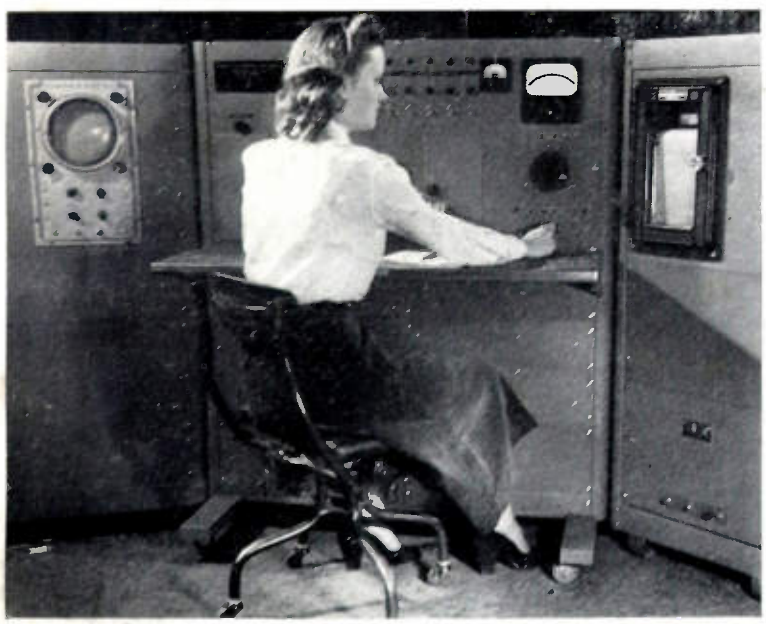
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

February, 1949

Volume 37 Number 2



Sylvania Electric Products Inc.

FLUORESCENT-SCREEN WHITENESS CONTROL

Lengthy and complex spectrophotometric measurements are replaced by a one-minute automatic operation. The above spectroradiometer draws a graph showing screen luminous output for all visible-light frequencies, thus aiding standardization of screen "whiteness."

PROCEEDINGS OF THE I.R.E.

Waves and Electron Section

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Convention Program and Summaries of Technical Papers in this issue.

The Institute of Radio Engineers

What to See at

The Radio Engineering Show

March 7-10, 1949 at Grand Central Palace, New York

200 Exhibits of More Than Radio-Electronic Equipment

Six Million Dollars Worth of Components, Tools and Materials

Firm	Booth
Aerovox Corp., New Bedford, Mass. Molded tubular capacitors, Oil and Wax Impregnated Capacitors, Electrolytic Capacitors and Mica Capacitors.	124
Aircraft-Marine Prod., Inc., Harrisburg, Pa. Solderless Terminals, Solderless Wiring Devices, Machines and Tooling for their application.	262
Airtron, Inc., Linden, N. J. Flexible and rigid waveguides, wavematch switches and coaxial cables and aircraft ignition shielding.	291
Alden Products Co., Brockton, Mass. Facsimile equipment.	320 & 321
Allegheny Ludlum Steel Corp., Pittsburgh, Pa. Metals, Core Materials and Magnets.	47 & 48
Allied Control Company, Inc., New York, N. Y. Relays.	280
Alpha Metals, Inc., Brooklyn, N. Y. Tri Core Rosin filled solder, Tri Core "Energized" rosin filled solder, Tri Core leak pruf acid filled solder.	207
Altec Lansing Corp., New York, N. Y. Loudspeakers, amplifiers, transformers, radio tuners, Intermodulation Test Equipment and Microphones.	B
American Lava Corp., Chattanooga, Tenn. Ceramic parts for radio and wire communications, television and radar, electronic components, control equipment industrial heating devices.	55
American Phenolic Corp., Chicago, Ill. Radio components, Coaxial cable, rf connectors, television antennas, AN connector, industrial sockets and connectors, Plastics.	111 & 112
Amperex Electronic Corp., Brooklyn, N. Y. Transmitting, Industrial, high power, special purpose, electromedical, radiation counter, etc.,—tube types.	10 & 11
Amplifier Corp. of America, New York, N. Y. Magnetic Tape Recorders and accessory equipment, direct coupled amplifiers, regulated power supplies.	259
Arnold Engineering Company, Chicago, Ill. Powdered Molybdenum Permalloy and Deltamex cores, sintered and cast Alnico, Cunico, Cunife, and other permanent magnet materials.	47 & 48
Antara Products, New York, N. Y. Carbonyl Iron Powder for use in cores and coils.	27 & 28
Arrow Electronics, Inc., New York, N. Y. Television Assembly's projection television kit and set plus individual circuits and components for this unit.	314
Astatic Corporation, Conneaut, Ohio Phonograph pickups, pickup cartridges, microphones, recording heads and accessories.	219
Atomic Instrument Co., Boston, Mass. Nuclear Measurement Apparatus.	335
Audak Company, New York, N. Y. Electronic recorders and reproducers.	43
Audio Development Company, Minneapolis, Minn. Transformers, Amplifiers, Filters, Jacks, Plugs, and Patch Cords, Audiometers.	205
Audio Devices, Inc., New York, N. Y. Audiodiscs (recording discs) Audiopoints (recording and playback points).	233
Audio Engineering, New York, N. Y. Magazines.	94A
Audio Equipment Sales, New York, N. Y. Call letter plates, the Audio rule, jack strips, plattertags, low-voltage filtered power supplies, line equalizers, jacks, plugs and patch cords.	A
Automatic Electric Sales Corp., Chicago, Ill. Relays, stepping switches, and other telephone type remote control components.	265
Ballantine Labs., Inc., Boonton, N. J. Sensitive Electronic voltmeters, Geiger-Mueller Counter Tubes.	100
Barker & Williamson, Inc., Upper Darby, Pa. Test Equipment, Coils, Capacitors and Components.	O
Barnes & Noble, Inc., New York Technical Books.	315 & 316
Barry Corporation, Cambridge, Mass. Vibration and Impact Isolators, standard aircraft mounting bases.	293
Bendix Aviation Corp., Eclipse-Pioneer Div., Baltimore, Md. Special Purpose Tubes, Synchros, Small Servos, Gyros.	15, 16, 17
Bendix Aviation Corp. (Radio Div.), Baltimore, Md. Type NA-3 VHF Omni-Range Receiver, Type TA-18 VHF Transmitter, Type ARN-6 (Commercial Equivalent), Type TG-18 Fixed Station Transmitter.	15, 16, 17
Bendix Aviation Corp., Scintilla Magneto Div., Sidney, N. Y. Electrical Connectors, Electric Filters, Plastics, Bus-K-Nects, Ceramics, switches and radio shielding.	15, 16, 17
Bendix Aviation Corp., Red Bank Div., Red Bank, N. J. Dynamotors, Regulated Dynamotors, Fractional H.P. Motors.	15, 16, 17
Berkeley Scientific Co., Richmond, Calif. Nuclear Instruments, Meters.	333
Bird Electronic Corp., Cleveland, Ohio RF Wattmeters and terminations, coaxial switches, VHF and UHF Antennae and filters.	268
Boonton Radio Corp., Boonton, N. J. 202-B FM Signal generator, 203-B Univerter, 160-A and 170-A Q-Meters, 211-A Signal Generator, 140-A Beat Frequency Generator.	277 & 278
W. H. Brady Co., Milwaukee, Wis. "Quik-Label" self adhesive wire markers, name plates, inspection labels, terminal markers, printed roll tapes.	301
William Brand & Co., New York, N. Y., Chicago, Ill. Varnished and Plastic Electrical Tubings and Sleevings. Plastic insulated Hook-Up and Hi-Voltage Wires and Cables.	84B & 85
British Industries Corp., New York, N. Y. Ersin Multicore Solder, Garrard Record Changers.	274
Brooks & Perkins, Inc., Detroit, Mich. Deep drawn magnesium radio and radar boxes and covers, misc. stampings, innumerable fabricated mfg. parts pertaining to aircraft and airborne equip.	122

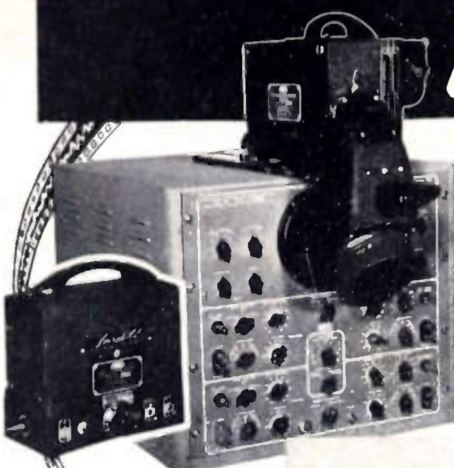
(Continued on page 6A)

PROCEEDINGS OF THE I.R.E., February, 1949, Vol. 37, No. 2. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of contents will be found following page 32A

For permanent oscillograph records . . .

it's DUMONT all the way!



START with Type 314 Oscillograph-record Camera. Especially designed for cathode-ray oscillograph photography. Maximum convenience. Either single-image or continuous recording. Variable speed, electronically controlled. Cat. No. 1366-E, with f/1.5 lens, $1,155.00. Cat. No. 1217-E, with f/2.8 lens, $980.00.

Or with Type 271-A Oscillograph-record Camera (not illustrated). Single-image. Fixed-focus, f/3.5 lens. Cat. No. 1216-E, with mounting, $162.50.

DEVELOP your films with Type 2512 Motor-driven Processing Unit. Utterly simple. Accommodates up to 100 feet of 35mm. film. Cat. No. 1372-E, $231.00.

FINISH with Type 2514 Portable Drying Rack. Holds up to 200 feet of 35 mm. film. Motor-driven. Provided with heater. Easy rewinding. Unit may be folded up. Carrying case supplied. Cat. No. 1375-E, $232.00.

DU MONT CATALOG ITEMS . . .

HIGH AND LOW VOLTAGE OSCILLOGRAPHS: Amplifier frequency response selectable from d-c to 10 mc. Writing rates up to and exceeding 400 inches/microsecond. Deflection factors as low as 0.01 rms volt/inch. Prices from $127.20 to $6,073.75.

OSCILLOGRAPH POWER SUPPLIES: Up to 25,000 volts output for application as final accelerating potential to cathode-ray tubes.

PROJECTION LENS: Projects oscillograph traces on screen up to 30 feet distant. Excellent for demonstrations and lectures. Applicable to high-voltage oscillographs. Type 2542, Cat. No. 1431-E, $103.50.

VOLTAGE CALIBRATOR: For quantitative measurements. Operates with any oscillograph. Type 264-A, Cat. No. 1240-A, $39.50.

ELECTRONIC SWITCH: Provides a time-sharing system for oscillograph presentation of two separate traces. Type 185-A, Cat. No. 1072-A, $105.00.

LOW-FREQUENCY LINEAR TIME-BASE GENERATOR: Du Mont Type 215 accessory extends low-frequency range of the time-base of oscillographs. Cat. No. 1189-A, $215.00.

SPECIALIZED EQUIPMENT: Type 279 Dual-beam Cathode-ray Oscillograph presents two entirely separate traces. Cat. No. 1386-E, $1,294.50. • Type 275-A Cathode-ray Polar Coordinate Indicator employs a circular time-base. Cat. No. 1250-E, $390.00. • Calibrated scales; constant-voltage transformer; transducers; magnetic shields, etc.

DU MONT CATHODE-RAY TUBES: A full line of cathode-ray tubes. A choice of phosphors suited to your particular needs.

In oscillography, Du Mont can supply you with the *tube*, the *oscillograph*, the *accessory*. Note partial list of standard items.

And if your oscillograph needs are *extremely special*, even to the extent of exceeding the broad range of our standard equipment, Du Mont can now place at your disposal the services of our Instrument Model Shop which is equipped to design, develop and manufacture non-standard cathode-ray equipment, or to modify existing equipment and designs. Consult us.

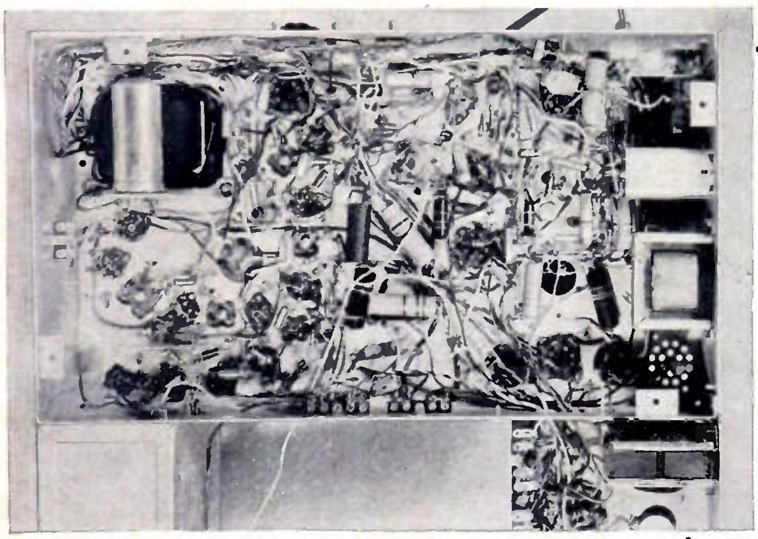
Detailed literature on request. Equipment demonstrations arranged — no obligation.

DUMONT *for Oscillography*

ALLEN B. DU MONT LABORATORIES, INC., PASSAIC, N. J.
CABLE ADDRESS: ALBEEDU, NEW YORK, N. Y., U. S. A.

IT **PAYS** TO SPECIFY **Hi-Q** COMPONENTS

for Hi-Quality T. V. Performance



RCA Victor television receiver Table Model 8T241 uses many **Hi-Q** *capacitors for Uniform, Dependable reception.*

A FEW OF THE COMPANIES USING
Hi-Q ELECTRONIC COMPONENTS

STROMBERG-CARLSON
ZENITH RADIO
RCA VICTOR
Westinghouse
Sparton
CROSLEY
Bendix Radio
Motorola
Tele-tone
Packard-Bell
Admiral
DUMONT
GENERAL ELECTRIC

Hi-Q components are specified by over 200 leading manufacturers. Space does not permit listing all of our valued customers.

● Not only RCA Victor, but practically every manufacturer of television sets looks to **Hi-Q** Electrical Reactance Corporation for components of *Dependable* quality ...*Precision* tested and *Uniform* capacity.

Today's tremendous demand for high quality **Hi-Q** components is being met at three (3) modern plants equipped with the most modern machines helping supply the needs of the fast growing electronics industry.

Our competent engineering staff is available for consultation with your engineers in the design of newly developed circuits.

Booth No. 54 at I. R. E.

Hi-Q COMPONENTS BETTER 4 WAYS

PRECISION Tested step by step from raw material to finished product. Accuracy guaranteed to your specified tolerance.

UNIFORMITY Constancy of quality is maintained over entire production through continuous manufacturing controls.

DEPENDABILITY Interpret this factor in terms of your customers' satisfaction . . . Year after year of trouble-free performance. Our Hi-Q makes your product better.

MINIATURIZATION The smallest BIG VALUE components in the business make possible space saving factors which reduce your production costs . . . increase your profits.

Hi-Q *Electrical Reactance Corp.*
FRANKLINVILLE, N. Y.

Plants: FRANKLINVILLE, N. Y.—JESSUP, PA.—MYRTLE BEACH, S. C.
Sales Offices: NEW YORK, PHILADELPHIA, DETROIT, CHICAGO, LOS ANGELES

What to SEE at the 1949 Radio Engineering Show

Directory of 200 Exhibits

(Continued from page 14A)

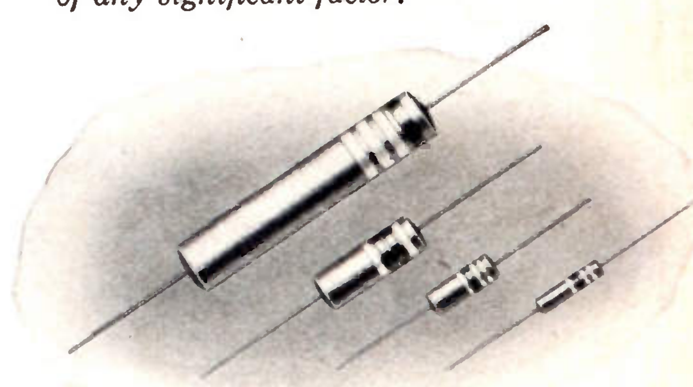
Firm	Booth
Browning Labs., Inc., Winchester, Mass. Cathode Ray Oscilloscopes, Tuners, FM-AM Tuners, Frequency Meters.	G
Brujac Electronic Corp., New York, N. Y. High Voltage Power Supplies, Electronic Microammeters, Impedance bridges, Composite Video Signal Generators, Video Sweep Generator.	239, 240
Brush Development Co., Cleveland, Ohio Recording Equipment, Industrial Instruments and acoustic devices.	F
Bud Radio, Inc., Cleveland, Ohio Sheet Metal Housings, Variable Condensers, Air-Wound Coils, plugs, jacks, chokes, and other misc. radio and electronic components.	231
Bundy Tubing Co., Detroit, Mich. Copper Brazed steel tubing for antennas. Also antennas made by various manufacturers from bundyweld tubing.	123
Burlington Instrument Co., Burlington, Iowa AC and DC Electrical Indicating Instruments, AC and DC portable instruments, Voltage regulators, synchronizers.	229
Bussmann Manufacturing Company, St. Louis, Mo. Fuses and Fuse Holders for the radio and electronic trade.	309
Caldwell-Clements, Inc., New York, N. Y. "Tele-Tech" television and telecommunications engineering magazine and "Radio and Television Retailing" merchandising and servicing trade magazine.	286
Cambridge Thermionic Corp., Cambridge, Mass. Terminals, Terminal Board Assemblies, Electronic Hardware, IF and RF Chokes and Coils, Earth Shock and Underwater Pressure Gauges.	288
Camburn, Inc., Woodside, N. Y. TV and FM Antennas and Accessories, Auto Antennas.	213
Canadian Marconi Co., Montreal P. Q., Canada Nuclear instruments, scalar kick-sorter and component assemblies.	336
Cannon Electric Development Co., Los Angeles, Calif. Connectors Multiple Contact Electronic.	210
Capitol Radio Engineering Inst. Inc., Washington, D. C. Complete library of CREI Text material.	267
Carter Parts Company, Chicago, Ill. Electrical Controls.	255
Centralab, Div. of Globe Union, Inc., Milwaukee, Wis. Printed Electronic circuits, ceramics, capacitor, control switches, components.	234 & 235
Chatham Electronics Corp., Newark, N. J. Tubes, Radiological and Chemical Display.	120 & 121
Chicago Rivet & Machine Co., Bellwood, Ill. Rivets, Steel and Brass, Tubular and Split, Automatic Rivet Setting Machines.	214 & 215
C. P. Clare & Co., Chicago, Ill. Relays, including hermetically sealed relays, stepping switches, lever keys, push keys, turn keys.	258
Clarostat Mfg. Co., Inc., Dover, N. H. Controls (wire wound and carbon) wire wound resistors, power rheostats, television parts.	221
Cleveland Container Company, Cleveland, Ohio A complete line of Cosmalite coil forms and television deflection yoke sleeves.	220
Sigmund Cohn Corp., New York, N. Y. Radio tube filament and grid wires, platinum metals and their alloys, electroplated wires, rhodium plating solution, wollaston process wire.	284
Collins Radio Co., Cedar Rapids, Iowa Airborne transmitters and receivers, Broadcast AM and FM transmitters and speech equipment, specialized mechanical accessories for radio equipment, VHF ground station transmitters and receivers.	75–80
Communication Products Co., Inc., Keyport, N. J. Coaxial Aircore Transmission Line and accessories, Rotary Band switches, Antennas for Mobile Service, Dehydrators.	70 & 71
Communications, New York, N. Y. Communications magazine.	289
Condenser Products Co., Chicago, Ill. Capacitors, power supplies, pulseforming networks.	334
Continental Carbon, Inc., Cleveland, Ohio Nobleloy Metal Film, Carbon Film, Carbomite resistors, suppressors, wire wound resistors, low power capacitors.	200A
Cornell-Dubilier Electric Corp., So. Plainfield, N. J. Capacitors.	72 & 73
Daven Co., Newark, N. J. Audio, video and RF attenuators, rotary switches, measuring equipment, resistors.	94B & 95
Bryan Davis Pub. Co., Inc., New York, N. Y. Communications Magazine, Service magazine.	289
DeMornay Budd, Inc., New York, N. Y. Microwave Test equipment and plumbing, radar components, microwave spectroscopy components.	313
Tobe Deutschmann Corp., Norwood, Mass. Capacitors, Filters, Test and Measuring Equipment.	45
Dial Light Co. of America, Inc., New York, N. Y. Pilot light assemblies, warning signal lights, sockets for min. bayonet and candelabra screw base lamps.	46
Distillation Products, Inc., (Eastman Kodak Co.) Rochester, N. Y. High vacuum tube exhaust equipment, high vacuum rotary coater for aluminizing cathode ray or other tubes.	227 & 228
Allen B. DuMont Labs., Inc., Clifton, N. J. Cathode-ray instruments, cathode-ray tubes.	125–128
Allen B. DuMont Labs., Inc., Clifton, N. J. Television transmitting and studio equipment.	244–248
Eitel-McCullough, Inc., San Bruno, Calif. Transmitting type vacuum tubes and vacuum capacitors, vacuum switches.	36
El-Tronics, Inc., Philadelphia, Pa. LS-100 Decade Scaler, LS64 Binary Scaler, RM4 Counting Rate Meter, SM 3 Portable GM Survey Meter, LS-1 Non Scaling Counter.	327
Electrical Reactance Corp., Franklinville, N. Y. Ceramic capacitors, resistors, choke coils.	54

