable performance has been built up over the last quarter of a century. In 1925, Warner and Pike[2] described their work on low-power types, such as the UV210, UV211, UV204A, and UV851. Since then, many more low-power types have been made available commercially to satisfy the ever-expanding application requirements.

As might be expected, many attempts have been made to use thoriated-tungsten filaments in high-power tubes, the earliest being about 1926, as far as the author is aware, but without any reported success until about 1945.[3] This early experience gave rise to the general belief that such filaments could not be used in tubes operating above 5,000 volts because of deactivation due to high-speed ion bombardment. Statements in recent literature,[4] and in books on vacuum tubes published as late as 1948,[5] indicate that this belief is still generally accepted.

Developments in radar since 1937 stimulated further work with thoriated-tungsten filaments, and their use in high-peak-power pulse-oscillator tubes during the last war has been described in a number of papers.[6–10] A recent article[11] describes the use of thoriated-tungsten filaments in high-voltage rectifiers for X-ray service.

II. Initial Work

A critical review of thoriated-tungsten-filament theory and design, early in 1940, indicated no fundamental reason why such filaments could not be used in high-power tubes, provided adequate attention was paid to mechanical strength and processing.

The RCA-207 type was selected for the initial tests. Thoriated-tungsten was used for the filament instead of pure tungsten, and barium-type getters were incorporated to assure a suitably low pressure. Two of these tubes operated in a 10- to 18-mc telegraph transmitter at a plate voltage of 10 kv for 7,953 hours before failure of one tube due to low emission. Throughout life the grid currents were very sensitive to filament input, the best performance being obtained at 517 watts instead of the usual 1,144 watts, a reduction of 55 per cent.

III. Design Considerations

A. Filament

The filament-design factors presented in this paper are based on a compromise between successful structures using pure-tungsten filaments, field experience with low-power thoriated-filament tubes, and the results of numerous life tests run at various temperatures. No attempt was made to obtain the optimum balance between power saving, mechanical strength, and life.

* Decimal classification: R331. Original manuscript received by the Institute, May 19, 1950; revised manuscript received, December 20, 1951.
† Tube Department, RCA, Lancaster, Pa.

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[2] J. C. Warner and O. W. Pike, "The application of the *x-I* filament to power tubes," Proc. I.R.E., vol. 13, pp. 589–609; 1925.

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[4] O. W. Pike, "Cathode design," *Communications*, vol. 21, pp. 5–8, 28; October, 1941.

[5] K. R. Spangenburg, "Vacuum Tubes," McGraw-Hill Book Co. Inc., New York, N. Y.; 1948.

[6] H. A. Zahl, J. E. Gorham, and Glenn F. Rouse, "A vacuum-contained push-pull triode transmitter," Proc. I.R.E., vol. 34, pp. 66W–69W; February, 1946.

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[9] I. E. Mouromtseff, "A quarter century of electronics," *Elec. Eng.*, vol. 66, pp. 171–177; February, 1947.

[10] J. E. Gorham, "Electron tubes in world war II," Proc. I.R.E., vol. 35, pp. 295–301; March, 1947.

[11] Z. J. Atlee, "Thoriated tungsten filaments in rectifiers," *Elec. Eng.*, vol. 68, no. 10, p. 863; October, 1949.

The self-supported type of multistrand structure shown in Fig. 1 was chosen because of its simplicity and reliable performance both in this country and in England.[12] The strands are connected in a series-parallel arrangement so that the heating current flows in opposite directions in adjacent strands. This effects the best balance of the magnetic forces. In addition, it permits alternate strands to be held in mechanical alignment by means of tie-wire assemblies located at any suitable number of horizontal planes. These may be seen in Fig. 1(a). The same wire size was used as for the pure-tungsten filament to assure adequate mechanical strength, and incidentally, to permit the use of available jigs and tools. The result is a simple filament structure having no springs, sliding members, center supports, or guide insulators.

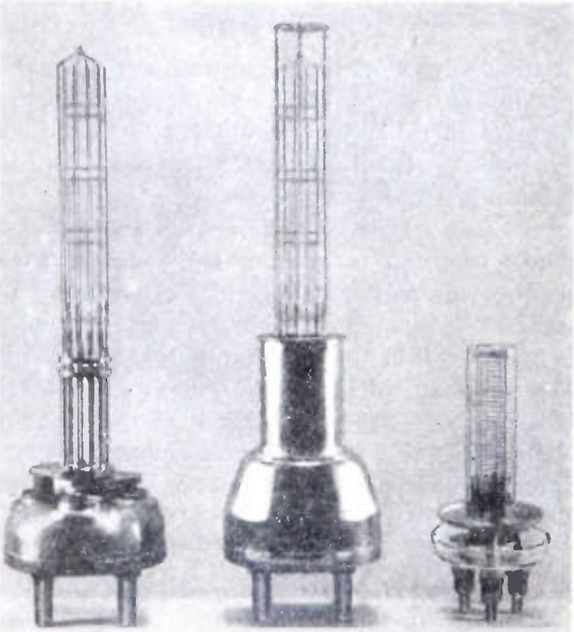


Fig. 1(a–c)—Self-supported, multistrand, thoriated-tungsten filament assemblies (with and without grid assemblies).

In order to realize the full emission capabilities, a thoriated-tungsten filament must be flashed at approximately 2,800 degrees K for a brief period to reduce some of the thorium oxide to metallic thorium. The temperature is then reduced to 2,200 degrees K or less for a period of time to permit diffusion of the thorium to the surface where a monatomic layer is formed.[13] The operating temperature (usually 1,950 to 2,000 degrees K) must be such as to maintain a stable balance between the production of thorium atoms and their loss from the wire surface due to evaporation and to ion bombardment.

In practice, a thoriated-tungsten filament is carburized to reduce the rate of thorium evaporation and also to reduce the susceptibility to deactivation by ion bombardment.[14,15] Carburization is accomplished during manufacture by heating the filament to a high temperature in a hydrocarbon vapor or gas, such as benzine, toluene, or acetylene. The heated filament reacts gradually with the carbon in this atmosphere forming a shell of tungsten carbide around a core of tungsten, as shown in the magnified cross section in Fig. 2. This procedure has been described by Horsting.[16] The percentage of the filament cross section converted to tungsten carbide is called the "per cent carbide." During operation, the filament gradually decarburizes, and a point is finally reached where the emission can no longer be maintained.


Fig. 2—Magnified cross section of a carburized, thoriated tungsten filament.

Unfortunately, a carburized filament is considerably more brittle than an uncarburized filament. Thus, while it would be desirable to carburize a filament completely for long life, high strength would dictate no carburization at all. Therefore, a compromise is made whereby only part of the cross section is converted to the carbide. This is one reason for retaining the large wire diameter in the work being described.

Field experience and life tests have indicated that loss

[12] J. Bell, J. W. Davies, and B. S. Gossling, "High-power valves: construction, testing, and operation," *Jour. IEE* (London), vol. 83, no. 500, pp. 176–207; August, 1938.

[13] I. Langmuir, "Thoriated tungsten filaments," *Jour. Frank. Inst.*, vol. 217, no. 5, pp. 543–569; May, 1934.

[14] M. R. Andrews, "The evaporation of thorium from tungsten," *Phys. Rev.*, vol. 33, pp. 454–458; 1929.

[15] S. Dushman, "Thermionic emission," *Rev. Mod. Phys.*, vol. 2, p. 449; 1930.

[16] C. W. Horsting, "Carbide structures in carburized thoriated-tungsten filaments," *Jour. Appl. Phys.*, vol. 18, pp. 95–102; January, 1947.

of carbide and loss of emission are closely related, with operating temperature a very important factor. The carbide loss curve shown in Fig. 3, with temperatures expressed in terms of emission efficiency, was obtained from data taken on tubes of the 50-watt type having 8.5 mil filaments. This curve along with those on Fig. 4,

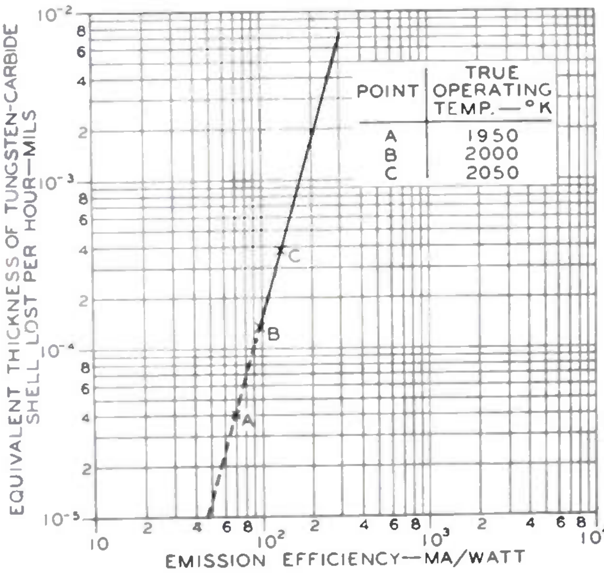


Fig. 3—Carbide loss versus operating temperature.

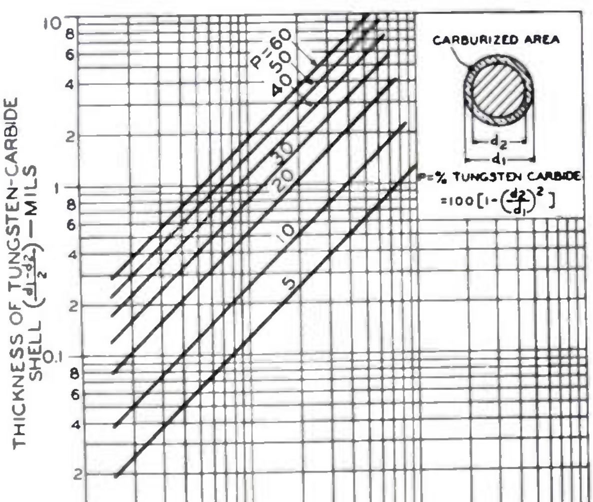


Fig. 4—Carbide shell thickness versus wire diameter and per cent carbide.

showing the relationship between carbide shell thickness, wire size, and per cent carbide, makes it possible to determine the life of any thoriated-tungsten filament. The curves of Fig. 4 were calculated on the assumption that the carbide was distributed uniformly over the wire surface. Use of these curves to predict the emission life of other tube types involving different wire sizes and carbide content have given surprisingly good results considering the structural and processing differences involved. Predictions to date on high-power types have been somewhat conservative. This is rather to be expected in view of the inherent differences between a tube completely enclosed in a glass bulb and one in which the active structure is enclosed in a metal envelope. As more experience is gained, some corrections may be necessary, but in the meantime these data are presented as additions to existing design information on thoriated-tungsten filaments.[17–19]

B. Grid

Merely changing the filament material and reducing its operating temperature by 500 to 600 degrees K is not sufficient to produce good thoriated-filament high-power tubes. Any material which enhances the emission from the filament would act similarly on the grid surface. Thus, the problem becomes one of finding a material or surface treatment which will withstand the processing temperatures and will be difficult to activate with thorium present. Some methods of attack on this problem have been described in the literature.[10,20] Early work at RCA indicated that a platinum surface was most desirable for high-power tube use. The mechanical weakness and high cost of pure platinum was overcome by using a molybdenum wire clad with a very thin layer of platinum. While this material is still quite expensive, the fact that it can be welded readily and cleaned by chemical and electrochemical means, that it does not oxidize, and that it has proven reliable over a long period of time has justified its continued use.

C. Anode

Little can be said in the way of comparison of anode materials, as only OFHC (oxygen-free high-conductivity) copper was used in all of these high-power tubes. This material has been used extensively in the electronic-tube industry for many years. Undoubtedly, its purity has contributed to the successful use of thoriated-tungsten filaments in high-power transmitting tubes.

D. Getter

Although several tubes (still operating after many thousands of hours of service) were made without any getter other than that provided by well-processed tube elements, particularly the anode, it was deemed advis-

[17] B. T. Barnes, "Properties of carbonized tungsten," *Jour. Phys. Chem.*, vol. 33, pp. 688–691; 1929.
[18] H. A. Jones and I. Langmuir, "The characteristics of tungsten filaments as a function of temperature," *Gen. Elec. Rev.*, vol. 30, pp. 310–319; June, 1927.
[19] H. J. Dailey, "Designing thoriated tungsten filaments," *Electronics*, vol. 21, pp. 107–109; January, 1948.
[20] H. E. Sorg and G. A. Becker, "Grid emission in vacuum tubes," *Electronics*, vol. 18, pp. 104–109; July, 1945.

able to provide additional means of assuring a high degree of vacuum. A small sheet of zirconium suitably located within the tube structure and heated by filament radiation was used for this purpose. The gettering properties of zirconium at elevated temperatures have been described in the literature.[21]

E. Processing

Proper processing is of utmost importance in building reliable vacuum tubes. Adequate cleanliness requires complete degreasing of parts and appropriate chemical and electrolytic cleaning, followed by complete removal of all traces of cleaning solutions. Uniform filament carburization requires suitable control of gas flow, hydrocarbon content, temperatures, and shielding. Care is required at final sealing to minimize oxidation of the metal parts. High-temperature outgassing of tube parts and limitation of maximum gas pressures are important factors in the exhaust process. The final pressure with the tube hot should be of the order of 10^{-5} to 10^{-6} mm of mercury.

IV. Types Developed

Development work on higher-power tubes with thoriated-tungsten filaments, started in 1942 and continued after the war, culminated in the commercial announcement of the RCA-5671, RCA-5770, and RCA-5771 types. Representative ratings are shown in Table I.

V. Advantages of the Thoriated-Tungsten Filament

The advantages to be gained by using thoriated-tungsten filaments in high-power transmitting tubes may be enumerated as follows:

(a) The 60- to 70-per cent reduction in heating power results in lower costs for filament transformers, controls, and other components as well as lower operating costs. A comparison of the thoriated-tungsten and equivalent pure-tungsten filament types is shown in Table II below.

[21] W. M. Raynor, "The use of zirmet for gas absorption," *Foote Prints*, vol. 18, no. 2, pp. 22–24; 1947.

TABLE II

Tube Type	Filament				Power Saving	
	Material	Volts	Amperes	Kilowatts	Kilowatts	Per Cent
9C21, 9C22	W	19.5	415	8.1	—	—
5770, 5671	Th W	11.0	285	3.1	5.0	61.7
880	W	12.6	320	4.0	—	—
5771	Th W	7.5	170	1.3	2.7	67.5

(b) The reduction in glass and seal temperatures provides greater safety factor for high-frequency operation and makes possible the elimination of forced-air cooling in certain low-frequency cases.

(c) The narrower temperature range covered in the heating and cooling cycle tends to simplify the mechanical problems.

(d) The hum level in broadcast transmitters is reduced as much as 8 db.

(e) There is less compression of the plate characteristics in the high positive-grid region.

(f) The anode dissipation can be increased by an amount equal to the reduction in filament input.

(g) The decreased number of filament leads on some types results in a lower input capacitance.

VI. Operating Experience

Several years of field experience has proven that thoriated-tungsten filaments are in every respect as reliable in high-power transmitting tubes as in the low-power types. Over 30,000 hours of trouble-free operation has been obtained in 50-kw broadcast service, with no indications of imminent failure. Operation in various industrial applications at plate voltages up to 17 kv and at frequencies up to 13 mc continues equally successful after more than 22,500 hours of service.

VII. Acknowledgments

Progress in any development is made possible by the combined efforts of many people. It is, therefore, with great pleasure that the author acknowledges the help and encouragement of the many engineers and technicians at RCA and in the field who have contributed to the success of this development.

TABLE I
Tabulation of Maximum CCS Ratings for Typical Services

Type	Class of Service	Frequency (mc)	DC Plate Voltage (volts)	Plate Current (amp)	Grid Current (amp)	Plate Input (kw)	Plate Dissipation (kw)	Typical Performance Power Output (kw)
RCA-5671	Class C Telegraphy (Plate-Modulated)	up to 10	12,500	4.5	—	55.0	17.0	40.0
RCA-5770	Class C Telegraphy	up to 20	17,000	9.0	1.25	150.0	50.0	105.0
		(1.6 to 25)	12,500	6.0	0.8	60.0	22.5	44.0
RCA-5771	Class C Telegraphy	(below 1.6)	15,000	6.0	0.8	67.5	22.5	53.0

Signal Distortion by Directional Broadcast Antennas*

CLIFFORD H. MOULTON†, ASSOCIATE, IRE

Summary—Directional broadcast antenna systems are inherently capable of producing signal distortion. One type of distortion, described by Doherty,[1] involves the frequency sensitivity of the antenna input impedance and relationships between this impedance and the transmitter and transmission lines. A second type of distortion results from the directional radiation characteristics of the array, and is caused by differences in the response conditions of a directional array for the carrier and each sideband frequency.

Sources of Signal Distortion

The waveform of the modulation envelope is responsible for the receiver audio-signal waveform in a double sideband AM system. The modulation envelope must therefore remain unchanged during transmission if signal distortion is to be avoided. Any change in the relative phase or amplitude of a signal component may result in a change in the waveform of the modulation envelope, which may cause audio distortion.

The directional pattern of an array is a function of the transmitted frequency, and is necessarily different at the sideband frequencies from that at the carrier frequency. These radiation-pattern differences cause changes in the amplitudes and relative phases of the signal components arriving at the receiving point and hence changes in the modulation envelope. The effect of altering the phases or amplitudes of the signal components in any particular manner may be determined by adding the components vectorially and obtaining the distorted modulation envelope.

Certain combinations of phase shifts and amplitudes result in large amounts of modulation-envelope distortion. One such condition occurs in directions where the high-frequency sideband amplitudes are increased with respect to the carrier amplitude. When the transmitter is modulated 100 per cent with a high-frequency tone, the sideband amplitudes are then greater than required for 100-per cent modulation of the carrier, producing a type of overmodulation.

At low audio modulating frequencies the response conditions for the carrier and sideband frequency components are essentially identical. With high audio modulating frequencies and relatively low carrier frequencies, however, the antenna bandwidth may be sufficiently low to result in severe changes in the relative phases and amplitudes of the signal components. In antenna-pattern null directions the strong carrier-frequency radiation fields of the antennas may almost completely cancel, but the cancellation may rapidly approach reinforcement for a sideband component as its frequency is removed from the carrier frequency. Carrier-frequency null directions are therefore most likely to be accompanied by increased high-frequency sideband power. In directions of maximum carrier power the converse conditions are likely to occur, resulting in reduced high-frequency sideband power. The audio-frequency response characteristics of a directional array will therefore be a function of the receiving direction.

It is significant that although audio distortion components may be found in the receiver output when modulation-envelope distortion exists the array itself does not introduce new frequency components in the transmitted signal but merely alters the amplitudes or phases of existing components. The receiver second detector is responsible for the addition of the large number of distortion components found in the receiver output.

Experimental Measurements

The simplest method of evaluating the audio distortion and frequency-response changes produced by directional broadcast arrays appears to be by direct measurement rather than by calculations or by the use of scale models. Two standard broadcast stations with directional antenna systems were therefore selected as test stations. One of these, station A, employed a two-tower 550-kc array with shunt feed. The other, station B, employed a three-tower 1,280-kc array with series feed.

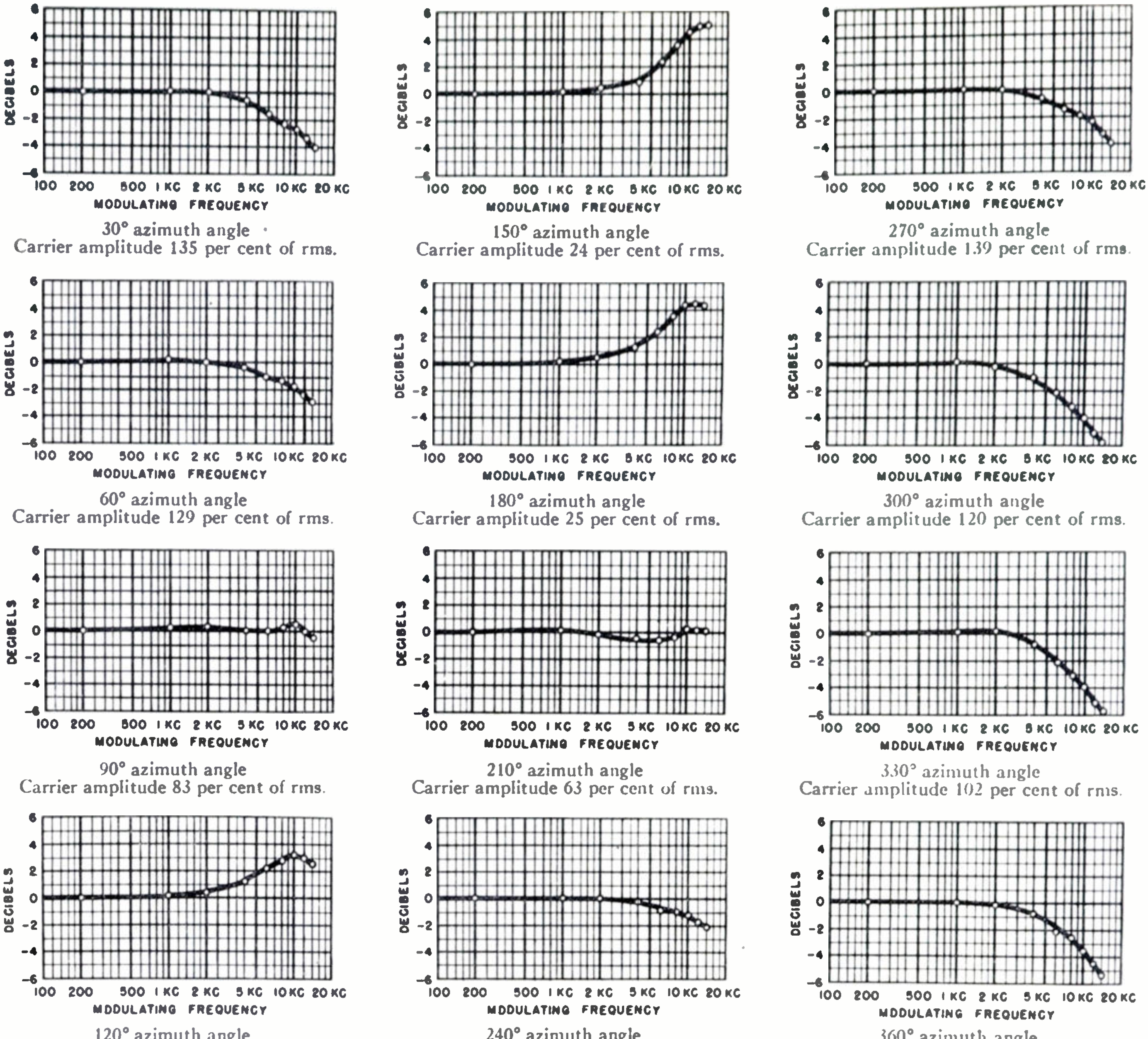
The signal distortion and frequency-response data which follow were obtained during the test period from 1 to 6 A.M. A battery amplifier with a whip antenna and ground rod was carried into completely open spaces. The received test signal was amplified at carrier frequency by broad-band amplifiers, and sent to a mobile unit by coaxial cable. After further amplification at the truck, the signal was then distributed to a Tektronix model 511-AD oscilloscope for photographing waveforms and to a Hewlett Packard model 330-B noise and distortion meter for distortion- and frequency-response measurements. Receiving locations were chosen outside of the antenna induction-field regions and at random distances from the antenna systems.

Frequency-response data are for 50-per cent modulation at the transmitter. This modulation percentage allows a few decibels of sideband power increase before the point of severe distortion, and yet prevents noise and interference from becoming too objectionable. Two-

* Decimal classification: R326.4×R148.11. Original manuscript received by the Institute, January 22, 1951; revised manuscript received, October 1, 1951.

† Tektronix, Inc., P. O. Box 831, Portland 7, Ore.

[1] W. H. Doherty, "Operation of AM broadcast transmitters into sharply tuned antenna systems," Proc. I.R.E., vol. 37, p. 729; July, 1949.



30° azimuth angle
Carrier amplitude 135 per cent of rms.

150° azimuth angle
Carrier amplitude 24 per cent of rms.

270° azimuth angle
Carrier amplitude 139 per cent of rms.

60° azimuth angle
Carrier amplitude 129 per cent of rms.

180° azimuth angle
Carrier amplitude 25 per cent of rms.

300° azimuth angle
Carrier amplitude 120 per cent of rms.

90° azimuth angle
Carrier amplitude 83 per cent of rms.

210° azimuth angle
Carrier amplitude 63 per cent of rms.

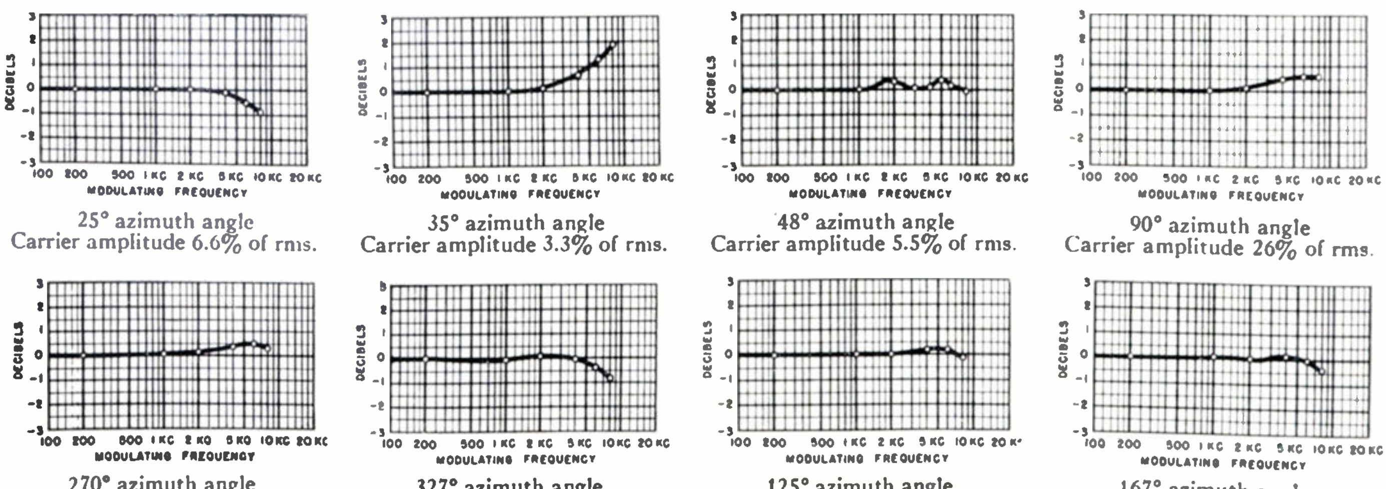
330° azimuth angle
Carrier amplitude 102 per cent of rms.

120° azimuth angle
Carrier amplitude 33 per cent of rms.

240° azimuth angle
Carrier amplitude 117 per cent of rms.

360° azimuth angle
Carrier amplitude 112 per cent of rms.

Fig. 1—Audio-frequency response curves for station A at various azimuth angles. Data are for 50-per cent modulation at the transmitter and random distances from the antenna system.



25° azimuth angle
Carrier amplitude 6.6% of rms.

35° azimuth angle
Carrier amplitude 3.3% of rms.

48° azimuth angle
Carrier amplitude 5.5% of rms.

90° azimuth angle
Carrier amplitude 26% of rms.

270° azimuth angle
Carrier amplitude 161% of rms.

327° azimuth angle
Carrier amplitude 132% of rms.

125° azimuth angle
Carrier amplitude 12% of rms.

167° azimuth angle
Carrier amplitude 21% of rms.

Fig. 2—Audio-frequency response curves for station B at various azimuth angles. Data are for 50-per cent modulation at the transmitter and random distances from the antenna system.

kc, square-wave audio modulation was employed for a series of waveform photographs at various receiving locations because the phases and amplitudes of a large number of audio-frequency components could be observed simultaneously. Test data at station B were limited to 8 kc because of modulator overload.

Results

Audio-frequency response curves at various azimuth angles are shown for station A and station B in Figs. 1 and 2, respectively. The high-frequency audio response is greater in the null directions than in the maximum directions in both cases. The zero decibel level for each direction is taken as the received audio level at 200 cycles for 50-per cent modulation at the transmitter output. The differences in the response curves at station B are much less than at station A, partly because station B operates at a carrier frequency 2.33 times as high as station A. The audio spectrum represents a much smaller percentage of the carrier frequency at 1280 kc than at 550 kc. It is necessary to correct the data of station B to 550-kc conditions, in order to compare them directly with the data of station A, by dividing the modulating frequencies of station B by 2.33. Tests at station B extend to 8 kc, which becomes 3.44 kc when corrected.

The maximum spread in response curves, Δ, expressed in decibels, is plotted as a function of audio modulating frequency for both stations in Fig. 3. Another curve of corrected Δ for station B is given in Fig. 3 after correcting the audio frequencies to 550 kc. From the latter curve it is seen that Δ would be higher for station B than

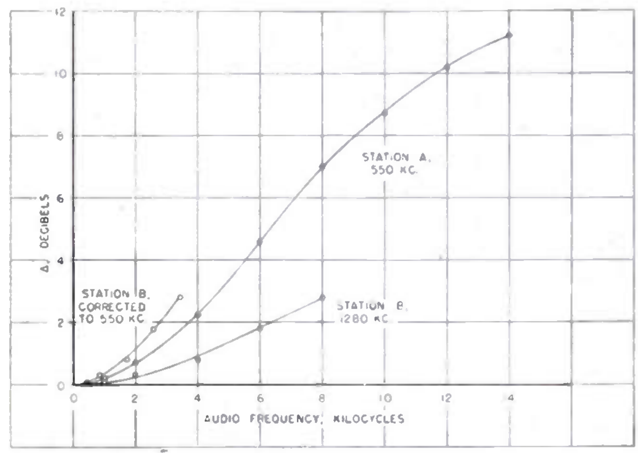
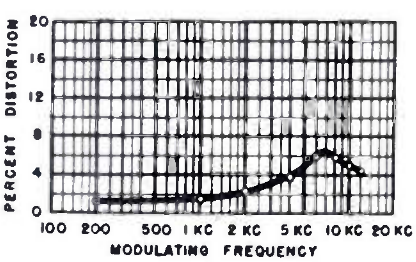
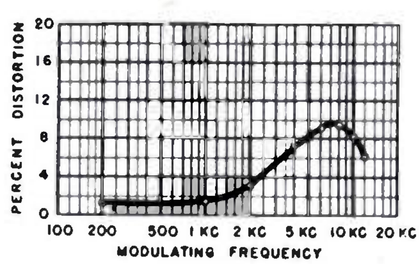


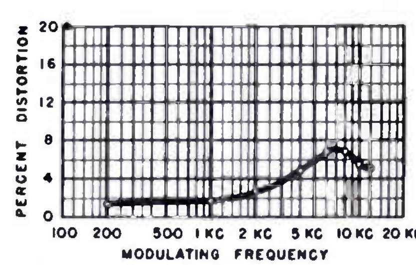
Fig. 3—Maximum response-curve differences, Δ, versus modulating frequency for 50-per cent modulation at the test stations.



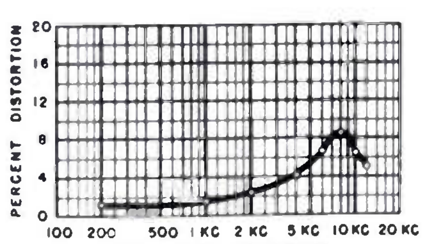
30° azimuth angle
Carrier amplitude 135 per cent of rms.



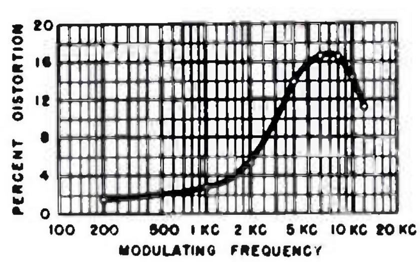
135° azimuth angle
Carrier amplitude 24 per cent of rms.



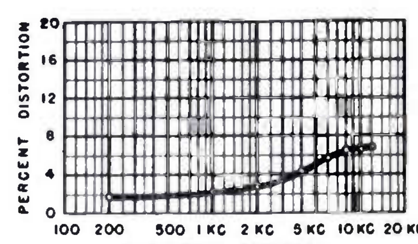
255° azimuth angle
Carrier amplitude 124 per cent of rms.



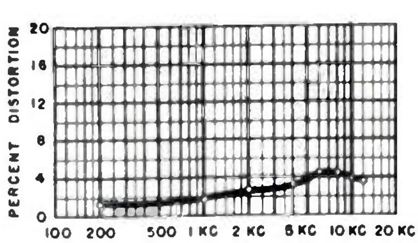
75° azimuth angle
Carrier amplitude 110 per cent of rms.



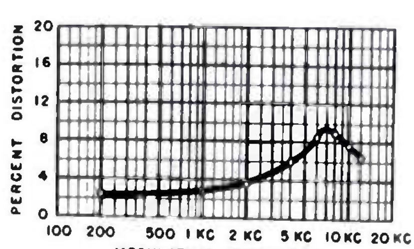
165° azimuth angle
Carrier amplitude 24 per cent of rms.



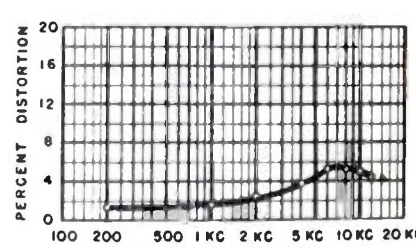
300° azimuth angle
Carrier amplitude 120 per cent of rms.



120° azimuth angle
Carrier amplitude 33 per cent of rms.



210° azimuth angle
Carrier amplitude 63 per cent of rms.



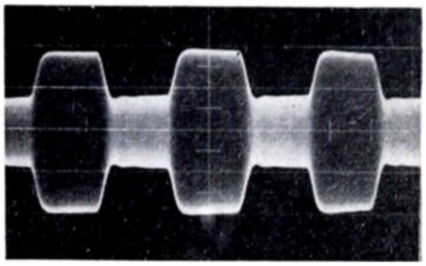
345° azimuth angle
Carrier amplitude 104 per cent of rms.

Fig. 4—Station A audio-frequency harmonic distortion as a function of modulating frequency at various azimuth angles. The transmitter was modulated 50 per cent.

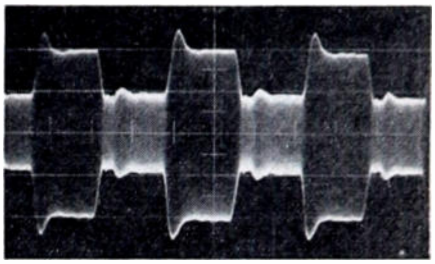
station A if both were to be scaled to operate at 550 kc.

Harmonic distortion of the received signal with 50-per cent sinusoidal modulation at the transmitter output is shown in Fig. 4 as a function of modulating frequency and azimuth angle at station A. The highest distortion percentages are found in the null directions. Similar data could not be taken at station B because of interference and noise in the null directions.