(Continued on page 22A)

Proper balance can be

for resistors

Proper balance can be mighty difficult . . . but not for IRC resistors. Basically engineered for balanced *performance in every important characteristic, each IRC resistor type offers outstanding features for specific applications—without sacrifice of any significant factor.*

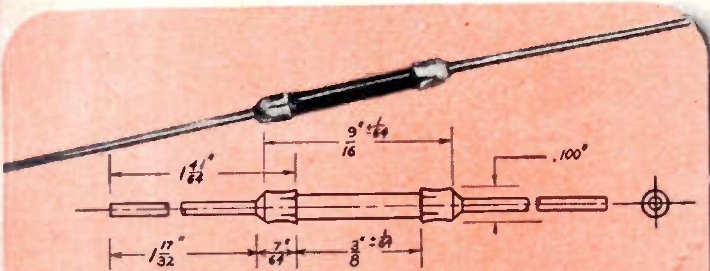


New, ADVANCED Type BT Resistors, for example, are uniformly superior in every important JAN-R-11 requirement. At ⅓, ½, 1 and 2 watts they meet JAN-R-11 specifications for fixed composition resistors. Balanced in every characteristic, small IRC ADVANCED BT's are particularly suited to high ambient temperatures and rigorous television circuits. 12-page Bulletin B-1 gives all the performance facts. Use the convenient coupon.

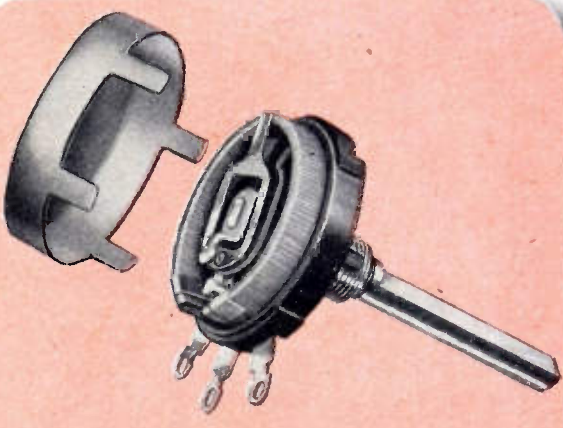
difficult

For close tolerance requirements, IRC Precisions offer a fine balance of accuracy and dependability. Extensively used by leading instrument makers, they excel in every important characteristic. 1% accuracy is standard. Noise level is inherently low, and windings are fully protected against high humidity. Available in a wide selection of ranges and types, as described in Bulletin D-1.

too!



Miniature MPM resistors are IRC engineered for high frequency applications. Their frequency characteristics are outstanding, but absolute balance has been maintained with all other significant electrical characteristics. Thin resistance film is permanently bonded to ceramic rods. Cupped ends of wire lead terminals are cemented to resistor bodies to form axial pigtails. Rated at ¼ watt, Type MPM's are available in resistance values from 10 ohms to 1.0 megohms. Write for Technical Data Bulletin F-1.



IRC Type W Wire Wound Controls are so carefully balanced, your customers can actually feel the difference. With center tap they are widely used as vertical and horizontal centering controls in television receivers. Design provides maximum adaptability to most rheostat and potentiometer applications within 2-watt power rating. Type W Controls have a 1¼" diameter, and 9/16" depth behind panel. Spiral Spring Connector provides positive electrical connection. Bulletin A-2 gives details. Write for your copy.

All standard IRC resistors are readily available in nominal quantities from your local distributor's well-stocked shelves. This is IRC's Industrial-Service Plan

at work, assuring you 'round-the-corner service on your small order requirements. We'll be glad to send you the name of your nearest IRC Distributor.

IRC

Wherever the Circuit Says —∿∿—

Power Resistors • Precisions • Insulated Composition Resistors • Low Wattage Wire Wounds • Rheostats • Controls • Voltmeter Multipliers • Deposited Carbon Precistors • Voltage Dividers • HF and High Voltage Resistors

INTERNATIONAL RESISTANCE COMPANY

405 N. Broad Street, Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

INTERNATIONAL RESISTANCE COMPANY
405 N. Broad St., Philadelphia 8, Pa.
I want to know more about the IRC Resistors checked below —

- ☐ Advanced Type BT's
- ☐ Precision Resistors
- ☐ MPM High Frequency Resistors
- ☐ Type W Controls
- ☐ Also send name and address of our IRC Distributor

Name ....

Title ....

Company ....

Address ....

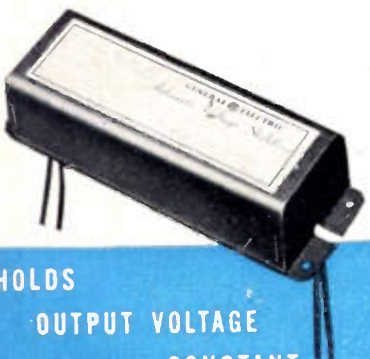
....

Digest

TIMELY HIGHLIGHTS ON G-E COMPONENTS

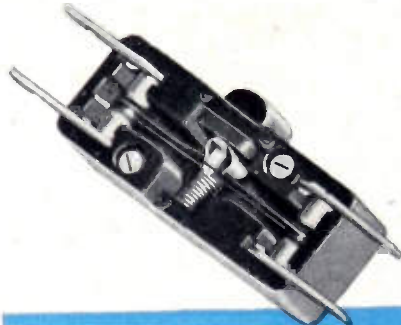
SIMPLIFY CONTROL WIRING WITH THESE TERMINAL BOARDS

Easy-action hinged covers protect control wiring, help give your product a neat appearance. Hook-ups are easy with the hard-gripping connectors. Simply strip the wire end, screw down the connector on the bare wire. Blocks are durable, too, constructed of strong Textolite with reinforced barriers between poles to insure against breakage. Marking strips are reversible—white on one side, black on the other. These terminal boards are available with 4 to 12 poles, 2 inches wide, 1¼ inches high. Send for bulletin GEA-1497C.



HOLDS OUTPUT VOLTAGE CONSTANT

This latest addition to G.E.'s line of automatic voltage stabilizers comes in 15-, 25-, and 50-va ratings. Output is 115 volts, 60 cycles. The small size of the unit makes it particularly applicable to shallow-depth installations in many types of equipment. You may have a job for this unit which will give you automatically stabilized output voltage at a low cost. There are no moving parts, no adjustments to make; long service is assured. Check bulletin GEA-3634B for more information about this and other G-E voltage stabilizers.



LOOKING FOR LIGHTWEIGHT SWITCHES?

Switchettes* are designed for applications which require a manually operated electric switch in a limited space. Though small, these switchettes are lightning fast in action and are built to withstand severe service. A wide variety of forms and terminal arrangements makes them particularly useful where special circuit arrangements are necessary. Switchette shown above has one normally open and one normally closed circuit, transferable when button is depressed. Check bulletin GEA-4888.

*Switchette is General Electric's trade name for these small snap switches.

FOR YOUR COOLING FANS

Here's a fractional-horsepower fan motor suitable for many uses because of its compact design, low servicing requirements, and extreme quietness. Long, dependable operation is assured by sturdy, totally enclosed construction. These Type KSP unit-bearing motors are of shaded pole type design with low starting torque characteristics especially applicable to fans. A continuous oil circulation system furnishes good lubrication. You can use simple, hubless, low-cost blades with the special mounting arrangement. Write for bulletin GEC-219.

General Electric Company, Section J642-19
Apparatus Department, Schenectady, N. Y.

Please send me the following bulletins:

- ☐ GEA-4996 Capacitor Pulse-forming Networks
- ☐ GEA-5102 Panel Instruments
- ☐ GEA-1497C Terminal Boards
- ☐ GEA-3634B Automatic Voltage Stabilizers
- ☐ GEA-4888 Switchettes
- ☐ GEC-219 Unit-bearing Fan Motor

NAME ....

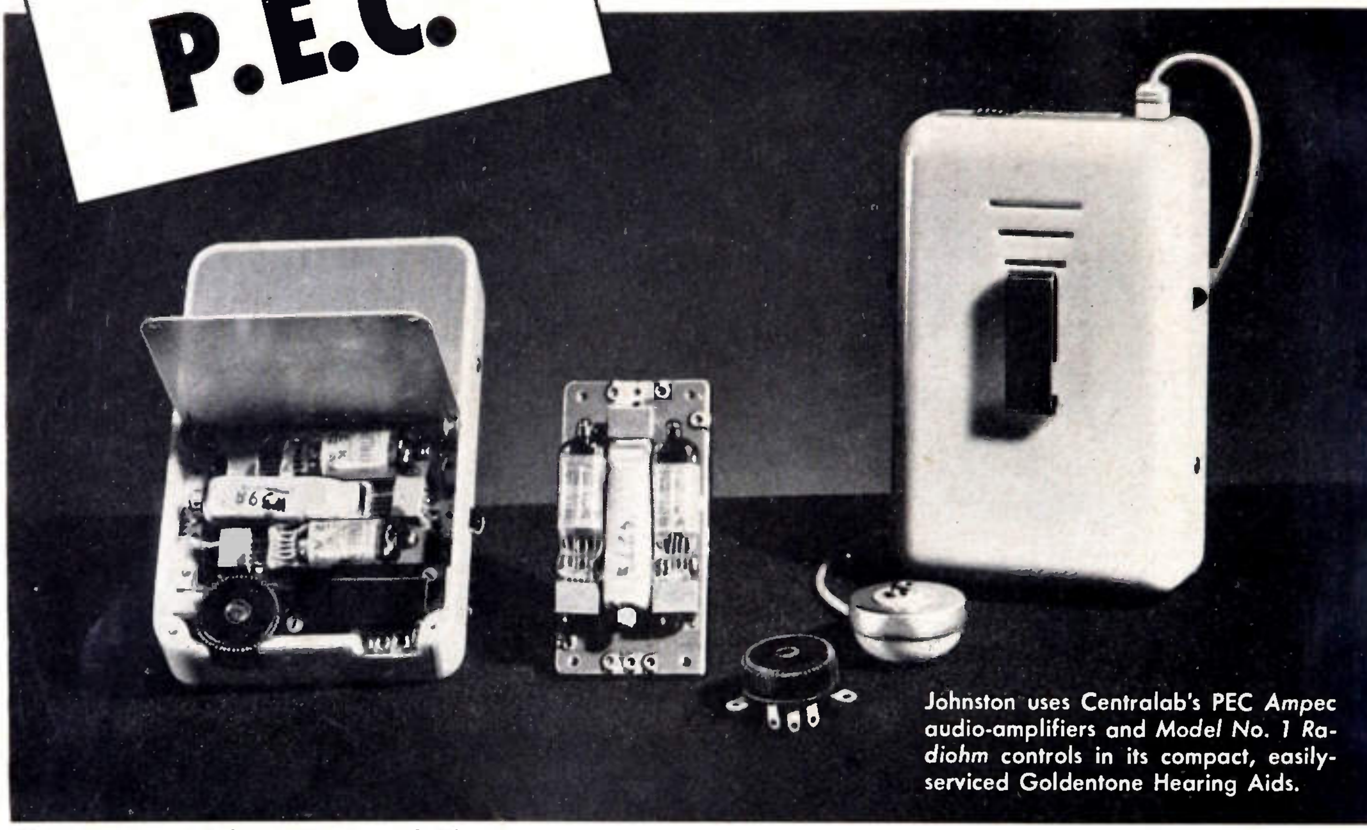
COMPANY ....

ADDRESS ....

CITY .... STATE ....

PROGRESS REPORT ON P.E.C.*

How Johnston uses CRL's "Printed Electronic Circuit" to build its light-weight, compact Goldentone Hearing Aid!



Johnston uses Centralab's PEC *Ampec* audio-amplifiers and *Model No. 1 Radiohm* controls in its compact, easily-serviced Goldentone Hearing Aids.

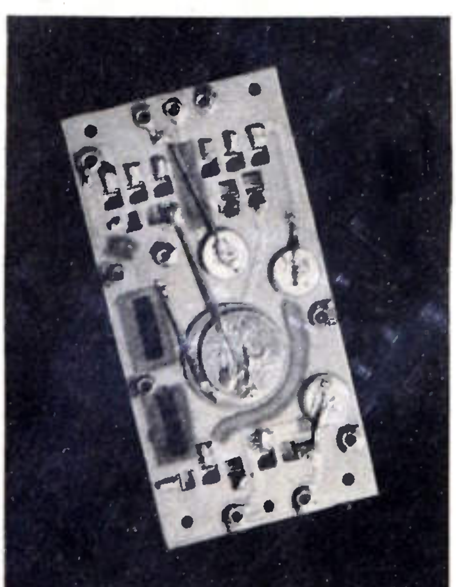
Models courtesy of Johnston Hearing Aid Mfg. Co.

*Centralab's *Printed Electronic Circuit* — Industry's newest method for improving design and manufacturing efficiency!

CUSTOMER comfort...greater output...dependable performance. That's what Johnston wanted for its new Goldentone. And that's what it got — with the help of Centralab's amazing P.E.C. Yes—*Ampec* made it possible to save space and material by reducing the number of components needed. It cut production time by eliminating many assembling operations. It improved performance by minimizing the chance of broken or loose connections and by resisting changes in temperature and humidity.

INTEGRAL CERAMIC CONSTRUCTION: Each *Printed Electronic Circuit* is an integral assembly of CRL *Hi-Kap* capacitors and resistors closely bonded to a steatite ceramic plate and mutually connected by means of metallic silver paths "printed" on the base plate.

This outstanding hearing aid development, illustrated above, was the result of close cooperation between Centralab and Johnston engineers. Working with your engineers, Centralab may be able to fit its *Printed Electronic Circuit* to your specific needs. Write for full information, or call your nearest Centralab Representative.



TYPICAL "AMPEC"—(actual size, back view) shows how you can get complete electrical circuits—tube sockets, capacitors, resistors and wiring—in one miniature Centralab amplifier unit.

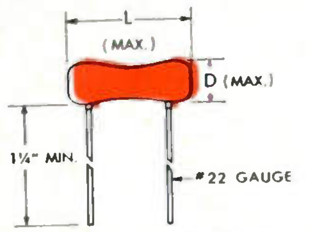
LOOK TO **Centralab** IN 1949!

Division of GLOBE-UNION INC., Milwaukee

New...

for positive identification...

ERIE Radial Lead Insulated CERAMICONS® now have distinctive *red bodies*



DIPPED PHENOLIC INSULATED

Style	Dia. "D"	Length "L"	Max. Cap.
331	.240	.460	715 MMF
332	.240	.710	1500 MMF
338	.312	.550	2000 MMF
337	.312	.937	4100 MMF
333	.315	1.250	5100 MMF
344	.415	1.213	8000 MMF
335	.415	1.650	.012 MFD
336	.415	2.025	.016 MFD

ERIE brings order out of confusion . . . by the simple expedient of giving ERIE radial lead, dipped phenolic coated Ceramicons distinctive red bodies.

In the past manufacturers have found it almost impossible to differentiate between the various makes of such condensers. The common brown body color has sometimes caused confusion in incoming inspection departments and in the final assembly lines. In addition, it has been difficult to fix responsibility for any service reports.

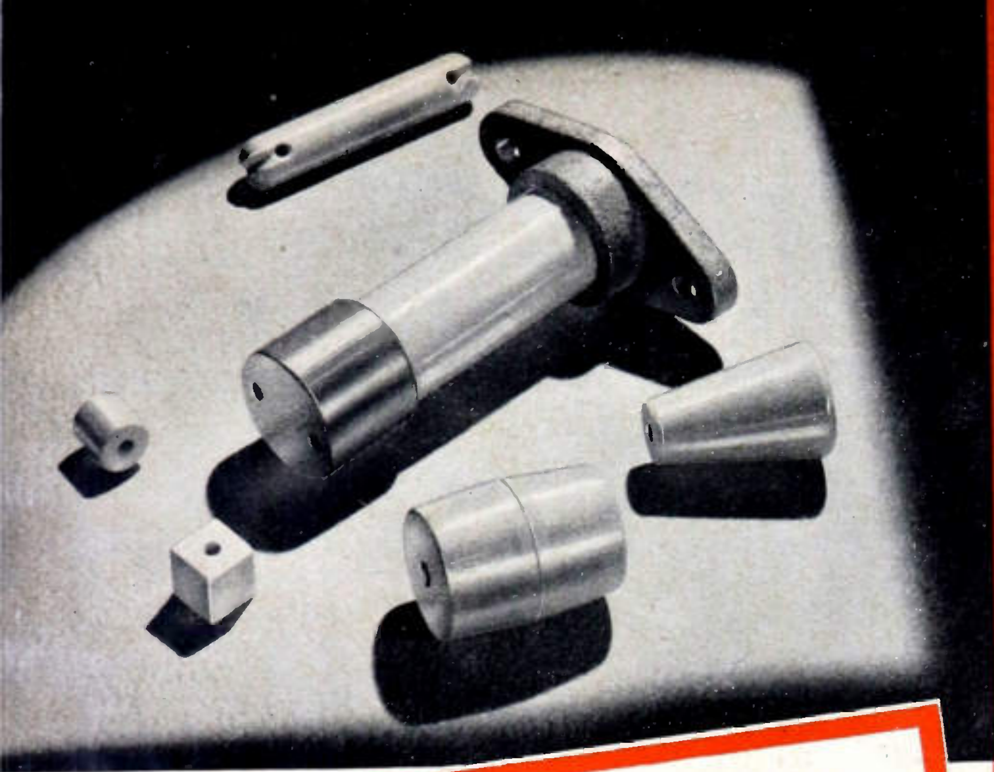
Now, ERIE Radial Lead Insulated Ceramicons are positively and unmistakably identified . . . and the red body also makes it easier to read all RMA color code dots. ERIE axial lead ceramicons will continue to have molded low-loss phenolic insulation.

When you see ceramic condensers with the red body color, you can be sure you have high-quality, dependable ERIE radial lead insulated Ceramicons which will "stay put" in your chassis for the life of the set.

Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND TORONTO, CANADA.



ALSIMAG

TRADE MARK REGISTERED U.S. PATENT OFFICE

STANDARD INSULATORS are now made in L-5 MATERIAL

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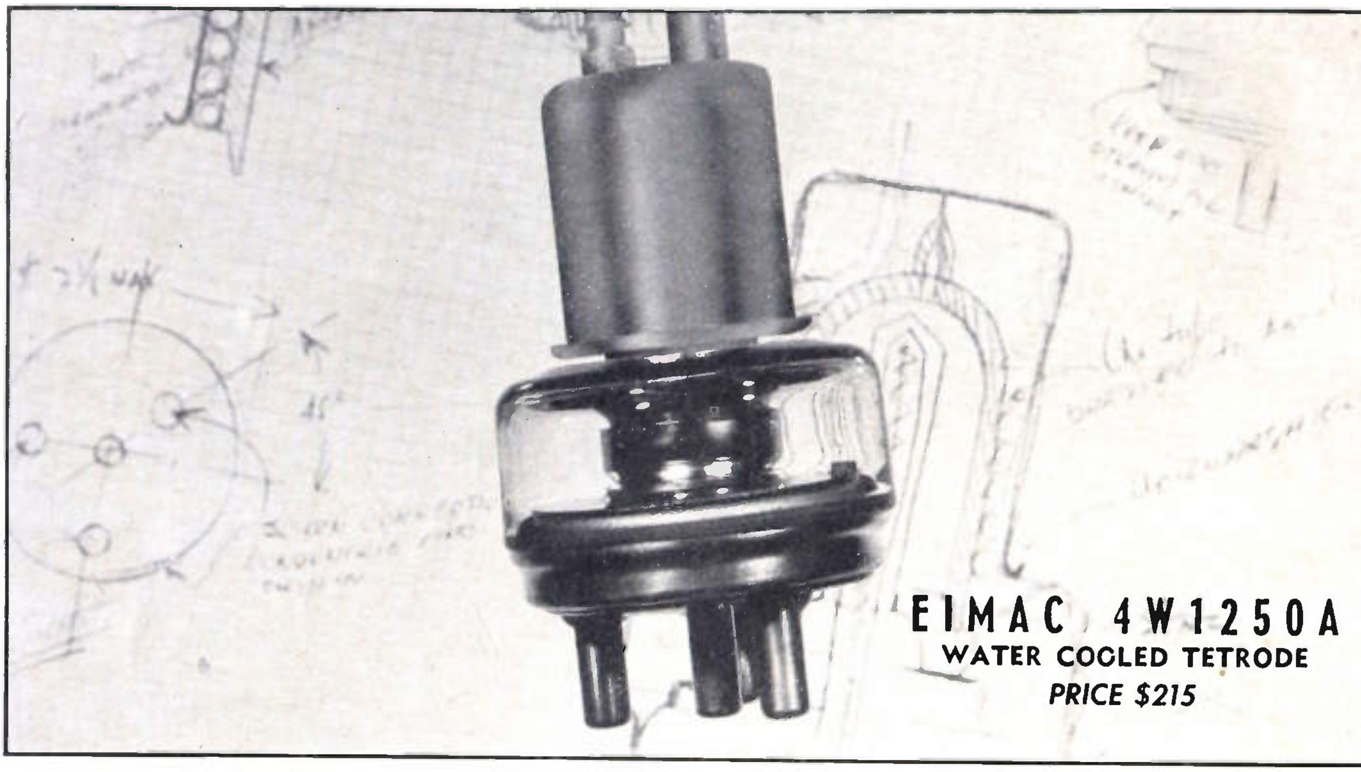

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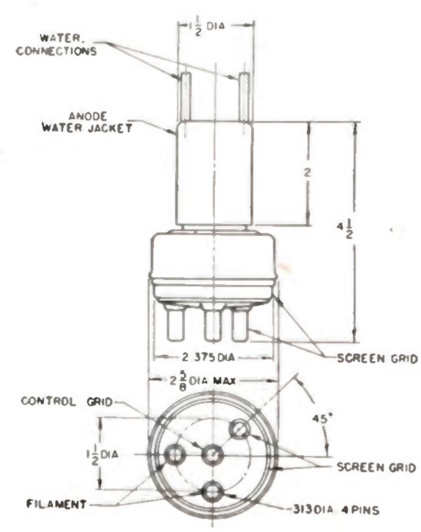

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Current	13.5 amperes
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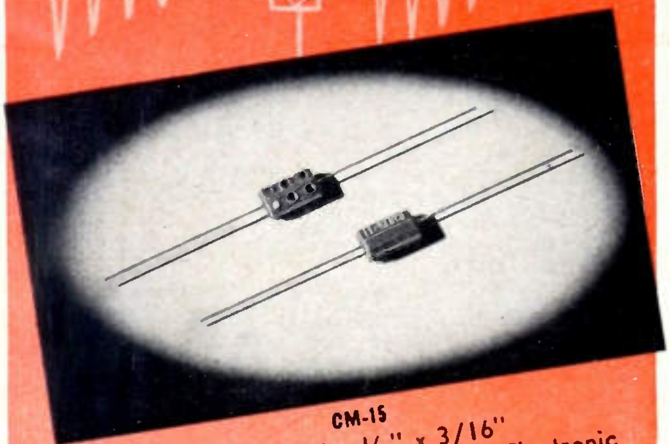

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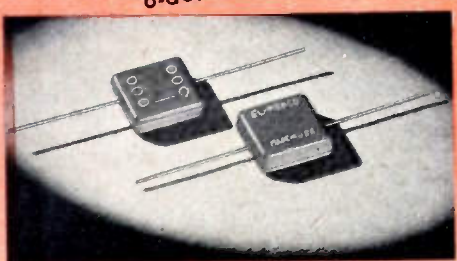

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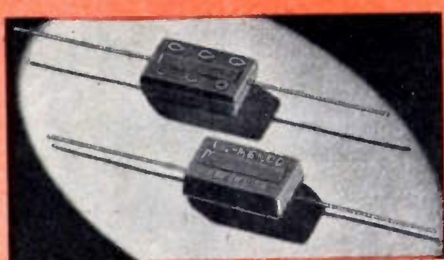

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Fansteel Metallurgical Corp., North Chicago, Ill. Tantalum, Tungsten, Molybdenum, Electronic Tube Components, Selenium rectifiers, electrical contacts.	292
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Freed Transformer Co., Inc., Brooklyn, N. Y. Transformers, reactors, wave filters, laboratory test equipment.	103
Furst Electronics, Chicago, Ill. Wow meter for microgroove turntables and wire recorders. Various types of regulated Power supplies.	R
General Ceramics & Steatite Corp., Keasbey, N. J. Ferrites (magnetic ceramic materials), steatites, titanates, zircon porcelains, wet process porcelains, rigid coaxial line.	37
General Electric Co., Apparatus Div. Capacitors, transformers (radio) voltage regulators, wire and cable, soldering irons, auxiliary relays manual switches, selenium rectifiers.	113–119
General Electric Co., Electronic Div. Miniature TV equipment, plug-in and portable amplifiers, precision test equipment, loudspeakers, cartridges, single sideband selector, tone arms.	113–119
General Electric Co., Tube Div. Transmitting tubes, industrial tubes, receiving and cathode ray tubes including the 16AP4 metal tube and the 12KP4 "Daylight" tube.	113–119
General Electronics, Inc., Paterson, N. J. Electron tubes, oscillators, rectifier, thyratrons and cathode ray.	72
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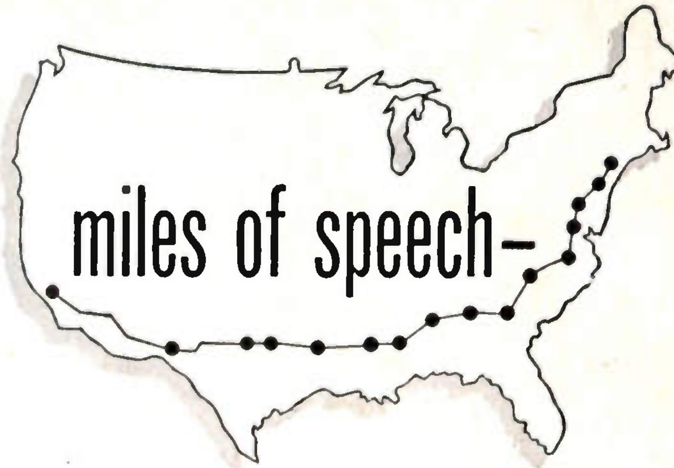

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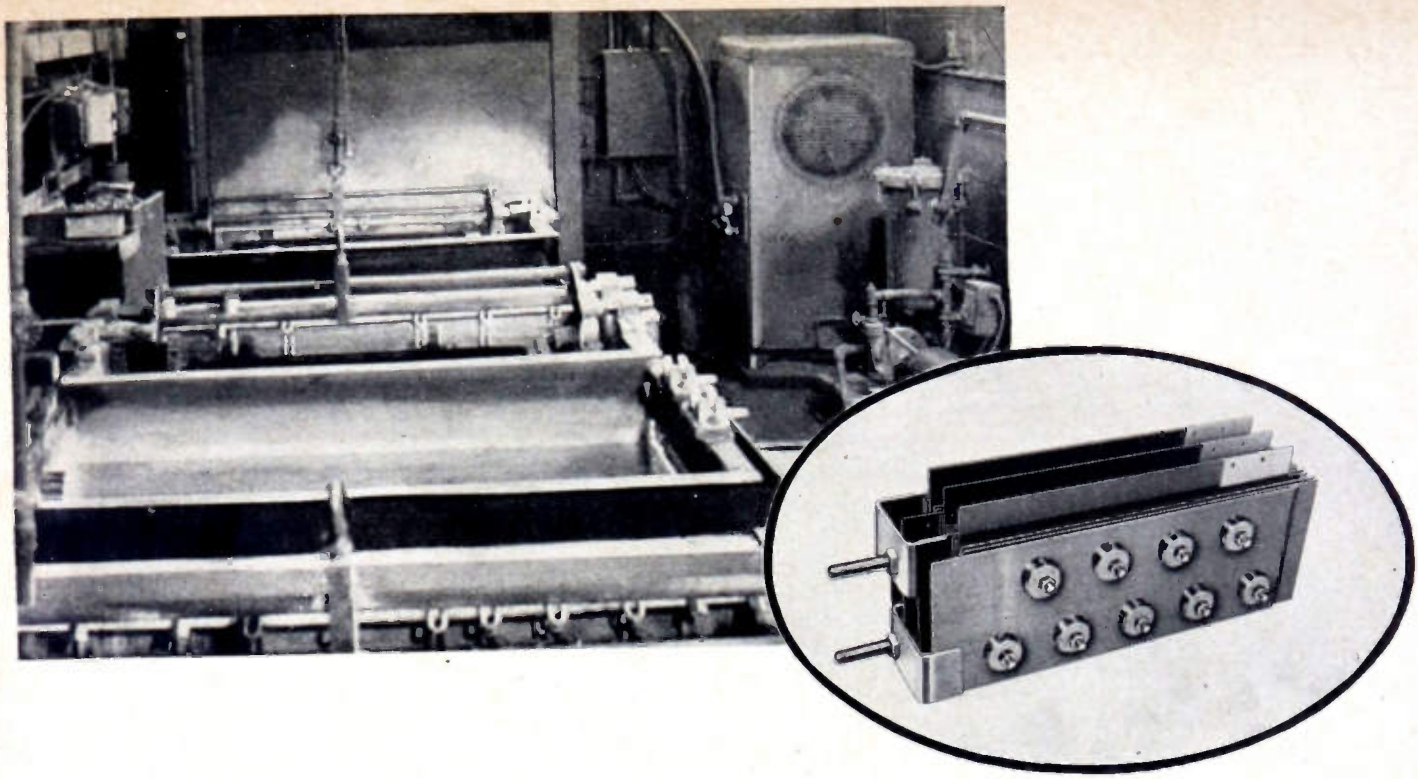

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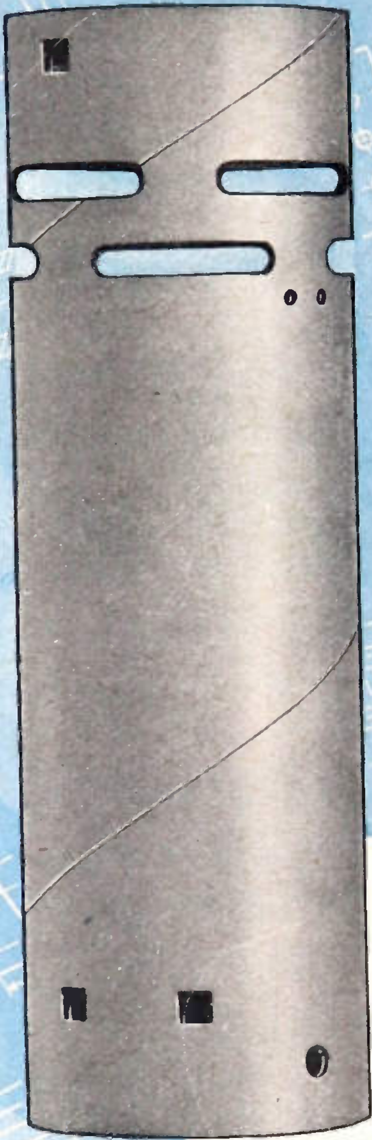

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Model	Load Range Volt-Amperes	*Regulation Accuracy
150	25-150	0.5%
250	25-250	0.2%
500	50-500	0.5%
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*Models available with increased regulation accuracy.

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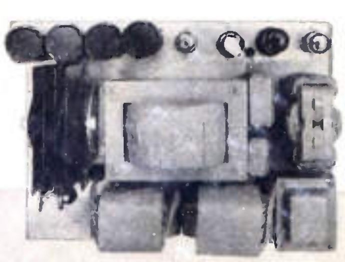

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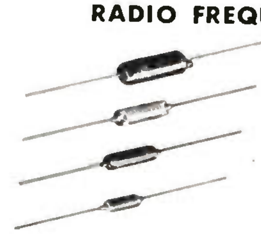

Single-layer wound on low power factor steatite or molded plastic cores. Seven stock sizes cover range 3 to 520 mc. Two units rated 600 ma; all others 1000 ma.

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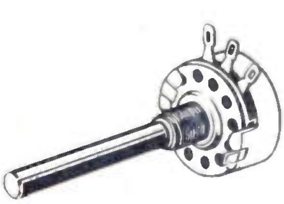

A 2-watt molded composition unit with good margin of safety. It is unaffected by heat, cold, or moisture. Resistance element is a thick, solid-molded ring.

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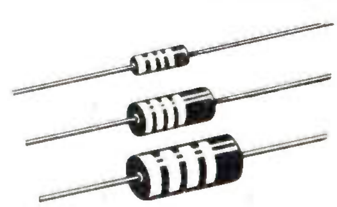

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PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

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VOLUME 37 — *February, 1949* — NUMBER 2

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

Arthur S. McDonald

Vice-President, 1949

Arthur S. McDonald, Chief Engineer of the Australian Overseas Telecommunications Commission, was born at Castle Donnington, Victoria, Australia, on March 6, 1891. Shortly after leaving secondary school he was apprenticed to a civil engineer, after which he gravitated to the mechanical engineering field. In 1910 he was appointed to the engineering branch of the Australian Postmaster General's department, and he was transferred to the radio section of that department on its inception in 1911.

From 1911 until 1915 Mr. McDonald supervised the erection of many of the first radio stations in Australia, and was honored in 1913 by the Federal Executive Council of the Australian Government for his work. In 1915 he transferred to the Naval Radio Service and two years later was named engineer for electrical equipment, responsible to the Naval Board for design and inspection of radio and electrical equipment.

Mr. McDonald was appointed chief engineer of Amalgamated Wireless, Ltd., in 1922, when governmental and certain private operational and manufacturing activities merged. There he supervised the company's broadcasting, fixed, and mobile short wave telegraph and telephone activities. In 1933 he was named chief engineer and assistant manager, and later assistant general manager. In the course of his duties he paid several visits to Europe and America in order to negotiate business transactions, and also to report upon radio activities in those continents.

With the growth of radio activities prior to and during World War II, Mr. McDonald's duties were greatly augmented, and, during this period, he directed the company's activities in radio research, construction, and operation for coast-guard, international radiotelegraph, and telephone services; and broadcasting and miscellaneous mobile radio services. He also helped to fill Naval, Military, and Air telecommunication contracts.

The Australian Overseas Telecommunications Commission was established in 1946 and the operational side of fixed and maritime mobile communications networks owned by Amalgamated Wireless was transferred to this Commission. At the same time, Mr. McDonald was named Chief Engineer.

Mr. McDonald became a member of the IRE in 1923. In 1941 he was advanced to Fellow rank in the IRE. He is a corporate member of the Australian Institution of Engineers and a Fellow and Past President of the Australian Institute of Radio Engineers.

Communications and electronic engineers owe a notable debt of gratitude to the inspired, capable, and indefatigable physicists who have discovered and explained the basic phenomena from which the corresponding branches of engineering have been developed. It is, therefore, particularly fitting that the members of the IRE shall do what they can to meet the specifications for new devices needed by their esteemed colleagues, the physicists, to accomplish further and important tasks. These requirements have been assembled for the following guest editorial by a member of the IRE Board of Editors and a research worker at the University of Illinois.—*The Editor.*

What Electronics Means to the Physicist

JAMES S. ALLEN

During the last fifteen years there has been a steady increase in the application of electronic techniques to the problems of experimental physics. In recent years, even the field of theoretical physics has been benefited directly through the use of electronic computers. Consequently, the physicist is vitally interested in the new products and techniques which result from the research and development originated by the electronic or radio engineer. It is certain that both the physicist and the engineer will be aided by an increased understanding of the problems which are of mutual interest to the two professions.