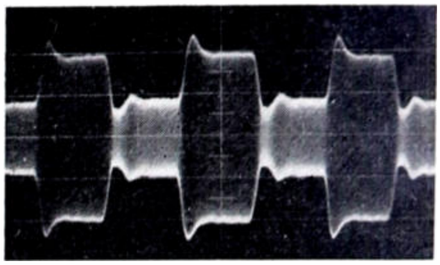
The responses of both stations to square-wave audio modulation are seen in Figs. 5 and 6. The waveforms of station A are markedly different in the null and maximum directions. The slow rises of the waveforms in the maximum directions show reduced high-frequency sideband power while the leading- and trailing-edge overshoots in the null directions show increased high-frequency sidebands. The peculiar waveform in the leading



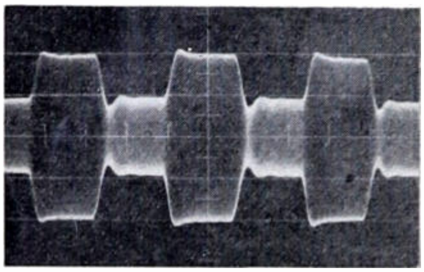
45° azimuth angle
Carrier amplitude 139 per cent of rms.



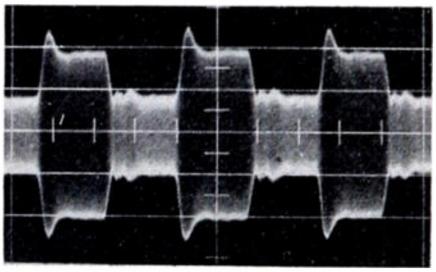
135° azimuth angle
Carrier amplitude 24 per cent of rms.



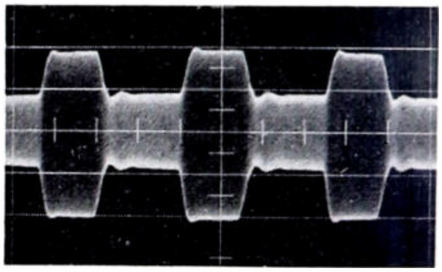
180° azimuth angle
Carrier amplitude 25 per cent of rms.



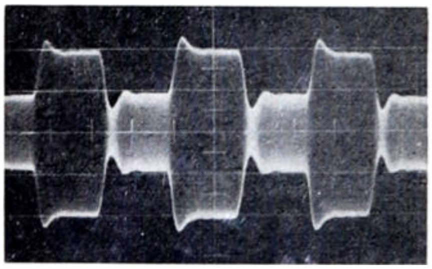
90° azimuth angle
Carrier amplitude 83 per cent of rms.



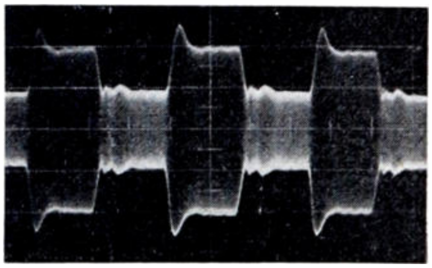
150° azimuth angle
Carrier amplitude 24 per cent of rms.



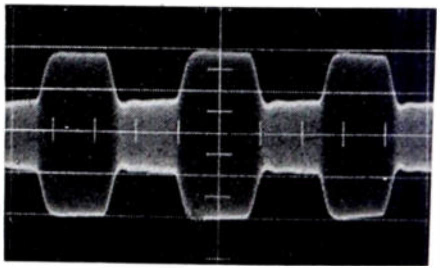
210° azimuth angle
Carrier amplitude 63 per cent of rms.



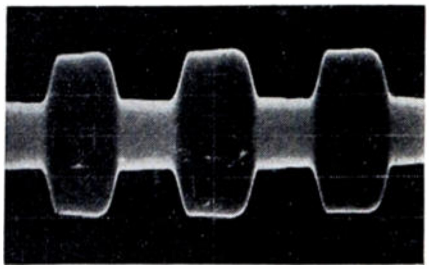
120° azimuth angle
Carrier amplitude 33 per cent of rms.



165° azimuth angle
Carrier amplitude 24 per cent of rms.

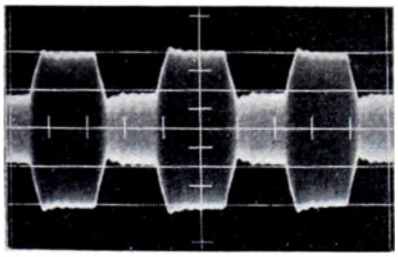


270° azimuth angle
Carrier amplitude 139 per cent of rms

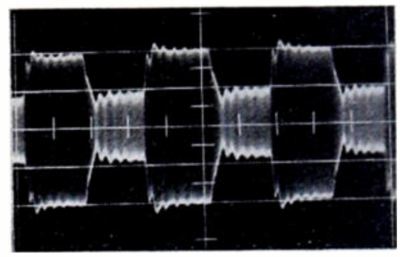


345° azimuth angle
Carrier amplitude 104 per cent of rms.

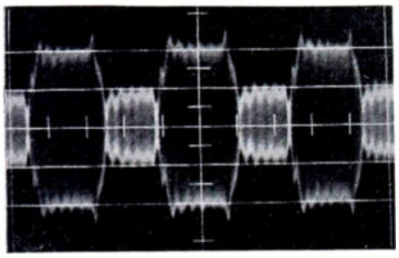
Fig. 5—Photographs of station A modulation-envelope oscilloscope traces at various azimuth angles for 2-kc square-wave modulation. The transmitter was modulated 50 per cent.



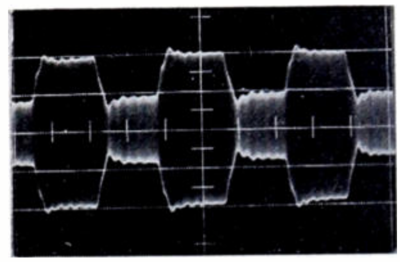
25° azimuth angle
Carrier amplitude 6.6 per cent of rms.



48° azimuth angle
Carrier amplitude 5.5 per cent of rms.

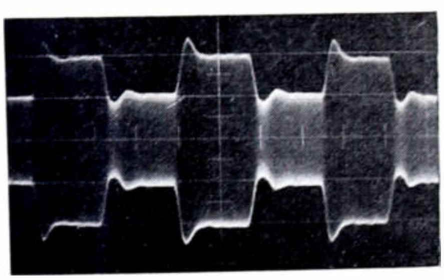


35° azimuth angle
Carrier amplitude 3.3 per cent of rms.

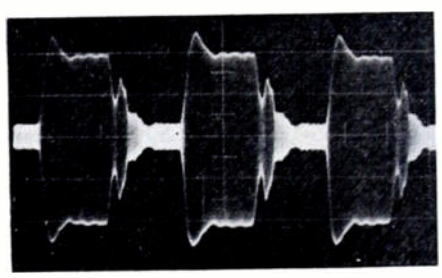


270° azimuth angle
Carrier amplitude 161 per cent of rms.

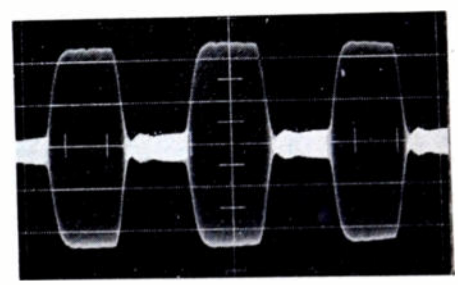
Fig. 6—Station B square-wave modulation photographs for various azimuth angles. The transmitter was modulated 50 per cent with a 2-kc square wave.



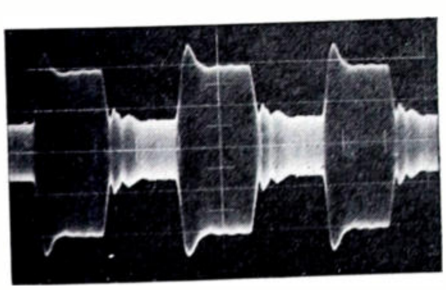
(a)
Modulation 40 per cent.



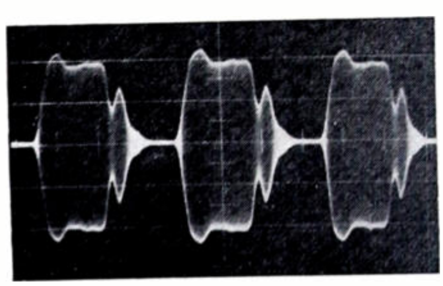
(c)
Modulation 85 per cent.



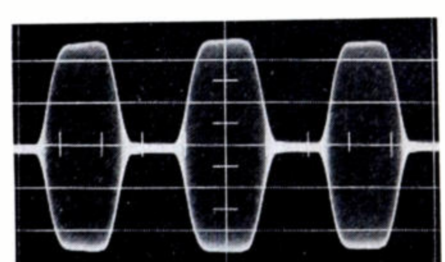
(e)
Modulation 85 per cent.



(b)
Modulation 60 per cent.



(d)
Modulation 98 per cent.



(f)
Modulation 95 per cent.

Fig. 7(a–d)—Station A square-wave response for various modulation percentages at an azimuth angle of 135° (carrier amplitude 23.8 per cent of rms).
Fig. 7(e and f)—Same at an azimuth angle of 255° (carrier amplitude 133.5 per cent of rms).

edge of the downward modulation half of the square wave in the null region of station A should be noted. The damped oscillation on each waveform of station B in Fig. 6 was present with a dummy antenna and was produced in the transmitter audio system. Noise and interference are responsible for the fuzziness of the null direction waveforms.

Fig. 7 shows the effect of modulation percentage on the transient response characteristics of station A for two receiving directions. A given percentage of modula-

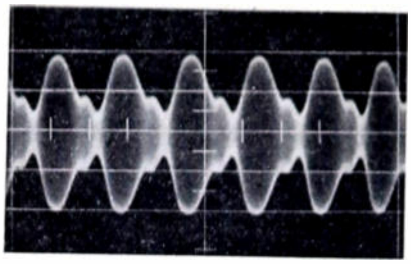
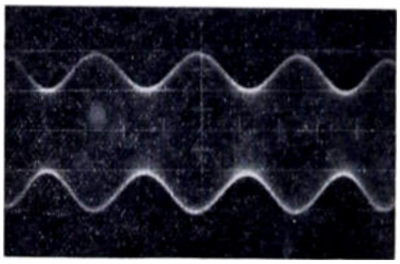
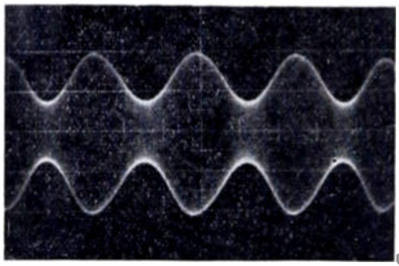


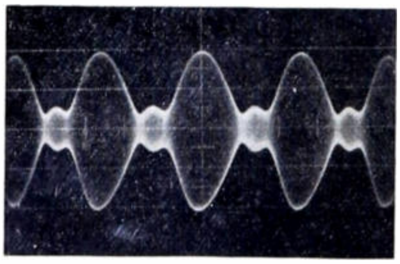
Fig. 8—Modulation envelope waveform at station A for 70-per cent modulation at 4 kc at an azimuth angle of 165° (carrier amplitude 23.8 per cent of rms).



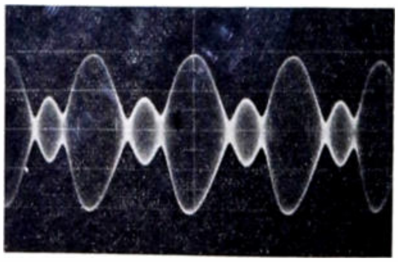
Transmitter modulation 20 per cent.



Transmitter modulation 40 per cent.



Transmitter modulation 60 per cent.



Transmitter modulation 80 per cent.

Fig. 9—Station A modulation-envelope photographs for 10-kc sinusoidal modulation at various modulation percentages. The receiving location azimuth angle was 135° (carrier amplitude 23.8 per cent of rms).

tion results in entirely different modulation envelopes in the two receiving directions. High percentages of modulation are accompanied by severe distortion in the null region.

An example of serious distortion in the null direction at 4 kc is shown in Fig. 8. Analysis of this waveform shows that the two sidebands are not of the same amplitude and have been shifted slightly in time phase with respect to the carrier. Waveforms of the type shown in Fig. 9 occur when the sidebands have similar amplitudes and are shifted from the normal phase position. The effect of modulation percentage on the modulation envelope of station A in the null direction with 10-kc sinusoidal modulation is seen from Fig. 9.

Discussion

Arrays with many elements, nigh-Q tuning networks, negative power elements, and deep nulls are more likely to have severe antenna distortion than simpler, lower Q systems. Stations operating at the low-frequency end of the broadcast band are much more subject to this distortion than those operating at the high-frequency end of the band.

Deep nulls should be avoided from a standpoint of signal distortion if service is to be rendered in the null directions.

Directional signal distortion in the horizontal plane would not be expected for single-element vertical antennas.

While this article has dealt principally with transmitting antenna signal distortion at broadcast frequencies, it is apparent that somewhat similar effects should also be expected for receiving antennas, for other radio frequencies, and for other types of modulation.

Conclusions

1. Signal distortion is observed in directional broadcast antennas, and is found to be a function of receiving direction.
2. Signal distortion results from changes produced by the directional antenna system in the magnitudes or relative phases of the signal components.
3. Directional signal distortion is accentuated by deep nulls, low-percentage antenna bandwidth, high audio-modulating frequency, and high percentage of modulation.

Acknowledgment

The author wishes to thank the co-operating radio stations for the use of their facilities, Professor Grant S. Feikert of Oregon State College for his guidance, and the many students and staff members of the Oregon State College Physics and Electrical Engineering Departments who assisted in this project.

Efficiency of Thermal Electron Emission*

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Summary—**The efficiency of thermal electron emission is defined as the total motional power imparted to emitted electrons over the total heating power of a cathode. Under the assumption that power loss is by radiation only, this efficiency is calculated theoretically and its optimal value determined. Theoretical results are compared with practical values, and hints are given pointing to obtaining a better efficiency.**

Introduction

Usually, with thermal electron emission, the emission current per unit of heating power is given for several types of cathodes. A better figure of merit is, in the author's opinion, given by the power imparted to the emitted electrons in relation to the heating power. *This power ratio will be called "efficiency of thermal electron emission."*

At an absolute temperature of T degrees Kelvin, the mean energy of the electrons emitted normally to a plane emission surface is kT, k being Boltzmann's constant (1.38×10^{-23} Joule per degree Kelvin). The corresponding energy at two mutually perpendicular directions both pointing tangentially to the emission surface is $kT/2$ each. Hence the total energy of emitted electrons is $2kT$, corresponding to a voltage $2kT/e$ where e is the amount of the electron charge (1.60×10^{-19} Coulomb). Hence, at an emission current of I amps, the power imparted to the emitted electrons amounts to

$$P_e = \frac{2kT}{e} I \text{ watts.} \qquad (1)$$

This is based on a Maxwellian velocity distribution of emitted electrons and a cosine angular distribution. The ratio of this power P_e to the heating power P_h of the cathode is the efficiency η of thermal emission

$$\eta = \frac{P_e}{P_h}. \qquad (2)$$

This efficiency will now be considered in several cases.

Ideal Cathode

If an emission cathode is of sufficient length in one direction, its supports being at the ends, the heating power P_h will be mainly spent for (a) power radiation, and (b) electron emission. The fraction of P_h spent for heat losses by conduction at the cathode's supports will, in this case, be relatively negligible. We shall now consider this ideal cathode and calculate its efficiency from the well-known equations of power radiation and of electron emission.

The power radiation per unit (cm²) of cathode surface is given by

$$P_r = c_n T^n; \qquad n \geqq 4. \qquad (3)$$

The coefficient c_n and the exponent n in this equation depend on the type of cathode surface. In the ideal case of so-called "black-body radiation," we have $n=4$ according to Stefan-Boltsmann's law and $c_4=5.75\times10^{-12}$ watts/cm² (degrees Kelvin).[4] Radiation of known cathode surfaces is considerably lower than this black-body radiation at every attainable temperature T. By the second law of thermodynamics, it cannot be higher for any surface under consideration. However, for a limited range of temperature, n may be >4, although $c_nT^n < c_4T^4$. Hence, if we consider black-body radiation, the corresponding efficiency will be *lower* than the practical figures for existing cathodes.

The electron-emission current I amps per unit (cm²) of cathode surface is, according to Dushman's equation,

$$I = AT^2 \exp\left(-\frac{e\phi}{kT}\right). \qquad (4)$$

The constant A is equal to 120 by theory in the case of metals, although measured values differ by as much as an order of magnitude. For nonmetals one may find values of A between 10^{-2} and 10^2. The value ϕ (volts) is characteristic of the emitter under consideration and may vary between about one and 5 volts with practical cathodes. The power $P_e=2IkT/e$ may be evaluated from (4).

The efficiency of thermal-electron emission is

$$\eta = \frac{P_e}{P_e + P_r} = \frac{1}{1 + P_r/P_e} = \frac{1}{1+r}; \qquad r = \frac{c_nT^n}{2\frac{k}{e}AT^3\exp\left(-\frac{e\phi}{kT}\right)}. \qquad (5)$$

Optimal Value of Efficiency

If the ratio r of (5) attains a minimum value, the efficiency is optimal. The only variable in the expression of r, for a given cathode surface, is the temperature T. Hence we put

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$$\frac{\partial r}{\partial T} = 0 = c_n(n-3)T^{n-4} - c_n T^{n-3}\frac{e\phi}{kT^2}.$$

From this equation we obtain the temperature T corresponding to a minimum value of r

$$kT = \frac{e\phi}{n-3}. \qquad (6)$$

By evaluation of the second differential quotient of r with respect to T, it is easily verified that this optimal temperature T according to (6) corresponds to a *minimum* value of r and, hence, to a maximum value of the efficiency η. Insertion of (6) into (5) yields

$$r_{\min} = \frac{c_n\phi^{n-3}\left(\frac{e}{k}\right)^{n-2}}{2.1(n-3)^{n-3}\exp(3-n)}. \qquad (7)$$

In the case of black-body radiation, if $n=4$ and $A=120$, we obtain

$$\underset{n=4}{r_{\min}} = 0.88 \times 10^{-5}\,\phi.$$

It is obvious that this value corresponds substantially to an efficiency $\eta=1$ if ϕ is in the usual range between one and 5 volts. Unfortunately, the temperature $T=e\phi/k$ cannot be attained for any known emitter without its destruction by evaporation. Hence this maximum value of η is only of theoretical interest.

Application to Practical Emitters

The value $r_{\min}$ corresponds to a lower temperature T, at equal ϕ, if n is higher. If $n=13$, and $\phi=1$, we would obtain a temperature T_{opt} corresponding to $r_{\min}$ of

$$T_{\text{opt}} = \frac{e}{10k} = 1.16 \times 10^3 \text{ degrees Kelvin,}$$

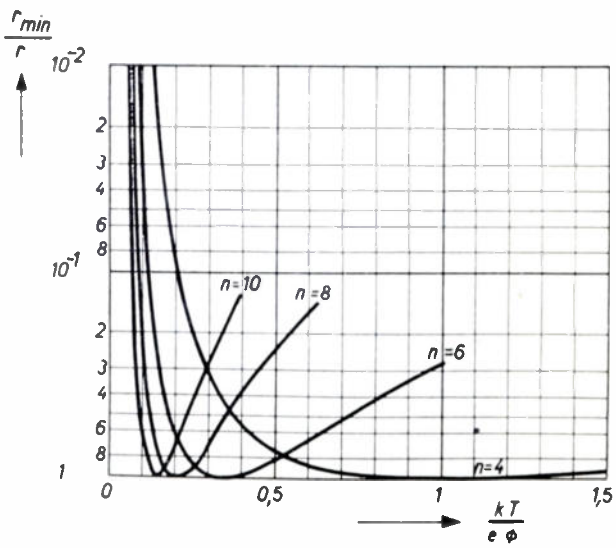

Fig. 1—Ratio $r_{\min}/r$ according to (8) as dependent on $kT/e\phi$, k being Boltzmann's constant, T the cathode's temperature in degrees Kelvin, e the electron charge, and ϕ the cathode's work potential.

which would, in some cases, be attainable without undue evaporation at the cathode's surface. It is obvious that such surfaces corresponding to high values of n would offer advantages from the point of view of efficiency of thermal electron emission.

In Figs. 1 and 2, the value $r_{\min}/r$ is shown as dependent on $kT/e\phi$ for $n=4$, 6, 8, and 10. The corresponding equation is

$$\frac{r_{\min}}{r} = \frac{\left(\frac{e\phi}{kT}\right)^{n-3}\exp\left(-\frac{e\phi}{kT}\right)}{(n-3)^{n-3}\exp(3-n)}. \qquad (8)$$

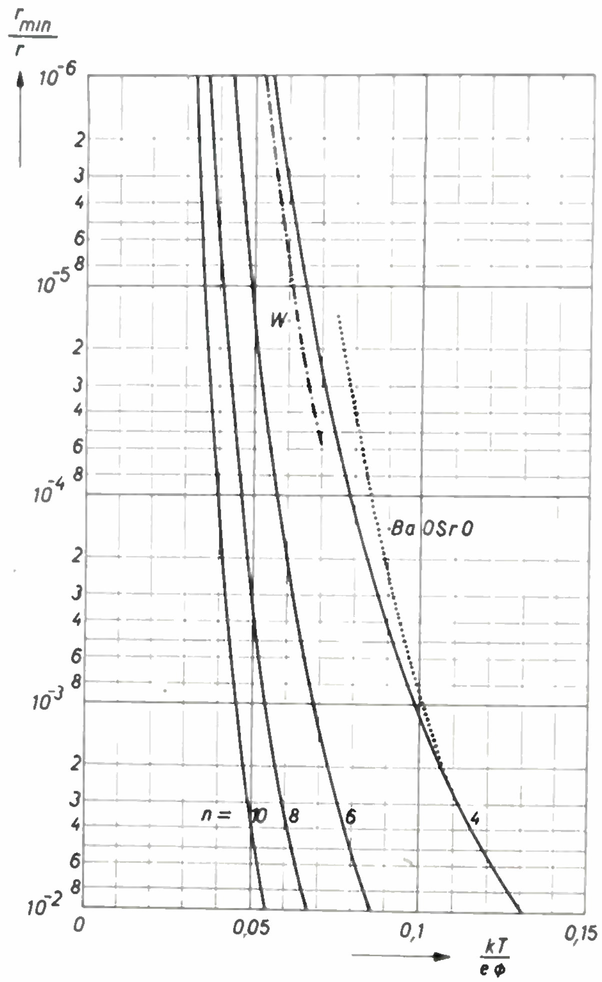

Fig. 2—Continuation of Fig. 1. The dot-dash curve corresponds to a tungsten cathode ($n \sim 4.75$) and the dotted curve to a barium-oxide, strontium-oxide coated platinum-iridium cathode.

The value of $r_{\min}$, according to (7), is shown in Fig. 3 as dependent on A and on ϕ in the case $n=4$. Moreover, in Fig. 2 a dotted curve corresponding to a barium-oxide-strontium-oxide cathode on a platinum-iridium base metal is shown. Herewith, A was given the value 10^{-2} and ϕ the value one volt. The dot-dash curve of Fig. 2 corresponds to tungsten ($A=100$, $\phi=4.54$ volts). The n values are approximately 3.98 in the dotted curve and approximately 4.75 in the dot-dash curve. For both

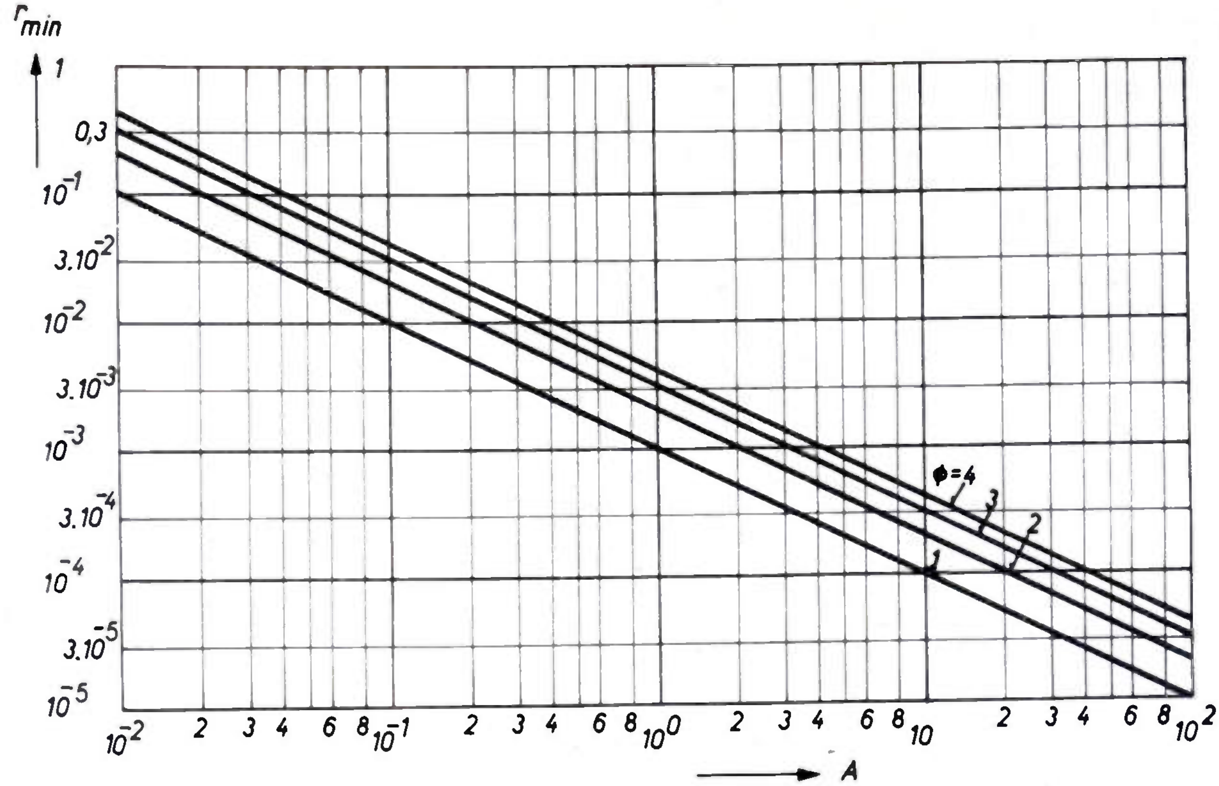

Fig. 3—Value of r_{min} according to (7) as dependent on A and on ϕ in the case $n=4$ (black-body radiation.)

curves, experimental values were used.[1,2] Calculating the efficiency η from these experimental data, we find for the oxide-coated cathode at 1,300 degrees Kelvin $\eta=3.5$ per cent, and for the tungsten cathode at 2,500 degrees Kelvin, $\eta=0.18$ per cent. Therefore, the former cathode has almost 20 times the efficiency of the latter one. The above temperatures were chosen so as to correspond to about equal vapor pressures in both cases.

The following table gives the values of r_{min} and of η_{max} if $A=100$, assuming $c^nT_n=c_4T^4/2$ at $T=1000$ degrees Kelvin.

TABLE I

n	6	6	10	10
ϕ	1	4	1	4
r_{min}	4.9×10^{-5}	3.13×10^{-3}	9.6×10^{-6}	1.58
η_{max}	~1	~1	~1	0.39

[1] H. A. Jones and I. Langmuir, "The characteristics of tungsten filaments as a function of temperature," *Gen. Elec. Rev.*, vol. 30, pp. 310, 319, 354–361, 408–412; 1927.

[2] R. W. King, "Thermionic vacuum tubes and their applications," *Bell Sys. Tech. Jour.*, vol. 2; April, 1923.

Conclusions

If the surface of a cathode radiates heat according to the law c_nT^n, assuming power loss by electron emission and heat radiation only, the maximum efficiency of thermal electron emission would occur at a temperature T corresponding to $T=e\phi/k(n-3)$ in which e is the electron charge, ϕ the work potential of the emitter, and k Boltzmann's constant. The corresponding maximum efficiency would be substantially unity (ratio of power imparted to emitted electrons in relation to heating power of cathode). With practical cathodes, due to n being in the neighborhood of 4, this maximum efficiency cannot be attained, as the cathode's surface would evaporate at the temperatures in question. Efficiency figures for tungsten and for oxide-coated platinum iridium are shown to be at about 1.8×10^{-3} and 3.5×10^{-2}, respectively. Cathode surfaces of much higher values of n at correspondingly favorable values of A (emission constant in Dushman's equation) and ϕ could eventually lead to much enhanced efficiency figures.

Acknowledgment

I wish to thank E. Bielek for his assistance in obtaining the numerical data of this paper.

Correspondence

The Detector as a Factor in Channel Capacity*

Shannon[1] has defined channel capacity as the maximum of the rate of transmission

$$R = H(x) - H_y(x),$$

with respect to all possible information sources used as input to the channel, where $H(x)$ is the information per symbol generated by the source and $H_y(x)$ is the equivocation.

It appears worthwhile to remark that in most realizable systems the rate of transmission depends not only upon the statistics of the signal and of the noise but upon the properties of the detector as well. (The term "detector" is used here to mean the collective apparatus which transforms the received signal into a message in its original form.)

This consideration is of practical interest where there is no great freedom to tamper with the message statistics as with radar, but where, instead, the detector must be tailored to fit the signal and noise statistics if the transmission rate is to be maximized.

In order to show the precise connection between the channel capacity and properties of the detector, let the message consist of discrete symbols $x_i(i=1, 2, \cdots, n)$, the corresponding symbols in the output messages being $y_i(i=1, 2, \cdots, n)$. If the successive symbols in the message are statistically independent with relative frequencies (with respect to the ensemble of messages), $p(x_i)$, then the average information per symbol in the original message is

$$H_x = -\sum p(x_i) \log p(x_i).$$

Because of the presence of noise, the symbol x_i in the transmitter input may produce one of several symbols y_j in the receiver output. Denoting by α_{ij} the conditional probability that x_i in the original message will appear as y_j in the final message, the relative frequencies of the symbol y_i (in the ensemble sense) are

$$p(y_i) = \sum_j p(x_j)\alpha_{ji},$$

while the conditional probabilities that the input symbol was x_j when the output symbol was y_i is

$$\beta_{ij} = \frac{p(x_j)\alpha_{ji}}{\sum_j p(x_j)\alpha_{ji}}$$

by Bayes' formula.

The average information per symbol (the equivocation) contained in a "message" which corrects every mistake in the received message is, accordingly,

$$H_E = -\sum_i p(y_i) \sum_j \beta_{ij} \log \beta_{ij}.$$

When corrected, the output message contains the same information per symbol as the original so that the information per symbol of the uncorrected output message is just

$$R = H_x - H_E.$$

The set of conditional probabilities α_{ij} depends, in fact, upon the behavior of the detector; and by proper design of the detector they can be given values which maximize R.

For definiteness suppose there is one degree of freedom in the received signal for each symbol in the message. Denote the j'th degree of freedom in the signal corresponding to the j'th symbol in the message (in order of transmission) by η_j. Then owing to random interference there will be a set of continuous conditional probability density functions $p(x_i; \eta_j)i=1, 2, \cdots$ for n_j, given that x_i was transmitted. The detector might be so designed that the output would be y_k corresponding to x_k whenever

$$\theta_{k-1} < \eta_j \leq \theta_k,$$

where the θ_k obviously satisfy the inequalities $(\bar{\eta})_k < \theta_k < (\bar{\eta})_{k+1}$ in which

$$(\bar{\eta})_k = \int_{-\infty}^{\infty} d\eta \eta p(x_k; \eta).$$

The values chosen for the θ_k clearly affect the rate of transmission since the set of conditional probabilities α_{ij} are given by

$$\alpha_{ij} = \int_{\theta_{j-1}}^{\theta_j} p(x_i; \eta) d\eta.$$

A maximum of R with respect to the θ_j will exist in most cases of interest.

The following simple, though unrealistic, case will serve as illustration. The message code contains two symbols 0 and 1 with relative frequencies p and $1-p$. The degrees of freedom η_j in the received signal have the normal conditional probability densities

$$p_0(\eta_j) = \frac{1}{\sqrt{2\pi}\,\sigma} e^{-\eta_j^2/2\sigma^2}$$

$$p_1(\eta_j) = \frac{1}{\sqrt{2\pi}\,\sigma} e^{-(\eta_j-u)^2/2\sigma^2}.$$

The symbol 0 appears as the j symbol in the output if $\eta_j < \theta$ and the symbol 1 if $\eta_j \geq \theta$. The ratio R/H_x, the fraction of information in the original message which is preserved during transmission and detection, is plotted as a function of the parameter $\epsilon \equiv \theta(\sqrt{2}\sigma/P)$ for various values of p and $P/\sqrt{2}\sigma$, which might be regarded as the square root of the signal-to-noise power ratio.

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* Received by the Institute. December 3, 1951.

[1] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. XXVII, pp. 407–410; July, 1948.

Television-Image Reproduction by Use of Velocity-Modulation Principles*

This is in answer to the letter from Thomas, published on page 1341 of the October, 1951 issue of the Proceedings, in which he criticizes the paper by Honnell and Prince.[1]

First of all, I wish to point out to Thomas that phase and frequency modulation are special cases of angle modulation, and that they may both be correctly called by that name. By analogy, there are several classes of velocity-modulation cathode-ray displays which we mention under the "General Analysis" in our paper. The criterion by which velocity-modulation reproduction of displays should be judged is the fact that the trace brightness is a function of the spot velocity rather than of the spot intensity. In my opinion, therefore, Thomas' suggestion that we should use the terms "displacement" or "position-change-modulation" is trivial.

Obviously, Thomas misunderstood the reference he cited.[2] Figs. 8.22 and 8.23 of this reference clearly show that the video signal originating at the iconoscope is integrated by the tube T_1 and the condenser C_1 before it is applied to the horizontal deflection plates of both the iconoscope and the kinescope. In the successful velocity-modulation systems cited in our references the video signal is integrated because in the velocity reproduction of a signal voltage the trace velocity is indisputably the derivative of the signal voltage applied to the deflection plates of the kinescope. In fact, in some of these systems the integrated video signal voltage was transmitted to the receiver. (All of these systems, of course, produce a distortion of the video signal voltage in the time domain.)

Thomas should study our equations (6), (7), and (8) more carefully. We state clearly that, "A response of particular importance is obtained when $G(t)$ is the integral of $F(t)$." With this condition imposed on (6), the corresponding velocity equation becomes

$$v = V_0 \pm k \cdot F(t),$$

and we have velocity modulation by Thomas' own definition. In several places in our paper we discuss the effects of signal-voltage integration on the resulting brightness response. We did not emphasize the signal-voltage versus velocity relationship because the spot velocity is only an intermediate step in the more important criterion of the brightness versus signal-voltage relationship.

In spite of Thomas' allegation to the contrary, our Fig. 2 is absolutely correct. It is the velocity reproduction of an integrated square wave (that is, a pyramid signal) superimposed on the horizontal sweep voltage which produces the alternate bright and dark bars of unequal widths as we show clearly in Fig. 3. Furthermore, all of our figures are correct velocity-reproduction displays of the signal voltages we show or describe.