Perhaps a few examples of the many electronic devices and techniques used by the physicist will serve to indicate their widespread use in this field of science. At present, there is an urgent need for wide-band pulse amplifiers able to handle the pulses of extremely short time duration produced by electron-multiplier tubes. The delay-line amplifier with tubes in parallel rather than in series may prove to be a solution of this problem. An improved counting circuit similar to the familiar "scale of two" scaling circuit is required. This device should be able to count pulses 10^{-8} to 10^{-9} seconds wide and should resolve two such pulses separated by an interval of time less than 10^{-7} seconds. The study of the magnetic properties of nuclei by the microwave absorption method has been made possible by the recent developments in the field of radar. In order to be successful in this field of research, the investigator should have a thorough understanding of the theory and application of microwaves to this problem. Through the combined efforts of engineers and scientists in other fields, large electronic computers, such as the ENIAC, have been developed. These machines should make possible computations never before considered practical or even possible, because of the length of time required when more conventional methods of computing are used. At present, there is a definite need for improved memory devices and also for simplified and highly reliable electron tubes for the basic counting circuits employed in these computers.

As the need for electronic devices, circuits, and techniques becomes more urgent in physics, and in other allied fields of research, the market for these products will be enlarged. It is hoped that this increased demand for the services of the engineer will result in new ideas and developments, which will aid both the physicist and the engineer.

Theory and Practice of Tropospheric Sounding by Radar*

ALBERT W. FRIEND†, SENIOR MEMBER, IRE

Summary—Studies of tropospheric sounding by pulsed radio waves have been extended from the original medium-frequency phase to include the use of modern high-powered, sharply beamed microwave radar systems. The apparent detection of discrete echoes from entire air-mass boundaries and from thick regions of relatively high values of dielectric gradient at vertical incidence is indicated by experimental results attained on medium frequencies during the period from 1935 to 1942. On certain occasions, the indicated heights embraced almost the entire troposphere. Sample results are illustrated.

More recently, continuous photographic records have been made by vertical beam-sounding techniques which involved the use of modified SCR-584 and AN/CPS-1 (MEW) microwave radar systems. In addition to the familiar scattering reflections from raindrops, snowflakes, and ice particles from heights up to about 50,000 feet, continuous records of "dot," "line," and other unusual weak echo signals are obtained in some form at all times, up to heights of at least 30,000 feet, when sufficient power and sensitivity are available in the radar system. Samples of these photographic recordings are shown, and a simultaneous radiosonde plot is provided for comparison. References are cited to show additional supporting experimental evidence which seems to indicate more strongly that the origins of many of these echo patterns are in the fine structures of the dielectric transition layers of air-mass boundaries and air-turbulence and mixing regions. Many of these effects appear to originate within definite strata, and some of the strongest and sharpest echoes involve surfaces of pronounced transitions of the water-vapor content of the air.

The theoretical approaches indicate that certain sharp echo effects may be produced almost entirely by dielectric transition and deviation layers of thickness of the order of one wavelength or less. This coincides with the observation of single discrete echoes from air-mass boundaries by medium-frequency sounding, and of scintillating and multiple-point or "dot" echoes from similar strata by microwave radar sounding. More pronounced reception of the undulating echo bands, which are apparently from continuous dielectric gradient regions, is indicated for medium and high frequencies, in contrast with the generally more discrete echoes observed with microwaves.

It is concluded that a set of perhaps three vertical beam systems for simultaneous and continuous sounding and recording of echoes from the troposphere could provide a most comprehensive supply of data by operation with high-gain antenna systems and maximum available power on, perhaps, 400, 4,000, and 40,000 Mc. The first of these respective frequencies would be used for the exclusive recording of major dielectric-variation phenomena, the second frequency would be used for the recording of pronounced dielectric fine-structure effects and for the penetration of dense precipitation regions to record their structures, and the third frequency would be used for the detection of the smallest concentrations of the most minute precipitate or dust particles.

* Decimal classification: R122.2×R116. Original manuscript received by the Institute, May 3, 1948; revised manuscript received, August 19, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 25, 1948. The research reported in this document was made possible, in large part, through support extended Cruft Laboratory, Harvard University, jointly by the Navy Department (Office of Naval Research) and the Signal Corps, United States Army, under ONR Contract N5ori-76, T.O. 1.

† Formerly, Cruft Laboratory, Harvard University, Cambridge Mass.; now, Radio Corporation of America, RCA Laboratories Division, Princeton, N. J.

Symbols and Abbreviations

a = radius of raindrop in equation (2)

a = constant defined in equation (81)

a = antenna efficiency factor in equation (96)

a_1 = constant of the transformation in equation (59)

A = constant of the Epstein transformation

$A = E_0/8 = 27.75\sqrt{P_{kw}G_t}$, as defined in equation (71)

A = total area of wave propagation cross section in equation (87)

A_0 = cross-sectional area of a small discrete reflecting region

A_1 = constant of the transformation in equation (59)

A_e = effective area of receiving antenna

A_p = area of paraboloidal antenna reflector

b_1 = constant of the transformation in equation (60)

B = receiver acceptance bandwidth before detector, in cycles per second

B_1 = constant of the transformation in equation (60)

$\beta = 4\pi/\lambda s = 4\pi f/c$

α, β, γ = constants in the hypergeometric differential equation

c = velocity of electromagnetic waves in free space

$\gamma = d\epsilon_r/dh$ = dielectric gradient

γ_0 = constant defined in equation (77)

γ_c = constant value of dielectric gradient

d = total distance from transmitter to reflecting surface and back to receiver

d_1 = term defined in equation (36)

d_2 = term defined in equation (36)

D = term in equation (18)

Γ = Gaussian gamma function

$e = 2.718$

$E_0 = 60I\sqrt{2G_t}$, as defined in equation (70)

E_r = reflected electric field-strength vector

ϵ = dielectric constant

$\epsilon_0 = (1/36\pi)10^{-9}$ dielectric constant of space

ϵ_r = relative dielectric constant

$\epsilon_r' = (\epsilon_r - 1)$

$\Delta\epsilon_r$ = a small difference of relative dielectric constant

$\exp(\ ) = e^{(\ )}$, the exponential

f_{Mc} = frequency in megacycles per second

$f_\gamma(h) = \partial\epsilon_r'/\partial h$

$F_1 = \cdots F_e \cdots$ solutions of the hypergeometric differential equations

F_d = dissipation factor to account for power-loss ratio of circuit from antenna

F_{en} = external noise power, per unit bandwidth, divided by kT_n

F_n = noise factor of receiver

F_n' = noise figure of receiver with antenna attached

$F(\alpha, \beta, \gamma, x)$ = hypergeometric function

$g(\xi) = k^2\epsilon(\xi)$

G_r = power-gain ratio of receiving antenna in comparison with an isotropic radiator

G_T = power-gain ratio of transmitting antenna in comparison with a half-wave antenna in free space

G_t = power-gain ratio of transmitting antenna in comparison with an isotropic radiator

η = a co-ordinate of the electric field E

h = altitude (height)

h_0 = a constant defined in equation (77)

h_1 = lower altitude of region considered

h_2 = upper altitude of region considered

$h_n = h_2$ = second height of integration

$h_p = h_2 - h_1 = 1/2c\tau_p$ as defined by equation (69)

$h_z = h_2 - h_1$ = range interval of integration

hg.d. e. = hypergeometric differential equation

I = current at the center of a half-wave transmitting antenna

$k = 1.37 \times 10^{-23}$ joules per °K = Boltzmann's constant

$k = \omega/c = 2\pi/\lambda$ = wave number

k_0 = radar observer's constant defined in equation (98)

$k_1 = a + j$ = constant defined in equation (82)

k_1 = radar detectability constant, controlled by presentation, integration, and other similar aids for increasing detectability

$k_2 = a - j$ = constant defined in equation (83)

k_2 = pulse-shape factor which equals unity for a rectangular pulse shape

K = term in equation (18)

K_1, K_2, K_3 = abbreviations for combinations of α, β, and $ln(\ ) = \log_e(\ )\gamma$, as defined by equation (15a)

L = layer thickness (as defined)

$L_{1/2}$ = symmetrical layer thickness between half-deviation points

$\kappa = 4\pi/\lambda s$ = constant of the Epsteinian transformation

$M = \Delta\epsilon_r$ = maximum relative dielectric-constant deviation in a symmetrical layer

M_0 = an initial increment of dielectric constant with respect to unity

μ = constant of transformation in equations (59) and (60)

n = refractive index

n_0 = index of refraction of raindrop

$N = \Delta\epsilon_r = \epsilon_{r2} - \epsilon_{r1}$ = total transition difference of relative dielectric constant in a monotonic transition layer

ξ = height (distance) out from the center of a layer or space co-ordinate

p = total air pressure in millibars

P_{kw} = transmitted power, in kilowatts

P_r = received power, in equation (90)

P_{rj} = received power, in equation (2)

P_t = transmitted power

$P(x)$, $Q(x)$ = abbreviations in the hypergeometric differential equation

$p(\xi) = ln(-x)$, as used in equation (17)

r = pulse-repetition rate of radar system

r_0 = constant of integration in the text with equation (14)

$r(\xi)$ = as defined in the text with equation (13)

R = radial distance from radar antenna to raindrop, in equation (2)

R = electric field-strength (voltage) reflection coefficient or ratio

$|R|$ = *magnitude* of electric field-strength reflection coefficient or ratio $|R|_\Delta$ = magnitude of electric field-strength (voltage) reflection coefficient of an incremental layer of thickness (Δh)

R_a = radiation resistance of a half-wave dipole in free space

ρ = radius of curvature of dielectric-variation curve

$\rho = 2\pi a/\lambda$ in equation (2)

S = relative layer thickness (see (30))

S_0 = power captured by an isotropic reflecting region

S_r = power returned to receiver location from reflector

σ_j = radar scattering cross section of particles

T = absolute temperature of the air in °K in equation (103)

T = electric field-strength transmission ratio

T_n = absolute temperature of a noise source in °K

t = time (seconds)

τ_p = time duration of pulse, in seconds

ϕ = pulse-shape function

v = velocity of propagation $= c/\sqrt{\epsilon_r}$

w = water-vapor mixing ratio, in grams per kilogram of dry air

x = a variable space co-ordinate

x, y = variable of the hypergeometric differential equation

z = variable of equation (39)

$\omega = 2\pi f$ = angular frequency

* = denotes the conjugate function

Examples of the Experimental Results

THE PROCESS of continuous sounding of the troposphere by radar stems from the original experiments which were performed on medium frequencies during the interval between 1935 and 1941.[1-6] These early experiments produced numerous instances of discrete echoes which appeared to coincide with strata of the atmosphere, wherein there were abrupt dielectric boundaries. Aside from these observations, there were many unexplained reflection phenomena of a random and rapidly varying nature.

A Medium-Frequency Experiment

A short series of tests was conducted[7-9] upon the nearly plane, pine barrens of south central New Jersey, in 1942, in an effort to reduce all possible experimental errors of the medium-frequency (2.398 Mc.) sounding system to an absolute minimum. These tests indicated the validity of the previous experiments. In addition, they appeared to confirm an earlier hypothesis of extended patterns of interfering wave reflections from deep strata of rather moderate dielectric gradients. The peak transmitted pulse power was 40 kw on 2.398 Mc.

An example of an echo pattern which combines these two effects is shown in Fig. 1, for June 23, 1942. From the ground up to an elevation of 1.0 km, a rather high value of dielectric gradient is indicated by the radiosonde (RAOB) data, as the result of a steep lapse in the water-vapor content of the air. A characteristic wave-interference pattern was indicated by the radio echo pattern.

From about 1.0 to 1.4 km, the RAOB indicated an extremely high dielectric gradient in passing from the lower moist stratum to the dry upper air mass. A pronounced radio echo was apparently reflected from this region. The extended thickness of this layer may have

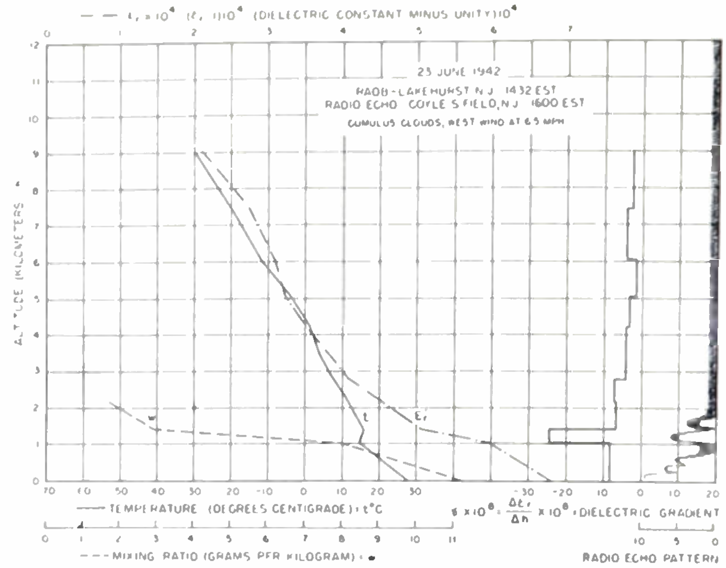


Fig. 1—First RAOB and radio echo plots; June 23, 1942.

produced the double-pulse effect, which is indicated in the pattern.

Later in the same day, after the passage of a minor frontal disturbance, the results shown in Fig. 2 were recorded. The more complicated dielectric gradient ex-

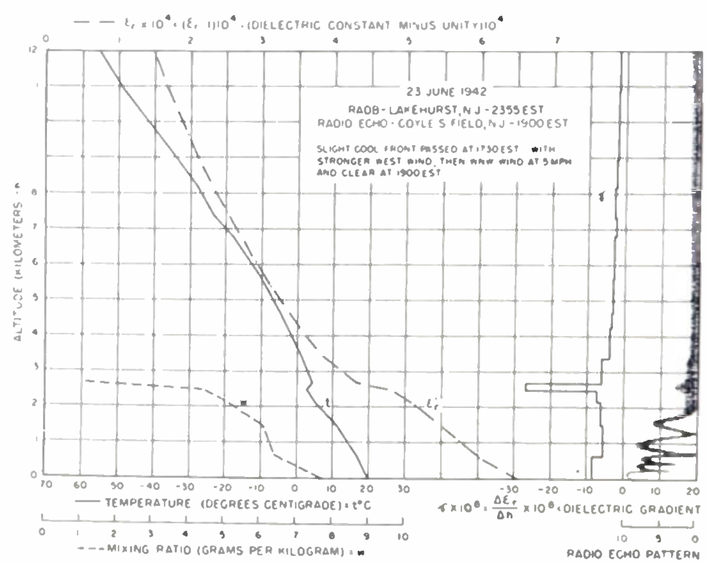


Fig. 2—Second RAOB and radio echo plots; June 23, 1942.

hibited below 2.5 km apparently gave rise to a much more complicated echo pattern, as indicated. The pulse echo from about 1.3 km may have been from an increase in dielectric gradient at about that level, which was only slightly indicated by the radiosonde. The very-well-defined upper boundary of the moderately moist layer of air, which had moved in below 2.5 km, produced a rather high value of dielectric gradient, which was apparently marked by a sharp echo pattern on the radio-wave-sounding indicator.

Figs. 1 and 2 represent only two of the thousands of such patterns which have been observed on this or similar sounding equipment since 1935. Their meanings

[1] R. C. Colwell and A. W. Friend, "Tropospheric radio wave reflections," *Science*, vol. 86, pp. 473–474; November 19, 1937.

[2] A. W. Friend and R. C. Colwell, "Measuring the reflecting regions of the troposphere," Proc. I.R.E., vol. 25, pp. 1531–1541; December, 1937.

[3] A. W. Friend, "Continuous determination of air-mass boundaries by radio," *Bull. Amer. Met. Soc.*, vol. 20, pp. 202–205; May, 1939.

[4] A. W. Friend and R. C. Colwell, "The heights of the reflecting regions in the troposphere," Proc. I.R.E., vol. 27, pp. 626–635; October, 1939.

[5] A. W. Friend, "Developments in meteorological sounding by radio waves," *Jour. Inst. Aero. Sci.*, vol. 7, pp. 347–352; June, 1940.

[6] A. W. Friend, "Further comparisons of meteorological soundings by radio waves with radiosonde data," *Bull. Amer. Met. Soc.*, vol. 22, pp. 53–61; February, 1941.

[7] These tests were conducted under a program of NDRC research, sponsored by J. A. Stratton, of the Massachusetts Institute of Technology, through the Radiation Laboratory.

[8] A. W. Friend, "Continuous tropospheric sounding by radar," Proc. I.R.E., vol. 36, pp. 501–503; April, 1948.

[9] A. W. Friend, "Reflections of vertically propagated electromagnetic waves from the troposphere." A thesis submitted to the faculty of the Graduate School of Engineering of Harvard University in partial fulfillment of the requirements for the S. D. degree in the field of communication engineering, Cambridge, Mass., August, 1947.

have become increasingly clear after years of observation, but the application of the mathematical analysis will bring the details into sharper focus.

The theory has indicated the desirability of conducting sounding tests on a frequency between 100 and 500 Mc. Numerous beginnings upon projects of this sort have been made since 1939, but each time the progress has been arrested by procurement problems or by obligations of higher priority. A program which was begun in 1946 is now being continued by others at Cruft Laboratory, Harvard University. It is hoped eventually to operate a paraboloidal antenna with a power gain of 300 times and a transmitter of 10^6 watts peak pulse power output.

Microwave Experiments

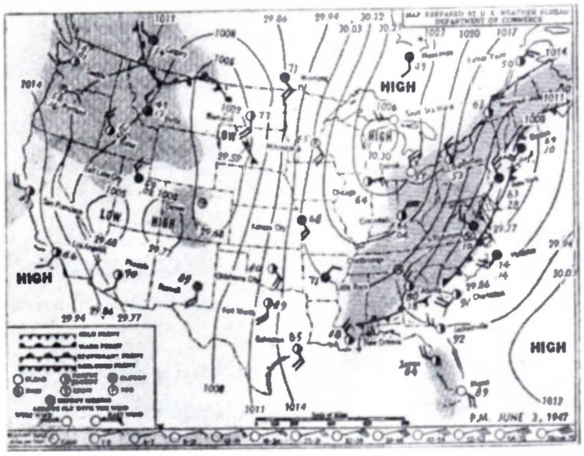
Recent tests with microwaves (2,800 Mc or 10.7 cm) indicate results which differ somewhat from those previously attained, but which may be related to the medium-frequency experiments by scaling down the depths of the strata of dielectric transitions or deviations and by considering small mixing-stratum eddies, instead of complete air-mass boundaries or large stratifications of the troposphere.

Patterns of apparent small-size reflection sources have been observed on plan-position-indicator (PPI) radar screens on clear days when very high concentrations of energy were used, with systems such as the AN/CPS-1 (MEW) microwave early-warning radar. Observations of these dot-echo phenomena on the PPI screen were reported in conversations during 1946 with A. G. Emslie, of Cruft Laboratory, Harvard University, and Lawrence Mansur, of the Cambridge Field Station, Air Materiél Command.

The author had been experimenting with vertical-beam tropospheric sounding on 2,800 Mc with a modified SCR-584 radar set. There were slight indications of echoes from the clear sky after all ground echoes had been eliminated, but it seemed that the energy concentration was insufficient. As a means for extending these tests, the United States Air Force allowed the addition of a vertical beam antenna to their modified experimental AN/CPS-1 radar system at Bedford Airfield (Mass.). The site of this set was not well arranged for suppression of the ground echoes, so, for lack of time, no effort was made to reduce the echo return from hills across the field, at about 6,900 to 10,200 feet (2.1 to 3.1 km). Numerous echoes of many types were observed from all unobscured ranges up to almost 50,000 feet (15.24 km), during all types of weather. The dot echoes which were previously observed on the PPI screen appear on continuous photographic records as stratified patterns of varying concentrations of dot echoes, at heights which corresponded with mixing boundaries between moist and dry masses of air, as measured by RAOB equipment at the United States Weather Bureau stations at Portland, Maine, and Nantucket, Mass.

There appear to be not less than three classes of echo-return mechanisms involved in explanations which include all of the various types of echo signals which have been observed by the author and by others,[10,11] within the frequency range between 1.6 and 30,000 Mc.

An example of a continuous sounding recorded from a modified SCR-584 radar system on June 3, 1947, is shown in Fig. 3. These echoes were of the very strong



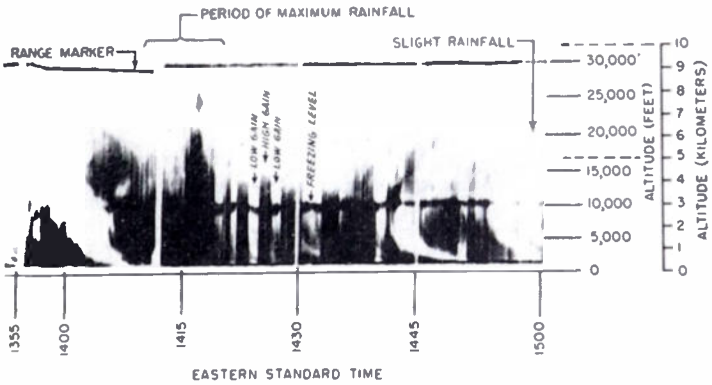


Fig. 3—Sounding record during heavy rainstorm, vertical-beam radar soundings on λ=10 cm with modified SCR-584 radar; June 3, 1947, at East Lexington, Mass.

type which are produced on microwaves by the scattering of the waves from particles of snow, ice, and water. Periodic reduction of the receiver gain allowed detection of the stratum just below the freezing isotherm where the melting snowflakes apparently produce larger amounts of reflection on account of the liquid water upon their surfaces.

An example of the performance of a higher-power and more concentrated beam of microwaves from a modified AN/CPS-1 (MEW) radar system is shown in Fig. 4. This chart, which was recorded on September 13, 1947, exhibits a number of various effects. The weather was hot and sultry, with a perfectly clear sky except for a few widely scattered, towering thunderheads. "Line" and "dot" echoes were recorded from the clear sky. A

[10] H. T. Friis, "Radar reflections from the lower atmosphere," Proc. I.R.E., vol. 35, pp. 494–495; May, 1947.

[11] W. B. Gould, "Radar reflections from the lower atmosphere," Proc. I.R.E., vol. 35, p. 1105; October, 1947.

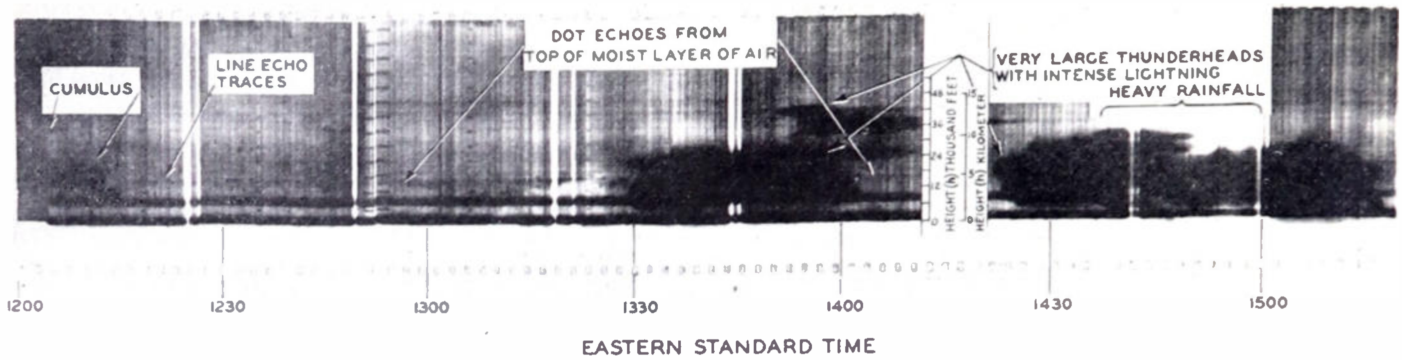


Fig. 4—Vertical beam sounding with modified AN/CPS-1 radar (2800 Mc) showing dot echoes and large thunderhead pattern taken at Bedford Airfield (Mass.) September 13, 1947, with the weather clear except for very large scattered thunderheads.

thunderhead, which arrived at 1330 EST, is indicated by the densely recorded signals from the scattering regions of the clouds and from the falling snow and raindrops.

A plot of the RAOB records from Nantucket, Mass., and Portland, Maine, is given in Fig. 5, along with interpretations of the recorded radar data, from Fig. 4. The main group of small dot echoes possessed a sharp upper boundary coincident with the top of a minor stratum of slightly moist air, at about 5 km altitude. Two somewhat clearer examples of dot echoes have been published by the author.[8,9]

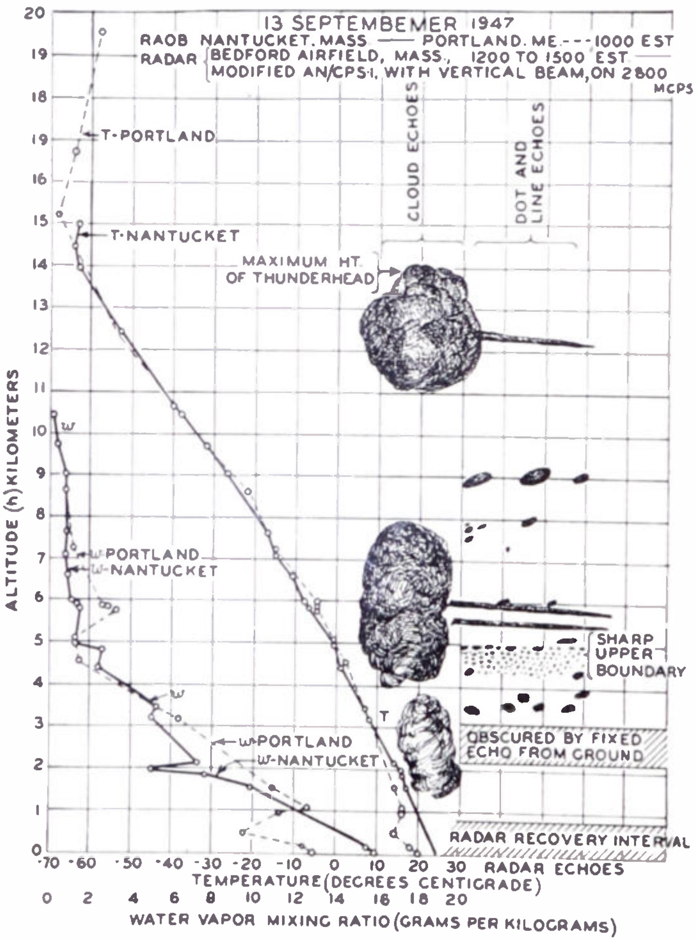


Fig. 5—RAOB plots and 2,800-Mc vertical-beam radar echoes; September 13, 1947

Particle-Scattering Echo Theory

The most obvious echoes which may be observed in the 2,000 to 30,000 Mc (microwave) spectrum are those produced by the effects of wave scattering from particles of water, snow, ice, or dust, which have been rather completely explained by Mie,[12] Stratton,[13] and J. W. and D. Ryde.[14–16] The latter authors have produced a solution which may be interpreted in terms of constants which make it accurate for water-drop sizes of less than 0.55 cm diameter for waves of 10 cm length (3,000 Mc), or more. It shows that the reflected power from a single raindrop is proportional to its equivalent radar cross section

$$\sigma_j = \frac{\lambda^2}{\pi}\left(\frac{n_0^2 - 1}{n_0^2 + 2}\right)\rho^6, \tag{1}$$

and that the received power is

$$P_{rj} = \frac{P_t A_p^2 \sigma_j}{9\pi R^4 \lambda^2} = \frac{P_t A_p^2 \rho^6}{9\pi^2 R^4}\left(\frac{n_0^2 - 1}{n_0^2 + 2}\right)^2$$
$$= \frac{64\pi^4 P_t A_p^2}{9R^4}\left(\frac{a}{\lambda}\right)^6\left(\frac{n_0^2 - 1}{n_0^2 + 2}\right)^2. \tag{2}$$

Possibly, the most interesting result which may be readily gleaned from these equations is that the reflected power from raindrops is proportional to the sixth power of the ratio of the radius of a raindrop to the wavelength of the radar wave (a/λ). In terms of the received voltage field strength (volts per meter, for instance), this becomes the third power of the ratio (a/λ).

The first obvious conclusion is that operation of a very-high-power tropospheric sounding system in the spectrum between 100 and 500 Mc should be relatively free from precipitation echoes. For instance, on 110 Mc

[12] G. Mie, "Beitrage zur Optik truber Medien, Speziell Kolloidaler Metallösungen;" *Ann. der Phys.*, ser. 4, vol. 25, pp. 377–445; 1908.

[13] J. A. Stratton, "Electromagnetic Theory," 1st edition, McGraw-Hill Book Co., New York, N. Y., 1941; pp. 563–573.

[14] J. W. Ryde, "Echo intensities and attenuation due to clouds, rain, hail, sand, and dust storms," General Electric Co., Ltd., Report No. 7831, October, 1941.

[15] J. W. Ryde and D. Ryde, "Attenuation of centimeter waves by rain, hail, and clouds," General Electric Co., Ltd., Report No. 8516, August, 1944.

[16] J. W. Ryde and D. Ryde, "Attenuation of centimeter and millimeter waves by rain, hail, fogs, and clouds," General Electric Co., Ltd., Report No. 8670, May, 1945.

the theoretical echo return by scattering should be about 85 db below that obtained with 2,800-Mc radar equipment of similar energy concentration and sensitivity.

Dielectric-Stratum Reflections

In the case of partial reflections from dielectric stratifications, the problem has not previously been so completely solved. The simplest assumption which one may make is that there is an infinite plane of division between two half-space regions each of which contains a homogeneous, isotropic medium which is a perfect nonconductor, which contains no free electrical charges, and which has a magnetic permeability of unity. It is further assumed that there is a difference in the specific inductive capacities (or dielectric constants) of these media, and that the bounding surface between the two regions is of infinitesimal thickness. A solution of this well-known problem may be found in the literature.[17] For the condition of normal incidence (vertical propagation toward a level stratum), it is found that the *voltage* reflection coefficient is

$$|R| = \frac{\sqrt{\epsilon_2} - \sqrt{\epsilon_1}}{\sqrt{\epsilon_2} + \sqrt{\epsilon_1}} = \frac{n_{12} - 1}{n_{12} + 1}, \tag{3}$$

where $|R|$ is the magnitude of the *voltage* reflection coefficient, ϵ_2 is the dielectric constant of one medium, ϵ_1 is the dielectric constant of the other medium, and n_{12} is the index of refraction of medium (1) with respect to medium (2). The dielectric constant ϵ is connected with the relative dielectric constant ϵ_r and the dielectric constant of space $\epsilon_0 = (1/36\pi)10^{-9}$ by the relation

$$\epsilon = \epsilon_r \epsilon_0. \tag{4}$$

From (4), (3) becomes

$$|R| = \frac{\sqrt{\epsilon_{r2}} - \sqrt{\epsilon_{r1}}}{\sqrt{\epsilon_{r2}} + \sqrt{\epsilon_{r1}}}. \tag{5}$$

In the atmosphere, $(\epsilon_r - 1)$ is never much greater than 7×10^{-4}, so that, within practical limits of accuracy, $\sqrt{\epsilon_{r2}} - \sqrt{\epsilon_{r2}} = (\epsilon_{r2} - \epsilon_{r1})/2$, and $\sqrt{\epsilon_{r2}} + \sqrt{\epsilon_{r1}} = 2$; thus (5) may be simplified to

$$|R| = (\epsilon_{r2} - \epsilon_{r1})/4 = \Delta\epsilon_r/4. \tag{6}$$

Thus, when a dielectric transition takes place within a small fraction of a wavelength, the reflection coefficient is simply one-quarter of the dielectric transition difference.

Reflection Coefficients of Monotonic Transition and Symmetrical Deviation Layers

The Wave Equation

It is desirable to find useful solutions for the reflection coefficients of rather thin strata wherein the dielectric property changes in gradual fashions which may approximate actual boundary-layer or mixing-stratum conditions. The assumptions of dielectric properties which vary as (1) functions of the height above the earth, (2) plane waves propagated vertically and incident upon horizontal strata, and (3) the absence of ionization or free charge of any sort, and hence no conduction or absorption effects, leads to wave equations of the type characterized by the expression

$$\frac{\partial^2 \eta}{\partial \xi^2} + k^2 \epsilon_r(\xi) \cdot \eta = 0 \tag{7}$$

wherein

$$k = \omega/c, \tag{8}$$

when the electric field varies in accordance with the harmonic time relation

$$E = \eta(\xi) e^{j\omega t} \tag{9}$$

as a function of the height ξ. A similar equation may be written for the magnetic field. It may be shown that the velocity of propagation is

$$v = c/\sqrt{\epsilon_r} = c/\sqrt{\epsilon_r(\xi)}. \tag{10}$$

From this, the index of refraction is

$$n = c/v = \sqrt{\epsilon_r} = \sqrt{\epsilon_r(\xi)}. \tag{11}$$

Except in the cases of some quite simple and trivial dielectric variational functions, (7) is, in general, soluble only in terms of certain approximation relations, which may considerably restrict the range of application of the solution. It is desirable to solve the wave equation (7) in a way which allows the results to be interpreted in as broad a manner as possible.

The Hypergeometric Differential Equation

In seeking a solution, one may investigate possible transformations of the variables. On account of the very general form of the Gaussian hypergeometric differential equation (hg.d.e.) it seems advisable to follow the method of Epstein[18] from this point, and to translate and introduce applicable portions of the very valuable extensions of Epstein's method which have been published by Rawer.[19]

The general form to be considered is given by the expression

$$0 = \frac{d^2y}{dx^2} = \frac{\gamma - (\alpha + \beta + 1)x}{x(1-x)} \frac{dy}{dx} - \frac{\alpha\beta}{x(1-x)} \cdot y$$

$$= \frac{d^2y}{dx^2} + P(x)\frac{dy}{dx} - Q(x) \cdot y \tag{12}$$

[17] See pp. 490–497 of footnote reference 13.

[18] P. S. Epstein, "Reflection of waves in an inhomogeneous absorbing medium," *Proc. Nat. Acad. Sci.*, vol. 16, pp. 627–637; October, 1930.

[19] Karl Rawer, "Electrische wellen in einer geschichten medium," *Ann. der Phys.*, vol. 35, pp. 385–416; July, 1939.

where α, β, and γ are three arbitrary constants. The variables ξ and η are next introduced in place of x and y, where $x=x(\xi, \eta)$ and $y=y(\xi, \eta)$. This transformation converts the wave equation in terms of ξ and η. It is required that all terms of the transformation which do not pass into the wave equation shall vanish. It follows from this that $\partial x/\partial\eta=0$ and $y=\eta\cdot r(\xi)$. Accordingly, there is the transformation

$$x = x(\xi); \qquad y = \eta\cdot r(\xi). \tag{13}$$

The substitution of transformations (12) and (13) proceeds in accordance with the sequence outlined by the author,[9] and in lesser detail by Rawer.[19] A solution is found for a value of $r(\xi)$ in terms of an integration constant r_0. Having obtained this value for r, a return is made to the transformation (13) of the hg.d.e. to obtain, from the balance of the equation, a wave equation of the same sort as (7). This may be rearranged and abbreviated[9] by a function $g(\xi)$, to yield

$$0 = \frac{d^2\eta}{d\xi^2} + g(\xi)\ \eta \tag{14}$$

wherein

$$g(\xi) = \left[\frac{1}{2} - \frac{d}{d\xi^2}\ln\frac{dx}{d\xi} - \frac{1}{4}\left(\frac{d}{d\xi}\ln\frac{dx}{d\xi}\right)^2\right] - \left(\frac{d}{d\xi}\ln x\right)^2 \cdot\left[K_1 + K_2\frac{x}{1-x} + K_3\frac{x}{(1-x)^2}\right]. \tag{15}$$

when

$$4K_1 = \gamma(\gamma-2);\ 4K_2 = 1-(\alpha-\beta)^2+\gamma(\gamma-2),$$
$$4K_3 = (\alpha+\beta-\gamma)^2 - 1, \tag{15a}$$

or

$$\alpha=\tfrac{1}{2}\left[1+\sqrt{1+4K_1}+\sqrt{1+4K_3}-\sqrt{1+4(K_1-K_2)}\right],$$
$$\beta=\tfrac{1}{2}\left[1+\sqrt{1+4K_1}+\sqrt{1+4K_3}+\sqrt{1+4(K_1-K_2)}\right],$$
$$\gamma=\sqrt{1+4K_1}+1. \tag{16}$$

These expressions are not uniquely determined here. They are to be employed in a later expression. Ultimately, it is to be proved that the definition

$$p(\xi) = \ln(-x) \tag{17}$$

is practical. Therefore, the equation

$$g(\xi) = \frac{1}{2}\left[\frac{d^3p}{d\xi^3}\Big/\frac{dp}{d\xi} - \frac{3}{2}\left(\frac{d^2p}{d\xi^2}\Big/\frac{dp}{d\xi}\right)^2\right] \quad (\text{“}D\text{ term”})$$
$$-\left(\frac{dp}{d\xi}\right)^2\left\{\left(K_1+\frac{1}{4}\right) - K_2\frac{\exp p(\xi)}{1+\exp p(\xi)} - K_3\frac{\exp p(\xi)}{[1+\exp p(\xi)]^2}\right\} \quad (\text{“}K\text{ term”}) \tag{18}$$

may be written.