The velocity reproduction we describe in our paper is correct in name and title in accordance with the customary usage of this term by the original investigators cited in the bibliography of the paper.

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* Received by the Institute. November 23, 1951.

[1] M. A. Honnell and M. D. Prince, "Television image reproduction by use of velocity-modulation principles," Proc. I.R.E., vol. 39, pp. 265–268; March, 1951.

[2] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, New York, N. Y., pp. 238–242; 1940.

Correspondence

A Note on Telepathic Communication*

In recent correspondence, Bibbero,[1] Hollmann,[2] and Stockman[3] have referred to telepathic-communication experiments of Cazzamalli, Loefgren, Von Ardenne, and others during the past few years. Assuming the existence of such telepathic waves as proposed, it becomes interesting to speculate on the nature of this telepathic radiation. If this radiation is of the electromagnetic type, it must be of extremely high frequency since the brain cannot be conceived of creating an electromagnetic wave of less than about 1,000 mc per second. An electromagnetic radiation of this frequency could only be detected over distances of line-of-sight. Also, low-energy electromagnetic radiation would be attenuated to negligible proportions by the skull itself. Thus, it seems that an electromagnetic type of telepathic radiation is ruled out.

It appears reasonable from present knowledge of nuclear physics to assume that the telepathic radiation could be corpuscular in nature. Before Young's famous experiment in 1801, proving the wave nature of light, the corpuscular theory of light was predominant. An extremely low mass particle having a very small charge or none at all could be imagined as the source of telepathic radiation. Since its mass is very small as compared to an electron, it would be propelled at nearly the speed of light, with a potential perhaps as recorded by Berger's encephalography which measures the fluctuation potential of the brain.

It appears that there is a particle in nuclear physics which fulfills most of these conditions. This particle is called a "neutrino." It was first postulated by Pauli in 1927 to explain the beta process in nuclear physics. According to Frank,[4] the rest mass of the neutrino is almost certainly less than one tenth that of an electron, and, so far as present measurements can show, it is not appreciably different from zero. In 1948 Curran, Angus, and Cockroft[5] investigated the mass of the neutrino at the University of Glasgow. They decided that on this evidence the neutrino mass must be less than two-thousandths part of that of an electron. Such a mass could be propelled essentially at the speed of light with a potential of perhaps only a few volts.

A neutrino has an extremely long mean free path and an extremely small cross section. Its cross section has been estimated at approximately 10^{-44} square cm. According to the well-known physicist, Enrico Fermi, the small cross section of the neutrino would permit it to cross our sun with little probability of being absorbed. Its mean free path in air has been estimated to be over 100,000 miles. Such a mean free path in solid rock is calculated to correspond to over 1,000 miles. This particle could travel through the earth apparently with only moderate attenuation.

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1712 New Hampshire Ave.
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* Received by the Institute, February 7, 1952.

[1] R. J. Bibbero, "Telepathic communication," PROC. I.R.E., vol. 39, pp. 290–291; March, 1951.
[2] H. H. Hollman, "Telepathic communication," PROC. I.R.E., vol. 39, p. 841; July, 1951.
[3] H. Stockman, "More on telepathic communication," PROC. I.R.E., vol. 39, p. 1571; December, 1951.
[4] F. C. Frank, "The mass of the neutrino," *Proc. Phys. Soc.* (London), vol. 59, p. 408; 1947.
[5] S. C. Curran, J. Angus, and A. L. Cockcroft, *Nature* (London), vol. 162, p. 302; 1948.

Contributors to Proceedings of the I.R.E.

For a photograph and biography of W. J. Albersheim, see page 358 of the March, 1952 issue of the PROCEEDINGS OF THE I.R.E.

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RAYMOND B. AYER

Raymond B. Ayer (A'26–M'40–SM'43) was born on June 29, 1905, in Plainville, Mass. He received the B.E.E. degree from Northeastern University in 1925. After graduation he joined the General Electric Company at Schenectady, N. Y., as a test engineer, transferring later to the research laboratory and, subsequently, to the vacuum tube engineering department. There he was engaged in rectifier and transmitting-tube development work.

In 1936 Mr. Ayer became associated with the engineering section of the tube department of RCA at Harrison, N. J., as a design and development engineer on high-power transmitting tubes. He transferred to the Lancaster, Pa., plant in 1943, when the power tube engineering activity was moved to this new location. Mr. Ayer is currently associated with super-power tube development work in the advanced development laboratory at Lancaster.

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JOHN M. BROOMALL

John M. Broomall (A'47–M'47) was born in Philadelphia, Pa., on August 28, 1916. He received a B.S. degree in electrical engineering from the University of Pennsylvania in 1941, and joined the signal department of the Pennsylvania Railroad upon graduation. At the start of the war he received a commission in the Navy, specializing in fire control and gunnery both ashore and afloat. In 1946, Mr. Bromall joined RCA Victor where he worked on sonar development for two years. This was followed by more than two years work on digital and analogue computers as a member of the staff at the Moore School of the University of Pennsylvania.

Since May of 1951, Mr. Broomall has been with the Brown Instruments Division of Minneapolis-Honeywell, engaged in the development of electrical control systems.

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R. W. CORKUM

Russell W. Corkum was born in East Boston, Mass., on January 6, 1925. He attended the U. S. Naval Academy, and transferred to Boston University where he received the A.B. degree in physics in 1949. He has since done graduate work in physics and mathematics at the latter place.

Mr. Corkum joined the Antenna Laboratory of the Air Force Cambridge Research Center in 1946. He worked in the Ground Antenna Branch and participated in research and development of artificial dielectrics, development of microwave lenses and optical systems using ordinary and artificial dielectrics, and design and testing of ground antennas.

Recently he has been assigned to the Plans Branch of the Electronics Research Division of the Center.

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H. J. DAVIS

Harold J. Davis was born in East Orange, N. J. on May 21, 1909. He received the S.B. in electrical engineering from the Massachusetts Institute of Technology in 1937. Before graduation he worked several years for the Bell Telephone system.

After several years working as a test set construction engineer and radiometric test engineer, he joined the New York Signal Corps Procurement Office. In 1943 Mr. Davis accepted a commission in the Signal Corps and was assigned at dif-

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ferent times as officer in charge at various plants and inspection areas in New Jersey, New England, Los Angeles, and Philadelphia.

In 1946 he joined Sylvania Electric Products, Fluorescent Lamp Division, as a design and quality engineer, and in 1948 he joined the Raytheon Manufacturing Company in Newton, Mass., as parts quality control engineer for the Special Purpose Tube Section.

Mr. Davis is a Senior Member of the American Society for Quality Control and the Boston Society for Quality Control. He holds a reserve commission as a captain, and he is active in an army research and development unit.

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J. Lewis Hathaway (SM'47) was born in Denver, Colorado, and participated in early broadcast engineering in that city. In 1929 he received the B.S. degree in electrical engineering from the University of Colorado and the same year joined the Development Group of the National Broadcasting Company. There he has since been engaged in diversified engineering development in the fields of radio and audio frequency as well as television.

J. L. Hathaway

While on leave of absence from 1941 to 1944, he served as special research associate at Harvard University, performing underwater sound development work.

In 1948 Mr. Hathaway was appointed Assistant Manager of Engineering Development, National Broadcasting Company.

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J. W. Herbstreit (A'40–SM'45) was born in Cincinnati, Ohio on September 7, 1917. He received the E.E. degree from the University of Cincinnati in 1939. From 1935 to 1940 he was associated with the Crosley Corporation in Cincinnati as transmitter engineer for stations WSAI, W8XAL, and 500-kw WLW.

J. W. Herbstreit

In 1940 Mr. Herbstreit joined the Federal Communications Commission National Defense Operations in Marietta, Ga., later acting as radio inspector at the Atlanta office. In 1941 he was transferred to the FCC Engineering Department in Washington, D. C. as radio engineer in the Safety and Special Services Division.

Mr. Herbstreit joined the Operational Research Staff in the Office of the Chief Signal Officer, Department of the Army, in 1942. There he made numerous operational radio-systems studies, including measurements of atmospheric noise levels and the attenuation of radio signals by jungles in Panama and the Southwest Pacific, measurements and analyses of experimental low frequency loran in the Western Hemisphere, and frequency requirements for low-power radio communications and navigation equipment.

In 1946 he joined the Central Radio Propagation Laboratory of the National Bureau of Standards for which he conducted research on cosmic radio noise and vhf and uhf propagation. He served as technical advisor to the International High Frequency Broadcast Conference held in Mexico City in 1948, and was responsible for the preparation of radio-propagation information used by the conference. He served in a similar capacity in 1950 when the second international conference was convened in Florence and Rapallo, Italy. Since 1949 he has been in charge of the tropospheric-propagation research work being conducted at the Bureau, now centered at Boulder, Colo.

Mr. Herbstreit is a member of Tau Beta Pi, Eta Kappa Nu, Phi Eta Sigma, American Association for the Advancement of Science, Engineering Society of Cincinnati, and is a registered professional engineer.

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Donald O. Holland was born in Thornburg, Iowa, on November 28, 1911. He attended the State University of Iowa where he received the B.A. and M.S. degrees in 1934 and 1936, respectively.

D. O. Holland

During 1935 and 1936 Mr. Holland did research work on the mean-free-path of molecules and the growing of zinc single crystals. From 1936 to 1937 he was head of the Physics and Chemistry Department at Upper Iowa University. He served on the staff of the State Geological Survey of Illinois in the physics division from 1937 to 1941. During the early part of World War II he was employed by the Navy Department in the protection of ships against magnetic mines.

Since 1944 Mr. Holland has been at Raytheon Manufacturing Company Receiving Tube Design Dept., where he has served as project engineer on government contracts in the development of ruggedized tubes and subminiature high-wattage tubes.

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Hans Erich Hollmann (A'48) was born on November 4, 1899, in Solingen, Germany. He majored in physics at the Technical University of Darmstadt, Germany, and received his Doctor's degree in 1928. In 1930 he accepted a position with the "Heinrich-Hertz Institute für Schwingungsforschung" in Berlin. From 1934 to 1936 he wrote the first encyclopedia on microwaves and vhf which contains a chapter on the field later to be known as radar. Dr. Hollmann has since occupied himself extensively with studies and research in microwaves, vhf diathermy, and electrocardiography.

H. E. Hollmann

Over a hundred papers have resulted from research in his "Laboratorium für Hochfrequenztechnik und Elektromedizin." The most outstanding have been published in the U. S. and England. During this time he was also consulting scientist for the "Telefunken," "Siemens," and other companies. He is credited with nearly 300 inventions in the fields of microwaves, magnetrons, klystrons, beam-riding, electrocardiography and others. In addition he was the director of the "Forschungsgesellschaft für Funk- und Tonfilmtechnik."

After the war, Dr. Hollmann became professor in charge of the applied physics department at the Friedrich Schiller University in Jena, Germany. Since 1947 he has been consultant at the Naval Air Missile Test Center, Pt. Mugu, California.

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Robert S. Kirby (M'50) was born in Lawrence, Kan., on July 31, 1920. He received the B.S. degree in engineering in 1943 from the U. S. Naval Academy. He served as an officer on Pacific Fleet destroyers from 1943 until 1946, his duties including those of communications officer, navigator, and executive officer afloat.

Robert S. Kirby

In 1947 Mr. Kirby joined the staff of the National Bureau of Standards, and has been engaged in propagation studies at the Central Radio Propagation Laboratory. In 1948 he acted in an advisory capacity on propagation tests of air-to-ground communications, conducted at the Naval Air Test Center, Patuxent River, Md. Since 1951 he has been working with the Laboratory in Boulder, Colo.

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Raymond E. Lafferty (A'41–M'47) was born in Brooklyn, N. Y. on July 12, 1918. He graduated from the RCA Institutes General Engineering School in 1939 and

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R. E. Lafferty

was employed by the Ferris Instrument Corporation in the same year. In 1940 he became associated with the Allen B. Du Mont Laboratories and later that year, accepted the position of chief engineer with radio station WSLB.

In 1942 Mr. Lafferty took a leave of absence in order to teach in the New York State Signal Corps Training School and then joined the Boonton Radio Corporation in 1943 as a project engineer engaged in the development of microwave signal generators. He enlisted in the U. S. Navy in 1944, and in 1946 returned to his position at WSLB.

Mr. Lafferty has been with the Engineering Development Group of the National Broadcasting Company since 1948 as a development engineer. He is a member of Delta Phi Omega and the Radio Club of America.

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Irving E. Levy was born in Malden, Mass., on August 6, 1919. He received the B.S. degree in chemical engineering from Tufts College in 1941. Following graduation he was employed by Raytheon Manufacturing Company as a process and development engineer. During the war he served in the U. S. Navy as assistant navigator on an aircraft carrier, after which he returned to Raytheon to work in the general engineering division. He has been primarily concerned with special projects in connection with the processing and development of special-purpose tubes.

Irving E. Levy

Mr. Levy is an active amateur with the call letters W1SFR. He is a member of the American Society for Quality Control and the American Society for Testing Materials, for which he serves on a subcommittee.

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C. H. Moulton (S'48–A'50) was born in Portland, Ore., on September 6, 1927. He received the B.S. degree in June, 1948 and the M.S. degree in June, 1950 from Oregon State College. He remained at this college as a graduate assistant and later as an instructor in electrical engineering until June, 1951. While at Oregon State he worked on an Air Force research project, "Generation of Millimeter Electromagnetic Waves." He is now a development engineer for Tektronix, Inc.

C. H. Moulton

Mr. Moulton is an associate of the AIEE, and a member of Phi Theta Kappa, Eta Kappa Nu, Sigma Pi Sigma, ASEE, and AAUP.

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K. A. Norton (A'29–M'38–SM'43–F'43) was born on February 27, 1907, in Rockwell City, Iowa. He received the B.S. degree in physics from the University of Chicago in 1928.

K. A. Norton

During 1929, Mr. Norton was with the Western Electric Company. From 1929 to 1930, he was in the radio section of the National Bureau of Standards, and at Columbia University the following year. He was associated with the technical information section of the FCC from 1934 to 1942.

Mr. Norton served from 1942 to 1943 as assistant director of the Operational Research Group in the Office of the Chief Signal Officer, from 1943 to 1944 as a radio and tactical countermeasures analyst in the Operational Research Section of the Eighth Air Force in England, and from 1944 to 1946 as a consultant in the Radio Propagation Section of the O.C.S.O.

Since 1946 Mr. Norton has been in the Central Radio Propagation Laboratory of the National Bureau of Standards as chief of the Frequency Utilization Research Section. He is also assistant chief in charge of Laboratory activities in Boulder, Colo.

Mr. Norton is a Fellow of the American Physical Society and of the American Association for the Advancement of Science, and a member of the American Institute of Electrical Engineers, the American Geophysical Union, the American Mathematical Society, the Institute of Mathematical Statistics, and the American Statistical Association.

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For a photograph and biography of Leon Riebman, see page 700, of June, 1951 issue of Proceedings of the I.R.E.

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Morris Schulkin (A'44) was born in Brooklyn, N. Y., on February 6, 1919. He received the B.A. degree, magna cum laude, in physics and mathematics from Brooklyn College in 1939, and the M.S. degree in physics from the George Washington University in 1948.

From 1940–1941, Mr. Schulkin worked at the United States Weather Bureau on upper air analysis, and from 1941–1945, he was associated with the Ordnance Development Division of the National Bureau of Standards. In 1945, he transferred to microwave propagation research in the Central Radio Propagation Laboratory of NBS, and then joined the Radio Astronomy Group at the Naval Research Laboratory, doing work in atmospheric absorption at radio frequencies.

Morris Schulkin

Mr. Schulkin was also an associate in physics at the George Washington University and an assistant in mathematics at the University of Maryland in 1947–1948, returning to the Central Radio Propagation Laboratory in 1948–1950. In this capacity he worked in co-operation with the FCC Ad Hoc Committee on TV and FM frequency allocation problems.

At present, Mr. Schulkin is head of the Sound Propagation and Oceanography Section of the United States Navy Underwater Sound Laboratory, at New London, Conn. He is a member of Pi Mu Epsilon, Sigma Pi Sigma, Sigma Xi, American Geophysical Union and Acoustical Society of America.

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For a photograph and biography of M. J. O. Strutt, see page 496 of the April 1952 issue of the Proceedings of the I.R.E.

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William C. White (A'15–M'25–F'40) was born on March 24, 1890, in Brooklyn, N. Y. He received the E.E. degree from Columbia University in 1912. He has been associated with the General Electric Company since graduation in a number of capacities in the field of electronics. At present he is electronics engineer of the Research Laboratory.

William C. White

Mr. White is a Fellow of the AIEE, and a member of Sigma Xi and Tau Beta Pi. He was Treasurer of the IRE in 1946, and a Director from 1943 to 1947.

He has served on the following Institute committees, among others: Executive, Papers Procurement, Board of Editors, Awards, Nominations, Membership, and Professional Recognition.

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For a photograph and biography of L. A. Zadeh see page 701 of the June, 1951 issue of the Proceedings of the I.R.E.

Left—IRE Secretary Haraden Pratt (left) greets radio pioneer John R. Binns, (right) who as a wireless operator in 1909, was the first to use radio to prevent a major sea disaster when he summoned aid for his faltering ship.

Right—President D. B. Sinclair (left) receives gavel of office from Past President I. S. Coggeshall at President's Luncheon.

IRE FORTIETH CONVENTION & RADIO GREATEST 29,000

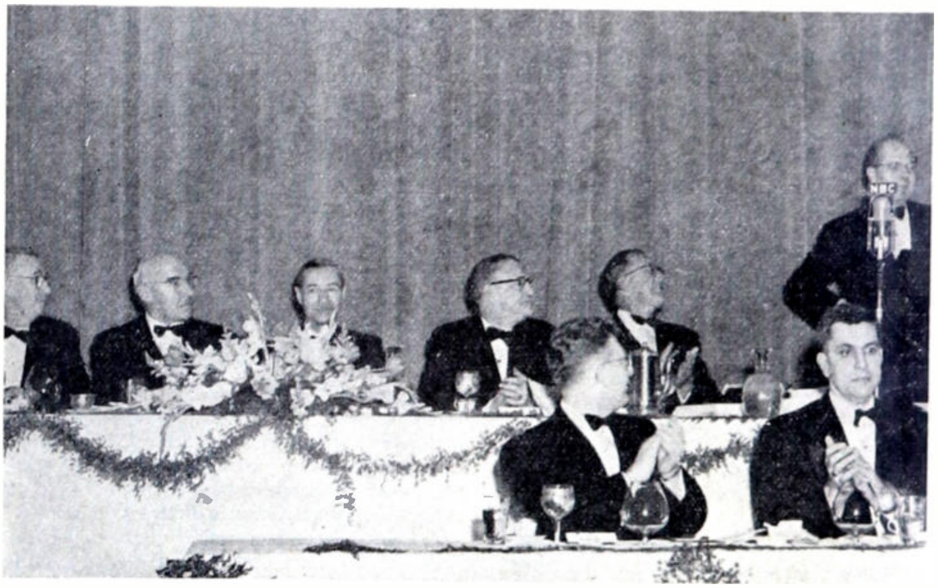
The 1952 IRE National Convention will long be remembered as the most successful conference in the forty-year history of the Institute. During March 3 through 6, 29,000 radio engineers and scientists from the United States and numerous foreign countries registered at the Waldorf-Astoria Hotel and Grand Central Palace in New York City to witness a comprehensive program of 217 technical papers and 356 engineering exhibits, keynoted by the slogan "Forty Years Sets the Pace." The record-shattering registration exceeded the 1951 figure by 6,000 and resulted in full attendance at the 43 technical sessions, the four-floor Radio Engineering Show, and the social events.

The convention opened on Monday morning with the Annual Meeting of the Institute at which IRE officers made their annual reports to the membership. In honor of the fortieth anniversary of the founding of IRE, the meeting featured an interesting and informative colloquy by Alfred N. Goldsmith and John V. L. Hogan, two of the founders of the Institute, entitled "The IRE: From Acorn to Oak," in which they discussed many of the early radio pioneers, the contributions they made to the radio field, and how they came to form The Institute of Radio Engineers.

Technical Sessions

The technical sessions, which were held at the Waldorf-Astoria Hotel, Grand Central Palace, and the nearby Belmont Plaza Hotel, covered advances in almost every phase of endeavor in the radio-electronics field (substantially as was reported in the February issue of the Proceedings). Of the 43 technical sessions, 13 took the form of IRE Professional Group symposia. The program of each symposium was carefully arranged to provide a complete and coordinated discussion of the subject by panels of experts. As can be seen from the list below the symposia covered a broad gamut of subjects of vital current interest to the radio engineer. The list of subjects is as follows: Sub-Audio Instrumentation, Management of Research and Development, New Developments in Telemetering, Television Broadcasting—Audio and Video Systems, Television Station Construction and Theater

SPEAKERS' TABLE AT THE IRE



Back row (L. to R.)—Austin Bailey, Convention vice chairman; R. F. Guy, Senior Past President, IRE; Haraden Pratt, Secretary, IRE; W. R. G. Baker, 1952 Medal of Honor; Charles E. Wilson, guest speaker; W. L. Everitt, toastmaster. *Front row*—IRE Award Winners—Jerome Freedman, H. W. Welch, Jr.

Defense Mobilizer Charles E. Wilson, guest speaker, addresses the IRE Annual Banquet.

William L. Everitt performs the duties of toastmaster for the Convention Banquet.

W. R. G. Baker discusses IRE Professional Groups at the President's Luncheon.

Right—IRE founders Alfred N. Goldsmith (left) and John V. L. Hogan. (right) present an interesting colloquy, "The IRE: From Acorn to Oak," during the recent Annual Meeting of the Institute at the Waldorf Astoria Hotel.

ANNIVERSARY ENGINEERING SHOW EVER HELD ATTEND

Left—Convention Guest Speaker Charles E. Wilson (left) watches D. B. Sinclair present the Medal of Honor to W. R. G. Baker (right)

Conversion, UHF Receivers, Magnetic Core Memory Devices for Digital Computers, The Integration of Electronic Equipment with Airframe Design, What's New in Mobile Radio, Reliability of Military Electronic Equipment, Transistor Circuits, and Digital Computers in Control Systems.

40TH ANNIVERSARY BANQUET

Back row (L. to R.)—D. B. Sinclair, President; I. S. Coggeshall, Junior Past President, IRE; A. N. Goldsmith, Editor, IRE; Glen McDaniel, President, RTMA; R. D. Bennett, spokesman, 1952 Fellows; G. W. Bailey, Convention chairman. *Front row*—IRE Award Winners B. D. Loughlin, William Shockley, Newbern Smith.

The subject of transistors was the highlight of the opening session on Monday afternoon and, in fact, was a high point of the convention. Interest in this session ran so high that it was repeated on Thursday morning. Papers were heard on transistor circuits, followed by a Signal Corps demonstration of a radio teletype converter employing transistors in place of tubes. Sessions were also held on instrumentation, engineering management, information theory, audio, and telemetering.

On Tuesday the predominant subject was television, with two sessions covering broadcasting and station construction, one session on color, and one general session. In addition, a special symposium on "Present Status of NTSC Color Television Standards" was held on Tuesday evening. Papers were also heard on instrumentation, circuit and information theory, medical electronics, and waveguides.

Television continued to share the spotlight on Wednesday with uhf receivers receiving considerable attention. Digital computers was also a prominent topic on the program with two symposia being presented on this important subject. Antennas, circuits, propagation, tubes, and navigation aids rounded out the day's activities.

Thursday's program included a number of interesting symposia on such subjects as reliability of military electronic equipment, mobile radio, and the relation between electronic equipment and airframe design. Television came in for further discussion, both in a symposium on uhf receivers and in sessions on high-frequency and cathode-ray tubes. The four-day program was completed with papers on antennas, feedback control, circuits, computers, and radio communication systems.

EXHIBITS

For the first time, the Radio Engineering Show filled all four floors of Grand Central

(*Cont.* on page 610)

1951 IRE President I. S. Coggeshall gives a final message during the recent Banquet.

Raymond F. Guy discusses uhf television during the Convention press conference.

Ralph D. Bennett responds for the 1952 Fellow Award winners during the Banquet.

Left to right—A. E. Hylas, Allen B. DuMont Labs.; W. V. Tyminski, Allen B. DuMont Labs.; R. A. Varone, Admiral Corp.; J. White, Zenith Radio Corp.; H. W. A. Chalberg, General Electric Co.; T. Murakami, RCA; Lewis Winner, *TV Engineering*; D. D. Israel, Emerson Radio & Phono. Corp.; E. O. Johnson, RCA; W. B. Whalley, Sylvania; A. M. Scandurra, Kollsman Instrument Corp.; M. F. Melvin, P. R. Mallory & Co.; John Bell, Zenith Radio Corp.; Max Beier, Zenith Radio Corp.

Palace where 356 manufacturers displayed $10,000,000 worth of their newest products, making it the largest showing of communications and electronics apparatus ever to be assembled.

The official opening of the Radio Engineering Show on Monday morning was preceded by a press conference which featured the latest developments in uhf television. A special exhibit, which was later put on public display, was on view containing the most recent uhf television equipment to come from the laboratories and factories of 15 firms. The display included uhf transmitting and receiving tubes, transmitting and receiving antennas, converters, tuners, test equipment, and a uhf receiver in actual operation. The display was supplemented by comments from leading experts in the uhf field, including R. F. Guy of NBC and R. W. Davis of Station WELI, New Haven, Conn. The press conference closed with a showing of a film describing RCA's experimental uhf station in Bridgeport, Conn.

Transistors and television, highlights of the technical sessions, were also in evidence at the Radio Engineering Show, with considerable interest focused on a CBS color television demonstration utilizing an RCA color tube. Military applications of electronic equipment were featured in a special Military Radio Exhibit comprising the most recent unclassified developments of 20 firms. Telemetering equipment, computers, audio apparatus and sound demonstration rooms, nuclear instruments, components and tubes, measuring devices of all descriptions—these were representative of only a few of the many types of systems, apparatus, parts, materials, and techniques which went to make up the most impressive and comprehensive exhibition in the radio industry.

SOCIAL EVENTS

The social activities of the convention received a send-off on the first evening of the convention when a "get-together" cocktail party was held in the spacious Grand Ballroom of the Waldorf Astoria. The success of the affair is attested to by the fact that 1,500 members and guests attended, providing an excellent opportunity for visitors from all parts of the country to renew old acquaintances and make new ones.

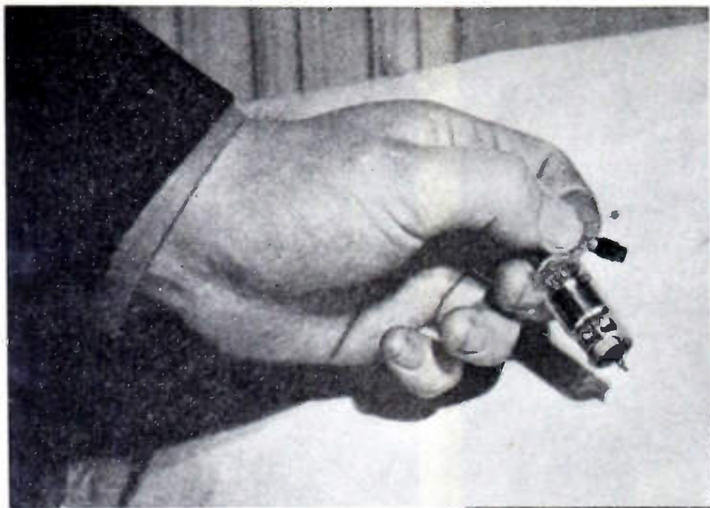
The Starlight Roof of the Waldorf Astoria was the scene of the traditional President's Luncheon on Tuesday. The luncheon featured W. R. G. Baker, Chairman of the Professional Groups Committee, who spoke on "The IRE Professional Group System." Former IRE President I. S. Coggeshall served as toastmaster to introduce his successor in office, President Donald B. Sinclair. An added feature was the presence of John R. Binns, President of Hazeltine Corporation, whose role in 1909 as a wireless operator aboard the steamship *Republic* in the first case where radio effected the saving of many lives at sea, is well known in the annals of radio history.

A capacity audience attended the IRE Fortieth Anniversary Banquet, held in the Grand Ballroom on Wednesday evening, to hear an address by Charles E. Wilson, who spoke in his capacity as Director of Defense Mobilization on the subject "In Strength Is Peace." William L. Everitt, dean of the college of engineering at the University of Illinois, ably performed the duties of toastmaster. The IRE awards for 1952 were presented by President Sinclair to the recipients (see pages 612–617, this issue) with W. R. G. Baker, General Electric Co., receiving the Medal of Honor, the Institute's highest award, and Ralph D. Bennett, U. S. Naval Ordnance Laboratory, responding on behalf of the 45 newly elected Fellows. The Morris Liebmann Memorial Prize was awarded to William Shockley, Bell Telephone Laboratories, and the Harry Diamond Memorial Award was presented to Newbern Smith, National Bureau of Standards. The Vladimir K. Zworykin Television Prize Award, presented this year for the first time, went to B. D. Loughlin, Hazeltine Corp. Awards were also presented for outstanding PROCEEDINGS papers, with H. W. Welch, Jr., University of Michigan, receiving the Browder J. Thompson Memorial Award and Jerome Freedman, Griffis Air Force Base, receiving the Editor's Award.

The social program was rounded out with an entertaining schedule of fashion shows, tours, matinees, and television shows for the wives of members and guests.

FIRST FLOOR EXHIBITS AT RADIO ENGINEERING SHOW, GRAND CENTRAL PALACE

Highlights of Annual Convention and Engineering Show



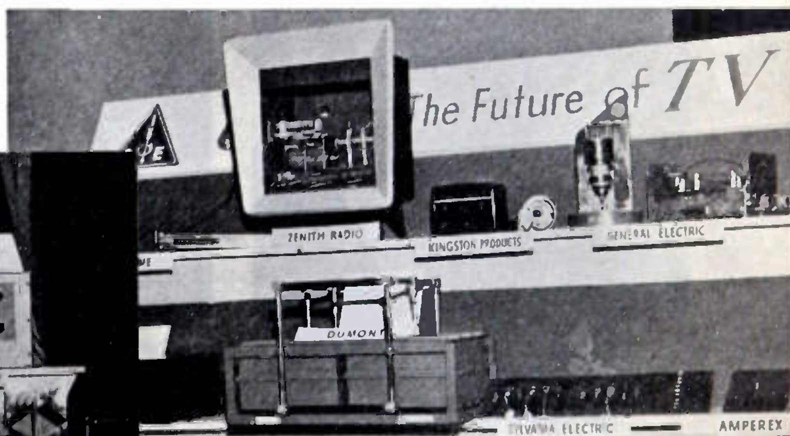
Above—Pea-sized transistor, dramatic new development in electronics, show with its case (middle, compared with subminiature tube which it replaces in Signal Corps equipment.

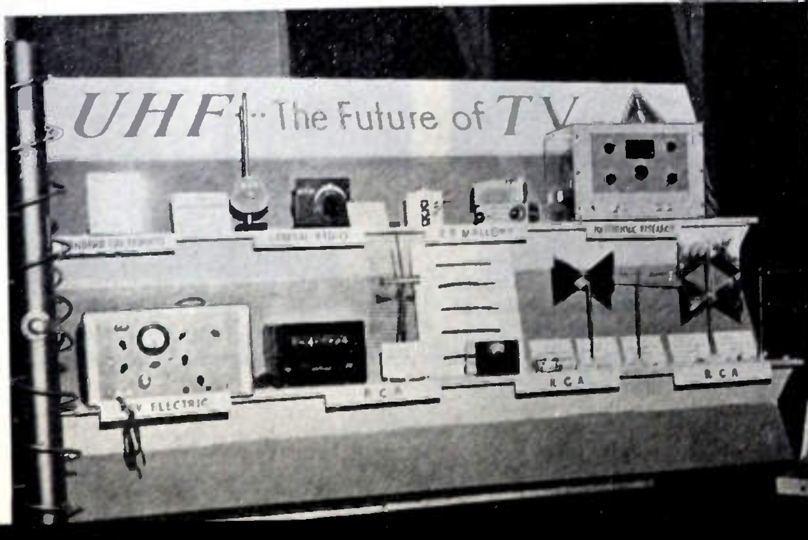
Above—IRE President D. B. Sinclair (left) receives from Publicity Committee Vice-Chairman Lewis Winner a statement sent by N. Y. State's Gov. Dewey highlighting the IRE Convention and the importance of radio engineering to defense.

Above—A short range FM transmitter (right) with battery (left) which can be hidden in the pocket allowing a performer complete freedom of motion during a lecture.

Above, right—A conventional bulky waveguide (right) is compared with the same component produced by Federal's new microwave wiring technique employing printed circuits.

Above—A uhf television transmitter tube is displayed at a special uhf exhibit held for the press at Grand Central Palace (see also below).





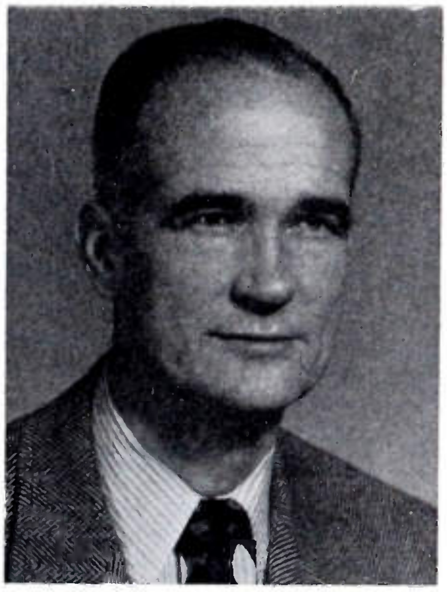
Left and above—A special uhf exhibit on public display at Grand Central Palace included transmitting and receiving tubes, converters, tuners, antennas, measuring equipment and a uhf receiver in operation.

IRE Awards, 1952

Medal of Honor

W. R. G. Baker

"In recognition of his outstanding direction of scientific and engineering projects; for his statesmanship in reconciling conflicting viewpoints and obtaining co-operative effort; and for his service to the Institute."



Morris Liebmann Memorial Prize

William Shockley

"In recognition of his contributions to the creation and development of the transistor."

Harry Diamond Memorial Award

Newbern Smith

"For his fundamental work during a period of many years on radio wave propagation, this work being the basis for the practical use of ionospheric observations in the operation of world-wide communication systems."

Vladimir K. Zworykin Television Prize Award

B. D. Loughlin

"For his outstanding contribution to the theory, the understanding, and the practice of color television."

Browder J. Thompson Memorial Award

H. W. Welch, Jr.

"For his paper entitled 'Effects of Space Charge on Frequency Characteristics of Magnetrons,' which appeared on pages 1434–1449 of the December, 1950, issue of the Proceedings of the I.R.E."

Editor's Award

Jerome Freedman

"For his paper entitled 'Resolution in Radar Systems,' which appeared on pages 813–818 of the July, 1951, issue of the Proceedings of the I.R.E."

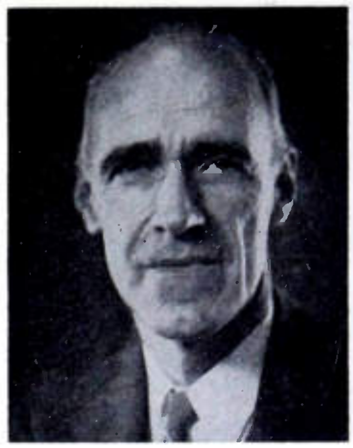
Fellow Awards

JOHN L. BARNES

"For contributions to mathematical theory and exposition in the field of transients in linear electrical and mechanical systems."

SEMI J. BEGUN

"In recognition of his contributions to the field of magnetic recording."



RALPH D. BENNETT

"For his contributions to the administration of research in Government service and to the measurement art as physicist, engineer, and educator."

LEO L. BERANEK

"For his contributions in research, teaching and administration in the fields of acoustics and speech communication."

HENDRIK W. BODE

"In recognition of his many contributions in the field of circuit and wave-filter theory."

LEON N. BRILLOUIN

"For his many teaching accomplishments and his contributions to the literature of physics and radio."

MARVIN CAMRAS

"In recognition of his contributions to the field of magnetic recording."

CHALON W. CARNAHAN

"For original contributions in the fields of frequency modulation, television, and electronic systems engineering."

SAMUEL P. CHRISTALDI

"In recognition of his contributions in the field of cathode-ray devices."

Fellow Awards

J. Lan Chu

"In recognition of his contributions to the theory and practice of waveguides and microwave antennas."

Howard P. Corwith

"For his leadership in the development of radio-telegraph, landline and submarine cable communications."

Arthur B. Crawford

"In recognition of his contributions in the field of high-frequency and microwave propagation."

Louis A. DeRosa

"For his contributions in the fields of electronic direction-finding techniques, acoustics, and aids to air navigation."

Lloyd T. DeVore

"For his contributions in the field of electronic systems engineering, and in the direction of research and development."

David W. Epstein

"In recognition of his contributions in the field of cathode-ray tube development and application."

Lester M. Field

"For many technical contributions to the electron tube art, and in particular, the development of new forms of traveling-wave tubes."

Ludlow B. Hallman, Jr.

"For his accomplishments in the development of electronic equipment for airborne communication and navigation."

Ralph N. Harmon

"For his contributions in the fields of broadcast antennas and transmitters and to electronic development in World War II."

Fellow Awards

HENRY E. HARTIG

"For his achievements as a teacher, his research in the field of acoustics, and his contributions to the underwater sound program during World War II."

JOHN K. HILLIARD

"For his contributions in the field of motion-picture and audio engineering, and in the advancement of standards."

RALPH S. HOLMES

"In recognition of his early contributions to the development of television and his work in the field of television standards."

CHARLES N. KIMBALL

"In recognition of his contributions and leadership in applying electronic techniques to a wide variety of industrial uses."

WINSTON E. KOCK

"In recognition of his contributions in the field of electromagnetic-wave lenses and antennas."

HAROLD B. LAW

"For his development of techniques and structural methods leading to practical storage- and television-tube designs."

LOUIS MALTER

"In recognition of his many contributions in the fields of vacuum and gas-filled tubes."

WILLIAM S. MARKS, JR.

"For his contributions and leadership in the field of military communications, and his pioneering work in radio-relay systems."

WILLIAM W. MUMFORD

"In recognition of his contributions in the field of high-frequency propagation and in the development of microwave components."

Fellow Awards

LEON S. NERGAARD

"In recognition of his contributions in the fields of ultra-high-frequency measurements, electron tubes and circuits."

HARRY NYQUIST

"In recognition of his fundamental contributions to physical and mathematical sciences in the field of communications."

J. A. OUIMET

"For his engineering contributions in the development and direction of radio and television in Canada."

HENRY W. PARKER

"In recognition of his contributions in the field of electron tube design."

DONALD W. PUGSLEY

"For his technical contributions and leadership in the development and design for production of commercial and military electronic equipment."

LAWRENCE R. QUARLES

"For his contributions in the field of engineering education, particularly in the teaching of communication and circuit theory."

JOHN D. RYDER

"For his contributions in industrial applications of electronic circuits and to education in radio and allied fields."

BERNARD SALZBERG

"For his contributions in the fields of electron tube development, circuit design, and military electronic systems."

HERMON H. SCOTT

"For his contributions to acoustic measurement and the reduction of noise in audio reproduction."

Fellow Awards

William G. Shepherd

"For his contributions to the development and design of electron tubes, particularly the reflex klystron."

Newbern Smith

"For his contributions to the theory and measurement of ionospheric radio-wave propagation and the development of methods of predicting ionospheric transmission on a world-wide scale."

Phillip H. Smith

"For his contributions to the development of antennas and graphical analysis of transmission line characteristics."

Constantin S. Szegho

"In recognition of his contributions to the development of cathode-ray devices."

Lester C. Van Atta

"In recognition of his contributions in the field of microwave antenna theory and design."

Russell H. Varian

"For his contributions in the field of applied physics and, particularly, in the field of velocity-modulated tubes."

John R. Whinnery

"For his contributions to the knowledge of electromagnetic theory and the application of that theory to microwave problems."

Jerome B. Wiesner

"For his contributions in the field of information theory and the administration of research on advanced techniques and concepts."

Charles E. Williams

"For his contributions to electronic and radio engineering, and for planning and development of Naval radio installations."

Institute News and Radio Notes

Technical Committee Notes

Under the Chairmanship of A. G. Jensen, the **Standards** Committee held their Annual Convention Meeting at the Waldorf-Astoria Hotel on March 6, 1952. Mr. Jensen reviewed the accomplishments of the Committee during his term as Chairman, stating that five Standards have been published in the PROCEEDINGS, and that four more have been approved by the Executive Committee for publication. Additional Standards will be ready for consideration by the Standards Committee in the near future. The Chairman mentioned the change of scope of the **Mobile Communications** Committee to that of a Systems Committee. The Committee on **Modulation Systems** has also changed its scope to incorporate "Information Theory." The Chairman discussed briefly other phases of progress including the formation of the new **Servo-Systems** Committee under the Chairmanship of W. M. Pease, and the establishment of several new subcommittees. J. W. McRae, the Standards Co-ordinator, commended the manner in which the Committee handled the problems of agreement on such difficult questions as methods of measurement and definitions and the patience and devotion of the Committee's work from which the whole Institute benefits. Dr. McRae said that Standards work was a long-established and very important part of Institute responsibility along with the Sections activities and the PROCEEDINGS. The Chairman then referred to the new rule of returning approved Standards to the originating committee for comment on the revisions suggested by the Standards Committee. Also mentioned was the practice, initiated at the beginning of the year, of sending Standards to the Definitions or Measurements Subcommittees and technical committee chairmen for comment before they are reviewed at the Standards Committee meetings. The IRE representation on CCIR (International Radio Consultative Committee) was discussed pertaining to the advantages of this arrangement from the standpoint of co-ordination. In reply to a question by Ernst Weber, Dr. McRae expressed the opinion that proposals for the formation of joint subcommittees should receive Executive Committeee approval.

The **Facsimile** Committee met on February 1, under the Chairmanship of R. J. Wise. Henry Burkhard pointed out to the Committee the results of some measurements he had made employing "acutance" principles (Eastman Kodak). Although the first results did appear somewhat discouraging with many obstacles in the path of a practical application of this principle, it was decided to pursue the matter further. The next item was a report by A. G. Cooley, who explained his purpose in proposing a bold rectangular-type face for letter-block charts. After some discussion, it was conceded that there was a basis for the use of this type face providing a sufficiently wide range of sizes was available, together with a tone strip, by which a sound basis for comparison of equipments or circuits could be obtained. However, it is becoming increasingly apparent that no single chart will completely satisfy all of the requirements of the wide variety of applications for which test charts are now used or are likely to be used in the future. I. H. Franzel's proposed definitions of recording processes were discussed briefly, but were not acted upon at this session.

Under the Chairmanship of P. C. Sandretto, the **Navigation Aids** Committee met on February 11. Discussion ensued concerning the list of navigation terms recently defined by the Committee. It is the opinion of the Committee that the definitions discussed at this meeting should be published without further delay. The Committee Chairman will inform the Standards Committee Chairman of the Committee's views in the matter. Also mentioned were a number of terms in the field of electronic navigation which had not been considered. Two lists of such terms had been made up and circulated recently to the Committee members. In addition, Harry Davis is preparing a list of radar terms which may be quite long. These additional terms will not be published with the completed list of navigation terms.

The **Measurements and Instrumentation** Committee met on February 20, under the Chairmanship of F. J. Gaffney. G. L. Fredendall, Chairman of the Subcommittee on Video Measurements, reported that very little active work had been done by his Subcommittee due to the fact that they were involved in the activities of the NTSC. Nevertheless, the Committee will continue with the program previously planned. Chairman J. G. Reid of the Subcommittee on Electronic Components, reported that the organization of the Symposium on Progress in Quality Components, jointly sponsored by the IRE, AIEE, and RTMA, was virtually complete. He informed the Committee of the details of the program. C. D. Owens, Chairman of the Subcommittee on Magnetic Measurements, proposed two directions of the work his Subcommittee is planning. These are concerned with (1) the use of magnetic materials in magnetic amplifiers for instrumentation use, and (2) high-frequency testing of magnetic materials. The Chairman reported on recent attempts to reactivate the Subcommittee on Interference Measurements. H. E. Dinger of the Naval Research Laboratories has been invited to serve as Chairman of this Subcommittee and the Committee is awaiting his decision.

The **Servo-Systems** Committee met on January 24, under the Chairmanship of W. M. Pease. The Committee agreed on the immediate encouragement of feedback-control papers at the local meetings of the IRE. Several topics and speakers were suggested. Professor Pease will check with the local IRE Committee regarding established procedures for planning local technical meetings. The work to date on terminology was discussed, and R. E. Graham has agreed to serve as Chairman of a Subcommittee on Terminology. The Committee feels the need for stronger liaison between their Committee and Subcommittee 4.7 of the Circuits Committee. The Committee has a number of people whom they will approach regarding their willingness to take part in Committee work.

Calendar of COMING EVENTS

IRE New England Radio Engineering Meeting, Copley-Plaza Hotel, Boston, Mass., May 10

IRE National Conference on Airborne Electronics, Hotel Biltmore, Dayton, Ohio, May 12–14

4th Southwestern IRE Conference and Radio Engineering Show, Rice Hotel, Houston, Tex., May 16–17

Radio Parts and Electronic Equipment Show, Conrad Hilton Hotel, Chicago, Ill., May 19–22

The IRE Professional Group on Industrial Electronics Symposium on Electronics and Machines, Palmer House, Chicago, Ill., May 22–23

Audio Engineering Society Audio Fair, Conrad Hilton Hotel, Chicago, Ill., May 23–24

AIEE-IRE Telemetering Conference, Los Angeles, Calif., August 26–27

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29

National Electronics Conference, Sherman Hotel, Chicago, Ill., September 29–October 1

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

1953 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 23–26

CBC Plans TV Programs for August

Regular TV programs of at least two or three hours duration will be broadcast daily from the CBC television stations, in both Toronto and Montreal, commencing in August of this year. The August dates are based on the possibility of having to use temporary antennas in case the steel for the transmitter towers has not been received and erected before August 1. Permanent antennas may not be available until October, 1952, for Toronto, and December, 1952, for Montreal. It is expected that the temporary antennas will be effective for distances up to thirty or forty miles from the transmitters.

Professional Group News

Electron Devices

In recognition of the needs of IRE members whose fields of activity are concerned primarily with electron tubes and semiconductor devices, the Institute has set up a new Professional Group to be responsible for the interests of its many members in that branch of radio. This is the Professional Group on Electron Devices.

The first meeting of the Administrative Committee of the Group was held at the Institute Headquarters, March 5, 1952, the purpose being to elect officers and to decide upon the way in which the Group shall begin its professional activity. Following is the membership of the Administrative Committee, including offices and business addresses:

G. D. O'Neil, Chairman
Sylvania Electric Products Inc.
Bayside, L.I., N.Y.

L. S. Nergaard, Vice Chairman
Radio Corporation of America
Princeton, N.J.

J. S. Saby, Secretary
Electronics Laboratory
General Electric Co., Syracuse, N.Y.

J. E. Gorham, Treasurer
Evans Signal Laboratory
Belmar, N.J.

H. F. Dart, Membership
Westinghouse Electric Corp.
2 MacArthur Ave., Bloomfield, N.J.

H. J. Reich, Publication
Dunham Laboratory, Yale Univ.
10 Hillhouse Ave., New Haven, Conn.

G. A. Espersen
Philips Laboratories, Inc.
Irvington-on-Hudson, N.Y.

C. E. Fay
Bell Telephone Laboratory, Inc.
Murray Hill, N.J.

J. J. Glauber
Philco Corporation
Church Road, Lansdale, Pa.

H. Q. North
Hughes Aircraft Co.
Culver City, Calif.

A. C. Rockwood
Raytheon Manufacturing Co.
55 Chapel St., Newton, Mass.

Karl Spangenberg
Office of Research Coordination
Stanford Univ., Stanford, Calif.

The Professional Group on Electron Devices will establish chapters in areas where interest is great enough to warrant so doing. At the meeting of the Administrative Committee, chapters were established for a number of locations. These chapters, and the chairmen who will complete their organization are as follows: New York City, G. A. Espersen; Philadelphia, J. J. Glauber; Los Angeles, H. Q. North; Boston, A. C. Rockwood; North Jersey, C. E. Fay; Princeton, L. S. Nergaard; Syracuse, J. S. Saby; Monmouth, J. E. Gorham. It is anticipated that additional chapters will be organized.

During 1952, the Group will sponsor a number of papers in the Proceedings of the I.R.E. and will conduct technical sessions at the IRE-RTMA Fall Meeting at Syracuse, N. Y., October 22, and the IRE West Coast Convention, Long Beach, Calif., Aug. 27–29. (June 6 is deadline for Convention papers.) It will also publish a Newsletter and will co-operate with a number of IRE Sections in holding meetings.

Members of the IRE who are especially interested in electron tubes and semiconductor devices are invited to join the Group. The annual assessment is $2.00, which should be sent to the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N.Y. There is no additional registration charge.

Group members interested in the formation of additional chapters should direct their inquiries to Secretary J. S. Saby, whose address is given above.

Electronic Computers

A Philadelphia Chapter of the Electronic Computers Group has been formed under the Chairmanship of I. L. Auerbach. Plans are being made for an interesting program of activities.

Transactions

The following Groups have mailed Transactions to their members since the IRE National Convention in March: Audio, Vehicular Communications, Antennas and Propagation and Airborne Electronics.

Symposium
on
Electronics and Machines
May 22–23, 1952—Palmer House, Chicago, Ill.

Sponsored by the IRE Professional Group on Industrial Electronics. A two day Symposium featuring the application of electronic methods and techniques in industry.

*** Contact Palmer House, Chicago, for hotel reservations.**

New Groups

Two new Groups have joined the ranks of the IRE Professional Groups: the Professional Group on Microwave Electronics, Chairman, Ben Warriner of General Precision Laboratories, Pleasantville, N. Y., and the Professional Group on Medical Electronics, Chairman, L. H. Montgomery of Vanderbilt University Medical School, Nashville, Tenn. An enthusiastic group is to be congratulated on an early program start in the Medical Electronics Group. Inquiries regarding these Groups are welcome.

The petition for a new Professional Group on "Communications" has been received and the formation of this Group will be announced at a later date.

Engineering Management

Interest in the Engineering Management field has grown in Philadelphia to the point where plans are being formulated for a local chapter of this Professional Group.

(*Continued on page 620*)

Transactions of IRE Professional Groups

The following issues of Transactions have been published by IRE Professional Groups and additional copies are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N.Y., at the prices listed below.

A listing of previously published issues appears on page 363 of the March, 1952 issue of Proceedings.

Sponsoring Group	Publication	Members of Sponsoring Group	IRE Members	Non-members*
Antennas and Propagation	PGAP-1; Sessions 2, 6, 11 of 1951 IRE Western Convention (152 pages)	$3.00	$4.50	$9.00
Antennas and Propagation	PGAP-2; Commissions 2, 3 of URSI-IRE meeting at Cornell University, October, 1951 (64 pages)	1.50	2.25	4.50
Audio	PGA-5; "Design Interrelations of Records and Reproducers," by H. I. Reiskind (8 pages)	0.30	0.45	0.90
Audio	PGA-6; editorials, technical papers, and news (42 pages)	0.80	1.20	2.40
Vehicular Communications	PGVC-1; National Meeting on Land Mobile Communications, Chicago, Ill., October, 1951 (148 pages)	4.40	6.60	13.20

* Public libraries and colleges can purchase copies at IRE Member rates.

Professional Group News, (cont.)

Information Theory

A very active Los Angeles Chapter of the Information Theory Group holds monthly meetings to which guest speakers are invited. Recent papers contributed were "Information Channel Capacity," "Transforms for Linear Time-Varying Systems."

Western IRE Convention and Electronic Show

The Los Angeles Section of the Professional Groups Committee, under the Chairmanship of L. C. Van Atta, is making arrangements for the Western IRE Convention, August 27–29, 1952, at Long Beach, Calif. It is expected that the following Groups will participate: Audio, Antennas and Propagation, Airborne Electronics, Broadcast Transmission Systems, Broadcast and Television Receivers, Electronic Computers, Information Theory, Instrumentation, and Radio Telemetry and Remote Control.

Professional Groups Committee

A "task force" on publication procedure has been appointed by the Professional Groups Committee under the Chairmanship of Austin Bailey. Guide sheets for use in the field will be issued to the Groups to assist them in planning their publications. Copies may be obtained from the IRE Editorial Department.

Airborne Electronics

George Rappaport has been elected Chairman of the Airborne Electronics Group for the year of 1952.

The 1952 National Airborne Electronics Conference, in Dayton, Ohio, May 12–14, was very successful and enjoyed a large attendance. More than forty technical papers were presented including the following subjects: "The Airlanes and Electronics," Unitization and Miniaturization," "Progress in Basic Components," "Capacitors," "Coils, Inductors, and Transformers," "Design and Production Methods," "Transistors," Aspects of Reliability," and "Electron Tubes."

Antennas and Propagation

A joint meeting was sponsored by the USA National Committee, URSI, and the Antennas and Propagation Group, on April 21–24, at the National Bureau of Standards in Washington, D. C. Papers were presented on the following subjects: radio measurement methods and standards, tropospheric radio propagation, ionospheric radio propagation, terrestrial radio noise, radio astronomy, radio waves and circuits (including general theory and antennas and waveguides), and electronics.

Professional Groups Manual

A revised copy of the Professional Groups Manual, Model Constitution, and Suggested Bylaws has been mailed to all members of the Administrative Committee for their guidance.

Instrumentation

The Progress in Quality Electronic Components Conference, sponsored jointly by the Instrumentation Group, AIEE, and RTMA, was held in Washington, D. C. on May 5–7, 1952. The following is a list of subjects covered during the conference: electronics today, basic materials, advances in miniaturization, progress in basic components, capacitors and inductors, miscellaneous components, design and production methods, transistors, aspects of reliability, and electron tubes.

Audio

J. J. Baruch of the Massachusetts Institute of Technology has been elected Chairman of the Audio Group.

The Boston Chapter of the Group on Audio holds meetings monthly at which interesting papers and discussions are included.

Abstracts and References Index To Be Published

As an added service to IRE members, the Executive Committee recently approved the publication in the Proceedings of the Annual Index of "Abstracts and References," which appear in each issue. This action will become effective with the 1952 Index, which will be published early in 1953.

Signal Corps Research Advisory Council Formed

The Army Signal Corps has announced the recent formation of a Signal Corps research and development advisory council, composed of leading figures in the scientific field in electronics and communications and including several prominent members of the Institute. The council held its first meetings April 9 and 10 in the Pentagon to discuss Signal Corps research and development problems.

Chairman of the council is Maj. Gen. George I. Back, Chief Signal Officer of the Army. Other members are: Assistant to the Chairman, Dr. W. R. G. Baker (Fellow, IRE), Vice-President and General Manager, Electronics Division, General Electric Co.; Dean A. F. Spilhaus of the University of Minnesota's Institute of Technology; Dr. A. G. Hill (Fellow, IRE), Director of the Research Laboratory of Electronics, Massachusetts Institute of Technology; Dr. James W. McRae (Fellow, IRE), Vice-President of Bell Telephone Laboratories; David B. Smith (Fellow, IRE), Vice-President, Philco Corp.; Ralph S. Holmes (Fellow, IRE), Director of the RCA Contractor Research Laboratory; and W. R. Hewlett (Fellow, IRE), Vice-President of Hewlett Packard Company, Palo Alto, Calif.

Correction

The IRE-RTMA Radio Fall Meeting will be held in Syracuse, N.Y., on October 20–22, 1952.

Notice to Contributors

The IRE has published Standards covering the proper usage of signs, symbols, and abbreviations in the radio-electronic field. In the interest of uniformity the attention of authors is directed to the availability of these Standards from the Institute of Radio Engineers, Inc., 1 East 79 St., New York 21, N.Y., as noted below.

48 IRE 21S1 Standards on Abbreviations, Graphical Symbols, Letter Symbols, and Mathematical Signs, 1948—$0.75.

49 IRE 21S1 Standards on Designation for Electrical, Electronic, and Mechanical Parts and their Symbols, 1949—$0.60. (Reprinted from Proceedings, pp. 118–124; February, 1950)

51 IRE 21S1 Standards on Abbreviations of Radio-Electronic Terms, 1951—$0.50. (Reprinted from Proceedings, pp. 397–400; April, 1951)

The attention of authors is also directed to IRE Standards covering definitions of terms in various fields. A full list of available Standards and their prices may be found opposite page 65A, in the December, 1951 issue of the Proceedings.

Officers Named for Electronics Conference

The election of officers has been announced for the 1952 National Electronics Conference, Incorporated, which will be held September 29–October 1, at the Sherman Hotel, Chicago, Ill. This eighth annual conference will be sponsored by the American Institute of Electrical Engineers, the Institute of Radio Engineers, the Illinois Institute of Technology, the Northwestern University, the University of Illinois (with participation by Purdue University), the University of Wisconsin, and the Society of Motion Picture and Television Engineers. The officers of the conference are as follows:

President, J. A. M. Lyon, Northwestern Technological Institute; Chairman, Kipling Adams, General Radio Company; Executive Vice President, J. D. Ryder, University of Illinois; Executive Secretary, Karl Kramer, Jensen Radio Company; Secretary, R. R. Jenness, Northwestern Technological Institute; Treasurer, Ralph Benedict, University of Wisconsin; Arrangements Committee Chairman, O. I. Thompson, DeForest's Training Incorporated; Proceedings Committee Chairman, C. E. Barthel, Jr., Illinois Institute of Technology; Exhibits Committee Chairman, R. M. Krueger, American Phenolic Corporation; Program Committee Chairman, R. M. Soria, American Phenolic Corporation; Procedures Committee Chairman, LeRoy Clardy, Swift and Company; Publicity Committee Chairman, S. R. Collis, Illinois Bell Telephone Company; Housing Committee Chairman, A. J. Ward, S and C Electric Company.

IRE People

Tore Anderson (S'47–A'49) has been appointed chief engineer of Airtron, Incorporated, Linden, N. J. He has been a member of that organization since 1947.

Mr. Anderson, a graduate of Oregon State College and Cooper Union, in electrical engineering, served as a lieutenant in the Engineer Corps during World War II. He worked at Belvoir on underwater mine detection devices and assisted in the development of a magnetomotor type detector for "frogmen." Following the war, he was with the Electrical Testing Laboratories and Intercontinent Engineering, where he gained valuable experience working on layout of aircraft communications and instrument landing systems.

Mr. Anderson holds patents in the field of electronics, and is an ardent "ham" with call number W2KVMVM. He is a member of the Tau Beta Phi and the American Institute of Electrical Engineers.

❖

Wallace J. Miller (SM'47), a professor of electrical engineering at the Georgia Institute of Technology, died recently. His age was 48.

Mr. Miller, a native of Kentucky, received his education at the Ohio Northern University, the United States Naval Academy, and the University of California, where he received the M.S. Degree in 1936.

Mr. Miller served as an officer with the United States Navy for a period of 20 years during which time he also was engaged in radio communications work. He served as an assistant director of the Naval Research Laboratory, Washington, D. C.

Mr. Miller was Vice Chairman of the IRE Atlanta Section in 1947–1948, and Chairman in 1950. The IRE Atlanta Section drew up a resolution whereby the members of that Section expressed their thankfulness for their association with Mr. Miller and their bereavement and sympathy to his family.

❖

Melvin B. Kline (A'43–M'44–SM'48) has been appointed as assistant engineer manager of the Allen B. DuMont Laboratories, Incorporated.

Mr. Kline received his B.S. degree in physics from the College of the City of New York and has done graduate work at Columbia University, the Polytechnic Institute of Brooklyn, and the Stevens Institute of Technology. He joined the engineering staff of the DuMont organization in 1941, and has been associated primarily with the development and design of cathode-ray oscillographs and related equipment. During World War II, he worked on loran and radar indicators. He has also served at the Du Mont New York television station, WABD, as video control and master control engineer.

Prior to his recent appointment, Mr. Kline has been head of the special projects section of the instrument engineering department of DuMont, responsible for the engineering of government projects as well as special instruments for commercial use.

❖

William G. Fockler (A'46) has been named as assistant engineer-manager of the Allen B. DuMont Laboratories, Incorporated.

Mr. Fockler received his B.S.E.E. degree from the West Virginia University in 1945, and started as a junior engineer at the DuMont Laboratories. He was then transferred to the position of senior engineer in the development section of the instrument division where he was responsible for the design of general purpose cathode-ray oscillographs and indicators requiring the use of high voltage supplies and cathode-ray tubes. Prior to his recent appointment he was the head of the development section responsible for the design of the commercial line of instruments and accessories.

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

❖

William H. Happe, Jr., (A'39) has been appointed as works engineer of the electronics division of the Curtiss-Wright Corporation, at Carlstadt, N. J.

W. H. Happe

Mr. Happe, who has been associated with research and manufacturing of electronics equipment since 1937, graduated with a B.E.E. degree from Brooklyn Polytechnic Institute. He joined the engineering and research division of RCA at Harrison, N. J., and during World War II, he was director of manufacturing for Research Enterprises Ltd., Toronto, which had charge of production of all Canadian radar equipment. He also directed radar development work with the National Research Council which worked on detection equipment for Canadian, British, and United States Forces. Prior to joining Curtiss-Wright, Mr. Happe was director of the vacuum-tube division of the Federal Radio and Telephone Corporation.

Mr. Happe is a member of the American Institute of Electrical Engineers, the Association of Professional Engineers of Ontario, Tau Beta Pi, and Eta Kappa Nu.

Earle D. Benson (A'49) has been appointed field engineering manager of the industrial equipment division of Tracerlab, Incorporated. He will be responsible for field engineering activities in the company's beta gauge program. For the past 12 years, he has been associated with the commercial engineering department of Sylvania Electric Products Incorporated.

Mr. Benson attended the Massachusetts Institute of Technology where he studied electrical engineering. He is a member of the American Institute of Electrical Engineers and the Illuminating Engineers Society.

❖

Marvin Hobbs (A'35–M'41–SM'43), electronics adviser to the chairman of the Munitions Board, has been appointed a member of the Electronics Production Board. The Electronics Production Board was established within the Defense Production Administration several months ago and is responsible for the over-all co-ordination of the electronics production program.

Mr. Hobbs' engineering and production experience in radio electronics and television industry extends over a period of 20 years. During World War II, he was associated with the War Production Board, and after the war, he worked as a consulting engineer in Chicago for a number of radio and television manufacturers. He has been serving in an executive capacity with the Department of Defense since 1950.

❖

P. S. Christaldi (S'35–A'40–SM'44–F'52) has been appointed assistant division manager of the Allen B. DuMont Laboratories, Incorporated.

Dr. Christaldi was born on November 26, 1914, at Philadelphia, Pa. He received the E.E. degree from Rensselaer Polytechnic Institute in 1935, and was a graduate fellow in physics there from 1935–1938, when he received his Ph.D. degree. He specialized in waveguide communications.

Dr. Christaldi joined the DuMont organization in 1938 as a development engineer engaged in developing commercial tubes and electronic instruments. From 1941–1947, he was the chief engineer, and during World War II, he played a leading part in the design, development, and production of military electronic equipment, as well as cathode-ray tubes, at DuMont. In 1947, Dr. Christaldi became the engineering manager of the instrument division, a post he held until his recent appointment. He is a nationally known expert on cathode-ray tubes and oscillographs.

Dr. Christaldi has been active on the IRE Committees of Admissions and Measurements and Instrumentation, and is the Secretary of the IRE Northern New Jersey Subsection. He is a member of Sigma Xi, and a Fellow of the Radio Club of America.

IRE People

Ralph S. Holmes (A'25–SM'45–F'52) has been named the director of the contract research laboratory, at the RCA Laboratory Division, Princeton, N. J.

Mr. Holmes was born in 1902, at Murray, Neb., and received the B.S.E.E. degree from the University of Nebraska, in 1923. In 1922, he was a student engineer with the Commonwealth Edison Company in Chicago, and in 1923, he joined the radio testing department of the General Electric Company. From 1925–1929, he worked on the design and development of broadcast receivers for General Electric, and then joined the engineering department of the RCA Manufacturing Company. In 1932, Mr. Holmes transferred to the RCA research division, in charge of research, development, and design of television receivers. In 1940–1943 he worked on the research and development of radar and television systems until he was appointed the manager of technical services of the RCA Laboratory Division. Prior to his recent promotion, Mr. Holmes was manager of the RCA contract projects, in Princeton.

❖

Bernard Tullius (A'43–M'51) has been appointed sales engineer for the transmitter division of the Allen B. DuMont Laboratories, Inc., at Clifton, N. J. His duties will include sales and technical counselor to clients, aiding them in planning, laying out and installing uhf and vhf transmitter equipment, co-ordinating transmitter design and construction, and supervising field work of many kinds.

Mr. Tullius has been associated, as transmitter and station engineer, with several Oklahoma City AM stations, and spent 5 years as chief engineer of Station KTOK, Oklahoma City. From 1942–1947, he was field engineer for Hazeltine Electronics, and has been an instructor in television at the Eastern School of Radio and Television. Before joining DuMont, he was senior engineer at radio engineering laboratories, in Long Island City.

Mr. Tullius is an amateur radio operator, transmitting under the call letters, W2UDW.

❖

G. Robert Mezger (A'37–VA'39) has been named engineering manager of the Allen B. DuMont Laboratories, Incorporated.

Mr. Mezger joined DuMont as a development engineer in 1936 after graduating from the Rensselaer Polytechnic Institute in electrical engineering. From 1936–1939, he was actively engaged in the development of cathode-ray instruments. He was the technical sales manager, from 1939–1941, when he went on active duty with the United States Navy. From 1941–1944, he was assigned to instrument development work at the David Taylor Model Basin, in Washington, D. C., and from 1944–1945, he was active in the design of naval radar equipment. In 1945, Mr. Mezger left the Naval Research Laboratory as a Commander to return to the DuMont Company.

He is a member of the American Institute of Electrical Engineers.

❖

Douglas H. Ewing (SM'46) has been appointed director of research services at the RCA Laboratories Division, Princeton, N. J.

Dr. Ewing, formerly director of development for the Air Navigation Board of the United States Government, was previously manager of advanced development for the engineering products department of the RCA Victor Division.

D. H. Ewing

A native of Indiana, Dr. Ewing received the B.A. degree at Butler University in 1935, and the M.S. and Ph.D. degrees from the University of Rochester, in 1937 and 1939, respectively. From 1939–1945, he was associated with the Radiation Laboratory at Massachusetts Institute of Technology, where he served as technical aid to the director and as chairman for the laboratory's overseas office. In 1945 he joined the staff of the engineering section of RCA and was engaged in the development of RCA's comprehensive air navigation system, Teleran.

Dr. Ewing is a Fellow member of the American Physical Society.

Harold L. Hazen (SM'48), head of the department of electrical engineering at the Massachusetts Institute of Technology, will become dean of the Institute's graduate school next July.

Dr. Hazen was born in Philo, Ill., in 1901, and received the B.S., M.S., and D.Sc. degrees from MIT, in 1924, 1929, and 1931, respectively. He worked in the Schenectady and Pittsfield laboratories of the General Electric Company as a research assistant, and subsequently obtained experience with both the American Telephone and Telegraph Company and the Raytheon Manufacturing Company. He joined the staff of MIT in 1925, and was named instructor in electrical engineering in 1926, and assistant professor in 1931. He became an associate professor in 1936 and was placed in charge of the department's graduate study and research from 1937–1938, and was promoted to full professor and head of the department of electrical engineering in 1938.

Dr. Hazen has co-operated in the design and construction of several important electrical devices at MIT, including the network analyzer, upon which power systems are reproduced in miniature for the solution of electrical engineering problems. He was also associated with the development of the differential analyzer which was designed by Vannevar Bush.

During the summer of 1951, Dr. Hazen headed a commission of 15 of the nation's leading engineering educators which visited Japan for consultation with the Japanese Ministry of Education. This commission was organized by the American Society for Engineering Education and the Unitarian Service Committee, Inc., at the request of the Supreme Commander of the Allied Powers.

Dr. Hazen served, during World War II, as Chief of Division 7 of the National Defense Research Committee, in the field of ordnance fire control. In 1948, he received the President's Certificate of Merit, for "outstanding services to his country."

Dr. Hazen is a fellow of the American Institute of Electrical Engineers, and a fellow of the American Academy of Arts and Sciences. He is a member of the Franklin Institute, American Society for Engineering Education, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

Sections*

Chairman	Section	Secretary
R. L. Burtner 3607 Harrison Ave. N. W. Canton 9, Ohio	Akron (4)	R. M. Byrne 316 Melbourne Ave. Akron 3, Ohio
W. L. Fattig Box 788 Emory University, Ga.	Atlanta (6)	H. W. Ragsdale 654 Cooledge Ave., N.E. Atlanta, Ga.
D. G. Little 408 Old Orchard Rd. Baltimore 29, Md.	Baltimore (3)	G. R. White Bendix Radio Div. Towson 4, Md.
L. W. McDaniel 3385 Timberwood Lane Beaumont, Tex.	Beaumont-Port Arthur (6)	C. B. Trevey 2555 Pierce St. Beaumont, Tex.
L. H. Stantz 168 Moeller Binghamton, N. Y.	Binghamton (4)	E. L. Pittsley Reynolds Rd. R. D. 1 Johnson City, N. Y.
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Books

The Conduction of Electricity Through Gases by K. G. Emeleus

Published (1951) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 91 pages +4-page appendix +3-page index +x pages. 37 figures. 4×6½. $1.50.

K. G. Emeleus is a professor of physics, at Queens University, Belfast, Ireland.

In this monograph the author outlines well-known phenomena pertaining to the conduction of electricity through gases which lend to quantitative study. Originally published in 1929, and revised in 1936, the current edition serves primarily as a background for early work and should, therefore, not be looked to as a source of new material, although a few late references have been added to the appendix.