In this fashion, following Rawer,[19] the hypergeometric differential equation (hg.d.e.) is converted into a wave equation with space-dependent dielectric properties, the form of which is determined by (15) or (18). A solution of this wave equation can then be obtained from known solutions of the hg.d.e.

A Solution of the Wave Equation

Gauss[20] has already given a solution of the hg.d.e. He first obtained the power series

$$F_1 = F(\alpha, \beta, \gamma, x)$$
$$= 1 + \frac{\alpha\beta}{1\ \gamma}x + \frac{\alpha(\alpha+1)\beta(\beta+1)}{1\cdot 2\cdot\gamma(\gamma+1)}x^2 + \frac{\alpha(\alpha+1)(\alpha+2)\beta(\beta+1)(\beta+2)}{1\ 2\ 3\cdot\gamma(\gamma+1)(\gamma+2)}x^3 + \cdots \tag{19}$$

which is the "hypergeometric series." It converges for $|x|<1$. In the equation family there is, however,

$$F_2 = x^{1-\gamma}F(\alpha-\gamma+1),\ \beta-\gamma+1,\ 2-\gamma,\ x), \tag{20}$$

which is also a solution. Both are linearly independent and, therefore, describe a fundamental system.[21] The hg.d.e. has three singular points, 0, 1, and ∞. For each of these points, it yields a fundamental system which converges in those regions.

With $x=0$:

$$F_1 = F(\alpha, \beta, \gamma, x);$$
$$F_2 = x^{1-\gamma}F(\alpha-\gamma+1),\ \beta-\gamma+1,\ 2-\gamma,\ x).$$

With $x=1$:

$$F_3 = F(\alpha, \beta, \alpha+\beta-\gamma+1, 1-x);$$
$$F_4 = (1-x)^{\gamma-\alpha-\beta}\cdot(\gamma-\alpha,\ \gamma-\beta,\ \gamma-\alpha-\beta+1,\ 1-x). \tag{21}$$

With $x=\infty$:

$$F_5 = x^{-\alpha}F(\alpha,\ \alpha-\gamma+1,\ \alpha-\beta+1,\ 1/x);$$
$$F_6 = x^{-\beta}F(\beta,\ \beta-\gamma+1,\ \beta-\alpha+1,\ 1/x).$$

The solution has already been given in the form of a series which is convergent only in a determined region. Each of these series defines an analytic function, however, which is a solution of the hg.d.e. throughout the region of convergence of the series. Now the analytic arrangement F_5 of the given series is considered to be extended to the indicated convergence region of F_1 and F_2. F_j shall henceforth designate the series and their analytic arrangement. The "natural" functions F_1 and F_2, and also a solution of F_5, are now available.

Since three solutions of a linear differential equation of the second order are always linearly dependent, there is a linear relation with constant coefficients, between

[20] F. Gauss, "Werke," *Gottingen*, vol. 3, pp. 125–207; 1876.
[21] Riemann-Weber, "Die partiellen Differentialgelichungen der mathematisch Physik," sixth edition, vol. 2; Braunschweig, 1919.

these three functions. This relation is valid for analytic functions and also for all values of x to which one may extend the three functions. Gauss has already stated these relations, according to Schlesinger,[22] in the form

$$F_5 = (-1)^{-\alpha}\frac{\Gamma(\alpha-\beta+1)\cdot\Gamma(1-\gamma)}{\Gamma(1-\beta)\cdot\Gamma(1+\alpha-\gamma)}F_1 + (-1)^{\gamma-1-\alpha}\cdot\frac{\Gamma(\alpha-\beta+1)\cdot\Gamma(\gamma-1)}{\Gamma(\gamma-\beta)\cdot\Gamma(\alpha)}F_2 \qquad (22)$$

where Γ represents the Gaussian gamma function.

From each solution y of the hg.d.e., through the transformations, a solution η of the wave equation may also be obtained for each F_j. There is also, in accordance with (13) and footnote references 19 or 9, and η_j in terms of the value of $r(\xi)$.

Epstein[18] has found that, by the special transformation

$$x = -Ae^{\kappa\xi}; \qquad \kappa > 0, \qquad (23)$$

which in the following discussion will be known as the "Epstein transformation," the relation (22) obtains a clear physical meaning. With the application of (23), the relation $dx/d\xi = \kappa\cdot x$ is obtained, and the asymptotes for large positive values of ξ (large negative values of x) are also obtained, so that

$$\eta_5 = r_0^{-1}\kappa^{-1/2}x^{(-1-2\alpha)/2}\cdot(1-x)^{(\alpha+\beta-\gamma+1)/2}\cdot F(\cdots 1/x)$$

$$\approx r_0^{-1}\kappa^{-1/2}(-1)^{(\gamma-1)/2-\alpha}A^{(\beta-\alpha)/2}\cdot\exp\left(\frac{\beta-\alpha}{2}\kappa\cdot\xi\right) \qquad (24a)$$

since consideration may be limited to the first term of the hypergeometric series $F(\cdots 1/x)$. Analogous developments are obtained for η_1 and η_2 in the neighborhood of $x=0$ (for very large values of ξ), to yield

$$\eta_1 = r_0^{-1}\kappa^{-1/2}(-A)^{(\gamma-1)/2}\cdot\exp\left(\frac{\gamma-1}{2}\kappa\xi\right). \qquad (24b)$$

and

$$\eta_2 = r_0^{-1}\kappa^{-1/2}(-A)^{(1-\gamma)/2}\cdot\exp\left(\frac{1-\gamma}{2}\kappa\xi\right). \qquad (24c)$$

When the constants of the hg.d.e. are chosen so that

$$\frac{1-\gamma}{j} \quad \text{and} \quad \frac{\alpha-\beta}{j} \qquad (24d)$$

are real and positive, the above asymptotic approximation for η_5 states the position dependence of a plane wave running in the direction of the ξ axis. The expression of the solution for the time-dependent wave equation is

$$\exp\left(-j\frac{\alpha+\beta}{2}\kappa\xi + j\omega t\right). \qquad (25)$$

Correspondingly, for large negative values of ξ, in the approximations of (24), it is revealed that η_1 propagates in the axial direction, and that η_2 is an oppositely propagating plane wave.

From the combination in (22), the three hypergeometric functions are (simply through expansion with $1/r$) an expression corresponding to the three solutions of the wave equation η_j. η_5 is thus a plane wave for large positive values of ξ, and when one follows it backward it appears (on the other side of the ξ-axis) as two waves combined together. A primary wave η_1 is split into two parts by passing through a region of varying dielectric constant. This results in a transmitted wave η_5 and a reflected wave η_2.

The relation (22) yields the reflected-wave ratio (reflection coefficient) R in terms of amplitude and phase. In the case of the transformation (23), following Epstein, $A=1$ is chosen from the corresponding values of the null points of ξ, to yield

$$R = \frac{\Gamma(\gamma-1)\cdot\Gamma(1-\beta)\cdot\Gamma(1+\alpha-\gamma)}{\Gamma(1-\gamma)\Gamma(\gamma-\beta)\Gamma(\alpha)}. \qquad (26)$$

The hypergeometric functions were first obtained (still multiplied by $1/r$) as a solution of the wave equation. The progress of a wave in a medium of varying dielectric properties may become too difficult to handle in ordinary terms for numerical computations, but now, by the aid of the gamma functions, it is possible to study the reflection coefficient alone.

It is evident from the derivation which was given for the reflection ratio that only the asymptotic case of the transformation function may be utilized. Therefore, these functions are to be used in the entire region of the Epsteinian sort only when they extend far outside of such an asymptotic region, on both sides. The plane-wave solution of (24) is, therefore, always to be retained as an asymptotic solution.

Only the asymptotic behavior of the transformation is thus determined. In the finite region it is still largely arbitrary. Of course, the conditions of the physical fundamentals must still be complied with, namely, that in (15) the dielectric constants (ϵ_r) shall be proportional to $g(\xi)$. In general, infinite regions shall, therefore, be excluded from $g(\xi)$. As in (15), setting of $x=0$ or 1 must be avoided, and moreover $dx/d\xi$ must not be permitted to vanish with infinite ξ (otherwise the derivation of the ln would not be finite). Also, $dx/d\xi$ must not become infinite, on account of the two terms in (15). Furthermore, it demands, from the physical fundamentals, that the entire axis shall be depicted by (13). To summarize, it can be required that

$x(\xi)$ shall be ***monotonic*** and that it shall ***represent*** the entire ξ *axis* or the portion of the x *axis* between 0 and $-\infty$. (27)

Here the use of the singular points 0 and ∞ is arbitrary. The same results can be obtained through the combination of 0 and 1 or 1 and ∞.

[22] L. Schlesinger, "Handbuch d. Theorie d. linearen Differentialgleichungen," vol. 1, p. 483; Braunschweig, 1919.

When any transformation which fulfills the requirements is employed, there is, therefore, obtained, for large negative values of ξ, perhaps $-A_0 \exp \kappa_0\xi$ and upon the other side $-A_\infty \exp \kappa_0\xi$, for large positive values of ξ. Again, (24) is maintained asymptotic, whereby now only the expression indices are still joined together. Again A_0 can be made equal to 1 through the choice of null points. Nevertheless, it is only in R that the various constants from (24) are shortened; and (26) remains valid in view of (27).

The Variation of the Dielectric Constant

In the previous paragraphs, formulas which were given for the solution of the wave equation (7) contained $k^2\epsilon_r(\xi)$ as the function $g(\xi)$. The position dependence of the dielectric constant is now to be considered.

a. *The Epsteinian Transformation of Equation* (23). First, in the simplest case, (18) gives $g(\xi)$, representing $k^2\epsilon_r(\xi)$. In the case of the transformation (23), from (17)

$$p = \kappa\xi. \tag{28}$$

That causes the D term of (18) to vanish, and, with $A=1$, the result is

$$\epsilon_r(\xi) = -\frac{\kappa^2}{k^2}\left[\left(K_1+\frac{1}{4}\right) - K_2\frac{e^{\kappa\xi}}{1+e^{\kappa\xi}} - K_3\frac{e^{\kappa\xi}}{(1+e^{\kappa\xi})^2}\right]. \tag{29}$$

The ξ axis is considered from the left to the right, representing by the "left" large *negative* values of ξ and by the "right" large *positive* values. In (29) ϵ_r has the value $-(\kappa^2/k^2)(K_1+\frac{1}{4})$ on the "left," and on the "right" $-(\kappa^2/k^2)(K_1+\frac{1}{4}-K_2)$. Between these lies the region in which ϵ_r varies. Its breadth is given approximately by $1/\kappa$. Therefore, a "*relative layer thickness*" S is introduced by the relation

$$S = 2k/\kappa = 4\pi/\lambda\kappa. \tag{30}$$

S is the layer thickness in wavelengths, so the breadth in meters is λS. The evident meaning of the constant κ in Epstein's transformation is apparent from this equation.

In the following considerations for tropospheric applications, it will always be assumed that ϵ_r is to remain very slightly greater than 1. An initial increment of ϵ_r above 1 is to be represented by M_0.

A suitable choice for K_1 is

$$K_1 = -\frac{1+S^2(1+M_0)}{4}. \tag{31a}$$

It is likewise advantageous to choose

$$K_2 = -\frac{S^2}{4}N, \quad \text{and} \quad K_3 = -S^2M. \tag{31b}$$

From (30) and (31), (29) becomes

$$\epsilon_r(\xi) = 1 + M_0 - N\frac{e^{\kappa\xi}}{1+e^{\kappa\xi}} - M\frac{4e^{\kappa\xi}}{(1+e^{\kappa\xi})^2}. \tag{32}$$

The second term represents a "monotonic transition" of the dielectric constant from $(1+M_0)$ to $(1+M_0-N)$. The third term represents a "symmetrical layer," which is of lower value than $(1+M_0)$ when M is positive, and of higher value when M is negative. $\epsilon_{r(\min)}=(1+M_0-N)$ and reaches a limit of 1 when $M=M_0$. M is, therefore, taken as the "layer deviation," and N as the "transition deviation." These two basic forms are shown in Fig. 6. If it is desired, these functions may be combined with linear constants. The meaning of the "layer thickness" (λS) is clear from the figure.

When the symmetrical layer is considered, the course of the dielectric-constant curve in the vicinity of the minimum (or maximum) may be determined approximately by means of the radius of curvature ρ of the curve of Fig. 6. It may be calculated from (32) as

$$\rho = M(\lambda S)^2/8\pi^2. \tag{33}$$

The value $L_{1/2}$ of the symmetrical layer thickness at the half-deviation points is approximately

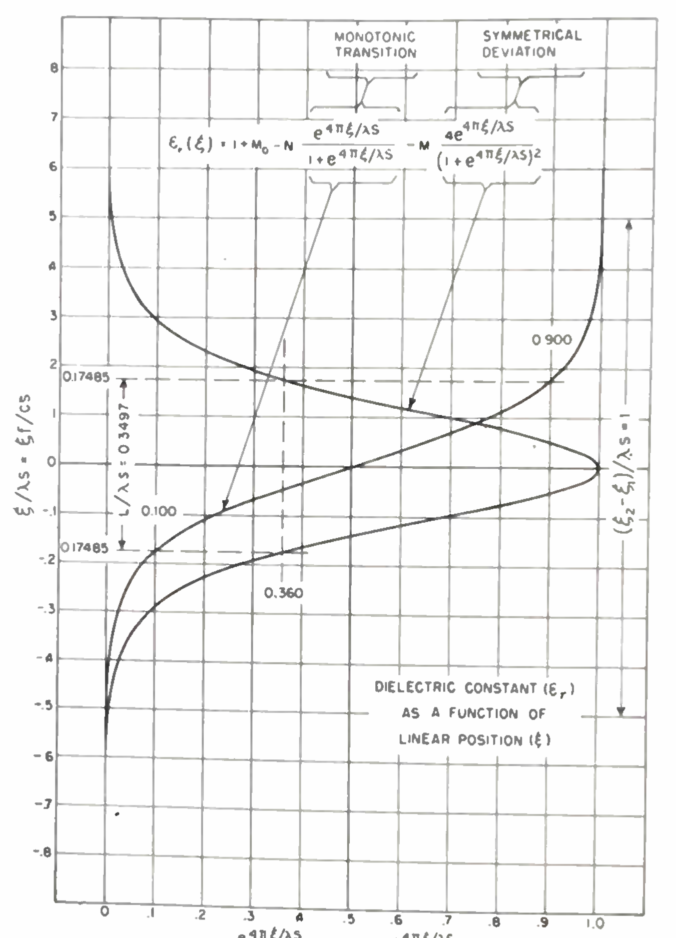


Fig. 6—Monotonic transition and symmetrical deviation curves.

$$L_{1/2} = 0.281\lambda S \quad \text{or} \quad L_{1/2}/\lambda S = 0.281, \qquad (34)$$

as used by Rawer.

In this application, the value to be called the "layer thickness" is the distance L between the points where the monotonic transition curve deviates by 10 per cent of the total deviation from each of its limiting values (ϵ_{r1} *and* ϵ_{r2} *in the dielectric deviation*). The transition and deviation curves have been normalized by plotting the transition and deviation fractions as functions of $\xi/\lambda S = \xi f/cS$. Therefore,

$$L = 0.3497\lambda S, \quad \text{or} \quad L/\lambda S = 0.3497, \qquad (35)$$

as indicated in Fig. 6. This definition makes L the distance between the equal-deviation points of the monotonic transition which encompass 0.800 of the total transition. In the case of the symmetrical layer, L represents the breadth (or thickness) of the deviation layer between the points where the deviation is 0.360 of the maximum amount.

Next, it is necessary to calculate the relations between the defined layer data and the constants of the hypergeometric differential equation (hg.d.e.). From the definitions (30) and (31), it follows that

$$\begin{aligned} jS\sqrt{1+M_0} &= \sqrt{1+4K_1}; \\ jS\sqrt{1+M_0-N} &= \sqrt{1+4(K_1-K_2)}; \\ \sqrt{1+4K_3} &= \sqrt{1+4S^2M} \\ &= 2(d_2+jd_1). \end{aligned} \qquad (36)$$

The last term defines d_1 and d_2 as the imaginary and real parts of the root, respectively. Substitution of these values into (16) yields

$$\begin{aligned} \alpha &= \frac{1}{2}+d_2+j\frac{S}{2}(\sqrt{1+M_0}-\sqrt{1+M_0-N})+jd_1, \\ \beta &= \frac{1}{2}+d_2+j\frac{S}{2}(\sqrt{1+M_0}+\sqrt{1+M_0-N})+jd_1, \\ \gamma &= 1+jS\sqrt{1+M_0}. \end{aligned} \qquad (37)$$

From (26), in view of (37),

$$R = \frac{\Gamma(jS\sqrt{1+M_0})\cdot\Gamma\left\{\frac{1}{2}-d_2-j\left[\frac{S}{2}(\sqrt{1+M_0}+\sqrt{1+M_0-N})+d_1\right]\right\}\cdot\Gamma\left\{\frac{1}{2}+d_2-j\left[\frac{S}{2}(\sqrt{1+M_0}+\sqrt{1+M_1-N})-d_1\right]\right\}}{\Gamma(-jS\sqrt{1+M_0})\cdot\Gamma\left\{\frac{1}{2}-d_2+j\left[\frac{S}{2}(\sqrt{1+M_0}-\sqrt{1+M_0-N})-d_1\right]\right\}\cdot\Gamma\left\{\frac{1}{2}+d_2+j\left[\frac{S}{2}(\sqrt{1+M_0}-\sqrt{1+M_0-N})+d_1\right]\right\}} \qquad (38)$$

where d_1 and d_2 are defined by (36).