In the seven chapters, a review is presented on such subjects as the glow discharge, excitation and ionization, mean free paths, initiation of the discharge, cathode phenomena, exploring electrodes, the negative glow, Faraday dark space, anode glow, the positive column of glow discharges, the action of magnetic fields, polarization of light excited by electron impact, pressure effects, plasma oscillations, and others. It is explained by the author that no attempt has been made to cover more recent work such as new probe methods and application of high-frequency techniques, since such inclusion would have involved major changes in the early text which admittedly was not intended. Nevertheless, it is difficult to understand this viewpoint, and, why an opportunity was not taken to at least add a chapter on the more recent developments, in order that the reader could acquire some perspective to all aspects of the field without looking up references.

The reviewer sees this book as a few hours of easy nonmathematical reading, being particularly convenient for "travel reading," and which could fill some need for undergraduate physics students or radio engineers majoring in the electron-tube field, and in early work, complementary to the more complete and up-to-date literature on the subject available in almost every technical library.

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Foundations of Wireless by M. G. Scroggie

Published (1951) by Iliffe & Sons, Ltd., Dorset House, Stamford St., London, S.E.1., Eng. 304 pages +8-page index+6-page appendix+9 pages. 236 figures. 5½×8½. Price, 12s. 6d., postage 8d.

This book, theoretical in character, starts with the elements of electricity, as applied to broadcasting. The emphasis is on receivers and includes Rf, If, audio, and power-supply parts. Also included is a chapter on the essentials of simple transmitters and a chapter on radiation and antennas. An 11-page chapter on cathode-ray tubes devotes about one and a quarter pages to the application of television, with an equivalent coverage of radar.

The text was apparently written for the nonradio man or those with no technical training. The first third of the book is on the physics of electricity and circuit elements. It is doubtful whether this book covers the elementary field sufficiently enough to satisfy the potential electrical engineer who has had no training either formally or practically. For the person who has studied physics, and not utilized it in his subsequent work, the book is of use as a refresher course for understanding the essentials covered in the latter parts of the book. For the amateur, the book will be found useful, depending upon the extent of his knowledge of radio circuit theory.

Beyond the first third of the book more detailed material is given than is needed by those who are interested only in how radio is operated. It contains quantitative information that is useful to one who wants to design his own receiver, but it does not give instructions on how to design it. The same is true of the chapter dealing with transmitters, which is sufficient for understanding the essential parts of a simple transmitter but inadequate for constructing or designing one. It would require a book at least twice its size to be a self-educating treatise of the same scope.

This is not a "how to make it" book. According to the publisher, "it provides the perfect introduction to the study of radio in all its branches." The reviewer can agree with this statement providing it is supplemented with a lecture course and some experimental work.

This fifth edition, the first being published in 1936, is largely rewritten with much added material. It is not as practical to a professional engineer, as it would be to an engineer introducing a son, in secondary school, to the theory of radio construction of a receiver or other simple amateur equipment.

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Waveguide Handbook Edited by N. Marcuvitz

Published (1951) by McGraw-Hill Book Company, Inc., 330 West 42 St., New York, 18, N.Y. 413 pages +6-page index +2-page glossary +6-page appendix +xiv pages. 140 figures. 6×9¼. $7.50.

N. Marcuvitz is an associate professor at the Polytechnic Institute of Brooklyn, Brooklyn, N.Y.

This is volume 10 of the MIT Radiation Laboratory Series. With this long awaited book, all 28 volumes in the series, with the exception of the index, have been published.

This book covers, mainly, equivalent circuit parameters for numerous microwave transmission structures. (A statement on the outside cover saying that cavities are included is incorrect). Burdensome three-dimensional electromagnetic field problems are thus reformulated as lumped-constant electrical network problems. The presentation of data for each microwave structure is given in the following pattern.

(1) A problem is briefly stated. (2) A figure then shows the microwave structure with pertinent dimensions labelled. (3) Another figure shows the equivalent network with appropriate parameters labelled. (4) Formulas are then presented which relate the equivalent network parameters to the microwave structure parameters. (Where theoretical results are not available, only experimental results are given). (5) The next paragraph describes the restrictions on the formulas, such as, the range (of wavelength) the formulas are valid, the limits of error, and the method employed to obtain the formula. This information contributes considerably to the usefulness of the formulas. (6) The results are then presented in numerical form, generally by graphs. The graphs, each of which usually cover an entire page, are clear and consistently well arranged so that values can be taken directly from them.

The microwave structures are divided into five groups with a chapter devoted to each group. The groups are: two-terminal, four-terminal, six-terminal, eight-terminal, and composite structures. Specific topics treated include the following: (1) Two-terminal—lines terminating into guides beyond cut off; lines terminating into space. (2) Four-terminal—obstacles, windows, posts, apertures, strips, arrays, and dents (of both zero and finite thickness) in various types of wave guides; junctions, bifurcations, changes in wageguide dimensions, corners, bends, etc. (3) Six- and eight-terminal—*Y* and *T* junctions; aperture-coupled and slit-coupled hybrid junctions, etc. (4) Composite structures—guides with dielectric slabs, cylinders, ridges, etc.

The first three chapters of the book are concerned with the theoretical considerations and the background material which form the basis for subsequent chapters. This material, which takes up 40 per cent of the book, includes the mathematical discussion of transmission-line theory, the electromagnetic field representations in all types of microwave transmission lines (with many illustrations, tables, and mathematical expressions for field components), and the theory and practice of representing a microwave structure by its equivalent network.

"The Waveguide Handbook,"—similar in nature to, but larger and more complete than the very useful Sperry handbook, "Microwave Transmission Design Data," published in 1944—thus presents in compact form a wealth of information on microwave structures. It is one of the most excellent and comprehensive books which has been published so far on this subject.

A formidable amount of effort must have been expended to make this book possible. It may in the future become the basis for a comprehensive microwave handbook that will be revised and made more complete every few years in order to bring the microwave tube and circuit worker most of the reference data he requires in one volume.

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Books

TV Factbook Number 14

Published (January 15, 1952) by the Television Digest, Wyatt Building, Washington, D.C. 112 pages. 11×8½. $5.00.

This semiannual Factbook is kept up to date with weekly addenda supplements reporting current TV actions of the FCC. The report includes such basic reference material as given in the following list.

1. Personnel and facilities data, with digests of rate cards, of all networks and of the 109 TV stations (including the Mexican-licensed outlet opposite Brownsville, Texas), together with complete listings of actual and projected TV stations in Canada, Mexico, Cuba, and South America.

2. Tabulation of the 479 applications for new TV stations now pending before the FCC (29 of them uhf), with present and proposed vhf and uhf channel allocations.

3. TV-radio production figures, sets-in-use estimates, and market data on TV and other areas of the United States. The TV Factbook has an enclosure of a 34×22-inch wall map (in color), showing present TV areas and actual and projected coaxial-microwave network routes.

4. Full text of the code of television practices, and lists of television program sources (film and live shows) and national sales representatives of TV stations. Also listed are organizations and publications dealing with TV, including labor unions, trade associations, technical groups, etc.

5. Complete directories of TV receiver manufacturers, the FCC, congressional committees handling TV-radio matters, consulting engineers, attorneys specializing in TV radio, NPA electronics division, and others.

This report can be acquired through the Television Digest, Wyatt Building, Washington, D. C.

Advanced Engineering Mathematics by C. R. Wylie, Jr.

Published (1951) by the McGraw-Hill Book Company, Inc., 330 West 42 St., New York 18, N. Y. 572 pages+10-page index + 58-page appendix + xiii pages. 201 figures. 6×9. $7.50.

C. R. Wylie, Jr., is professor of mathematics and chairman of the department of mathematics and astronomy, University of Utah, Salt Lake City, Utah.

This textbook covers most of the mathematics essential in the first years of graduate study. The subjects discussed are classical and can be briefly listed as follows: Ordinary differential equations, with particular emphasis on linear differential equations with constant coefficients; Fourier series, Fourier integrals and Laplace transformation; Partial differential equations and orthogonal sets; Bessel functions; theory of functions of complex variables; Vector analysis; and a long last chapter on numerical methods. The discussion on complex variables is extensively and clearly presented in six chapters covering a total of 140 pages.

A large number of examples are solved throughout the book, and theories are followed or preceded by illustrative problems. Over 950 problems of different degrees of difficulty are included, accompanied by answers and hints to their solution. With the exception of the chapter on the applications of linear differential equations to electrical and mechanical problems, the examples are usually of a mechanical nature, and electrical problems are seldom discussed.

In general, the style is clear and accurate, although we notice a statement rather strange in a text so carefully written, which refers to a basic topic in electrical engineering. For example, in discussing the use of complex notation in the solution of linear differential equations with constant coefficients, the author says, "A somewhat different approach to the determination of the particular integral, or steady state current, is sometimes taken by electrical engineers. Given either a cosine or a sine voltage, they write the basic differential equation in the form. . . . " (substituting exp $j\omega t$ to the real trigonometric function), " . . . trusting that real and imaginary terms will somehow retain their identity throughout the process." This last statement seems to imply that the correct final answer is due to a fortunate coincidence rather than to the linearity of the equation which, of course, is the essential point.

Except in a few instances, we feel that Professor Wylie has written an excellent textbook whose principal qualities are clarity and readability.

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Electromagnetic Problems of Microwave Theory by H. Motz

Published (1951) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 180 pages+2-page index+bibliography+vii pages. 56 figures. 4½×6¾. $2.00.

H. Motz is a research associate at the Microwave Laboratory, Stanford University, Stanford, Calif.

This short book, a new edition in the Methuen series of monographs on physical subjects, is intended to illustrate various mathematical methods of microwave analysis through the presentation of a number of thoroughly worked examples. These examples include the klystron and magnetron theory, field calculations in various waveguide and resonant-cavity structures, the impedance of an antenna inside a waveguide, and the shunt susceptance of transverse irises in waveguides.

Despite the few pages available, the treatment delves deeply into each problem and is definitely not superficial. However, the extremely large amount of information conveyed requires a condensation to an extent where the text is often difficult to follow, and where it places a large burden on a reader wishing to gain a full understanding of the material. The number of problems, treated in this brief book is necessarily small and the selection will, of course, please some readers while disappointing others.

The book is intended for students and research workers having a "fairly good" knowledge of electromagnetic theory. However, this reviewer believes that the book is too condensed and limited in scope to be useful to the average student or practicing engineer. An engineer already familiar with one or more of the problems presented in the book should find it of value as a review and as an introduction to new viewpoints of his field.

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Television Principles by Robert B. Dome

Published (1951) by McGraw-Hill Book Company, Inc., 330 W. 42 St., New York 18, N. Y. 281 pages+9-page index+xii pages. 171 figures. 6×9. $5.50.

Robert B. Dome is an electrical consultant at the General Electric Company, Syracuse, N. Y.

By using material prepared for an out-of-hours course in the General Electric Company, the author has necessarily prepared the principles of television engineering at an intermediate level which may not appeal to a large audience. Neither the novice nor the television engineer with a few years of experience will learn a great deal from this book. The graduate engineer with a solid background in the electronic and electric circuit fields, however, will find this book beneficial as an introduction to the special problems of television broadcasting under the present system in the United States.

Chapter by chapter the author discusses certain specific solutions to the problems arising in the successive circuit elements of a broadcasting system, beginning with camera tubes and progressing through video amplifiers, antennas, converters, IF amplifiers, and second detectors. Such orderly progression causes duplication; one example is the repetition of the discussion on tuned coupled circuits in connection with amplifiers for both receivers and transmitters. Since the emphasis in this book is placed on specific numerical designs rather than on fundamental theoretical principles, the reason for such repetition should be obvious.

Each chapter closes with a few well chosen problems concerned with the material just covered. The answers are also included. Only a few errors and inconsistencies, expected in a first edition, are to be found in this book. It is regrettable that somewhat more space was not allotted to specifying the general problems to which the author has given specific solutions.

For the serious student, the inclusion of more numerous and more recent references would have enhanced the value of this book. Subjects on amplifier design, antenna design, noise in amplifiers, etc., have been presented at great length in several texts issued since World War II. These texts not only supplement this book, but are important tools of the practical designer of television circuitry.

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Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with That Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February 1951, through January 1952, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London, S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

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.

ACOUSTICS AND AUDIO FREQUENCIES

519.272.11:[534.874.1+621.396.62 882
The Correlation Function in the Analysis of Directive Wave Propagation—H. Nodtvedt. (*Phil. Mag.*, vol. 42, pp. 1022–1031; September, 1951.) Correlation-function theory is useful for investigating directivity in the reception of signals with a broad spectrum. Conditions are analyzed for systems comprising (*a*) two nondirectional receivers, and (*b*) two directional receivers. Results are shown graphically and compared with those obtained by the phase-comparison method. Increase of signal bandwidth leads to reduction of side lobes, but the amount of the increase required for a substantial effect is too high for the method to be of practical importance in hf problems, though it is of interest in sound reception.

534+621.395.61/.62](083.71) 883
I. R. E. Electroacoustics Standards: Genesis of the Glossary of Acoustical Definitions—(Proc. I.R.E., vol. 39, p. 1567; December, 1951.) Corrections are given to I.R.E. Standards on Electroacoustics (2310 of 1951).

534.012 884
Longitudinal Modes of Elastic Waves in Isotropic Cylinders and Slabs—A. N. Holden. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 956–969; October, 1951.) The general properties of the longitudinal vibration modes are derived, making use of the close formal analogy between the dispersion equations for slabs and cylinders.

534.213 885
The Propagation of Sound Waves through a Medium with Very Small Random Variations in Refractive Index—T. H. Ellison. (*Jour. Atmos. Terr. Phys.*, vol. 2, no. 1, pp. 14–21; 1951.) "Mathematical techniques, developed for the theory of turbulence, are applied to the study of the propagation of waves in a medium in which the refractive index varies slightly with position in a random manner. Both a diffraction and a ray theory are used to relate the statistical properties of the wave to those of the refractive-index field, and the conditions for the validity of the ray theory are exposed. A particular result is, that, for a wave travelling in a statistically homogeneous medium where the scale of the variations in refractive index is suitably large compared with the wavelength but small compared with the path length, the variations in intensity produced by the medium are proportional to the cube of the path length for short paths, but directly proportional to it for long paths."

534.3:786.6 886
Organ Acoustics—F. Trendelenburg. (*Ricerca Sci.*, vol. 21, pp. 339–352; March, 1951.) A review of German research work covering (*a*) the various types of organ pipe and their characteristic frequency spectra, (*b*) the differences between the sound quality of organs of the Baroque period and of modern organs, (*c*) initial transients and their acoustic effect, (*d*) the relation between the acoustics of churches and organ music style.

534.321.9.012:[539.32:546.74 887
Frequency Dependence of Elastic Constants and Losses in Nickel—R. M. Bozorth, W. P. Mason and H. J. McSkimin. (*Bell. Sys. Tech. Jour.*, vol. 30, part 1, pp. 970–989; October, 1951.) An ultrasonic pulse method was used to measure the elastic constants of single crystals of Ni and their variation with magnetic field. Measurements were also made of the velocity and attenuation of elastic vibrations in well-annealed polycrystalline Ni rods over the frequency range 5–150 kc. The elastic constants vary with domain distribution in the case of demagnetized crystals. The small variation with magnetic field observed at the frequency used (10 mc) is due to a relaxation in the domain-wall motion due to micro-eddy-current damping. The average domain size was found to be about 0.04 mm.

534.85:681.84 888
New Lightweight Pickup and Tone Arm—L. J. Anderson and C. R. Johnson. (*Broadcast News*, pp. 8–14; May/June, 1951.) Illustrated description of RCA Type-MI-11874 pickup and Type-MI-11885 arm for use at broadcasting stations and suitable for disk speeds of 33⅓ or 78 rpm (ordinary groove) and 33⅓ or 45 rpm (fine groove).

534.862.4 889
The Reproduction of Magnetically Recorded Signals—R. L. Wallace, Jr. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1145–1173; October, 1951.) "It has been found experimentally and theoretically that introducing a spacing of *d* inches between the reproducing head and the recording medium decreases the reproduced voltage by 54.6 (d/λ) decibels when the recorded wavelength is λ inches. For short wavelengths this loss is many decibels even when the effective spacing is only a few ten-thousandths of an inch. On this basis it is argued that imperfect magnetic contact between reproducing head and recording medium may account for much of the high-frequency loss which is experimentally observed." Mathematical appendices deal with (*a*) the field due to a flat sinusoidally magnetized medium, (*b*) the field due to a round wire.

621.395.61/.62 890
Crystal-Transducer Response—B. J. Shelley. (*Electronic Eng.*, vol. 23, pp. 353–354; September, 1951.) Over the af range crystal transducers can have identical response curves for either direction of operation. By using two such crystals coupled to a common cavity the frequency response curves for the individual crystals can be obtained by combining the measured sum and difference response curves.

621.395.61:621.396.645.029.3 891
Construction of A.F. Amplifiers for Studio Equipment—Schiesser and Gathmann. (See 950.)

621.395.61:621.396.645 892
Equipment for Acoustic Measurements: Part 1—A Portable General-Purpose Microphone Amplifier using Miniature Valves—Shorter and Beadle. (See 946.)

621.395.616 893
New Capacitor Microphones for Broadcasting Studios—H. Grosskopf. (*Fernmeldetech. Z.*, vol. 4, pp. 398–402; September, 1951.) The bearing of acoustic and transmission conditions on the design of microphones is discussed. Two new capacitor types developed by the Northwest German Broadcasting Organization are described, viz., the M49 gradient type with continuously variable directivity, and the M50 pressure type.

621.395.616:534.612.4 894
Condenser Microphone Sensitivity Measurement by Reactance-Tube Null Method—H. E. von Gierke and W. W. von Wittern. (Proc. I.R.E., vol. 39, p. 1534; December, 1951.) Correction to paper abstracted in 2624 of 1951.

621.395.623.7:621.396.645 895
Cathode-Follower Loudspeaker Coupling—Fletcher and Cooke. (See 947.)

621.395.623.74:537.58 896
The Ionic Loudspeaker—S. Klein. (*TSF pour Tous*, vol. 27, pp. 278–281; September, 1951. Erratum, *ibid.*, no. 276, p. 342.) Description of a loudspeaker with no moving parts, developed from the thermionic cell previously described (593 of 1947 and 288 of February). The ion source consists of a Pt wire inserted in a small tube of refractory material, such as quartz or porcelain, covered with a positive-ion

emitting layer formed by a mixture of precipitated Pt, $AlPO_4$, Ir and graphite. This tube is fixed in the throat of an exponential horn of fused quartz with double walls, the space between the walls being evacuated. Surrounding the whole is a grounded cylindrical electrode. When a hf hv source (400 kc, 10 kv) is connected to the Pt wire, after about a minute the wire is maintained at a temperature of about 1,000°C, with consequent positive-ion emission from the coating on the surrounding tube and intense ionization of the air molecules in the throat of the horn. Modulation of the hf voltage thus results in modulation of the ionization of the air, and sound waves are emitted. The frequency/response curve given shows relatively small variations of level from 30 cps to over 10 kc. The device can also be used as a microphone. Circuit diagrams are given for the loudspeaker and microphone arrangements, with component details for the hf hv generator.

621.395.623.74:537.58 897
The Physical Principles of the Ionic Loudspeaker and Microphone—S. Klein. (*TSF pour Tous*, vol. 27, pp. 340–342; October, 1951.) Discussion, with numerical calculations, of (*a*) the effect of the high temperature on the air in the throat of the horn (896 above), (*b*) the effect of the accelerating voltage on the positive ions, (*c*) ionic mobility in the rarefied air in the horn throat, (*d*) the effect of frequency.

621.395.623.74:537.58 898
The Ionophone—M. Bonhomme. (*Toute la Radio*, pp. 251–255; October, 1951.) General description of Klein's ionic loudspeaker (896 above), with diagrams and photographs. See also *Wireless World*, vol. 58, pp. 2–3; January, 1952 (F.L.D.), and *Radio-Electronics*, vol. 23, pp. 44–45; November, 1951. (Aisberg and Bonhomme).

621.395.625.2 899
A Disk Reproducing Desk—R. D. Petrie. (*BBC Quart.*, vol. 6, pp. 105–112; Summer, 1951.) The basic requirements for this equipment are outlined and the electrical and mechanical design adopted to meet them is described. The output from the sapphire stylus pickup, after equalization to suit the frequency characteristic of the disk being played, is amplified and feeds a star mixer network. The output amplitude/frequency characteristic when playing a BBC record is flat to within ±1 db from 50 cps to 10 kc. The aluminum turntable is rim driven, speeds of 78 and 33⅓ rpm being available. Several mechanical features are incorporated to prevent damage when changing disks.

621.395.625.3 900
An Editing Machine for Magnetic Tape Recording—D. C. Yarnes. (*Broadcast News*, pp. 66–69; May/June, 1951.) Description of a professional-quality machine with facilities for the rapid marking, cutting and splicing of the tape.

621.395.625.3 901
Magnetic Processes in Tapes and Magnetic Heads of Magnetic Sound-Recording Apparatus—W. Puhlmann. (*Funk u. Ton*, vol. 5, pp. 65–75; February, 1951.) The hf bias technique for measurement of the static remanence characteristic of tape materials is described. Advantages of the powder-coated tape, and the selection of suitable magnetic materials are discussed. The "internal" demagnetization effected in the coated tape reduces the loss of sensitivity at high frequencies normally occurring after wiping. Losses due to the air gap in the magnetic head are calculated and noise is discussed with reference to the irregular distribution of magnetic particles.

621.395.665.1 902
Automatic Control of [volume] Contrast—Balz. (See 940.)

621.395.92.001.4 903
Measurements on Hearing Aids—F. Müller. (*Funk u. Ton*, vol. 5, pp. 361–368; July, 1951.) An account of the fundamentals of measurement technique regarding frequency range, amplification, sensitivity and distortion.

621.395.92.001.4 904
Testing of Hearing Aids—F. Müller. (*Funk u. Ton*, vol. 5, pp. 400–410; August, 1951.) An outline of all the measurements required to determine the characteristics of hearing aids, with illustrative examples.

789.983†:621.396.615.029.3 905
Recent Design Developments in Electronic-Organ Tone Generators—S. L. Krauss and C. J. Tennes. (*Proc. NEC* (Chicago), vol. 6, pp. 344–347; 1950. Pentode and triode circuits are described which generate simultaneously a sinusoidal and a pulse wave, this combination having been found particularly effective for the production of the tone qualities required for the different organ voices.

534.232 906
Die elektroakustischen Wandler (Electroacoustic Transducers) [Book Review]—W. Furrer. Publishers: J. A. Barth, Leipzig, 2nd ed., 1951, 221 pp., 22.30 Swiss francs. (*Tech. Mitt. schweiz. Telegr. Teleph Verw.*, vol. 29, p. 358; September 1, 1951.) Additions and corrections to the first edition, published in 1940, are included; the Giorgi system of units is used. Both sonic and ultrasonic devices are considered. The book will be useful for both students and practicing engineers.

ANTENNAS AND TRANSMISSION LINES

621.392:621-231.221 907
Microwave Rotating Joints—J. P. Grantham. (*Electronic Eng.*, vol. 23, pp. 332–335 and 377–381; September and October, 1951.) Most of the known types of rotating joint for circular-section waveguides and coaxial lines are described. Data are furnished which, together with information available in the works noted in the bibliography, should enable a rotating joint for a specific purpose to be designed with little experimental work. Coaxial-line joints are the most compact and easiest to design of all the types described; they are free from resonance effects, can cover very wide bands, and have adequate power-handling capacity. The "doorknob" transformer is very suitable for rotating-joint applications; a truncated-cone type of "doorknob" used with a semicircular piston has good wide-band properties without the use of matching irises. A form of circular-waveguide joint using resonant-ring H_{11}-mode suppressors and cavities is free from resonances, but is less compact than the coaxial-line joint, requires closer manufacturing tolerances, and is more susceptible to voltage breakdown.

621.392.09 908
Advances in the Theory of Waves on Wires—H. Kaden. (*Arch. elekt. Übertragung*, vol. 5, pp. 399–414; September, 1951.) The theory of Harms and Goubau for surface waves on wires is extended to apply to magnetically loaded wires produced either by coating the wire with ferrite or by inserting a ferrite core in the coiled wire. By using these methods of concentrating the field, the upper wavelength limit can be raised from 1 to 10 m. The field structure is characterized by a parameter called "critical radius," which increases with wavelength. Characteristic impedance is proportional to the logarithm of critical-radius/wire-radius. The concept of "equivalent coaxial line," i.e. that having the same characteristic impedance and attenuation constant as the wire, facilitates investigation of the coupling between parallel lines. Values of parameters are tabulated for particular lines at various wavelengths, and their performance is compared with that of coaxial lines.

621.392.21+621.392.26† 909
Simple Branch Connections of Lines and Waveguides—H. H. Meinke. (*Fernmeldetech. Z.*, vol. 4, pp. 385–388; September, 1951.) Whereas at frequencies below 100 mc circuit branching presents no special problems, at decimeter wavelengths certain unwanted reactances become appreciable, and at centimeter wavelengths the difficulties are even greater. The problem is surveyed for three-line junctions, series and parallel arrangements being distinguished; a selected bibliography is given.

621.392.26† 910
The Offset Wave-Guide Junction as a Reactive Element—L. D. Smullin and W. G. Glass. (*Jour. Appl. Phys.*, vol. 22, pp. 1124–1127; September, 1951.) These junctions are characterized by rugged mechanical construction and ease of adjustment. Simple analytical expressions are derived for some of the more important properties. The capacitive junction allows the practical construction of capacitively coupled resonant cavities; these have a more constant bandwidth over their tuning range than can be obtained with inductively coupled cavities.

621.392.26† 911
Theory of Space-Charge Waves in Cylindrical Wave-guides with Many Beams—P. Parzen. (*Elec. Commun.*, vol. 28, pp. 218–219; September, 1951.) The effect on gain and bandwidth due to the separation of the beams and the presence of the surrounding guide is derived approximately by an integral-equation method. The mechanism of energy transport is such that the input and output circuits required would be similar to those of a klystron rather than a traveling-wave tube. Cases given special attention are: (*a*) two thin annular beams within a circular guide, (*b*) a homogeneous mixture of two beams filling the guide.

621.392.26†:538.61 912
Magneto-optics of an Electron Gas with Guided Microwaves—L. Goldstein, M. A. Lampert and J. F. Heney. (*Elec. Commun.*, vol. 28, pp. 233–234; September, 1951.) Reprint. See 2911 of 1951.

621.392.26†:538.61 913
Magneto-optics of an Electron Gas with Guided Microwaves—L. Goldstein, M. Lampert and J. Heney. (*Phys. Rev.*, vol. 83, p. 1053; September 1, 1951.) Correction to paper abstracted in 2911 of 1951.

621.392.26†:538.61 914
Magneto-optics of an Electron Gas for Guided Microwaves: Propagation in Rectangular Wave Guide—L. Goldstein, M. Lampert and J. Heney. (*Phys. Rev.*, vol. 83, p. 1255; September 15, 1951.) The experimental arrangement was similar to that described in 2911 of 1951 for a circular waveguide. In the present case the guide without the electron gas supports only the TE_{10} mode at the frequency used, hence no polarization transformation can occur. Significant effects were observed at values of magnetic field such that electron gyro-frequency and signal frequency are nearly equal.

621.392.26†:621.392.09 915
The Propagation of Waves in Cylindrical Waveguides and the Hertzian Solution as Special Cases of the Propagation of Waves in Horns—H. Kleinwächter. (*Arch. elekt. Übertragung*, vol. 5, p. 439; September, 1951.) Explanatory note on 320 of March.

621.396.67 916
An Aerial for V.H.F. Broadcasting—G. D. Monteath. (*BBC Quart.*, vol. 6, pp. 122–128; Summer, 1951.) Illustrations and description of the antenna system installed at Wrotham. See also 2340 and 2643 of 1951 (Gillam).

621.396.67 917

The Measurement of Current Distributions along Coupled Antennas and Folded Dipoles—T. Morita and C. E. Faflick (Proc. I.R.E., vol. 39, pp. 1561–1565; December, 1951.) The theoretical analysis of coupled antennas is facilitated by representing the current in each element as a superposition of components in opposite directions, the components in one direction constituting the radiating currents which maintain the far-zone field, while those in the other direction are essentially non-radiating currents. Experiments on closely coupled antennas and folded dipoles are described which confirm the correctness of the above representation.

621.396.67 918

Theory of V-Antennas—R. King. (*Jour. Appl. Phys.*, vol. 22, pp. 1111–1121; September, 1951.) An integral equation for the current in an apex-driven symmetrical V-antenna is derived and solved by successive approximations. General formulas for the distribution of current and the impedance are obtained. The zero-order impedance is evaluated for a leg length $h=\lambda_0/4$, as a function of the angle between the legs; the results are compared with measured values, and experimental impedance curves for various angles of lateral and forward tilt are discussed.

621.396.67.029.62/.63 919

Omnidirectional Aerials with Horizontal Polarization for Wide Frequency Bands in the Metre- and Decimetre-Wave Region—K. Lamberts. (*Frequenz*, vol. 5, pp. 177–185; July, 1951.) Experimental investigation of two types of antenna consisting of four flat circular disks arranged (*a*) as two horizontal dipole elements crossed at right angles, (*b*) with the disks vertical and forming a square with length of side effectively $\lambda/2$. In case (*a*) the total input impedance is relatively constant but the radiation pattern is poor; the arrangement is not particularly suitable for wide-band operation. In case (*b*) the input impedance varies with frequency but the radiation characteristic is nearly circular for frequencies within ± 20 per cent of the center frequency.

621.396.671 920

On Zuhrt's Theory of Dipole Aerials—R. King: H. Zuhrt. (*Frequenz*, vol. 5, pp. 219–223; August, 1951.) King gives a critical analysis of Zuhrt's method of calculating the current distribution and input impedance of a symmetrical tubular radiator (2977 of 1950). It is pointed out that the current at the open end of the antenna does not vanish, but only changes direction. Evaluation of the Fourier series to only the third term limits the application of the method to antennas sufficiently thin and short or of length very near that for series resonance. Measurements substantiate these conclusions.

Zuhrt states that a correction for end effects is already included in the manuscript of a forthcoming book, and suggests a possible anomaly in the measurements mentioned by King.

621.396.676 921

Glide-Path Cavity Antenna for Jet Fighter Aircraft—L. E. Raburn. (*Tele-Tech*, vol. 10, pp. 32–33, 76 and 50–51 . . 91; September and October, 1951.) Radiation-pattern measurements on a one-tenth scale model were confirmed by results obtained with an experimental mock-up of a cavity installed in the lip of the air intake. The antenna is horizontally polarized, receives signals of frequency 329–335 mc from any forward direction, and has a voltage SWR of over 5:1.

621.396.677 922

Corrugated End-Fire Antennas—D. K. Reynolds and W. S. Lucke. (*Proc. NEC* (Chicago), vol. 6, pp. 16–28; 1950.) The effect of the surface parameters in modifying the guided-wave propagation over a corrugated surface is discussed in relation to the attainment of the optimum end-fire radiation pattern. The effects on field pattern and frequency bandwidth of various methods of driving such antennas are compared. Practical forms of corrugated antenna are described and theoretical and measured field patterns are compared.

621.396.677.3† 923

General Formulation for Calculation of Shaped-Beam Antennas—A. S. Dunbar. (*Jour. Appl. Phys.*, vol. 22, p. 1217; September, 1951.) Outline of a method based on Chu's synthesis of a shaped beam by the controlled variation of phase in an aperture having a known amplitude distribution. Using the method of stationary phase the far-field intensity is formally derived and illustrated by the case of a line source with cylindrical reflector.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.3† 924

Noise Figure of the Magnetic Amplifier—N. R. Castellini. (*Proc. NEC* (Chicago), vol. 6, pp. 52–58; 1950.) An expression for the noise figure is derived by using an equivalent circuit which takes into account winding and leakage inductances and stray capacitance. The effect of the Barkhausen and thermal-noise currents is considered. The noise figure varies inversely as the amplification factor and operating frequency. The calculated mean output current at threshold agrees well with experimental values.

621.314.3† 925

Magnetic Amplifiers with Orthonol-Tape Cores—W. A. Geyger. (*Proc. NEC* (Chicago), vol. 6, pp. 59–67; 1950.) Performance figures for these amplifiers, which use multi-layer toroidal windings on Orthonol-tape cores, show that far superior characteristics and greater power output are obtained when this rectangular-hysteresis-loop core material is used.

621.314.3† 926

Transient Response of Magnetic Amplifiers—L. A. Finzi, D. P. Chandler and D. C. Beaumariage. (*Elec. Eng.* (N. Y.), vol. 70, p. 809; September, 1951.) Summary of 1951 A.I.E.E. Great Lakes District Meeting paper.

621.314.3† 927

The Time Constant of a Magnetic Amplifier—H. Goldstein. (*Funk u. Ton*, vol. 5, pp. 76–78; February, 1951.) An approximate value of the time delay, τ, for build-up in the dc windings, is given by $\tau=\beta. V/\omega$ where β is the ratio of the currents in the ac circuit with and without dc premagnetization, and V the amplification factor.

621.318.572 928

The Counting of Random Pulses—E. H. Cooke-Yarborough. (*Jour. Brit. IRE*, vol. 11, pp. 367–380; September, 1951. Discussion, *ibid.*, vol. 11, p. 594; December, 1951. 1951 Radio Convention (London) paper, surveying electronic and electromechanical methods of counting pulses. Numerous references are given to published circuits.

621.318.572:621.385.38 929

The Thyratron as a Close-Differential Relay—J. J. Baruch. (*Proc. NEC* (Chicago), vol. 6, pp. 87–93; 1950.)

621.318.572:621.396.615.17 930

Linear Analysis of Electronic Switching—R. Ahmed. (*Indian Jour. Phys.*, vol. 24, pp. 281–290; July, 1950.) The multivibrator circuit is considered; assumptions are made leading to a simple equivalent circuit which can be treated by linear analysis, enabling the switching time to be calculated with fair accuracy. Criteria are established for designing fast switching circuits.

621.318.572:621.396.615.17 931

A Study of the Switching Action in a Multivibrator Circuit: Part 1—B. M. Banerjee. (*Indian Jour. Phys.*, vol. 24, pp. 361–370; August, 1950.) Description of cro equipment for displaying the rapid changes in tube electrode voltages which occur during conduction transfer from one tube to the other of a multivibrator. Oscillograms obtained with different values of the circuit parameters are reproduced.

621.319.4 932

Electrical Charge Storage in Polystyrene Capacitors—L. A. Matheson and V. J. Caldecourt. (*Jour. Appl. Phys.*, vol. 22, pp. 1176–1178; September, 1951.) Measurements on charge storage in polystyrene-film capacitors show that charge "soakage" is very small and that charges may be retained for periods of the order of 100 years. The specific resistance of the insulation approaches 10^{22} Ω cm over a period of months.

621.319.45 933

Electrolytic Capacitors—G. W. A. Dummer. (*Wireless World*, vol. 57, pp. 510–512; December, 1951.) Principles of operation are stated, and the advantage of the high capacitance/volume ratio is stressed. Present-day types are classified according as they use a plain-foil, etched-foil or sprayed-gauze anode. Methods of increasing anode surface area and the use of new materials, especially tantalum, have been investigated.

621.392.5 934

A Note on the Initial Excitation of Linear Systems—L. A. Zadeh. (*Jour. Appl. Phys.*, vol. 22, pp. 1216–1217; September, 1951.) The formula for the system response, previously given for the case where the initial values of input and output voltages and their derivatives are specified at the instant immediately following the application of the input voltage (2668 of 1951), is modified so as to apply to the case where the initial values are specified at the instant immediately preceding the application of the input voltage.

621.392.5 935

Graphical Study of Quadripoles at Constant Frequency—J. Scherer. (*Rev. gén. élect.*, vol. 60, pp. 359–370; September, 1951.) A relatively simple method is developed from the geometrical properties of the nomographic transformation relating the input impedance of a quadripole to the load impedance, making use of the basic open-circuit and short-circuit impedances or admittances and of the constructions of elementary plane geometry. Of the numerous possible applications of the method, the determination of the impedance of an iterative network and method of transforming a quadripole are discussed in particular.

621.392.5:512.831 936

Quadripoles and Matrices—H. G. Möller. (*Elektrotechnik* (Berlin), vol. 5, pp. 426–430; September, 1951.) The use of matrix methods in circuit analysis is illustrated with reference to transformers, Lecher-wire systems, filter networks, delay circuits, loudspeakers, amplifier tubes and detectors.

621.392.52 937

The Reduction of Branched Circuits and Its Application to Ladder Arrangements of Quadripoles—F. M. Pelz. (*Funk u. Ton*, vol. 5, pp. 57–64; February, 1951.) Klein's method of determinantal transformation (2999 of 1950) is used to derive a network transformation of star-delta type. Simplification of the reduced circuit is achieved by introducing a given relation between the values of the elements of the original circuit. Application of the method to the design of a band-pass filter is illustrated.

621.392.52 938

Determination of Input Impedance and Other Properties of Networks—E. Green. (*Marconi Rev.*, vol. 14, pp. 141–155; 4th Quarter, 1951.) Parameters of many classes of network can be expressed in terms of frequency response where this is known exactly. Formulas given by Dishal (3369 of 1949) in relation to band-pass networks are used for this purpose.

621.392.52 939

The Transfer Impedance of Recurrent Π and T Networks: Part 2—J. B. Rudd. (*AWA Tech. Rev.*, vol. 9, pp. 67–72; September, 1951.) An expression is derived for the transfer impedance of an *N*-section chain of symmetrical Π or T sections with equal resistance terminations. The impedance can be expressed in terms of Tchebycheff polynomials of the first and second kinds. Expressions are also derived for the insertion loss and insertion phase shift of ladders containing *N* sections with purely reactive elements. Part 1: 657 of 1947.

621.395.665.1 940

Automatic Control of [volume] Contrast—G. Balz. (*Funk u. Ton*, vol. 5, pp. 79–90; February, 1951.) The theory of the basic hexode volume-expansion circuit is developed, with particular attention to the time constants and the characteristic of the control circuit. In an automatic compression or expansion circuit designed to simulate the operation of a manual system, an additional control voltage depending on the program volume is derived from an auxiliary slow-acting control circuit and applied to the regulator tube in series with the usual control voltage. In the case of recorded sound the additional control is derived from an auxiliary pickup operating in advance of the main pickup.

621.396.6.015.7 941

Speed of Response of the Cathode-Coupled Clipper—P. F. Ordung and H. L. Krauss. (*TV Eng.* (N. Y.), vol. 2, pp. 22–24, 32 and 21, 27; April and June, 1951.) When essentially rectangular pulses were applied to a cathode-coupled clipper circuit, the build-up time of the output voltage was about 50 mμs, either with or without regeneration. Analysis of the effect of variation of the circuit constants indicates that the limiting speed is determined primarily by the time constant of the output circuit. With a sinusoidal input voltage this type of clipper circuit should operate reasonably well at frequencies up to 5 mc.

621.396.611.1 942

Response of a Circuit to a Linear-Frequency-Sweep Voltage—J. Marique. (*Onde élect.*, vol. 31, pp. 313–315; July, 1951.) The universal response curves of Hok (671 of 1949) are transformed into a set of curves which can be applied more readily in practice. These are based on circuit bandwidth corresponding to a 3-db fall in response, and show response as a function of sweep velocity. As this increases the current maximum decreases and occurs later, while the pass band (3 db below the maximum) increases.

621.396.611.1:681.142 943

The Study of Oscillator Circuits by Analog Computer Methods—Han Chang, R. C. Lathrop and V. C. Rideout. (*Proc. NEC* (Chicago), vol. 6, pp. 286–294; 1950.)

621.396.615.18 944

Modified Locked-Oscillator Frequency Dividers—P. G. Sulzer. (Proc. I.R.E., vol. 39, pp. 1535–1537; December, 1951.) The circuit described consists simply of an LC oscillator with a resistor inserted between the oscillatory circuit and the tube. It is capable of operating over a 7-to-1 range of anode voltage when dividing by a factor of 30.

621.396.645 945

Cascading Cathode Followers to provide High Impedance-Transformation Ratios—S. E. Smith and W. J. Kessler. (*Proc. NEC* (Chicago), vol. 6, pp. 129–135; 1950.) The input resistance and impedance-transformation ratio of a cathode-follower circuit can be increased considerably by connecting a second circuit in series so that the load resistor for the first stage is the grid resistor for the second stage. The effective load resistance of the first stage is then considerably greater than the dc circuit resistance. The derivation of the amplification equations is outlined and examples of the application of the circuit are described.

621.396.645:621.395.61 946

Equipment for Acoustic Measurements: Part 1—A Portable General-Purpose Microphone Amplifier using Miniature Valves—D. E. L. Shorter and D. G. Beadle. (*Elec. Eng.*, vol. 23, pp. 326–331; September, 1951.) The amplifier is designed to have the highest possible signal/noise ratio with minimum distortion at the maximum output of 400 mw. Details of the circuits, the mechanical layout and the performance are given.

621.396.645:621.395.623.7 947

Cathode-Follower Loudspeaker Coupling—E. W. Fletcher and S. F. Cooke. (*Electronics*, vol. 24, pp. 118–121; November, 1951.) Excellent square-wave response from 10 cps to 40 kc is obtained from a balanced push-pull cathode-follower stage with four Type-6AS7G double triodes per side (all tubes in parallel), with the low frequency and high-frequency loudspeakers in a dividing network connected directly between the cathodes. The audio power to the speakers is approximately 6w and the power gain is 37 db. With ten 6AS7G tubes per side the audio power is increased to about 33w.

621.396.645.018.424† 948

Design of Wide-Band Amplifiers—H. Behling. (*Frequenz*, vol. 5, pp. 209–217 and 246–249; August and September, 1951.) Over-all amplification *V* as a function of the number of stages *n* is compared for three types of amplifier. The relation is expressed in terms of a parameter *A* dependent on the slope and capacitance of the tubes used and on the required total bandwidth. For a prescribed amplification and bandwidth, the stagger-tuned circuit requires the least number of stages. The V/n characteristics for the single-tuned amplifier are the least favorable since they show a pronounced maximum value for *V*. The double-tuned amplifier shows a similar tendency to a maximum value of *V*, depending on the value of *n*, but the attainable amplification is much greater than for the corresponding single-tuned amplifier and this type affords the best compromise between difficulties of practical adjustment and optimum design. Negative-feedback and stagger-damped circuits are also considered and a practical example is discussed.

621.396.645.029.3 949

A Symmetrical Audio Amplifier with Over-All Negative Feed-Back—W. G. Whittleston. (*N. Z. Jour. Sci. Tech.*, *B*, vol. 32, pp. 25–29; September, 1950.) Amplifiers are described incorporating phase-splitter circuits comprising cathode-coupled triode pairs, with signal input applied to the grid of one triode and negative-feedback voltage applied to the grid of the other. Stability is better than that of the Williamson amplifier (2715 of 1947).

621.396.645.029.3:621.395.61 950

Construction of A.F. Amplifiers for Studio Equipment—H. Schiesser and H. Gathmann. (*Elektrotech. Z.*, vol. 72, pp. 523–525; September, 1951.) A review of developments in the design of microphone amplifiers.

621.396.645.22 951

Medium-Frequency Amplifiers with Reduced Phase Distortion—J. Laplume. (*Onde élect.*, vol. 31, pp. 357–362; August/September, 1951.) Analysis of tuned-transformer and stagger-tuned coupling circuits, showing phase- and amplitude-response diagrams. See also 681 and 1328 of 1949.

621.396.645.371 952

Note on Negative-Feedback Amplifiers—U. Kirschner. (*Frequenz*, vol. 5, pp. 223–230; August, 1951.) The matrix method of representation of a tube circuit (2666 of 1951) is applied to determine the amplification, degree of stability, attenuation and phase distortion of a selective feedback amplifier. Such a circuit is represented as an 8-terminal network and methods of dealing with this are described.

621.396.645.372:621.385.5 953

Miniaturizing Pentode Amplifiers by Positive Feedback—W. B. Anspacher. (*Proc. NEC* (Chicago), vol. 6, pp. 103–111; 1950.) Analysis of a 2-stage pentode amplifier circuit without bypass capacitors, the resulting degeneration being nullified by means of positive feedback between the two screen grids. An expression for the feedback resistance is developed in terms of the amplification factors obtained on removing in succession the screen-grid and cathode bypass capacitors.

621.316.8 954

Bauelemente der Nachrichtentechnik. Teil 2: Widerstände [Book Review]—H. Nottebrock. Publishers: Schiele & Schön, Berlin, 1949, 216 pp., 7.50 DM. (*Arch. elekt. Übertragung*, vol. 4, p. 188; May, 1950.) Gives practical information on the various types of resistors for telecommunication applications, including nonlinear types.

621.396.645+621.396.62 955

Verstärker und Empfänger (Amplifiers and Receivers) [Book Review]—M. J. O. Strutt. Publishers: Springer, Berlin, 2nd revised ed. 1951, 422 pp., 53.55 Swiss francs. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 29, p. 359; September 1, 1951.) A textbook covering all types of amplifiers and receivers, with thorough treatment of fundamental aspects.

GENERAL PHYSICS

537.525 956

High-Frequency Gas-Discharge Breakdown—S. C. Brown. (Proc. I.R.E., vol. 39, pp. 1493–1501; December, 1951.) "High-frequency discharge breakdown is controlled by the process of electron diffusion and, besides the theory of its behavior, the physical limitations of tube size, gas pressure, and frequency for this type of breakdown are given. The particular case of hydrogen is cited. The effects of superimposing a small dc field and a magnetic field on the ac field are also discussed."

537.525 957

The Mechanism of Positive-Ion Collection by a Spherical Probe in a Dense Gas—R. L. F. Boyd. (*Proc. Phys. Soc.*, vol. 64, pp. 795–804; September 1, 1951.) A theoretical treatment of the use of an electronegative spherical probe with dimensions greater than the ionic and electronic mean free paths shows this to be a possible means of determining ionic density in a dense gas discharge, where Langmuir's probe technique is inapplicable. The method depends on a knowledge of the radius of the space-charge sheath surrounding the probe. It is suggested that this is difficult to determine experimentally.

537.525 958

Form of Transient Currents in Townsend Discharges with Metastables—J. P. Molnar. (*Phys. Rev.*, vol. 83, pp. 933–940; September 1, 1951.) The form of the current is calculated for a Townsend discharge stimulated by a pulsed light beam, with particular reference to the current component initiated by metastable effects.

537.525.5:621.385.1 959

Studies of Externally Heated Hot-Cathode Arcs: Part 1—Modes of the Discharge—L. Malter, E. O. Johnson and W. M. Webster. (*RCA Rev.*, vol. 12, part I, pp. 415–435; September, 1951.) A study of the modes of discharge which take place as current increases in a gas-discharge cylindrical diode with inner hot cathode. The anode-glow, ball-of-fire, Langmuir and temperature-limited modes are considered and data are presented for plasma and floating potentials, electron temperature, plasma density and glow distribution for these forms of discharge.

537.533:537.525 960
Studies of γ-Processes of Electron Emission Employing Pulsed Townsend Discharges on a Millisecond Time Scale—J. P. Molnar. (*Phys. Rev.*, vol. 83, pp. 940–952; September 7, 1951.) An experimental investigation of the relative amounts of cathode electron emission due to ions, photons, and metastable atoms in a pulsed Townsend discharge in argon.

537.533.71 961
On the Reflection of Electrons by Metallic Crystals—L. A. MacColl. (*Bell Sys. Tech. Jour.* vol. 30, part 1, pp. 888–906; October, 1951.) The reflection coefficient is calculated for electrons incident normally on a plane face of a metallic crystal, assuming that the potential energy of an electron is a sine function of distance inside the crystal and obeys the classical image-force law outside the crystal.

537.568 962
Concerning the Mechanism of Electron-Ion Recombination: Part 2—M. A. Biondi. (*Phys. Rev.*, vol. 83, pp. 1078–1080; September 1, 1951.) Experiments on samples of He and Ne containing 0.1 per cent Ar indicate that the recombination involves an electron and a molecular ion and that the mechanism is probably dissociative recombination. Part 1: 658 of April (Biondi and Holstein).

538.11:621.318.2 963
The Air-Gap Field of Permanent Magnets—L. Kneissler. (*Elektrotech. u. Maschinenb.*, vol. 68, pp. 393–398; September 1, 1951.) Theoretical investigation of the fundamental consequences following from the assumption that the field of a permanent magnet results from a system of elementary currents.

538.311+621.317.441 964
Axially Symmetric Systems for Generating and Measuring Magnetic Fields: Part 1—M. W. Garrett. (*Jour. Appl. Phys.*, vol. 22, pp. 1091–1107; September, 1951.) A systematic discussion of (*a*) axially symmetric magnetic fields, both central and remote from the origin, (*b*) search coils reporting the field and gradient at a single point, and (*c*) mutual inductors. Universal error-contour maps are derived for the central field or gradient in systems having errors of second, fourth or sixth order, and for hybrid types. Source systems include circular filaments, cylindrical or plane circular current sheets, and thick solenoids of rectangular or notched section. Source constants derived for the particular source type are combined into a set of over-all coefficients that express the field constants for a complete system. Rapid methods are given for computing the source constants and from them all the field derivatives. Tabular aids and reference formulas are given, and the rates of convergence of series for central and remote fields, and for mutual inductors, are discussed. Special systems are briefly described.

538.312:538.56 965
Theory of Multipole Radiations—P. R. Wallace. (*Canad. Jour. Phys.*, vol. 29, pp. 393–402; September, 1951.) Analysis free from approximations is given for the radiation from a given oscillating system of charges and currents. The energy flux and density and the angular-momentum density are expressed in terms of "rotation operators" the expansion of which into eigenfunctions corresponds to separation into electric and magnetic multipoles of all orders.

538.56 966
Application of the Radiation from Fast Electron Beams—H. Motz. (*Jour. Appl. Phys.*, vol. 22, p. 1219; September, 1951.) Correction to paper abstracted in 2411 of 1951.

538.566 967
Total Reflection of an Electromagnetic Wave with Metallic Reflection of the Evanescent Wave—S. Gibellato. (*Nuovo Cim.*, vol. 6, pp. 344–359; September 17, 1949.) Experiments using waves of length 3.43 cm are described in which a plane metal surface is arranged parallel to the surface of a paraffin prism at which total reflection occurs. The evanescent refracted wave is thus reflected and influences the polarization of the totally reflected wave. The variation of polarization with distance between the two reflecting surfaces is plotted.

538.566:535.42 968
Rigorous Theory of the Diffraction of Plane Electromagnetic Waves at a Perfectly Conducting Circular Disk and at a Circular Aperture in a Perfectly Conducting Plane Screen—W. Andrejewski. (*Naturwiss.*, vol. 38, pp. 406–407; September, 1951.) Curves are given which have been calculated from the formulas previously derived by Meixner and the author (see 2767 of 1950). The values obtained by the Levine-Schwinger approximation method are included for comparison. The agreement is only satisfactory for values of $ka<2$, where a is the radius of the disk and k the wave number.

538.566:535.42 969
On the Half-Plane Diffraction Problem—F. G. Friedlander. (*Quart. Jour. Mech. Appl. Math.*, vol. 4, part 3, pp. 344–357; September, 1951.) The diffraction of an arbitrary disturbance is considered. The solution, for all points of space, can be obtained by a modification of the method developed by Hadamard for Cauchy's problem. For the diffraction represented by Hadamard's "elementary solution" a simple expression is found whose Laplace transform agrees with the Green's functions derived from Sommerfeld's two-valued solutions of the wave equation.

538.566:535.42 970
Study of Diffraction by Plane Screens and Application to Lenses for Hertzian Waves—J. C. Simon. (*Ann. Radioélect.*, vol. 6, pp. 205–243; July, 1951.) The general problem of diffraction by a perfectly conducting plane screen with any form of aperture is solved by a perturbation method. Diffraction at a single aperture is studied by considering an auxiliary source in the aperture; approximate solutions are obtained for a circular hole (*a*) small compared with the wavelength, (*b*) of the order of the wavelength. The case of a plane wave incident normally on a screen with equal holes uniformly spaced is related to that of propagation in a rectangular waveguide with two walls of zero impedance and two of infinite impedance. Phase variations in the case of oblique incidence are related to ordinary-wave and extraordinary-wave propagation in crystal optics. An equivalent refractive index <1 can be obtained by using several parallel screens. Measurements indicate that accurate phase correction may be made as in the Schmidt optical system. Application of these principles in the design of an antenna for 3.6 kmc is described.

538.566:537.526.6 971
Waves and Plasma—W. C. Schumann. (*Elektron Wiss. Tech.*, vol. 5, pp. 279–288; September, 1951.) The mode of oscillation and characteristics of waves propagated in plasma layers are discussed. Different combinations of em waves and stationary or moving plasma are considered in turn. 39 references.

539.153:546.28 972
Contribution to the Theory of Impurity Centers in Silicon—W. Baltensperger. (*Phys. Rev.*, vol. 83, pp. 1055–1056; September 1, 1951.) Detailed analysis of the energy of an electron bound to a phosphorus impurity atom in silicon, with components due to the field of the ion, the dipole field of the polarized atoms, and the periodic field of the crystal. The result is in agreement with experiment and with the simple theory assuming a hydrogen-like atomic binding.

541.183.5:621.385.833 973
The Use of the Field Emission Electron Microscope in Adsorption Studies of W on W and Bu on W—J. A. Becker. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 907–932; October, 1951.)

548.24:546.431.824–31 974
Detwinning Ferroelectric Crystals—E. A. Wood. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 945–955; October, 1951.) "Unstrained single crystals of $BaTiO_3$ can be detwinned under the influence of an electric field at elevated temperature, but strained crystals cannot. It seems probable that this is also true of crystals in a polycrystalline body such as a ceramic."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:621.396.9 975
Variation of Meteor-Echo Rates with Radar-System Parameters—McKinley. (See 995.)

523.72 976
Photon Counter Measurements of Solar X-Rays and Extreme Ultraviolet Light—H. Friedman, S. W. Lichtman and E. T. Byram. (*Phys. Rev.*, vol. 83, pp. 1025–1030; September 1, 1951.) Results obtained in a V-2 rocket experiment indicate that solar soft X-rays are important factors in *E*-layer ionization, that H_α radiation penetrates the atmosphere to well below the *E* layer, and that O_2 is rapidly changed to O at heights above 100 km.

523.72:621.396.822 977
Absence of Hydrogen Radiation of Wavelength 21 cm in the Sun—C. de Jager, M. Minnaert and C. A. Muller. (*Nature* (London), vol. 168, p. 391; September 1, 1951.) Using the same equipment as for investigations of the galaxy [982 below (Muller and Cort)] no indication of the hydrogen 1420 mc line could be detected in the solar rf radiation. An explanation of this is given.

523.85:621.396.822 978
An Accurate Determination of the Positions of Four Radio Stars—F. G. Smith. (*Nature* (London), vol. 168, pp. 555–556; September 29, 1951.) The results are given of recent determinations of the positions of four intense radio stars, using an interferometer previously described [121 of February (Ryle et al.)] and also a new interferometer at wavelengths of 3.7 and 1.4 meters. Declination was determined from the periodicity of the record, and by two new methods involving star transit times.

523.854:621.396.822 979
A Search for Long-Period Variations in the Intensity of Radio Stars—M. Ryle and B. Elsmore. (*Nature* (London), vol. 168, p. 555; September 29, 1951.) Interferometer observations were made on a wavelength of 3.7 meters on nearly every day for 18 months. It is concluded that none of the observed radio stars (approximately 100) varies by more than 0.1 magnitude with any period shorter than 1,000 days. See also 121 of February (Ryle et al.).

523.854:621.396.822 980
Galactic Radiation at 18.3 Mc/s—C. A. Shain. (*Aust. Jour. Sci. Res., Ser. A.*, vol. 4, pp. 258–267; September, 1951.) The observations are expressed as equivalent-temperature contours and cover a zone centered on declination $-34°$. The contours are similar in shape but higher in value than those found at 100 mc. The ratio of maximum to minimum temperature is less than at 100 mc even when the difference in antenna directivity is taken into account. Ionospheric absorption is higher and lower than expected near the F_1- and *E*-layer critical frequencies respectively. The collision frequency

at the height of maximum ionization in the *E* layer is $<10^4$/second.

523.854:621.396.822 981

Radio from Galactic Hydrogen at 1420 Mc/s—H. I. Ewen and E. M. Purcell. (*Nature* (London), vol. 168, p. 256; September 1, 1951.) Using a microwave radiometer which measured the apparent radio-temeprature difference between two spectral bands 17 kc wide and 75 kc apart, the 1,420 mc line was first detected in March 1951. The results of subsequent observations are briefly discussed.

523.854:621.396.822 982

The Interstellar Hydrogen Line at 1420 Mc/s and an Estimate of Galactic Rotation—C. A. Muller and J. H. Oort: J. L. Pawsey. (*Nature* (London), vol. 168, pp. 357–358; September 1, 1951.) Using equipment which measured the difference between the radiation received over two frequency bands 25 kc wide and 110 kc apart, curves showing the intensity of galactic radiation as a function of right-ascension have been obtained. From the observed characteristics of the 1,420 mc line, galactic rotational velocities are deduced which agree with values calculated from a schematic model of the galactic system.

551.510.535 983

The Negative-Ion Concentration in the Lower Ionosphere—D. R. Bates and H. S. W. Massey. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 1–13; 1951.) The processes leading to the formation and destruction of negative ions in the upper atmosphere are considered in detail. The daytime equilibrium concentration of negative ions in the *D* and *E* regions does not seem to be sufficiently high to be readily reconcilable with the dynamo theory of the *L* (lunar) and *S* (solar) magnetic variations. Thus the calculated transverse conductivity of the *E* region is found to be such that the magnitude of the *L* current can apparently only be explained if the local tidal motion is 6,000 times greater than at ground level, but this is not supported by radio observations. The transverse conductivity of the *D* region also appears to be less than that demanded by present theory. The fade-out enhancement of the electron concentration in the lower ionosphere is briefly discussed; the possibility that it arises from photo-detachment by Lyman (χ) radiation from solar flares cannot be excluded.

551.510.535 984

Comparison of the Results of Measurements of Ionospheric Absorption made at Two European Stations—K. Rawer. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 38–50; 1951. In French.) Absorption measurements at Slough show that (*a*) the seasonal variation of the noon absorption follows a cos $\chi^{3/4}$ law (χ is the sun's zenith angle), (*b*) the influence of the solar cycle is important, (*c*) day-to-day fluctuations are considerable. Daily values mesaured at Slough and at Freiburg (distance 400 km) mostly show a positive correlation, with coefficient about 0.4. A new method of correlation is proposed.

551.510.535 985

The Contributions of the D and E Regions in Measurements of Ionospheric Absorption—K. Bibl and K. Rawer. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 51–65; 1951. In French.) Observations of ionospheric absorption made on different frequencies reveal an important contribution of selective absorption occurring in the *E* layer, which can be calculated by assuming a parabolic variation of the electron density of the *E* layer and an exponential variation of the collision number. Using the results of this calculation, a new evaluation of the measurements at Slough and Freiburg has been made for a period of 24 months. A graphical method is described which gives the separate contributions of the *E* and *D* layers. The values of *D*-layer absorption thus obtained are lower than those given by the usual method, where the influence of the *E* layer is neglected. In spite of a considerable dispersion, this evaluation leads to the adoption of a fairly constant value for the influence of the *E* layer. It is proposed to use this value in future for reduction of absorption measurements; the corresponding *E*-layer collision number is certainly $<10^6$/second, a value in good agreement with recent temperature and pressure data at 125-km height. See also 984 above (Rawer).

551.510.535 986

Absorption of Solar Energy by Oxygen Molecules in the E Layer—R. Penndorf. (*Jour. Met.*, vol. 7, pp. 243–244; June, 1950.) Results of calculations of energy absorption for heights ranging from 90 to 125 km are plotted; the curve has a maximum at about 102 km and resembles closely that previously given for the number of quanta absorbed (2224 of 1949).

551.510.535 987

Determination of the Number of Collisions in the Ionosphere E and F Regions—K. Rawer, K. Bibl and É. Argence. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 667–669; September 17, 1951.) The collision frequencies were deduced by the classical method from the observed amplitudes of multiple reflections. For the *F* layer measurements were made at four frequencies, during the night, to eliminate the influence of the *D* layer, from November 1949 to April 1950 and during February 1951. For the *E* layer measurements were made at five frequencies, at midday, over a period of two years. Values of 2×10^2 per second and 6–7×10^3 per second respectively were found for the collision frequencies at the middle layers of the *F* and *E* regions. The double focusing effect due to the variation of curvature of the *F* layer is taken into account.

551.510.535 988

The Temperature of the Upper Atmosphere—D. R. Bates. (*Proc. Phys. Soc.*, vol. 64, pp. 805–821; September 1, 1951.) Thermal equilibrium in the *F* layers is studied. The most important energy-loss processes are conduction in the gas and emission by the magnetic dipole connecting the two levels of the ground term of atomic oxygen. Total heat loss is greater than is consistent with the energy gain attributed to ionizing photons. It is possible that the estimate of this energy gain, based on radio measurements, may be much too small. Such measurements give only a lower limit to the electron production rate. The suggestion is tentatively made that heat is supplied to the upper atmosphere mainly by nonobserved ionization.

551.510.535 989

Ionospheric Behaviour in the F_2 Region at Singapore—B. W. Osborne. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 66–78; 1951) Observations made at Singapore from November 1948 have shown the interdependence of F_2-layer maximum ionization density and height with the seasonal occurrence of thick-layer effects. Anomalous behavior was frequent during the morning hours of the December solstice. Layer height changes after sunset and the associated occurrence of the equatorial scatter phenomenon are discussed. Consideration is given to the interpretation of vertical-incidence virtual-height measurements under Singapore conditions of a thick *E* region.

551.510.535:551.594.6 990

Calculation of the Height of the Lower Layers of the Ionosphere from the Waveform of Atmospherics—F. Schindelhauer, A. Schrader and C. Horinger. (*Z. Met.*, vol. 5, pp. 277–284; September, 1951.) Ray-geometry methods of calculation introduced by Laby et al. (2184 of 1940) and developed by Schonland et al. (35 of 1941) are outlined. Oscillographic records of atmospherics obtained between September 1948 and September 1950 at Potsdam indicate reflection at heights of 35, 65 and 80–100 km, corresponding respectively to the *C*, *D* and lower *E* layers. Graphs are presented showing the variation of these heights with time and with solar activity. Sources of disturbance other than lightning discharges are considered responsible for some of the atmospherics. The *C* layer may correspond to the upper boundary of the ozone layer.

551.594.1 991

Diurnal Variations of Atmospheric Electricity and Air-Mass Exchange in the High Alps. The Atmospheric-Electrical Conditions on the Jungfraujoch (3472 m)—H. Israël, H. W. Kasemir and K. Wienert. (*Arch. Met. Geophys. Bioklimatol. A.*, vol. 3, pp. 357–381; August 10, 1951.) Observations made during periods of serveral weeks at different seasons are reported and discussed. In summer, the conditions are of continental type, with potential gradient varying in the opposite sense to conductivity and vertical current. In autumn the conditions tend to oceanic type, with conductivity varying in the opposite sense to potential gradient and vertical current. These results are explained on the basis of the movements of air masses between upper and lower levels.

551.594.22 992

The Work of the Bernard Price Institute of Geophysical Research, 1938–1951—B. F. J. Schonland. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 42, part 8, pp. 241–254; August, 1951.) Some account of the Institute's research on lightning, and of wartime radar developments, is included.

LOCATION AND AIDS TO NAVIGATION

621.39.001.11:621.396.9 993

Information Theory and the Design of Radar Receivers—P. M. Woodward. (Proc. I.R.E., vol. 39, pp. 1521–1524; December, 1951.) "Deals with the problem, frequently encountered in radar, of extracting simple numerical information from a noisy wave form. It is suggested that the only ideal way of doing this is to use the principle of inverse probability and convert the wave form into a probability distribution for the quantity sought. The method is applied to the problem of determining the time delay of a periodically modulated rf wave form in the presence of white Gaussian noise when the undelayed wave form without noise is exactly known. As a result, the matched predetection filter of Van Vleck and Middleton (1919 of 1947) is automatically specified, and the theory of ideal detection is briefly indicated."

621.396.9 994

Terrestrial Radar—A. Flambard. (*Onde élect.*, vol. 31, pp. 261–270 and 320–328; June and July, 1951.) Review of development, with descriptions of European and American equipment.

621.396.9:523.5 995

Variation of Meteor-Echo Rates with Radar-System Parameters—D. W. R. McKinley. (*Canad. Jour. Phys.*, vol. 29, pp. 403–426; September, 1951.) Observations made with crossed-polarization radar systems do not support the suggestion that an ionized meteor trail may act as a strong filter-polarizer of the incident radio wave. Experiments to determine the variation of normal meteor echo rates with transmitter power, antenna gain, and radio wavelength, all confirm Lovell's scattering formula, provided account is taken of the effective broadening of the scattering pattern of the meteor trail with increased wavelength.

621.396.9:526.9 996

Recent Lorac Developments—J. E. Hawkins. (*Proc. NEC (Chicago)*, vol. 6, pp. 218–226; 1950.) Description of a new method of radio surveying which has been tested experimentally in the Gulf of Mexico area, using frequencies near 1,772 and 1,798 kc. Intersecting hyperbolic interference patterns are produced by the radiations from the two pairs of transmitters; "fixes" are obtained directly from the readings of two phase meters in the mobile receiver.

621.396.93.089.6 997
The Calibration of Aircraft Direction-Finders with Particular Reference to Site Selection—J. H. Moon. In 143 of February, for "vol. 15" please read "vol. 14."

621.396.932/.933].2×621.317.789 998
Direction Finder and Flow Meter for Centimeter Waves—K. Morita. (Proc. I.R.E., vol. 39, pp. 1529–1534; December, 1951.) The direction of arrival of the wave is indicated by the minimum response of a dipole located at the focus, and along the axis, of a small paraboloidal reflector. The performance of an experimental model for 10-cm waves is described.

The flow meter indicates directly the active power flow of the wave in magnitude and direction, even when there are standing waves. An energy flow of 1μW/cm^2 at a wavelength of 10 cm has been measured.

621.396.932/.933].2 999
Selecting Critical Components for Matched-Channel Radio Receiving Systems—H. D. Webb. (*Proc. NEC (Chicago)*, vol. 6, pp. 206–217; 1950.) In some df systems receivers are necessary with two or more channels matched in phase and gain characteristics. Matching may be made less critical by using wide-band IF amplifiers having selective filters, or wide-band rf amplifiers with suitably selected bandwidths, or a combination of the two. Selective grading and matching is proposed as an economical means of providing component with the close tolerances requisite for such receivers.

621.396.933 1000
The Civil Aeronautics Administration V.H.F. Omnirange—H. C. Hurley, S. R. Anderson, and H. F. Keary. (Proc. I.R.E., vol. 39, pp. 1506–1520; December, 1951.) Detailed description of a system operating in the frequency band 112–118 mc and producing two signals, one providing a reference phase, while the phase of the other varies directly with the magnetic bearing of an aircraft from the ground station. A phase comparator in the aircraft enables the pilot to determine his magnetic bearing with respect to the station, and to select and fly on a course on any desired bearing. The accuracy of the system, which has been in continuous operation for more than three years, is within about 1.5°.

621.396.933 1001
Rotating Radio Beacon with Angle-Dependent Frequency for Position Finding—H. H. Rust. (*Arch. elekt. Übertragung*, vol. 5, pp. 421–424; September, 1951.) A position-finding system designed for maximum ease of operation on board ship or aircraft uses a rotating beacon in which either carrier or modulation frequency is a function of angle, so that every direction of radiation is associated with a definite frequency. With two such beacons at a suitable separation, position can be determined without ambiguity.

621.396.933:621.317.755 1002
A Video Oscillograph for Testing Distance Measuring Equipment—I. A. Hood. (*AWA Tech. Rev.*, vol. 9, pp. 73–97; September, 1951.) Account of the design, construction and performance of apparatus developed specially for testing the equipment described by Lindsay, Blom and Gilchrist (1659 of 1951).

621.396.933.23 1003
Flight-Path Control—D. L. Markusen. (*Proc. NEC (Chicago)*, vol. 6, pp. 227–237; 1950.) Analysis of the dynamic-stability problems concerned in automatically guiding an aircraft along a path in space as defined by such radio beams as ILS (Instrument Landing System) or omni-range. A control system in which the bank angle is proportional to the off-course error is shown to be inherently unstable, but may be stabilized over a limited range of gain by the addition of a simple phase-adjusting network or by use of a heading reference. The design of such a stabilized system is a compromise between gain, damping, and response to noise; the analytical work has been confirmed by flight tests.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 1004
Calibration of Ionization Gauges for Various Gases at Low Pressures—S. Wagener and C. B. Johnson. (*Jour. Sci. Inst.*, vol. 28, p. 278; September, 1951.)

535.215 1005
An Outline of Some Photoconductive Processes—A. Rose (*RCA Rev.*, vol. 12, part 1, pp. 362–414; September, 1951.) Analysis of photoconductive processes, with the introduction of the concept of a steady-state Fermi limit, to facilitate comparison with experimental results. Explanations are found for (a) current-versus-light curves with exponents between 0.5 and 1.0, (b) the lack of reciprocity between sensitivity and speed of response, and (c) the low values of currents excited in thin evaporated insulating films by light or electron bombardment. The ratio of life time of a free carrier to the observed time-constant is proposed as a figure of merit for a photoconductor.

535.215:546.23 1006
Photoconductivity in Amorphous Selenium—P. K. Weimer and A. D. Cope. (*RCA Rev.*, vol. 12, part 1, pp. 314–334; September, 1915.) Sustained photocurrents are obtained from films of amorphous Se which are highly insulating in the dark. Hole conduction is predominant over electron conduction, the range of the holes in the Se exceeding 10^{-8} cm. The spectral response is a maximum at 4,000–4,500 Å and extends into the far ultraviolet. The maximum response does not coincide with the optical absorption edge, as is usually the case; the absorption edge occurs at much longer wave lengths (6,000 Å). The time constant for the rise and decay of photocurrent is $<50\mu s$. Space-charge-limited currents are observed with excess light for low fields across the material. Evidence for primary and secondary photocurrents is discussed. The application of the television scanning method to photoconductive measurements is described and its advantages and limitations are discussed.