The Monotonic Solution

In the case of the "monotonic transition," alone, $M=0$. Equation (38) may be modified for the simple monotonic state by setting $M=0$ and assigning a value other than zero to N. (When $N=M=0$ there is no transition.) In this case it is to be assumed that $N \leqq M_0$. Also, since $M=0$, it is found, from (36), that $d_2=\frac{1}{2}$ and $d_1=0$. Then, from (37), from the product

$$\Gamma(z)\cdot\Gamma(1-z) = \pi/\sin \pi z, \qquad (39)$$

and, since N is considered to be real, if only the magnitude is of interest, multiplication by the conjugate yields

$$|R|^2 = R\cdot R^*, \qquad (40)$$

from which the *magnitude* of the voltage gradient reflection coefficient, (38) becomes

$$|R| = \frac{\sinh\left[\frac{\pi}{2}S(\sqrt{1+M_0}-\sqrt{1+M_0-N})\right]}{\sinh\left[\frac{\pi}{2}S(\sqrt{1+M_0}+\sqrt{1+M_0-N})\right]}. \qquad (41)$$

The physical conditions under which application of (41) is intended indicate that $10^{-3} \geqq M_0 \geqq N$, so an excellent approximation is attained by setting

$$|R| = \frac{\sinh(\pi SN/4)}{\sinh\left[\pi S\left(1+\frac{M_0}{2}-\frac{N}{4}\right)\right]} \doteq \frac{\sinh(\pi SN/4)}{\sinh(\pi S)}, \qquad (42)$$

since

$$1 \gg \left(\frac{M_0}{2}-\frac{N}{4}\right).$$

When $(\pi SN/4) < 0.25$, an error of less than 1 per cent is involved in the relation

$$|R| = \frac{\pi SN}{4\sinh(\pi S)}, \qquad (43)$$

and, when $\pi S < 0.25$, it is found that, within 1 per cent,

$$|R| = \frac{N}{4} = \frac{\epsilon_{r2} - \epsilon_{r1}}{4} = \frac{\Delta\epsilon_r}{4}. \tag{44}$$

It may be noted that this is precisely the relation attained in (6) by more elementary considerations, from the viewpoint of Fresnel's equations. The Fresnel formulas assumed an abrupt boundary between two media of differing dielectric constants (ϵ_{r1} and ϵ_{r2}), while the Epstein transformation was employed in this case to treat a continuous monotonic transition from a dielectric (ϵ_{r1}) to another dielectric (ϵ_{r2}). In arriving at the final approximation (44), the layer thickness coefficient S was allowed to diminish until the transition occurred within a small fraction of a wavelength. The Epsteinian transformation is thus seen to provide a considerably more general solution of the dielectric-constant transition problem.

From (35),

$$S = 2.860L/\lambda$$

or

$$S = 9.538 \times 10^{-3} Lf_{Mc} = 2.860L/\lambda. \tag{45}$$

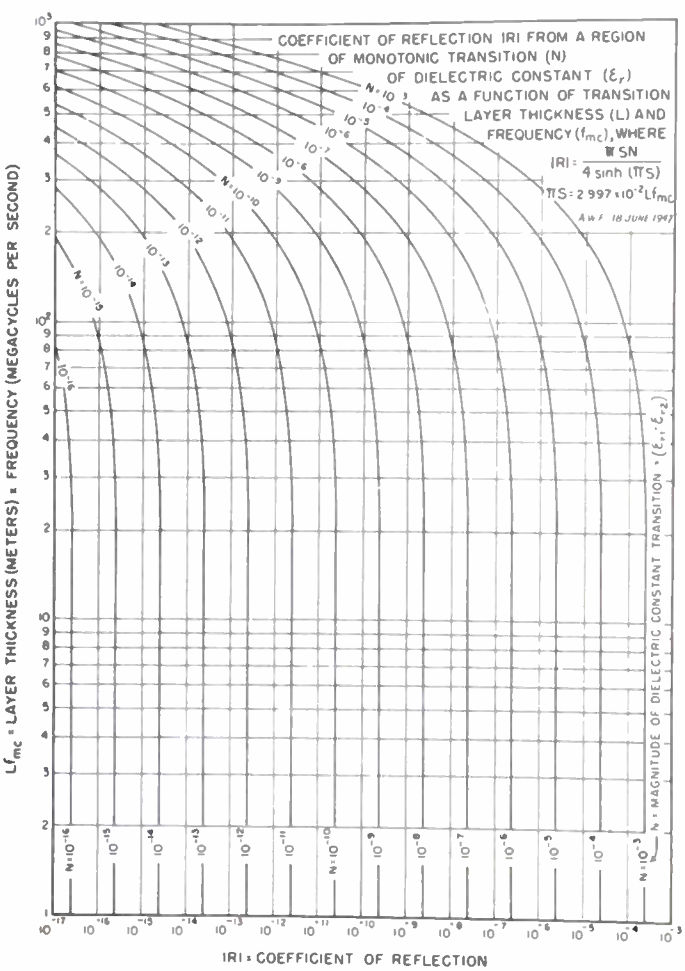


Fig. 7—Coefficient of reflection $|R|$ from a region of monotonic transition N of dielectric constant ϵ_r as a function of transition layer thickness L and frequency f_{Mc}, where $|R| = \dfrac{\pi SN}{4\sinh(\pi S)}$; $\pi S = 2.997 \times 10^{-2} Lf_{Mc}$.

From (45),

$$\pi S = 2.997 \times 10^{-2} Lf_{Mc}. \tag{46}$$

Therefore, the condition ($\pi S < 0.25$), assumed for (44) to make the error less than 1 per cent, leads to the conditions

$$Lf_{Mc} < 8.34 \quad \text{or} \quad L/\lambda < 0.0228. \tag{47}$$

When $L/\lambda = 1/3$, the true reflection coefficient, from (41), is $|R| = 0.332(\Delta\epsilon_r/4)$. When $L/\lambda = 1/10$, the exact value is $R = 0.887(\Delta\epsilon_r/4)$. This specifies an error of only 12.3 per cent for the $\Delta\epsilon_r/4$ value, when 80 per cent of a dielectric transition takes place within a distance equivalent to one-tenth of a wavelength.

The reflection coefficient $|R|$ in Fig. 7, computed for a dielectric transition in accordance with (41), is plotted versus Lf_{Mc}, with $N = (\epsilon_{r1} - \epsilon_{r2}) = \Delta\epsilon_r$ as a parameter. For the values of Lf_{Mc} less than 10, this family of curves represents the expression $|R| = \Delta\epsilon_r/4$ of (6) and (44). The ranges of N between 10^{-3} and 10^{-16} and of $|R|$ between 10^{-3} and 10^{-17} are believed to be completely adequate for dealing with all normal conditions which may be of interest.

If it is desired, the range of the chart of Fig. 7 may be easily extended by merely utilizing an appropriate change of the numbers of the N and $|R|$ scales when $Lf_{Mc} < 10^3$. When $Lf_{Mc} > 10^2$, (41) may be simplified to the form

$$|R| = 2\exp(-\pi S)\sinh(\pi SN/4), \tag{48}$$

and, when also $(\pi SN/4) > 2.40$, to

$$|R| = \exp\left[\pi S\left(\frac{N}{4} - 1\right)\right]. \tag{49}$$

The Symmetrical-Layer Solution

In the case of the "symmetrical layer" alone, $N = 0$. Equation (38) may be modified, under this condition and with (45), to yield

$$|R| = \frac{\Gamma(jS\sqrt{1+M_0})}{\Gamma(-jS\sqrt{1+M_0})} \cdot \frac{\Gamma\left[\frac{1}{2} - d_2 - j(S\sqrt{1+M_0} + d_1)\right] \cdot \Gamma\left[\frac{1}{2} + d_2 - j(S\sqrt{1+M_0} - d_1)\right] \cos\pi(d_2 + jd_1)}{\pi}. \tag{50}$$

The Case of $4S^2M > 1$

Again, if only the magnitude is of interest, one may set $|R|^2 = R \cdot R^*$, with (45) again introduced to eliminate the gamma functions. Then, if $4S^2M > 1$, so that $d_2 = 0$,

$$|R| = (\cosh^2 \pi S\sqrt{1+M_0} - \tanh^2 \pi S\sqrt{M - 1/4S^2} \cdot \sinh^2 \pi S\sqrt{1+M_0})^{-1/2}, \tag{51a}$$

when

$$d_1 = -j\tfrac{1}{2}\sqrt{1 - 4S^2M} = \tfrac{1}{2}\sqrt{4S^2M - 1},$$

$$\text{with } 4S^2M > 1, \tag{51b}$$

or, because $M_0 \ll 1$, (51a) may be written as

$$|R| = (\cosh^2 \pi S - \tanh^2 \pi S\sqrt{M - 1/4S^2} \cdot \sinh^2 \pi S)^{-1/2}. \tag{51c}$$

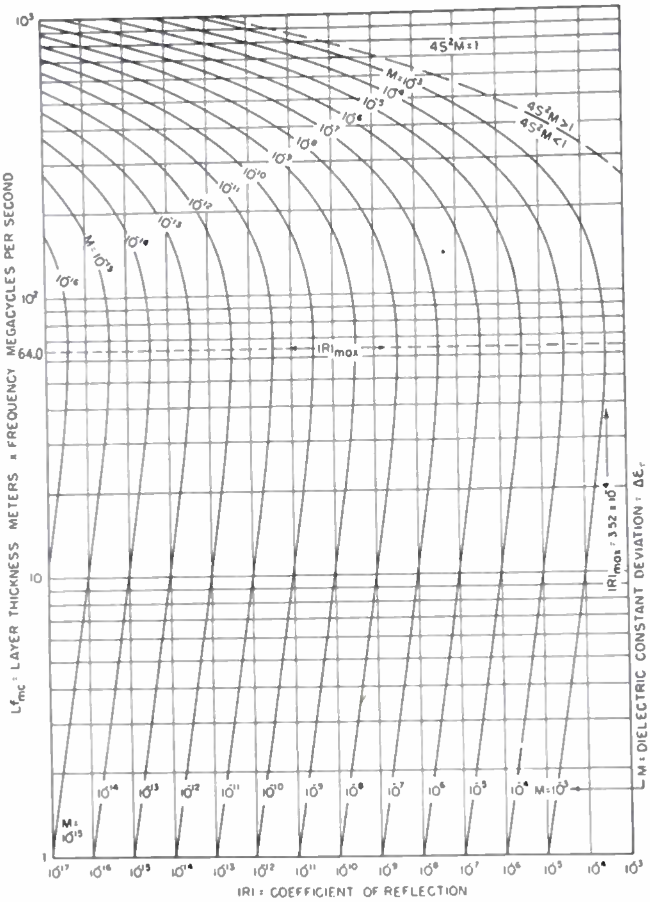


Fig. 8—Coefficient of reflection $|R|$ from a region of symmetrical deviation M of dielectric constant ϵ_r as a function of layer thickness L and frequency f_{Mc}, where $|R| = \{\cosh^2 \pi S + \tan^2 [(\pi/2\sqrt{1-4S^2M})\cdot]\sinh^2 \pi S\}^{-1/2}$, and $\pi S = 2.997 \times 10^{-2} Lf_{Mc}$, or if $4S^2M \ll 1$, $|R| = \pi S^2M/\sinh \pi S$.

The Case of $4S^2M < 1$

Suppose now that $4S^2M < 1$, so that $d_2 \neq 0$ and $d_1 = 0$. Then from (50), in view of (39), with the conjugate, as before, and by trigonometric conversion, the resulting equation is

$$|R| = \left(\cosh^2 \pi S\sqrt{1 + M_0} + \tan^2 \pi S\sqrt{\frac{1}{4S^2} - M} \cdot \sinh^2 \pi S\sqrt{1 + M_0}\right)^{-1/2} \tag{52a}$$

when

$$d_2 = \tfrac{1}{2}\sqrt{1 - 4S^2M}, \quad \text{with} \quad 4S^2M < 1, \tag{52b}$$

or, because $M_0 \ll 1$, (52a) may be written as

$$|R| = \left[\cosh^2 \pi S + \tan^2\left(\frac{\pi}{2}\sqrt{1 - 4S^2M}\right) \cdot \sinh^2 \pi S\right]^{-1/2}. \tag{52c}$$

The Case of $4S^2M = 1$, or $(d_1 = d_2 = 0)$

When $4S^2M = 1$, $d_1 = d_2 = 0$ and (51) and (52) become identical, so that

$$|R| = 1/\cosh \pi S\sqrt{1 + M_0}, \tag{53a}$$

or, since $M_0 \ll 1$,

$$|R| = 1/\cosh \pi S. \tag{53b}$$

The computation of the reflection coefficient $|R|$ is divided between two mathematical domains, as indicated by (51) and (52), so that (53) describes the intervening boundary.

Equations (51), (52), and (53) may be related to the layer thickness and frequency by (46), which was employed with the "monotonic transition." This applies the conditions of (35) and Fig. 6, so that the layer thickness L represents the distance, in meters, between the two points which represent 36 per cent of the maximum dielectric-constant deviation.

A chart which was evaluated from (52c) is reproduced in Fig. 8. It will be noted that the dashed line drawn diagonally across the upper right corner of the chart represents the boundary of (53). The values demanded for Lf_{Mc} and M in the present application require the use only of (52).

Useful Symmetrical Layer Solutions for the Troposphere

In evaluating (52c) for problems relating to the troposphere, it may be reduced to simpler forms in certain portions of the computation process. The evaluation of (52c), over the range of $10^3 \geq Lf_{Mc} \geq 1$ and $10^{-3} \geq M \geq 10^{-16}$, requires the use of the complete equation; but, except for the range of values $10^3 > Lf_{Mc} > 300$, where $M\sqrt{>10^{-4}}$, it is permissible to use the approximation $\sqrt{1 - 4S^2M} = 1 - 2S^2M$, where $4S^2M \ll 1$. For extension of the chart where $Lf_{Mc} > 10^3$ (where there is possibly a limited region of usefulness), this approximation has only a small zone of application. With the exception of this region, it is also found that

$$\cosh^2 \pi S \ll \tan^2 \frac{\pi}{2}\sqrt{1 - 4S^2M} \cdot \sinh^2 \pi S,$$

so that

$$|R| = 1/\tan \frac{\pi}{2}(1 - 2S^2M) \cdot \sinh \pi S. \tag{54}$$

In the region where $1.7124 \geq (\pi/2)(1 - 2S^2M) \geq 1.43$, when $2S^2M \ll 1$, $\sin (\pi/2)(1 - 2S^2M) = 1$ (within 1 per cent), so

$$|R| = \sin \pi S^2M/\sinh \pi S. \tag{55}$$

When $\pi S^2M \geq 0.248$, $\sin \pi S^2M = \pi S^2M$, within 1 per cent. This is true when $4S^2M \ll 1$, so the equation

$$|R| = \pi S^2 M/\sinh \pi S \quad (56)$$

may be used for plotting nearly all of the chart of Fig. 8.

This simple relation shows that, for all practical purposes, the reflection coefficient $|R|$ is directly proportional to the maximum deviation of the dielectric constant (M) of a symmetrical layer, and that it also varies in proportion to $\pi S^2/\sinh \pi S$.

It is found that when $\pi S < 0.25$, when $Lf_{Mc} < 8.34$, or when $L/\lambda < 0.227$, within 1 per cent, $\sinh \pi S = \pi S$, and in terms of (45),

$$\begin{aligned} |R| &= SM = 9.538 \times 10^{-3} MLf_{Mc} \\ &= 2.860\, ML/\lambda. \end{aligned} \quad (57)$$

This relation indicates that, if the layer thickness is very much less than a wavelength, the reflections from the top and bottom of the dielectric deviation layer tend to cancel one another and so to reduce the coefficient of reflection. If the layer thickness is much greater than one wavelength, the change of dielectric constant per wavelength is small, and so again the reflection coefficient is small. There is a value of the ratio of layer thickness to the wavelength (L/λ) which produces a maximum coefficient of reflection,

$$|R|_{max} = 0.352M \quad (58)$$

when $S = 0.6096$, so that $Lf_{Mc} = 64.0$ or $L/\lambda = 0.213$. This reflection coefficient is 42 per cent greater than the maximum value which may be obtained from a monotonic transition layer with a dielectric transition N equal to the maximum deviation M.

It is possible to combine the monotonic deviation and the symmetrical deviation conditions in the theory by the assignment of simultaneous values, other than zero, to both N and M.

Other Transformations

Other transformations are possible in addition to that disclosed by Epstein. They allow the insertion of different laws of variation of the dielectric constant. Rawers[19] has suggested the use of

$$p = \kappa\xi + A_1 \frac{e^{\mu\xi}}{a_1 + e^{\mu\xi}} \quad (59)$$

and

$$p = \kappa\xi + \frac{B_1}{\mu} \ln (1 + b_1 e^{\mu\xi}) \quad (60)$$

where A_1, B_1, a_1, b_1, and μ are constants of the transformation.

It is possible to compute extensive families of charts to indicate the effects produced by the many possible variations and combinations of the constants in these three transformations alone. It is believed that for the initial investigations, however, the essential data are mostly all contained in the charts of Figs. 7 and 8. More nearly precise matching of the theoretical variations of the dielectric properties with the actual variational conditions in the troposphere may modify the precise value of the computed coefficient of reflection, but probably not its order of magnitude. The present investigation is considered chiefly in terms of orders of magnitude of reflection coefficients, rather than their exact values. One has only to consider the present lack of precision and fine structure of RAOB data, or even of original recorded radiosonde data, to realize that comparisons of very small error are possible only with regard to very-large-scale effects.

An Integral Solution for the Pulse Echo Field Strength of the Basic Assumptions

Previous consideration has been given to a condition wherein a plane of infinite extent separated two half-spaces, each of which contained a nonconducting dielectric medium, without free charges. It was found that, if an electromagnetic wave was propagated normal to the separating plane, and if the difference between each dielectric constant and unity was very small in comparison with unity, the magnitude of the electric field-strength reflection coefficient of the wave could be written

$$|R| = (\epsilon_{r2} - \epsilon_{r1})/4 = \Delta\epsilon_r/4. \quad (61)$$

When the monotonic transition was considered, to ascertain the result when the boundary of transition from one medium to the other was *not* abrupt, the same expression was found to be applicable if the transition layer was less than one-tenth wavelength thick.

The relative dielectric constant of the air in the troposphere is found to vary as a function of any three space co-ordinates and time. It is observed that the variation in the vertical direction is usually of much greater magnitude than in any direction within horizontal planes, and so the troposphere may be considered as a horizontally stratified dielectric region.

The dielectric constant ϵ_r is to be considered as a function of height h, so that

$$\epsilon_r = f_\epsilon'(h) \quad (62)$$

and the reflected electric field is

$$E_r = f_e(h). \quad (63)$$

The vertical dielectric gradient γ is defined by

$$\gamma \equiv \frac{\partial \epsilon_r}{\partial h} = f_\gamma(h). \quad (64)$$

A point of consideration may be moved upward, normal to a horizontal plane surface, from height h_1 to h_2 so that $h_2 - h_1 = \Delta h$, the change of height. The dielectric constant changes simultaneously by an amount $\Delta\epsilon_r = \epsilon_{r1} - \epsilon_{r2}$. It may be assumed that a number of boundaries of relative dielectric-constant difference $\Delta\epsilon_r$

are each separated by equal height increments Δh. A coefficient of reflection $|R|_\Delta$ will, therefore, apply at each boundary. The reflection coefficient is always small ($|R| \ll 1$), so the change in amplitude of the upward (forward) propagating wave may be neglected. The reflected signal is, therefore, composed of the sum of the various reflected waves, each with a different delay time. The dielectric constant is assumed to be very nearly equal to unity, so that the speed of wave propagation may be considered equal to that of electromagnetic waves in empty space insofar as time delays are concerned.