535.215:[546.482.21×546.482.31 1007
Some Aspects of the Photoconductivity of Cadmium Sulfide—R. W. Smith. (*RCA Rev.*, vol. 12, part 1, pp. 350–361; September, 1951.) Measurements of photocurrent, time constant, Hall effect and potential distribution for CdS and CdSe crystals are described, and are discussed with reference to quantum yield.

535.215:546.863.221:621.397.611.2 1008
Properties of some Photoconductors, principally Antimony Trisulfide—S. V. Forgue, R. R. Goodrich and A. D. Cope. (*RCA Rev.*, vol. 12, part 1, pp. 335–349; September, 1951.) A study of those properties of red Sb_2S_3 which make it promising as a target material for television pickup tubes. Plots of the variation of dark current with voltage and of photocurrent with illumination are given, and spectral response, sensitivity and quantum efficiency are discussed. The effects of impurity, heat treatment and method of preparation are considered.

537.311.3+538.632]:546.723-31 1009
Electrical Properties of αFe_2O_3 and αFe_2O_3 containing Titanium—F. J. Morin. (*Phys. Rev.*, vol. 83, pp. 1005–1010; September, 1951.) Electrical conductivity, Hall effect and Seebeck effect were measured for polycrystalline samples of αFe_2O_3 and of αFe_2O_3 containing 0.05–1.0 per cent Ti (n-type impurity). In one set of samples there was a 0.6 per cent excess of Fe (n-type impurity) and a 0.6 per cent deficiency of Fe (p-type impurity) in the second set. Carrier-concentration results indicate that each added Ti ion contributes approximately one electron to the conduction process. Mobilities are found to be <2.0 cm per v/cm. suggesting that conduction involves electrons in the d level of Fe.

537.311.33 1010
Some Results concerning the Partial Differential Equations describing the Flow of Holes and Electrons in Semiconductors—R. C. Prim, III. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1174–1213; October, 1951.) The subject equations are investigated with the aim of establishing some general properties of the flow fields. The results include a number of geometrical characteristics of the vector fields, a suggested reformulation of the partial differential equations restricting carrier concentration and electrostatic potential, and a family of solutions in closed form for the steady-state no-recombination case.

537.311.33:546.289 1011
Hot Electrons in Germanium and Ohm's Law—W. Shockley. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 990–1034; October, 1951.) Ryder's data (1938 of 1951) on the mobility of electrons in electric fields up to 4×10^4 v/cm are analyzed. The mobility decreases because of the influence of scattering by optical modes and because of increases in electron energy. Electron temperatures estimated at 4,000°K have been produced in specimens having atomic-vibration temperatures of 300°K. The critical drift velocity above which there are deviations from Ohm's law is about 2.6×10^6 cm.

538.652:669.157.82 1012
The Magnetostriction of Single Crystals of Iron-Silicon Alloys—W. J. Carr and R. Smoluchowski.)*Phys. Rev.*, vol. 83, pp. 1236–1243; September 15, 1951.)

546.23:[535.323+535.343 1013
Optical Properties of Selenium—J. J. Dowd (*Proc. Phys. Soc.*, vol. 64, pp. 783–789; September 1, 1951.) The refractive index of amorphous Se and the absorption coefficient of amorphous and single-crystal Se were measured over a wide frequency range. The edge of the absorption band for amorphous Se is at about 0.60μ, corresponding to a band spacing of 2.05 ev. A discrepancy between the value obtained for the infrared refractive index and that deduced from the dielectric constant of amorphous Se indicates a further absorption band in the infrared.

546.28:548.55 1014
Measurement of the Elastic Constants of Silicon Single Crystals and their Thermal Coefficients—H. J. McSkimin, W. L. Bond, E. Buehler and G. K. Teal. (*Phys. Rev.*, vol. 83, p. 1080; September 1, 1951.) Measurements of velocities of propagation for both shear and longitudinal waves were made at frequencies in the range 8–12 mc, and the three independent elastic constants and their temperature coefficients were evaluated. The results are tabulated.

546.74:539.32]:534.321.9.012 1015
Frequency Dependence of Elastic Constants and Losses in Nickel—Bosorth, Mason and McSkimin. (See 887.)

621.315.61 1016
Electro-Ceramics, with Special Reference to Pyrophyllite—N. E. Hyde. (*Elec. Eng.*, vol. 23, pp. 336–340; September, 1951.) A brief review of the origin and history of ceramics, and of the processes involved in their manufacture, together with an account of the

more important characteristics of pyrophyllite ceramics recently developed. Mechanical impact strength is higher than that of any other other electrical porcelain, volume resistivity is high at elevated temperatures, shrinkage after firing is very small, and the ease of machining the unfired material makes it particularly suitable for the rapid construction of a wide range of shapes.

621.315.61:539.23 1017
The Electric Tunnel Effect across Thin Insulator Films in Contacts—R. Holm. (*Jour. Appl. Phys.*, vol. 22, p. 1217; September, 1951.) Correction to paper abstracted in 2455 of 1951.

621.315.612.8:546.817.831.4 1018
Antiferroelectric Structure of Lead Zirconate—E. Sawaguchi, H. Maniwa and S. Hoshino. (*Phys. Rev.*, vol. 83, p. 1078; September 1, 1951.) Oscillation and powder photographs, and polarization-microscope examination of a small untwinned crystal, indicate an orthorhombic structure and that the material is antiferroelectric.

621.315.612.8:546.817.831.4:537.228.1 1019
Piezoelectric Effect in Lead Zirconate—S. Roberts. (*Phys. Rev.*, vol. 83, p. 1078; September 1, 1951.) Piezoelectric effects in $PbZO^3$ ceramic disks were found to be only just detectable. This is explained by the structure of the material discussed by Sawaguchi, Maniwa and Hoshino (1018 above).

621.315.614/.617:53.093/.096 1020
Deterioration of Organic Polymers—B. S. Biggs. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1078–1102; October, 1951.) A general review of deterioration processes in polymers. Changes with time of the properties of such materials as rubbers, plastics, textiles, and varnishes are usually the result of chemical reaction with components of the atmosphere. The mechanisms of these reactions and some methods of preventing or retarding them are discussed.

669.3:621.314.632.1 1021
Carbon, Oxygen, and Sulfur Content of Chilean Coppers as Related to Cuprous-Oxide Rectifiers—C. C. Hein and W. H.. Hickam. (*Jour. Appl. Phys.*, vol. 22, pp. 1192–1195; September, 1951.)

MATHEMATICS

681.142 1022
A Quarter-Square Multiplier using a Segmented Parabolic Characteristic—B. Chance, F. C. Williams, Chia-Chih Yang, J. Busser and J. Higgins. (*Rev. Sci. Instr.*, vol. 22, pp. 683–688; September, 1951.) Description of equipment with a delay time <40 μs and accuracy within ±1 per cent.

681.142 1023
An Analog Computer for Indeterminate Mechanical Structures—J. P. Corbett and J. F. Calvert. (*Proc. NEC* (Chicago), vol. 6, pp. 315–332; 1950.)

681.142 1024
A Versatile Small-Scale Analog Computer—J. T. Carleton. (*Proc. NEC* (Chicago), vol. 6, pp. 308–314; 1950.) Description of a small computer which can be used either by itself or in conjunction with the larger Anacom equipment (2547 of 1949).

681.142:512.3 1025
Circuit for Generating Polynomials and Finding their Zeros—F. W. Bubb, Jr. (Proc. I.R.E., vol. 39, pp. 1556–1561; December, 1951.)

517.564.3(083.5) 1026
Tables of the Bessel Functions of the First Kind of Orders Seventy-Nine through One Hundred and Thirty-Five [Book Notice]—Staff of Computation Laboratory of Harvard University. Publishers: Harvard University Press, 1951, $8.00. (Proc. I.R.E., vol. 39, p. 1579; December, 1951.)

MEASUREMENTS AND TEST GEAR

621.3.018.4(083.74):621.396 1027
[British] Standard Frequencies—(*Wireless World*, vol. 57, p. 501; December, 1951.) A revised schedule of transmissions from Rugby is announced, as follows: 60 kc, 1,029–1,130 and 1,429–1,530 GMT; 5 mc, 0544–0615 GMT; 10 mc, 0628–0700 GMT.

621.3.087.45:[551.594.11+551.594.13 1028
An Apparatus for Simultaneous Registration of Potential Gradient and Air-Earth Current (Description and First Results)—H. W. Kasemir. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 32–37; 1951.) Description of equipment in which the capacitors and resistors are chosen so that minor variations of gradient and current are suppressed, in order to give a clear display of the diurnal variations.

621.317:621.396.82 1029
Technique for the Measurement of Interference Voltages—J. Pfister. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 29, pp. 321–328; September, 1951, In German.) Reference is made to standards formulated by the C.I.S.P.R. for the measurement of interference voltages. An evaluation is made of the output from the selective IF circuit of the measuring apparatus, comprising two cascaded critically coupled bandpass filters, in response to an input of given transient character, viz. ac pulse, ac step, dc pulse or dc step. The voltage obtained in the following detector stage is then determined. The results are compared with those obtained by the same calculation procedure for a filter with an ideal narrow passband, and are discussed in relation to the method of calibrating the apparatus.

621.317:621.396.822.029.3 1030
The Generation and Measurement of Low-Frequency Random Noise—R. R. Bennett and A. S. Fulton. (*Jour. Appl. Phys.*, vol. 22, pp. 1187–1191; September, 1951.) Methods are discussed for determining the important characteristics of low-frequency noise, such as the mean value, spectral density, amplitude distribution and autocorrelation function. Particular attention is devoted to the length of time necessary to establish satisfactory estimates of the properties of low-frequency noise. A brief description is given of a noise generator with a uniform power spectrum from zero to 25 cps.

621.317.335.2.029.5† 1031
Measurement of Capacitances at High Frequency—H. H. Emschermann and O. Zinke. (*Arach. Tech. Messen.*, no. 188, pp. T100–T101.; September, 1951.) Principles of measurement, range and accuracy of various circuits are noted. In the simple current voltage measurement, error can be kept below 1 per cent. Arrangements of the double voltage divider for large and small capacitances, and bridge circuits for measurements down to 1pF at 20 mc and to 10^{-4}pF at 1 mc, are shown. Application of the double-T bridge is described for impedances in which the resistive component is large.

621.317.335.3†:621.365.55† 1032
Measuring Dielectric Properties during H. F. Heating—E. Mittlemann. (*Proc. NEC* (Chicago), vol. 6, p. 79; 1950.) Summary only. Description of method and apparatus, with experimental results.

621.317.335.3.029.64† 1033
Measuring the Dielectric Constant and the Loss Angle of Solids at 3000 Mc/s—M. Gevers. (*Philips Tech. Rev.*, vol. 13, pp. 61–70; September, 1951.) Dielectric constant and loss angle are determined from the detuning and reduction of quality factor of a cavity resonator (E_{010} mode) when the dielectric is introduced. Quality factors as high as 17,190 (98 per cent of the theoretical value) have been obtained by careful surface finishing of the resonators. A reflex klystron is used as oscillator and a silicon crystal as detector. Measured values for various materials are tabulated and formulas and graphs are given which simplify the determination of the required quantities.

621.317.361 1034
Frequency Checking of Mobile Equipment—M. H. Diehl and C. J. Statt. (*Electronics*, vol. 24, pp. 138–184; November, 1951.) Outline of a method of making fine frequency adjustments. A standard input frequency of 10 kc triggers a multivibrator circuit producing 0.2-μs pulses with a recurrence frequency of 3333⅓/seconds which are fed to a mixer together with the frequency under test. Channel spacing is such that for correct adjustment of the frequency under test, one mixer output frequency is 1⅔ kc. This is used in conjunction with a frequency of 1⅔ kc derived from the 10-kc standard input to produce a stationary ellipse pattern on a cro when the transmitter frequency is exactly adjusted.

621.317.382 1035
New Method for Measurement of Active Power of H. F. Generators—F. Alf (*Elektrotech. Z.*, vol. 72, pp. 541–543; September 15, 1951.) The grid rectification inherent in the operation of the oscillator tube is used to obtain a dc proportional to the power output. Ordinary commercial watt meters are used, and measurements are accurate to within 5–7 per cent. Results are independent of frequency up to frequencies at which transit-time effects occur. Application is to generators for industrial heating.

621.317.441+538.311 1036
Axially Symmetric Systems for Generating and Measuring Magnetic Fields: Part 1.—Garrett. (See 964).

621.317.7 1037
Self-Balancing Instruments and their Application to Nucleonic Measurement—R. S. Medlock and W. A. Kealy. (*Jour. Brit. IRE*, vol. 11, pp. 393–405; September, 1951.) 1951 Radio Convention (London) paper. General principles of operation of mechanical and electronic self-balancing recording instruments, and standard applications of these instruments to measurements associated with atomic-energy projects, are described. Circuits are illustrated which permit simple addition, subtraction, multiplication, division, derivation of roots and powers, differentiation and integration.

621.317.7:621.396.932/.933 1038
Remote Monitoring of Naval and Air Service Transmissions—J. Marique. (*Onde élect.*, vol. 31, pp. 331–341; August/September, 1951.) Report of monitoring measurements carried out in Brussels by the C.C.R.M. (Centre de Contrôle des Radiocommunications des Services mobiles), including frequency checking, field-strength measurement, panoramic recording of frequency bands occupied, and spectrum analysis. Equipment is described and results are shown diagrammatically of a statistical analysis of frequencies in use, deviations from allocated frequency, and beacon field strengths.

621.371.71 1039
A Combined Current Indicator and Integrator—W. A. Higinbotham and S. Rankowitz. (*Rev. Sci. Instr.*, vol. 22, pp. 688–690; September, 1951.) An instrument suitable for measuring the beam current in a particle accelerator and covering the range 0.001 to 500 μA full scale.

621.317.71.029.6 1040
Measurement of Current at High Fre-

quency—O. Zinke. (*Radio franç.*, pp. 6–18; September, 1951.) Description of photoelectric, rectifier and thermocouple instruments. Design and application of the latter are particularly discussed; rating and characteristics of different elements are tabulated (*a*) for vacuum-type thermo-elements rated up to 200 ma, (*b*) for Miniwatt-S types rated up to 300 ma.

621.317.725 1041
Valve Voltmeter without Calibration Drift—M. G. Scroggie. (*Wireless World*, vol. 58, pp. 14–18; January, 1952.) Description of adaptor, of infinite input resistance and zero output resistance, for use with any dc voltmeter.

621.317.733 1042
A Self-Tracking Bridge Detector for Audio-frequency Bridges—J. L. Upham, Jr. (*Rev. Sci. Instr.*, vol. 22, pp. 659–664; September, 1951.) A tuned bridge detector with constant bandwidth of 5cps is described. It is automatically tuned to the output frequency of the associated signal generator. Its sensitivity and inherent noise are such as to permit detection of a 0.5-μv signal across a 1-MΩ input over the frequency range 100 cps-20 kc. The detector may be used for maintaining bridge balance during slow variations of the measured impedance, the variations being recorded in terms of the movements of the balancing controls.

621.317.74:621.397.2 1043
Television Streaking Test Set—R. K. Seigle. (*Electronics*, vol. 24, pp. 96–99; November, 1951.) Description of equipment for point-to-point testing of television transmission systems such as coaxial cables, rf links, etc. The apparatus requires blanking and synchronization inputs, and generates stepped wave forms at video frequency which, when fed into properly adjusted television receiving equipment, produce a sharp rectangular picture. The height, width and location of the rectangle can be varied. Maladjustment of the receiving equipment is manifested as a smearing of the rectangular picture.

621.317.755 1044
A Scanning Method for Simultaneous Display of Several Phenomena with a Single-Beam Oscillograph—R. Classen, F. W. Gundlach and F. Lentze. (*Arch. Tech. Messen*, p. T106; September, 1951.) Description of the equipment, with details of the pulse-generator circuits.

621.317.755 1045
A Cathode-Ray Oscillograph for Impulse Testing—W. G. Fockler. (*Proc. NEC* (Chicago), vol. 6, pp. 391–399; 1950.)

621.317.755:535.88 1046
A Portable Projection Oscilloscope—V. Wouk. (*Proc. NEC* (Chicago), vol. 6, pp. 380–390; 1950.) Description of equipment producing an oscillogram 14×11 inches which can be viewed in rooms with normal lighting.

621.317.755:621.317.6.029.3 1047
The Production Model of the Automatic A. F. Response Curve Tracer—G. L. Hamburger. (*Jour. Brit. IRE*, vol. 11, pp. 165–201; May, 1951.) Description of the production instrument developed from the experimental model previously described (148 of 1949). Layout and circuit details are given and operation of circuits is analyzed. The curves are displayed against a logarithmic-coordinate framework and can be either viewed for about one minute or photographed.

621.317.755.087.6 1048
A Six-Channel Cathode-Ray Recording Oscillograph—W. D. Tilton, Jr. (*Proc. NEC* (Chicago), vol. 6, pp. 373–379; 1950.)

621.317.755.088 1049
Astigmatism Correction for Oscilloscopes—H. O. Hoadley. (*Rev. Sci. Instr.*, vol. 22, pp. 706–708; September, 1951.) Astigmatism in oscilloscopes produces different degrees of sharpness for horizontal and vertical lines. It may be corrected by altering the average potential of one pair of deflector plates relative to the other. Circuits for balanced and unbalanced signal inputs are given. Some commercial oscilloscopes provide for astigmatism correction by varying the potential of the second anode, and this method is recommended for oscilloscopes using dc amplifiers.

621.317.757 1050
A High-Resolution Spectrum Analyzer—T. Miller and D. Sims. (*Proc. NEC* (Chicago), vol. 6, pp. 513–517; 1950.) Description, with circuit details, of an analyzer for use at frequencies of the order of 50 kc. A continuous automatic recording system gives records with a frequency range of 3 kc and resolution to 20 cps. Typical records of sideband structure are shown for carriers with modulation frequencies of 250 and 20 cps respectively.

621.317.761:538.569.4.029.64/.65 1051
New Techniques in Microwave Spectroscopy—W. E. Good. (*Proc. NEC* (Chicago), vol. 6, pp. 29–37; 1950.) A description is given of a sensitive microwave spectroscope, with particular reference to the low-noise input circuit and the characteristics of the Si-crystal detector. The absorption lines to be measured are displaced in frequency by application of an 85-kc alternating voltage to the absorbing gas. The resultant modulation is detected by the crystal, amplified by a high-gain 85-kc amplifier, and displayed on a cro. Lines of frequency up to 40 kmc can be measured with an accuracy to within one part in 10^6. The frequency-measurement equipment, for which a detailed circuit diagram is given, uses a Si crystal for generation of a series of harmonic frequency markers spaced at intervals of 500 mc, interpolation being effected by means of a calibrated receiver.

621.317.78:535.214 1052
The Measurement of Microwave Power by Radiation Pressure—A. L. Cullen. (*Engineering* (London), vol. 172, pp. 377–378; September 21, 1951.) The method used is analogous to that for the classical measurements of radiation pressure of light. Microwave radiation is directed on to a reflector surface within the vertical leg of a waveguide T system, this arrangement ensuring that reflected power is not returned to the source. It is emphasized that the method is absolute. See also 2771 of 1951.

621.317.789+621.396.932/.933].2 1053
Direction Finder and Flow Meter for Centimeter Waves—Morita. (See 998.)

621.317.799:621.396.61:621.396.931 1054
Mobile-Transmitter Testing Set—G. J. Kent. (*Electronics*, vol. 24, pp. 106–109; November, 1951.) Outline description of test equipment for PM or FM transmitters operating in the frequency bands 30–44 mc and 152–175 mc. With minor modifications, AM transmitters operating at any frequency from 540 kc to 110 mc can also be tested. Measurements of rf power output, af sensitivity, signal/noise ratio, and speech intelligibility, can be made in a few minutes.

621.396.615.14:621.396.621.001.4 1055
Microwave Generator with Crystal Control—W. F. Marshall. (*Electronics*, vol. 24, pp. 92–95; November, 1951.) Based on 1950 National Electronics Conference paper (*Proc. NEC* (Chicago), vol. 6, pp. 504–512; 1950.) A portable 3,100-mc signal generator for field work. The initial oscillation is crystal-controlled near 50 mc within a band of width 1¾ mc. After multiplication to 300 mc the signal is passed to a Si crystal, where *S*-band harmonics are produced. Preselection is required to utilize a specific output harmonic. Substituting control crystals in two channels gives a frequency range of up to 600 mc without changing other circuit components.

621.396.615.14.029.62/.63 1056
A 20 to 1000 Mc/s Sweep Oscillator—J. E. Ebert and H. A. Finke. (*Proc. NEC* (Chicago), vol. 6, pp. 499–503; 1950.) The full range is covered in a single continuous tuning adjustment. The tank circuit is a conventional λ/4 coaxial-line resonator at the higher frequencies; as its plunger moves back for the lower frequencies, the exposed section of the inner conductor gradually changes from a solid rod to a helix of increasing pitch. The frequency sweep is effected by rotation of a specially shaped capacitor plate close to the high-impedance end of the tank cavity.

621.397.61.001.41 1057
Measuring Television Transmitter Amplitude Characteristics—Ruston. (See 1133.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

548.0:537.228.1].001.8 1058
Piezoelectric Crystals as Sensing Elements of Pressure, Temperature, and Humidity—E. A. Roberts and P. Goldsmith. (*Elec. Eng.* (N.Y.), vol. 70, pp. 776–780; September, 1951.) Summary of 1951 A.I.E.E. Great Lakes District Meeting paper. Pressure may be measured either by its effect on the *Q*-value of a crystal or by the change in crystal frequency due to differential air loading. Temperature is measured by its effect on frequency, and humidity by the change of frequency due to the deposit of moisture on the crystal surface. Types and cuts of crystal for optimum results are discussed.

621.52:681.142 1059
Simulation—Its Place in System Design—H. H. Goode. (PROC. I.R.E., vol. 39, pp. 1501–1506; December, 1951.) Discussion with particular reference to the application of analogue and digital computers.

621.3.012.8:629.11.012.8 1060
Construction of an Electrical Analogue of a Motor-Car Suspension System—R. Lansard. (*Onde élect.*, vol. 31, pp. 307–312; July, 1951.)

621.316.7 1061
Automatic Control—A. Tustin. (*Nature*, (London), vol. 168, pp. 400–406; September 8, 1951.) Report of conference organized by the Department of Scientific and Industrial Research, July, 1951. Research aspects of the subject were discussed, and the relation between work in different fields was emphasized.

621.316.722.1 1062
Automatic Control of Inaccessible Terminal Voltages—R. L. Cosgriff and E. H. Gamble. (*Proc. NEC* (Chicago), vol. pp. 434–442; 1950.)

621.317.083.7+621.395.44]:621.316.1 1063
Telemetering Systems and Channels for a Large Interconnected Power System—G. K. Duff. (*Elec. Eng.*, (N.Y.), vol. 70, pp. 796–801; September, 1951.) Essential text of 1951 A.I.E.E. Summer General Meeting paper.

621.365.54† 1064
Induction Heater Control System—R. W. Ketchledge. (*Bell Lab. Rec.*, vol. 29, pp. 405–409; September 1951.) A photometric comparison method, based on the radiation from the heated workpiece is used for controlling the operation of the power source in brazing and soldering work. Powers up to 40 kw at 10 kc are thus controlled.

621.38.001.8 1065
Industrial Metal-Detector Design—C. R. Schafer. (*Electronics*, vol. 24, pp. 86–91; November, 1951.) Discussion of principles of design for specific purposes; description and circuit diagrams of two RCA metal detectors.

621.383:621.384.6 1066
Photoelectric Control Circuits for the Ion Source of a Pressure-Type Electrostatic Generator—L. O. Herwig. (*Rev. Sci. Instr.*, vol. 22, pp. 668–672; September, 1951.)

621.384.3:621.383.27† 1067
An Image Converter to Extend the Useful Range of Photo-multipliers to Longer Wavelengths—E. R. Holiday and W. Wild. (*Jour. Sci. Instr.*, vol. 28, pp. 282–283; September, 1951.) A spectrographic application is discussed in which a photomultiplier whose upper wavelength limit is 0.63 μ is to be used for measurements up to 1.2 μ. The adaptation is effected by interposing an infrared-sensitive image converter with Ni-treated ZnS screen between source and multiplier.

621.384.6† 1068
Linear Electron Accelerator to One Million Volts—A. T. Starr, G. King, and L. Lewin. (*Elec. Commun.*, vol. 28, pp. 186–194; September, 1951.) Problems arising in the design of various types of linear accelerator are discussed. A description of one using an E_0 guide with irises is given and its performance is outlined: a velocity corresponding to 1–2 Mev is achieved in a meter-length of guide. A possible line of development for a 10-Mev accelerator is briefly considered.

621.384.6†:621.319.339 1069
The Electrostatic Accelerator as a Source of Ionizing Energy—J. G. Trump. (*Elec. Eng.*, (N.Y.), vol. 70, pp. 781–787; September, 1951.) Essential text of 1951 A.I.E.E. Summer General Meeting paper. Discussion of application of es accelerators in nuclear research, for production of very high voltage X-rays, and for sterilization by means of high-energy electrons.

621.384.612.2†:621.396.6 1070
Radio Frequency for a Synchrocyclotron—A. J. Poté. (*Electronics*, vol. 24, pp. 100–105; November, 1951.) Description of the construction of the oscillator and modulator units, with the connections to the dee, of a FM cyclotron. Vacuum-type capacitors were found effective for insulation of the dee, on which an average voltage of 8.5 kv was obtained with an oscillator anode power input of 4.5 kw.

621.385.833 1071
Inorganic Replication in Electron Microscopy—C. J. Calbick. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 798–824; October, 1951.)

621.385.833 1072
A Theory of the Nearly Symmetrical Independent Electrostatic [electron] Lens—É. Regenstreif. (*Ann. Radioélect.*, vol. 6, pp. 244–267; July, 1951.) Explicit formulas are established for the fundamental optical properties of the elliptical lens in terms of its geometrical and electrical structure. See also 2793 of 1951.

621.385.833 1073
An Experimental Study of the Illuminating System of the Electron Microscope—M. L. De. (*Indian Jour. Phys.*, vol. 24, pp. 303–308; July, 1950.) Observations were made of the variation of electron-crossover size with grid-filament spacing for the electron microscope at Calcutta [1747 of 1949 (Dasgupta et al.)]. Results are discussed in relation to the quality of the final image.

621.385.833 1074
Tge Aberrations of Magnetic Electron Lenses due to Asymmetries—P. A. Sturrock. (*Phil. Trans. A*, vol. 243, pp. 387–429; July 6, 1951.) The relation between machining defects of the objective lens in an electron microscope and the resulting aberrations is investigated mathematically. A computational procedure is developed for fixing tolerances for a proposed lens design.

621.385.833 1075
On a New Test Method for Spherical Aberration of Electron Lenses—O. Klemperer. (*Proc. Phys. Soc.*, vol. 64, pp. 790–794; September 1, 1951.) "The focus of rays from a lens with spherical aberration appears as a spot surrounded by a discrete halo ring if a diaphragm with fine circular aperture is placed across the beam in front of the focus. The diameter of the halo allows an estimate of the magnitude of the aberration involved. The geometry of rays forming the halo is explained here by schematic drawings. Practical application of the halo test is illustrated by examples. In particular, the negative spherical aberration produced by an electronic space charge in a saddle field lens is demonstrated."

621.385.833 1076
The Electron-Optical Characteristics of Electrostatic Systems with Apertures Deviating from the Idealized Form—B. M. Rabin, A. M. Strashkevich and L. S. Khin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 438–447; April, 1951.) The fields of rectangular and elliptic apertures were investigated experimentally using an electrolyte tank. Formulas are derived for determining the potential distribution with great accuracy

621.385.833:541.183.5 1077
The Use of the Field Emission Electron Microscope in Adsorption Studies of W on W and Ba on W—J. A. Becker. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 907–932; October, 1951.)

621.387.424† 1078
Construction of the External-Cathode Geiger Counter—M. L. MacKnight and R. L. Chasson. (*Rev. Sci. Instr.*, vol. 22, pp. 700–701; September, 1951.) Description of the method adopted for quantity production.

621.387.424† 1079
A Circular Geiger-Mueller Counter—G. Ensell and S. D. Chatterjee. (*Rev. Sci. Instr.*, vol. 22, p. 700; September, 1951.) Description of the method of construction.

621.38.001.8 1080
Electronics [Book Review]—J. Millman and S. Seely. Publishers: McGraw-Hill, New York, 2nd ed. 1951, 559 pp., $7.25. (Proc. I.R.E., vol. 39, p. 1578; December, 1951.) ". . . primarily intended as a textbook to set the ground work for specialized courses in communications, electron-tubes, industrial electronics, etc., it would be a welcome addition to the reference shelf of most practicing engineers."

PROPAGATION OF WAVES

538.566:551.510.535 1081
The Production of Harmonics in the Ionosphere—K. Försterling and H. O. Wüster. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 22–31; 1951, In German.) Fuller discussion than that previously given (721 of 1951.)

621.396.11 1082
The Source of Long-Distance Backscatter—W. G. Abel and L. C. Edwards. (Proc. I.R.E. vol. 39, pp. 1538–1541; December, 1951.) By comparing the delay time of the leading edge of the back scatter with that of the response from a beacon transponder, the source of back scatter was shown to be the ground at and beyond the skip distance. The scatter forming the leading edge of the pattern sometimes arrived from directions off the principal axis of the antenna, and preceded the amplitude peak due to scatter along the axis.

621.396.11:551.510.535 1083
The Paths of an Electromagnetic Signal in the Ionosphere—É. Argence. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 607–608; September 10, 1951.) The approximate formulas for the refractive index of an ionized medium derived in 3094 of 1951 are used as a basis for further investigations of wave paths in the plane of the magnetic meridian. Analysis is given for the ordinary and the extraordinary ray for vertical incidence, and for the ordinary ray for oblique incidence. Three types of path are possible in each case.

621.396.029.62 1084
Propagation of V.H.F. via Sporadic E—T. W. Bennington. (*Wireless World*, vol. 58, pp. 5–9; January, 1952.) Clouds of abnormally dense ionization occurring within the normal *E* region can enable waves of frequency up to 100 mc to be propagated over oblique paths. Temperate-zone-type records of sporadic-*E* obtained by vertical-incidence measurements at Slough, Fraserburgh, De Bilt, Lindau, Freiburg, Domont and Poitiers are analyzed. It appears possible to trace the growth and movement of these clouds from observations at a number of stations.

621.396.11.029.62 1085
U.S.W. Propagation in the 30—100-Mc/s Range—J. Grosskopf. (*Fernmeldetech. Z.*, vol. 4, pp. 411–414 and 441–451; September and October, 1951.) Field-strength measurements made by the German Post Office over distances within the range of vision are analyzed. Factors investigated include the influence of ground reflections near transmitter or receiver on the shape of the field, the influence of undulations or hilliness on attenuation, and the diffracting effect of obstacles. Irregularities in the field-strength/distance curves are traces to relatively simple causes such as double reflections. Diffraction measurements indicate that in general the classical diffraction formulas are inapplicable, and that the analysis of the propagation process must take account of the nature of the terrain.

621.396.8.029.55 1086
Reception of Transatlantic Signals of Frequencies near 30 Mc/s—J. Maire. (*Ann. Radioélect.*, vol. 6, pp. 197–204; July, 1951.) Results of systematic reception tests made near Paris from 1937 to 1940 and from 1948 onwards are shown diagrammatically and discussed. The transmissions recorded were from the New York region, including harmonics of commercial transmissions, from WWV and amateur stations. Regularity of reception throughout the above two periods is related to the sunspot cycle, the time of sunset at the mid-point of the path, and the predicted muf. Reception of Buenos Aires transmissions on about 27.5 mc was, with minor exceptions, consistently good during much of the daytime from 1946 to 1950, and was almost completely free from echo disturbances.

621.396.812.3 1087
Further Statistics of Fade-Outs—D. Stranz. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 79–82; 1951.) Investigation, for the year 1949, of the weak absorption effect of the corpuscular cone emitted by the sun, previously detected by the statistical method of sample days, gave a result very similar to that for 1948 (3146 of 1950).

621.396.812.3:551.510.535 1088
Some Random Fading Records with Short-Wave Signals—P. M. Das and S. R. Khastgir. (*Indian Jour. Phys.*, vol. 24, pp. 277–280; July, 1950.) Rapid variations of the intensity at Dacca of 4.84-mc signals from Calcutta did not agree with Rayleigh's random-scattering formula. Possible explanations of the discrepancy based on diverse reflections from the ionosphere are discussed. See also 981 of 1951 (Khastgir and Das).

RECEPTION

621.396.62+534.874.1]:519.272.11 1089
The Correlation Function in the Analysis of Directive Wave Propagation—Nodtvedt. (See 882.)

621.396./397].621 1090
The Design of a Combined Television and

Radio Receiver—A. B. Bamford. (*Jour. Telev. Soc.*, vol. 6, pp. 253–263; July/September, 1951.) Details are given of a circuit designed to facilitate the production of a range of receivers providing complete coverage of the television channels and the medium- and long-wave broadcasting bands, using a common chassis and as many common sub-assemblies as possible.

621.396.621 1091
Radio Feeder Unit—J. F. O. Vaughan. (*Wireless World*, vol. 57, pp. 480–484; December, 1951.) Detailed description of a pretuned receiver providing switched selection of four stations, three medium-wave and one long-wave. The circuit comprises two rf stages, diode detector and af amplifier to compensate for losses due to tone-control circuits, with separate af tube for pickup input. The output is suitable for feeding to a high-quality amplifier.

621.396.621 1092
Sensitive T.R.F. Receiver—S. W. Amos and G. G. Johnstone. (*Wireless World*, vol. 57, pp. 452–456; November, 1951.) Details are given of the circuit and adjustment of a 3-tube receiver with 'amplified agc,' for medium and long waves.

621.396.621:621.396.619.13 1093
Design for an F.M. Receiver—J. G. Spencer (*Wireless World*, vol. 57, pp. 440–444 and 487–490. November and December, 1951.) Detailed description of a simple 7-tube receiver for the 90-mc band, designed to be comparable in cost with a medium-price AM receiver while realizing the improvement in background noise associated with FM. A triple-diode-triode tube serves for both discriminator and first audio amplifier. The IF-alignment method is described and performance figures are given.

621.396.621.54 1094
The Ionodyne, invented for the Ionophone; will it replace the Superheterodyne?—M. Bonhomme. (*Toute la Radio*, no. 160, pp. 291–295; November, 1951.) Discussion of a proposed receiver circuit for use with the ionophone (896 above). Two oscillators are used, one corresponding to the local oscillator of the normal superheterodyne, while the second is tuned to the frequency of the received carrier wave, by which it is accurately synchronized. This enables the depth of modulation of the input signal to be varied, a variable proportion of the input signal and of an unmodulated voltage from the synchronized oscillator being applied to the mixer together with the local-oscillator voltage. The arrangement thus resembles Tucker's synchrodyne circuit (525 and 526 of 1948), but in the ionodyne the synchronization is effected in phase, whereas it is in antiphase in the synchrodyne.

621.396.822 1095
The Power Spectrum of a Narrow-Band Noise Passed through a Nonlinear Impedance Element—J. L. McLucas and R. C. Raymond. (*Jour. Appl. Phys.*, vol. 22, pp. 1211–1213; September, 1951.) Noise centered at 1.6 mc, with the spectral shape of an error function with standard deviation of 3.8 kc, was passed through a Type-1N34 crystal diode. The observed spectral distributions of output power up to the sixth harmonic are in good qualitative agreement with Middleton's theory (3238 of 1948).

621.396.822 1096
On the Theory of Random Noise, Phenomenological Models: Parts 1 and 2—D. Middleton. (*Jour. Appl. Phys.*, vol. 22, pp. 1143–1163; September, 1951. Correction, *ibid.*, vol. 22, p. 1326; November, 1951.) Various models of electron noise are considered: (*a*) nonoverlapping periodic noise waves, (*b*) nonoverlapping nonperiodic disturbances, (*c*) Poisson noise. An attempt is made to determine explicitly the second-order probability density in the important stationary cases. For the first two models this is impracticable except in the simplest cases, while for Poisson noise an explicit treatment is possible for impulsive random noise, nearly normal random noise and normal random noise. In part 1 the general probability density is formally obtained; in part 2 the distribution density for nearly random noise is derived and the first and second (second-order) moments and energy spectral distributions are determined.

621.396.62+621.396.645 1097
Verstärker und Empfänger (Amplifiers and Receivers) [Book Review]—(See 955.)

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 1098
Entropy of a Transmission System—J. A. Ville. (*Câbles & Trans.* (Paris), vol. 5, pp. 189–198; July, 1951.) Entropy is interpreted in terms of the number of binary code signals required for the transmission of an intelligible message. A single information unit is expressed as the sum of three terms representing information, redundancy and error. Limiting values of entropy for known probability conditions are evaluated.

621.39.001.11 1099
Information Theory and Most Efficient Codings for Communication or Memory Devices—L. Brillouin. (*Jour. Appl. Phys.*, vol. 22, pp. 1108–1111; September, 1951.) Shannon's theorem about the capacity of a channel is discussed; the most efficient coding is the one yielding the most probable distribution of the code symbols. A specific rule is obtained for this most probable distribution. Most efficient coding is essential for communication channels or memory devices in large-scale computers.

621.394.441 1100
Two New Voice-Frequency Telegraph Systems—H. Gardère. (*Câbles & Trans.* (Paris), vol. 5, pp. 199–225; July, 1951.) Technical and economic aspects of multiplex telegraphy are reviewed. Critical discussion of various proposed systems indicates that those using PHM or FM are preferable. A description is given of a PHM system.

621.395.44+621.317.083.7]:621.316.1 1101
Telemetering Systems and Channels for a Large Interconnected Power System—G. K. Duff. (*Elec. Eng.* (N.Y.), vol. 70, pp. 796–801; September, 1951.) Essential text of 1951 A.I.E.E. Summer General Meeting paper.

621.395.44:621.316.1 1102
Co-ordination of a Power-Line Carrier Network—G. E. Burridge and A. S. Jong. (*Elec. Eng.* (N.Y.), vol. 70, p. 803; September, 1951.) Summary of 1951 A.I.E.E. Summer General Meeting paper.

621.396.619.11/.13:621.396.8 1103
Comparative Tests of Communication by Amplitude Modulation and Frequency Modulation carried out by the Société française Radioélectrique—(*Ann. Radioélect.*, vol. 6, pp. 287–288; July, 1951.) In districts where traffic and industrial electrical interference are very considerable, FM was found definitely superior to AM but where traffic and interference were small the two systems were found nearly equal in range and quality of modulation.

621.396.619.16 1104
Cross-Talk Considerations in Time-Division Multiplex Systems—S. Moskowitz, L. Diven and L. Feit. (*Elec. Commun.*, vol. 28, pp. 209–216; September, 1951.) Reprint. See 740 of 1951.

621.396.65 1105
Long-Distance Telephone Links and the Microwave Copenhagen-Skamlebaek Installation—G. Pedersen. (*Teleteknik* (Copenhagen), vol. 2, pp. 153–155; July, 1951.) General considerations with reference to the Danish Post Office 24-channel link, operating on 20-cm wavelength.

621.396.65 1106
Radio-Telephone Communications in Jutland—H. Schouboe-Madsen. (*Teleteknik* (Copenhagen), vol. 2, pp. 155–160; July, 1951). An account of the 24-channel Aarhus-Hammel link, which uses pulse-phase modulation and operates on frequencies between 1,400 and 1,500 mc, and a description of emergency radio telephone equipment for servicing islands when cable breakage interrupts communications.

621.396.65 1107
24-Channel Pulse-Modulated Microwave Telephony Equipment of Danish Manufacture—L. Christensen. (*Teleteknik* (Copenhagen), vol. 2, pp. 161–167; July, 1951.) Description of equipment for the Aarhus-Hammel and Copenhagen-Skamlebaek links.

621.396.65 1108
Notes on an Automatic Radio-Frequency Repeater System—J. A. Craig. (PROC. I.R.E., vol. 39, pp. 1524–1529; December, 1951.) Discussion of the basic principles of rf relay systems and description of the system and equipment for the broadcasting network covering the whole of Cuba.

621.396.65.029.64 1109
The TD-2 Microwave Radio Relay System—A. A. Roetken, K. D. Smith and R. W. Friis. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1041–1077; October, 1951.) The relay system is designed to supplement the coaxial-cable telephone system for long-distance communication and to provide facilities for wide-band signals. The system uses FM and provides twelve wide-band channels, six in each direction, spaced 40 mc apart in the 3,700–4,200-mc band. Each channel may be used to provide a large number of message circuits 4 kc wide or a single video circuit 4 mc wide. The repeater stations are located 25–30 miles apart in line-of-sight steps and only the main stations, situated every few hundred miles, are attended. Details of equipment are given. See 450 of 1951 (Clutts).

621.396.65.029.64 1110
An Unattended Broad-band Microwave Repeater for the TD-2 Radio Relay System—R. W. Friis and K. D. Smith. (*Elec. Eng.* (N. Y.), vol. 70, pp. 976–981; November, 1951.) Essentials of 1951 A.I.E.E. Summer General Meeting paper. Description of the equipment for converting the incoming 4-kc signals to an IF band centered at 70 mc (where 75 per cent of the required gain is provided) and for reconversion to a microwave band offset 40 mc from the carrier frequency of the incoming signals. See also 1109 above (Roetken, Smith and Friis).

621.396.6:621.396.97 1111
Precision A.M. Frequency Monitor—R. S. McKinney. (*Broadcast News*, no. 64, pp. 48–50; May/June, 1951.) Description and circuit of RCA Type-BW-11A broadcasting monitor. The transmitter signal, after passage through an untuned wideband amplifier, is heterodyned with an accurate reference signal of frequency 1kc below the assigned frequency.

621.396.665.1 1112
Instantaneous Compandors on Narrow-Band Speech Channels—J. C. Lozier. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1214–1220; October, 1951.) The conditions are examined for the distortionless transmission of compressed speech over a system with a passband no wider than that occupied by the uncompressed speech. The analysis indicates that more severe requirements must be imposed on the attenuation and phase characteristics of the system when this reduced-bandwidth mode of operation is used.

621.396.712 1113
The Operation of Broadcasting Studios, and New Equipment of the S.F.R. (Société française Radioélectrique)—Cordonnier and Bernard. (*Ann. Radioélect.*, vol. 6, pp. 268–285; July, 1951.) Economic and technical aspects of studio and control-room arrangements are discussed, particularly personnel requirements, size of apparatus, ease of operation and running costs. A comprehensive description is given of studio arrangements and types of equipment proposed by the S.F.R. to satisfy the essential conditions of both quality and efficiency.

621.396.931 1114
Radio for Taxis—(*Wireless World*, vol. 57, pp. 491–493; December, 1951.) Description and discussion of two-way AM radiotelephone system for taxicabs in London. Operation is in the frequency band 100–184 mc. Only one fixed station, located on very high ground is used.

621.396.932/.933:621.317.7 1115
Remote Monitoring of Naval and Air Service Transmissions—Marique. (See 1038).

621.396.932:623.98(44) 1116
Radiocommunication in the [French] Navy—P. David. (*Onde élect.*, vol. 31, pp. 297–306; July, 1951.) Survey of the problems of equipping a naval unit for long- and short-range communication, navigation, telecontrol and enemy interception. Selection of frequencies, disposition of apparatus in the vessel, arrangement of antennas, and recent improvements in df technique are discussed, reference being made to relevant published papers.

SUBSIDIARY APPARATUS

621-526 1117
Nonlinear Techniques for Improving Servo Performance—D. McDonald. (*Proc. NEC* (Chicago), vol. 6, pp. 400–421; 1950.

621-526 1118
Servo Mechanisms—A. L. Whiteley. (Proc. I.R.E., part I, vol. 98, pp. 289–297; September, 1951.) A review of progress. 46 references.

621.314.5 1119
Comparative Representation of Various D.C./A.C. Converters—H. Tiller (*Arch. elekt. Übertragung*, vol. 5, p. 439; September, 1951.) Correction to paper noted in 504 of March.

621.314.6.012.8 1120
The "Transrector", the Quadripole Equivalent of a Rectifier—H. Marko. (*Frequenz.*, vol. 5, pp. 196–203; July, 1951.) A method of calculating the approximate impedance characteristics of a rectifier circuit is developed. The rectifier is represented as an ideal transformer connecting the ac and dc parts of the circuit; the resistances of these parts determine the turns ratio to be assumed. Simple transformer theory is then applied. Applications to a mains rectifier, a modulator and a meter rectifier illustrate the method.

621.314.632.1 1121
The Copper Oxide Rectifier—W. H. Brattain. (*Rev. Mod. Phys.*, vol. 23, pp. 203–212; July, 1951.) The conductivity of the oxide layer in Cu_2O rectifiers can be explained on the basis of the usual energy-band representation of semiconductors only by assuming the presence of some donor-type impurities in addition to the usual acceptor type. Applying the Schottky theory of the space-charge exhaustion layer, the dependence of the capacitance of the rectifier on bias voltage shows that the density of ion charge in the rectifying layer is of the same order of magnitude as the difference between the donors and acceptors found from the conductivity, thus checking the theory. Analysis of the dc characteristic and its dependence on temperature indicates that the Cu-Cu_2O interface is not uniform, but acts like a patchy surface over which the potential maximum varies in magnitude.

621.316.543.2.029.64:621.392.26† 1122
A High-Speed K-Band Switch—M. W. Long. (Proc. I.R.E., vol. 39, pp. 1566–1567; December, 1951.) Description and performance characteristics of a 3-way rotary switch for 1.25-cm waves, with application to rapid scanning.

621.316.722.1:621.314.3 1123
Magnetic-Amplifier Voltage Regulator—J. L. Wolff. (*Proc. NEC*, (Chicago), vol. 6, pp. 45–51; 1950.) Description of equipment which uses a magnetic amplifier as the control element. Regulation is effected at 160–400v to within 0.5 per cent for load currents from 0 to 500 ma (line voltage constant), or for line-voltage changes of ± 10 per cent and load currents from 0 to 300 ma.

TELEVISION AND PHOTOTELEGRAPHY

621.396/.397].621 1124
The Design of Combined Television and Radio Receiver—Bamford. (See 1090.)

621.397.335:621.317.35 1125
Producing and Interpreting the Pulse Cross—D. M. Launer. (*TV Eng.* (N.Y.), vol. 2, pp. 12–15, 29 and 21, 29; September and November, 1951.) An oscillographic technique developed by Loughren Bailey (1937 of 1941) is adapted for examining the phasing and duration of synchronizing signals transmitted with the television picture signal. The "pulse-cross" pattern is obtained by modulating the intensity of the scanning beam and adjusting the phase of the synchronizing pulses so that they appear near the center of the raster. Photographs of patterns thus obtained are shown and interpreted.

621.397.5 1126
The Evaluation of Picture Quality with Special Reference to Television Systems: Part 1—L. C. Jesty and N. R. Phelp. In 295 of February, for "vol. 15" please read "vol. 14".

621.397.5 1127
The Evaluation of Picture Quality with Special Reference to Television Systems: Part 2—L. C. Jesty and N. R. Phelp. (*Marconi Rev.*, vol. 14, pp. 156–186; 4th Quarter, 1951.) Previous experimental results (275 of February) show useful correlation with system performance and provide a basis for an explanation of the relations between the limits of resolution, brightness and signal/noise ratio (i.e. graininess). Further experimental data are needed at low values of contrast. A fundamental performance figure is proposed, measured in terms of the number of quanta per picture element per picture required for a signal/noise ratio of unity. For comparing systems, both the viewing distance and the ratio viewing-distance/picture-height must be specified. The quality of 35-mm motion pictures can be obtained in a 600-line television system with spot wobbling. Application of the results to the determination of television standards is discussed.

621.397.5 1128
Flicker in Television Pictures—J. Haantjes and F. W. de Vrijer. (*Philips Tech. Rev.*, vol. 13, pp. 55–60; September, 1951.) Flicker in television pictures is compared with that in motion-picture projection. The origin of flicker and the related properties of the eye are discussed. Experiments show that with a suitable phosphor for the cathode-ray screen and a frame frequency of 50 c, a high-light luminance of 200 cd/m² is permissible without causing troublesome flicker.

621.397.5:535.623 1129
Oscillating Color Sequence in Color TV—R. G. Peters. (*TV Eng.* (N.Y.), vol. 2, pp. 18–19; September, 1951.) The account of the essential features of the technique described by Laughlin (826 of April).

621.397.5:535.767:621.398 1130
Stereo-Television in Remote Control—H. R. Johnston, C. A. Hermanson and H. L. Hull. (*Proc. NEC* (Chicago), vol. 6, pp. 170–177; 1950.) Description of an experimental system developed as a viewing system for remote manipulation in atomic-energy research.

621.397.5:535.88 1131
Large Screen Television in the Festival of Britain Telekinema—T. M. C. Lance. (*Jour. Telev. Soc.*, vol. 6, pp. 266–271; July/September, 1951.) The general problems connected with the production of bright television pictures of diagonal up to 26 feet and with methods of distributing programs to a large number of cinemas are outlined. A description is given of instantaneous projection equipment for both 405- and 625-line pictures and of its associated power supply and remote-control units. A cinema specially designed to house this equipment, as well as film projectors, is described. The projection box is built under the front of the balcony to keep the television projector on the center line of the metallized-fabric directional-viewing screen. The later has an illuminated surround of fixed low intensity to enhance the contrast of the television picture. The foyer, which can be used as a studio for closed-circuit work, has one wall of glass allowing the public to see into the projection room.

621.397.6 1132
Considerations on Television Pickup Technique: Part 2—G. Goebel. (*Fernmeldetech. Z.*, vol. 4, pp. 403–406; September, 1951.) The discussion of studio problems presented in part 1 (3483 of 1940) is continued, with particular reference to the illumination required when using an iconoscope or supericonoscope camera tube.

621.397.61.001.41 1133
Measuring Television Transmitter Amplitude Characteristics—J. Ruston. (*Tele-Tech.* vol. 10, pp. 30–31 . . . 72; September, 1951.) Description of a method of measurement in which the frequency of a test voltage applied to the video input terminals of the transmitter is swept over a 5-mc range. A cro is used to display the desired overall transmitter response. An inductive probe inserted in the output transmission line of the transmitter picks up a signal which is mixed with the "swept" signal by means of a crystal diode.

621.397.611.2 1134
The Vidicon—Photoconductive Camera Tube—P. K. Weimer, S. V. Forgue and R. R. Goodrich. (*RCA Rev.*, vol. 12, part 1, pp. 306–313; September, 1951.) Reprint. See 2040 of 1950.

621.397.611.2:535.215:546.863.221 1135
Properties of some Photoconductors, principally Antimony Trisulfide—Forgue, Goodrich and Cope. (See 1008.)

621.397.62 1136
A Time-Selection Circuit for Frame Sync-Separation—W. R. Luckett. (*Elec. Eng.*, vol. 23, p. 343; September, 1951.) Description, with diagram, of a circuit that is stable in operation and does not require critical adjustment of component values. The basic principle is the selection of pulses that occur within a fixed period after the occurrence of a reference pulse.

621.397.62 1137
A Single-Valve Line-Scan and E.H.T. Generator—C. H. Banthorpe. (*Elec. Eng.*, vol. 23, pp. 349–352; September, 1951.) A brief review of single-tube circuits is given.

Their advantages of compactness, cheapness and stability are offset by poor linearity and interdependence of controls. By suitable design and the use of modern components these disadvantages can be minimized. The design of a suitable circuit is described in some detail.

621.397.62 1138

Paris Television Show—A. V. J. Martin. (*Wireless World*, vol. 57. pp. 459–460; November, 1951.) Of the 87 different receivers shown, the majority were for the 819-line standard, and six were for both the 819-line and the 441-line standard. See also *Radio franç.*, pp. 18–22; October, 1951.

621.397.62:621.396.662 1139

Utilization of Printed Components in a Television Tuner—D. Mackey and E. Sass. (*RCA Rev.*, vol. 12, part 1, pp. 293–302; September, 1951.) "Printed coils are found to be practical in a turret-type television tuner which has individual coil strips. The coils on these strips are produced by means of a relatively unconventional photoetching process. Possible economic and electrical advantages of using the printed coils are indicated, and a circuit employing the coils is described."

621.397.62:621.396.68 1140

Ringing-Choke E.H.T. Systems—W. T. Cocking. (*Wireless World*. vol. 57, pp. 444–447 and 513–516; November and December, 1951.) Systems are described in which line-scan sawtooth voltage is applied to the grid of a pentode with a RLC circuit connected to its anode; voltage oscillations produced when the anode current cuts off are passed to a rectifier of either half-wave or voltage-doubler type. Because the high-voltage and scanning circuits are separate, optimum design and voltage regulation are facilitated.

621.397.621:621.314.2 1141

Study of a Line [-scan] Transformer considered as a Pulse Transformer—H. Gilloux. (*Radio Franç.*, pp. 1–5; September, 1951.) Design calculations are made for a transformer for the timebase circuit of a 441-line or 819-line scanning system. The critical value of leakage inductance is determined by considering a pulse transformer operating with pulse duration twice the flyback time. The transformer windings comprise five series-connected primary sections and four interleaved parallel-connected secondary sections. Additional windings may be incorporated to supply a voltage-doubler circuit for the hv required for the cathode-ray tube.

621.397.621.2 1142

The Davisson Cathode Ray Television Tube using Deflection Modulation—A. G. Jensen. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 855–866; October, 1951.) Description of an experimental cathode-ray tube, nearly 5 feet long, designed by C. J. Davisson and used during 1937 in demonstrations of television transmission from New York to Philadelphia by coaxial cable. See also 4462 of 1938 (Strieby).

621.397.82 1143

Internal Television-Receiver Interference—B. Amos and W. Heiser. (*Electronics*, vol. 24, pp. 122–125; November, 1951.) Minimum interference from harmonics of the sound and video channel intermediate frequencies occurs when 21.75 mc is used for the sound IF. The possible harmonics due to the video detector are analyzed and optimum frequencies given for intercarrier-sound and 41-mc operation.

621.397.82:621.365.55† 1144

Curing Industrial TVI—P. S. Rand, A. J. Riley and J. J. Lamb. (*QST*, vol. 35, pp. 29–33; September, 1951.) Account of methods for preventing radiation from rf heating installations. Essential requirements are complete shielding and rf filtering for all leads.

621.397.5 1145

Elements of Television Systems [Book Review]—G. E. Anner. Publishers: Prentice-Hall, New York, 1951, 771 pp., $10.35. (Proc. I.R.E., vol. 39, p. 1579; December, 1951.) "... written primarily for physicists and engineers rather than for those interested in the very elementary aspects of the subject."

621.397.621 1146

Theory and Design of Television Receivers [Book Review]—S. Deutsch. Publishers: McGraw-Hill, New York, 1951, 521 pp., $6.50. (Proc. I.R.E., vol. 39, pp. 1577–1578; December, 1951.) "The author has attempted to cover a large field and, on the whole, has done so in an excellent fashion."

TUBES AND THERMIONICS

621.383 1147

Relations for Barrier-Layer Photocells and Theory derived therefrom of the Barrier-Layer Photo- and Rectifier-Effect—P. E. Weber. (*Optik*, vol. 8, pp. 302–310; July, 1951.) From the variation of the no-load terminal voltage and of the short-circuit current of barrier-layer photocells with the intensity of illumination an equivalent circuit is derived and its resistance values are determined, assuming a constant emf. This circuit consists of a voltage source with internal resistance inversely proportional to the radiation intensity, and a constant resistance across the terminals. These resistances are estimated. From this equivalent circuit, theory is developed according to which the barrier layer consists of oxygen atoms bound to copper atoms and, in order that it may be conductive, one electron of the N-shell must be taken from an unbound copper atom of the Cu_2O layer by photoionization or polarization.

621.383.4:546.482.31 1148

Photoconductive Cells of Cadmium Selenide—E. Schwarz. (*Proc. Phys. Soc.*, vol. 63, pp. 624–625; August 1, 1950.) Polycrystalline layers of CdSe have been produced by methods previously described (1102 of 1949). Preliminary measurements on cells with such layers indicate a wide range of sensitivity. The properties of the cells are governed by the amount and form of oxygen present in the CdSe layer. Experiments support the view that adsorption of oxygen on the grain boundaries is an essential condition for the production of photocells with a high quantum yield. See also 3578 of 1949.

621.383.4:546.482.31 1149

Photoconductive Cells of Cadmium Selenide—E. Schwarz. (*Proc. Phys. Soc.*, vol. 64, pp. 821–822; September, 1, 1951.) Data are tabulated for an improved cell (1148 above) of high sensitivity for low applied voltage (6v). Minimum detectable energy for a bandwidth of 1cps is of the order of 2×10^{-12}w for a wavelength of 0.7μ. The sensitivity range extends from the X-ray region to about 1.4μ in the infrared, and the permissible current through the cell is of the order of 20–30 ma.

621.385 1150

Rare Metals in Electron Tubes—D. A. Wright. (*Jour. Brit. IRE*, vol. 11, pp. 381–392; September, 1951.) 1951 Radio Convention (London) paper. The part played by rare metals in overcoming difficulties encountered in the manufacture of different types of electron tubes is discussed. Typical applications mentioned are: electrodes for high-temperature operation; getters; high-emission electrodes; grids with reduced emission; materials for brazing and soldering.

621.385.029.631.64 1151

Effect of Hydrostatic Pressure in an Electron Beam on the Operation of Traveling-Wave Devices—P. Parzen and L. Goldstein. (*Elec. Commun.*, vol. 28, pp. 228–232; September, 1951.) Reprint. See 2580 of 1951.

621.385.029.63/.64:621.396.822 1152

Noise Measurements on a Traveling-Wave Tube—B. N. Agdur and C. G. L. Åsdal. (*Acta polyt.* (Stockholm), no. 86, 9 pp.; 1951.) Noise measurements were carried out on a tube operating at a frequency of about 10 kmc. Measured values of gain as a function of beam current (1–5 ma) are in fairly good agreement with Rydbeck's theory (2962 of 1948), according to which the gain should vary as $I_b^{1/4}$, where I_b is the beam current. The observed gain was about 12 db for a beam current of 3 ma. Noise factor increased from 16 db at $I_b=1$ma to about 19 db at $I_b=5$ ma. This change is of the same order as that predicted by present theories, which yield noise-factor values apparently much too high.

621.385.029.64/.65 1153

Periodic-Waveguide Traveling-Wave Amplifier for Medium Powers—G. C. Dewey, P. Parzen and T. J. Marchèse. (*Elec. Commun.*, vol. 28, pp. 220–227; September, 1951.) Reprint. See 2047 of 1951.

621.385.029.64 1154

Experimental Observation of Double-Stream Amplification—B. N. Agdur. (*Acta. polyt.* (Stockholm), no. 86, 13 pp.; 1951.) Description of the construction and performance of a double-stream microwave tube giving a gain of 30 db at 3 kmc. The two helices, wound with 30 turns of 0.4-mm Mo wire, are 25 mm long with a diameter of 8 mm, the intervening interaction space being of length 20 cm. Accelerating voltages of 250–290v were found suitable. Beam current was 30 ma. The change of gain characteristics with relative modulation of the beams is in fair agreement with the theory of Rydbeck and Forsgren (2866 of 1951). The gain/cathode-voltage-difference curve obtained is generally similar to that given by Haeff (1825 of 1949).

621.385.032.213.2 1155

Thermionic Data for some Capillary Metal Cathodes—H. Katz and K. L. Rau. (*Frequenz*, vol. 5, pp. 192–196; July, 1951.) The variations of the emission from a thoria cathode coating on a tungsten core with step variations of the heater current are shown graphically and explained by means of Richardson's law. With a metallic thorium emitting surface on a porous tungsten base the operating temperature can be considerably lower than for the thoria cathode; at 1,370°C. a current density of 1a/cm² was obtained. Emission characteristics for a metallic barium layer are shown, the cathode construction being modified to provide an auxiliary heating coil for the stock material.

621.385.032.216 1156

The Work Function for Oxide Cathodes—G. Jähnig. (*Funk u. Ton*, vol. 5, pp. 95–100; February, 1951.) Treatment of the thermodynamics of electron emission by Fermi statistics. The influence of cathode activation on work function is discussed.

621.385.032.216 1157

The Effect of Ion Bombardment on the Emission from Oxide-Coated Cathodes—P. A. Redhead. (*Canad. Jour. Phys.*, vol. 29, pp. 362–369; September, 1951.) The decay of emission was measured under different sets of conditions. It is caused by a reduction of the number of impurity centers in the cathode coating and by sputtering of the cathode surface by heavy ions. The life of an oxide cathode is not reduced even when the current density is increased by 20 times, provided there is no positive-ion bombardment.

621.385.032.216:621.386 1158

A Study of Oxide-Coated Cathode by X-Ray Diffraction Method—E. Yamaka. (*Jour. Appl. Phys.*, vol. 22, pp. 1087–1088; August, 1951.) Correlated measurements of the crystal size and thermionic emission of BaO, (BaSr)O and (BaSrCa)O as affected by previous heat treatment.

621.385.1:537.525.5 1159
Studies of Externally Heated Hot-Cathode Arcs: Part 1—Modes of the Discharge— Malter, Johnson and Webster. (See 959.)

621.385.15:537.533.8 1160
Development of a Secondary-Electron Multiplier with Aluminium-Oxide-Caesium Emitters—D. M. Khorosh. (*Zh. Tekh. Fiz.*, vol. 21, pp. 397–404; April, 1951.) Experiments were conducted with oxidized Al foil treated with Cs vapor. The coefficient of secondary emission of the surfaces is 6–7.5. The characteristics are free from any anomalies due to primary electron emission. The construction and production of five-stage electron multipliers using these surfaces is described. The amplification obtained is 900–1,600 for 300 v per stage and 2,000–2,500 for 500 v per stage.

621.385.2/.3:546.817.231 1161
Crystal Diode and Triode Action in Lead Selenide—C. A. Hogarth. (*Proc. Phys. Soc.*, vol. 64, pp. 822–823; September 1, 1951.) Single crystals of both *p*- and *n*-type PbSe were examined; whisker contacts of tungsten and phosphor-bronze were used. With rectifier connection, peak inverse voltages were usually between 4 and 7v and rectification ratios up to 300:1 were recorded for small signals, values of 50:1 being obtained without difficulty. With transistor connection, power gains of 1.5 to 2 could be obtained, but current gain greater than unity was not observed, the usual value being about 0.3.

621.385.2:537.315.6 1162
A Method of Calculating the Space Distribution of Potential in a Diode—H. Bonifas. (*Onde élect.*, vol. 31, pp. 363–369; August/September, 1951.) Two expressions are derived which involve the ratio u/α, where u is the initial electron emission velocity normal to the cathode and α the velocity corresponding to cathode temperature. Numerical integration of these expressions determines the curve representing the distribution of potential between the anode and cathode. Calculation for a particular case gives results in good agreement with experimental values.

621.385.2:537.525.92 1163
The Space-Charge Smoothing Factor—C. S. Bull. (PROC. I.R.E., Part III, vol. 98, pp. 470–472; November, 1951.) Discussion on 2058 of 1951. See also 2871 of 1951.

621.385.2:537.533 1164
Electron Streams in a Diode—F. Gray. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 830–854; October, 1951.) "A general solution of the electron-stream equations is developed for a parallel plane diode, under the assumption that the electron velocity is single valued. This solution contains all particular solutions. It serves to unify the wave theory and the particle theory of electron flow, and it is an approximation for multi-velocity streams over a wide range of conditions."

621.385.3/.4 1165
The Development of Electron Tubes for a New Coaxial Transmission System—G. T. Ford and E. J. Walsh. (*Bell Sys. Tech. Jour.*, vol. 30, part 2, pp. 1103–1128; October, 1951.) The fundamental problem in the development of tubes for wide-band systems was to devise means of obtaining closer grid-cathode spacings without sacrificing life performance. The closer spacings have been made possible by the use of rigid control-grid supports which can be wound with wire of very small diameter under tension. A flat winding is produced which can be mounted very close to a flat cathode. Construction and operation details are given of three new tubes developed for the L3 coaxial-cable system: Type-435A and Type-436A tetrodes and Type-437A triode.

621.385.3/.5].032.24:621.317.311 1166
Method for the Determination of Extremely Small Grid Currents in Valve—H. Köppen. (*Elektrotechnik* (Berlin), vol. 5, pp. 431–433; September, 1951.) Causes and effects of grid current, especially in output tubes, are discussed. Grid current due to thermionic emission caused by heat from the cathode may be 10^{-10} to 10^{-7}a. A method of production testing for this is suggested, using a galvanometer in the balanced anode circuit of a screened-filament electrometer tube, the grid of which is connected through a switch to the grid of the tube on test. Tests on five EF14 tubes are reported.

621.385.3:546.289 1167
Crystal Triodes—T. R. Scott. (*Elec. Commun.*, vol. 28, pp. 195–208; September, 1951.) Reprint. See 2592 of 1951.

621.385.3:546.289 1168
The Junction Transistor—D.G.F. and R.K.J. (*Electronics*, vol. 24, pp. 82–85; November, 1951.) A review of recently published material. The junction transistor consists of a single crystal of Ge having a *p*-type section enclosed between *n*-type end pieces, to which the emitter and collector connections are made, the base connection being made to the *p*-type section. Advantages over the point-contact transistor include (*a*) improvement in noise figure of 20–30 db, the equivalent of an absolute noise figure of between 20 and 10 db at 1 kc, (*b*) greater gain (40 to 50 db per stage), (*c*) better electrical and mechanical stability. This type of transistor has a larger barrier capacitance than the present point-contact type, resulting in reduced gain at hf. By means of impedance mismatching the frequency response can be made uniform to at least 1 mc, with some sacrifice of gain.

621.385.3.029.55 1169
Manufacture of a High-Frequency Transmitting Tube—(*Elec. Commun.*, vol. 28, pp. 171–185; September, 1951.) Fully illustrated description of the various processes in the manufacture of Type-F-5918 high-power water-cooled hf triodes. These tubes have an anode dissipation of 70 kw and operate at frequencies up to 22 mc. In push-pull class-C telegraph service, two tubes will give an output of 400 kw. In a class-C anode-modulation amplifier the output from two tubes is >200 kw.

621.385.4:537.525.92 1170
Space-Charge and Ion-Trapping Effects in Tetrodes—K. G. Hernqvist. (PROC. I.R.E., vol. 39, pp. 1541–1547; December, 1951.) In the space between the screen grid and the anode in an evacuated tetrode, a potential minimum is formed by the space charge. At low current levels this effect is small and most secondary electrons from the anode pass to the grid. For higher currents the secondary electron flow is reduced, while at still higher currents a virtual cathode is formed. If residual gas is present, the space-charge potential minimum traps positive ions until an equilibrium state is reached with formation of a plasma region near the anode. By using a square cut-off pulse, it has been shown experimentally that ion trapping occurs, is built up in a finite time and alters the division of current between screen grid and anode in a predicted manner.

621.385.832 1171
A Gun for Starting Electrons Straight in a Magnetic Field—J. R. Pierce. (*Bell Sys. Tech. Jour.*, vol. 30, part 1, pp. 825–829; October, 1951.) "In a simple electron gun consisting of a cathode and two apertured planes held at different potentials, the apertures act as electron lenses. When the gun is immersed in a uniform axial magnetic field the aperture spacings and potentials can be chosen so the emerging electrons have no radial velocities."

621.385.832:537.534 1172
Hollow Cathode for Positive-Ion Studies in Cathode-Ray Tubes—C. H. Bachman, H. Eubank and G. Hall. (*Jour. Appl. Phys.*, vol. 22, pp. 1208–1210; September, 1951.)

621.385.832:621.318.572 1173
New Electronic Tubes Employed as Switches in Communication Engineering: Part 1—Contact Tubes—J. L. H. Jonker and Z. van Gelder. (*Philips Tech. Rev.*, vol. 13, pp. 49–56; September, 1951.) Discussion of the principles and description of the construction of experimental "contact" tubes which have much greater switching speed than em relays and in which no cleaning of contacts is required. A secondary-emission output electrode is used, so that the input and output signals are in phase. See also 1801 and 3213 of 1950 (Jonker).

621.385.832:681.142 1174
Improvements in Cathode-Ray Tube Storage: Application to a Parallel Type of Digital Computer—G. H. Perry. (*Nature* (London), vol. 168, pp. 372–373; September 1, 1951.) The dot-dash system of charge distribution of cathode-ray tube screen, when applied to the parallel computer, was found to have several undesirable features which are shown to be absent in the defocus-focus system, which is also suitable for serial storage.

621.386.1 1175
Special X-Ray Tubes—B. Combée and P. J. M. Botden. (*Philips Tech. Rev.*, vol. 13, pp. 71–80; September, 1951.) Details of three tubes for diagnosis, contact therapy and endotherapy respectively.

621.396.615.141.2 1176
Enhanced Emission from Magnetron Cathodes—R. L. Jepsen and M. W. Muller. (*Jour. Appl. Phys.*, vol. 22, pp. 1196–1207; September, 1951.) Violations of the Hull cut-off condition considerably more pronounced than those previously reported were found at anode voltages and magnetic fields much larger than those used by earlier investigators. When enhanced emission occurred in magnetrons with pure-metal cathodes, it was found that maximum current limits existed for each value of the magnetic field. The experimental observations are combined into a self-consistent pattern for static magnetrons. It is suggested that the process causing cathode bombardment and enhanced emission is electronic interaction in the region of relatively dense space charge surrounding the cathode.

621.396.615.142.2 1177
Space-Charge Effects in Reflex Klystrons—M. Chodorow and V. B. Westburg. (PROC. I.R.E., vol. 39, pp. 1548–1555; December, 1951.) "Space charge in reflex tubes has an effect which causes considerable departure from the existing reflex theory. To a first order, it modifies the electronic admittance by a bunching effectiveness parameter, designated *F*, which, in general, takes some value between one and three. Several approximate methods have been used to estimate this *F* factor and the results are presented graphically for comparison with experimental measurements."

MISCELLANEOUS

061.4:621.396.933 1178
Aircraft Radio on Show—(*Wireless World*, vol. 57, pp. 448–450; November, 1951. Features and equipment shown at the exhibition of the Society of British Aircraft Constructors included the design of aircraft structures to accommodate or to serve as antennas; radio altimeters; vhf communication equipment; wide-band antenna amplifiers for reducing multiplicity of antennas at ground stations; distance-measuring equipment operating at 1 km.