When a simple harmonic wave amplitude is assumed, the reflected signal becomes

$$E_r = |E_0| \sum_1^n \left[\frac{|R_1|_\Delta e^{j\omega(t-2h_1/c)}}{2h_1} + \frac{|R_2|_\Delta e^{j\omega(t-2h_2/c)}}{2h_2} + \cdots + \frac{|R_n|_\Delta e^{j\omega(t-2h_n/c)}}{2h_n} \right], \quad (65)$$

or, by rearrangement and consideration of (6),

$$E_r = |E_0| \frac{e^{j\omega t}}{8} \sum_1^n \left[\frac{\Delta\epsilon_{r1}}{h_1} e^{-j4\pi h_1/\lambda} + \frac{\Delta\epsilon_{r2}}{h_2} e^{-j4\pi h_2/\lambda} + \cdots + \frac{\Delta\epsilon_{rn}}{h_n} e^{-j4\pi h_n/\lambda} \right]. \quad (66)$$

If this expression is divided and multiplied by Δh, on the right side, and if n is allowed to approach infinity as Δh approaches zero, while the total layer under consideration ($h_n - h_1 = h_p$) remains constant, (65) becomes

$$E_r = |E_0| \frac{e^{j\omega t}}{8} \lim \sum_{n=1}^{n=\infty} \left[\frac{\Delta\epsilon_{rn}}{\Delta h} \frac{1}{h_n} e^{-j4\pi h_n/\lambda} \right] dh, \quad (67)$$

and

$$E_r = |E_0| \frac{e^{j\omega t}}{8} \int_{h_1}^{h_n=h_2} \frac{d\epsilon_r}{dh} \frac{1}{h} e^{-j4\pi h/\lambda} dh$$

$$= |E_0| \frac{e^{j\omega t}}{8} \int_{h_1}^{h_2} \frac{\gamma}{h} e^{-j4\pi h/\lambda} dh \quad (68)$$

where $\gamma = f_\gamma(h)$ and $h_n = h_2$ represents any second height such that

$$h_2 - h_1 = h_p = \tfrac{1}{2} c\tau_p \quad (69)$$

where h_p is the vertical distance occupied by a pulse of wave energy of uniform amplitude in space.

Antenna radiation theory reveals that

$$|E_0| = 60I\sqrt{2G_T} = 60\sqrt{2 \times 10^3 P_{KW} G_T / R_a}$$

$$= 222\sqrt{P_{KW} G_T} \cdot \sqrt{2}, \quad (70)$$

or

$$|E_0|/8 = 27.75\sqrt{P_{KW} G_T} \cdot \sqrt{2} \equiv A\sqrt{2} \quad (71)$$

where A is a multiplier defined in (71).

In addition, a pulse-shape function ϕ may be used to multiply the terms under the integral sign. ϕ is assumed to be a function which specifies the relative amplitude of the voltage-gradient envelope of the radiated pulse in space. It is, thus, through the interdependence of time and space relations, considered at any instant as a function of space, or in this case as a function of altitude. If the radiated pulse is precisely rectangular, the function becomes a constant, and $\phi = 1$.

When A and ϕ are introduced into (68), and when the magnitude of the rms value of the electric field strength E_r is used, it becomes

$$E_r = A \int_{h_1}^{h_2} \frac{\phi\gamma}{h} e^{-j4\pi h/\lambda} dh \quad (72)$$

or, taking only the imaginary part,

$$E_r = -A \int_{h_1}^{h_2} \phi\gamma \frac{\sin \beta h}{h} dh \quad (73)$$

where

$$\beta = 4\pi/\lambda = 4\pi f/c. \quad (74)$$

The value of this integral expression is believed to reside chiefly in its adaptability for the insertion of desired pulse-shape functions $\phi(h)$ and dielectric-gradient functions $\gamma(h)$, and its possibilities for numerical or mechanical integration of otherwise difficult solutions.

Reflection of Rectangular Pulses from an Atmosphere of Constant Dielectric Gradient

If a rectangular pulse and a constant dielectric gradient are assumed, $\phi = 1$ and $\gamma = \gamma_c =$ a constant. Therefore, from (73),

$$E_r = -A\gamma_c \int_{h_1}^{h_2} \frac{\sin \beta h}{h} dh$$

$$= -A\gamma_c \int_{h_1}^{h_2} \frac{\sin \beta h}{\beta h} d(\beta h). \quad (75)$$

This expression may be expanded in terms of series[23] which are useful with large values of βh. In this application, it may be assumed that $\beta h \geqq 15.65$, so that the expression is accurate to within 1 per cent if only the first terms of the series are retained. It follows that

$$E_r = \frac{-A\gamma_c}{\beta} \left[\frac{\sin \beta h_1}{h_1} - \frac{\sin \beta h_2}{h_2} + \frac{1}{\beta} \left(\frac{\cos \beta h_1}{h_1^2} - \frac{\cos \beta h_2}{h_2^2} \right) \right]. \quad (76)$$

The envelopes of the maxima and minima of (76) are plotted as functions of frequency in Fig. 9, for the conditions $A = 27.75\sqrt{P_{KW} G_T}$, $P_{KW} = 10$ kw peak pulse power, $G_T = (1.5)^2$, $h_1 = 1000$ meters, $h_2 = 1500$ meters and $\gamma_c = -10^{-7}$. The value of γ_c was chosen as a repre-

[23] "Tables of Sine, Cosine and Exponential Integrals," vol. I, Federal Works Agency, Works Projects Administration for the City of New York, New York, N. Y., pp. XI and XII, 1940.

cent. This is true when $4S^2M \ll 1$, so the equation

$$|R| = \pi S^2 M / \sinh \pi S \quad (56)$$

may be used for plotting nearly all of the chart of Fig. 8.

This simple relation shows that, for all practical purposes, the reflection coefficient $|R|$ is directly proportional to the maximum deviation of the dielectric constant (M) of a symmetrical layer, and that it also varies in proportion to $\pi S^2/\sinh \pi S$.

It is found that when $\pi S < 0.25$, when $Lf_{Mc} < 8.34$, or when $L/\lambda < 0.227$, within 1 per cent, $\sinh \pi S = \pi S$, and in terms of (45),

$$|R| = SM = 9.538 \times 10^{-3} MLf_{Mc} = 2.860\, ML/\lambda. \quad (57)$$

This relation indicates that, if the layer thickness is very much less than a wavelength, the reflections from the top and bottom of the dielectric deviation layer tend to cancel one another and so to reduce the coefficient of reflection. If the layer thickness is much greater than one wavelength, the change of dielectric constant per wavelength is small, and so again the reflection coefficient is small. There is a value of the ratio of layer thickness to the wavelength (L/λ) which produces a maximum coefficient of reflection,

$$|R|_{max} = 0.352M \quad (58)$$

when $S = 0.6096$, so that $Lf_{Mc} = 64.0$ or $L/\lambda = 0.213$. This reflection coefficient is 42 per cent greater than the maximum value which may be obtained from a monotonic transition layer with a dielectric transition N equal to the maximum deviation M.

It is possible to combine the monotonic deviation and the symmetrical deviation conditions in the theory by the assignment of simultaneous values, other than zero, to both N and M.

Other Transformations

Other transformations are possible in addition to that disclosed by Epstein. They allow the insertion of different laws of variation of the dielectric constant. Rawers[19] has suggested the use of

$$p = \kappa\xi + A_1 \frac{e^{\mu\xi}}{a_1 + e^{\mu\xi}} \quad (59)$$

and

$$p = \kappa\xi + \frac{B_1}{\mu} \ln (1 + b_1 e^{\mu\xi}) \quad (60)$$

where A_1, B_1, a_1, b_1, and μ are constants of the transformation.

It is possible to compute extensive families of charts to indicate the effects produced by the many possible variations and combinations of the constants in these three transformations alone. It is believed that for the initial investigations, however, the essential data are mostly all contained in the charts of Figs. 7 and 8. More nearly precise matching of the theoretical variations of the dielectric properties with the actual variational conditions in the troposphere may modify the precise value of the computed coefficient of reflection, but probably not its order of magnitude. The present investigation is considered chiefly in terms of orders of magnitude of reflection coefficients, rather than their exact values. One has only to consider the present lack of precision and fine structure of RAOB data, or even of original recorded radiosonde data, to realize that comparisons of very small error are possible only with regard to very-large-scale effects.

An Integral Solution for the Pulse Echo Field Strength of the Basic Assumptions

Previous consideration has been given to a condition wherein a plane of infinite extent separated two half-spaces, each of which contained a nonconducting dielectric medium, without free charges. It was found that, if an electromagnetic wave was propagated normal to the separating plane, and if the difference between each dielectric constant and unity was very small in comparison with unity, the magnitude of the electric field-strength reflection coefficient of the wave could be written

$$|R| = (\epsilon_{r2} - \epsilon_{r1})/4 = \Delta\epsilon_r/4. \quad (61)$$

When the monotonic transition was considered, to ascertain the result when the boundary of transition from one medium to the other was *not* abrupt, the same expression was found to be applicable if the transition layer was less than one-tenth wavelength thick.

The relative dielectric constant of the air in the troposphere is found to vary as a function of any three space co-ordinates and time. It is observed that the variation in the vertical direction is usually of much greater magnitude than in any direction within horizontal planes, and so the troposphere may be considered as a horizontally stratified dielectric region.

The dielectric constant ϵ_r is to be considered as a function of height h, so that

$$\epsilon_r = f_\epsilon'(h) \quad (62)$$

and the reflected electric field is

$$E_r = f_\epsilon(h). \quad (63)$$

The vertical dielectric gradient γ is defined by

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A point of consideration may be moved upward, normal to a horizontal plane surface, from height h_1 to h_2 so that $h_2 - h_1 = \Delta h$, the change of height. The dielectric constant changes simultaneously by an amount $\Delta\epsilon_r = \epsilon_{r1} - \epsilon_{r2}$. It may be assumed that a number of boundaries of relative dielectric-constant difference $\Delta\epsilon_r$

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If a rectangular pulse and a constant dielectric gradient are assumed, $\phi = 1$ and $\gamma = \gamma_c =$ a constant. Therefore, from (73),

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The envelopes of the maxima and minima of (76) are plotted as functions of frequency in Fig. 9, for the conditions $A = 27.75\sqrt{P_{KW} G_T}$, $P_{KW} = 10$ kw peak pulse power, $G_T = (1.5)^2$, $h_1 = 1000$ meters, $h_2 = 1500$ meters and $\gamma_c = -10^{-7}$. The value of γ_c was chosen as a repre-

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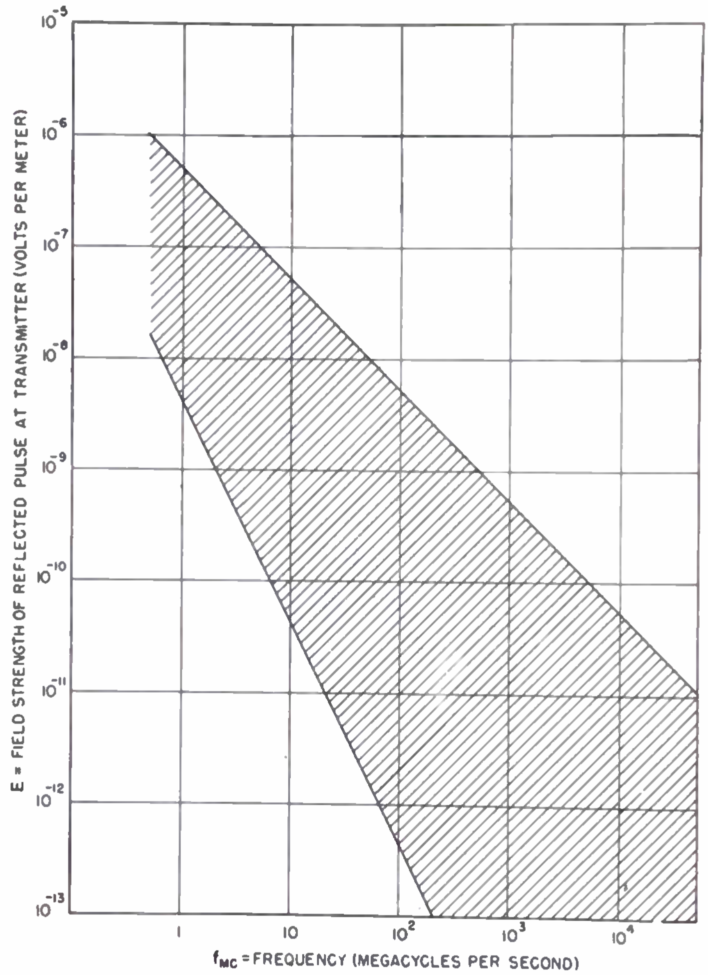

Fig. 9—Envelope of solution of $E=A\int_{h_1}^{h_2}\phi\gamma[(\sin\beta h)/h]dh$, rectangular pulse, $\phi=1$, constant dielectric gradients, $\gamma=\gamma_c=10^{-7}$, $A=27.75\sqrt{P_{KW}G_T}$, $P_{KW}=10kw$, $G_T=(1.5)^2$, $h_1=1000$ meters, $h_2=1500$ meters.

the reflection patterns. Precisely these effects are observed in the received patterns as they are displayed on the cathode-ray-indicator screen.

Reflection of Rectangular Pulses from an Atmosphere wherein the Dielectric Gradient γ Decreases Linearly with Respect to Altitude

If a rectangular pulse and a linearly decreasing dielectric gradient γ are assumed, $\phi=1$, and

$$\gamma = \gamma_0(h - h_0)/h_0, \tag{77}$$

where

h_0 = height in meters to point of intersection of assumed γ line with $\gamma=0$ (assumed average value, $h_0=12{,}000$ meters)

γ_0 = dielectric gradient at ground level (assumed average value, $\gamma_0=-8\times10^{-8}$, from smooth plots of RAOB data extended to ground level)

h = height in meters to any point below about $0.8\,h_0$ (γ actually approaches zero asymptotically).

Substitution of ϕ and γ from (77) into (73) yields

$$E_r = -\frac{A\gamma_0}{h_0}\int_{h_1}^{h_2}\left[\sin\beta h - h_0\frac{\sin\beta h}{h}\right]dh, \tag{78}$$

from which

$$E_r = \frac{A\gamma_0}{h_0\beta}(\cos\beta h_2 - \cos\beta h_1) + A\gamma_0[Si(\beta h_1) - Si(\beta h_2)]. \tag{79}$$

Then, by expansion in series, within 1 per cent if $\beta h \geqq 15.65$,

$$E_r = \frac{A\gamma_0}{\beta h}(\cos\beta h_2 - \cos\beta h_1) + \frac{A\gamma_0}{\beta}\left(\frac{\sin\beta h_2}{h_2} - \frac{\sin\beta h_1}{h_1}\right) - \frac{A\gamma_0}{\beta^2}\left(\frac{\cos\beta h_2}{h_2^2} - \frac{\cos\beta h_1}{h_1^2}\right). \tag{80}$$

sentative mean value, at low altitudes, for the Boston, Mass., region, by inspection of numerous curves of γ plotted from experimental radiosonde data. The other values are representative of average conditions of operation during the period of operation before the year 1940.

The stated value of γ_c is a mean for a smoothed curve of tropospheric dielectric gradient γ, so the received echo signal should appear as a curve with its amplitudes fluctuating about the computed mean value. In addition, the variations in the parameters of the equation should produce amplitude fluctuations as a function of time.

This latter effect could make a reflection from a constant-dielectric-gradient γ_c region appear as a group of numerous, closely spaced reflections. The maxima should reach the value of the upper envelope of Fig. 9. The minima may similarly approach the values of the lower envelope boundary, within very narrow zones. Meteorological variations and slight changes of the transmitter frequency may cause rapid fluctuations of

The data from the same series of radiosonde records yield parameters which make the plot of the maxima and minima envelopes of (80) almost the same as those of (76), as plotted in Fig. 9. The envelope curves are of similar shape and only a fraction of one order of magnitude less in value.

Reflection of Rectangular Pulses from an Atmosphere wherein the Dielectric Gradient γ Decreases Exponentially with Altitude

If a rectangular pulse and an exponentially decreasing dielectric gradient γ are assumed, $\phi=1$, and

$$\gamma = \gamma_0 e^{-ah} \tag{81}$$

where

γ_0 = dielectric gradient at the ground level with smoothed average exponential curves of γ. (In the Boston, Mass., region a representative value of $\gamma_0=-9\times10^{-8}$, with this curve.)

a = a constant with which the exponential relation yields a curve which best approximates the local weather conditions. (In this case ah should equal unity when $h=7000$ meters, so $a=1/7000 = 1.428\times10^{-4}$.)

Substitution of ϕ and γ, from (81), into (73) yields

$$\left.\begin{aligned} E_r &= -A\gamma_0\int_{h_1}^{h_2} e^{-ah}\cdot\frac{\sin\beta h}{h}\,dh, \\ &\text{or} \\ E_r &= \frac{jA\gamma_0}{2}\int_{h_1}^{h_2}\frac{1}{h}\left[e^{-h(a-j\beta)} - e^{-h(a+j\beta)}\right]dh \end{aligned}\right\}. \quad (82)$$

Then, if $k_1\equiv a-j$ and $k_2\equiv a+j$,

$$E_r = j\frac{A\gamma_0}{2}\left[\int_{h_1}^{h_2}\frac{e^{-k_1h}}{k_1h}\,d(k_1h) - \int_{h_1}^{h_2}\frac{e^{-k_2h}}{k_2h}\,d(k_2h)\right]. \quad (83)$$

Expansion in the power series of $\int(e^{-x}/x)dx$ leads to the solution

$$E_r = \frac{-A\gamma_0}{\beta}\left(e^{-ah_1}\cdot\frac{\cos\beta h_1}{h_1} - e^{-ah_2}\cdot\frac{\cos\beta h_2}{h_2}\right) - \frac{A\gamma_0}{\beta^2}\left[e^{-ah_1}\left(\frac{1}{h_1^2}+\frac{a}{h_1}\right)\sin\beta h_1 - e^{-ah_2}\left(\frac{1}{h_2^2}+\frac{a}{h_2}\right)\sin\beta h_2\right], \quad (84)$$

within 1 per cent, if the frequency in Mc is $f_{Mc}\geqq 0.034$, $a^2\ll\beta^2$ and $1\gg 2a/\beta^2h$, so that $h>500$ meters.

Plots of the maxima and minima envelopes of (84) lead to results which are likewise very similar to those from (76) (Fig. 9) and (79), but with values still a trifle smaller than in either of those cases, if similar operating conditions are assumed.

It is apparent from these three results that, within less than one order of magnitude, the simplest expression of (76) is almost as satisfactory for use in explaining the observed phenomena as are the more complicated arrangements.

From these data, one may observe that a 2.4-Mc sounding station, which radiates 10-kw peak pulses, should return an average peak threshold signal level of about 1.5×10^{-7} volt per meter from the region between 1.0 and 1.5 km altitude. This signal should be detectable by means of sensitive, medium-bandwidth pulse receivers.

The Maximum Envelope Gradient Reflection

Equation (76), from which Fig. 9 was plotted, may be reduced to the ***approximate maximum envelope*** expression

$$E_{r\,(\max)} = 662\sqrt{P_{KW}G_T}\,\gamma_c\left(\frac{1}{h_1}+\frac{1}{h_2}\right)\Big/ f_{Mc}. \quad (85)$$

This simple equation may be adopted as an indicator of the ***limiting maximum order of magnitude*** of tropospheric reflections from ***regions of slightly changing dielectric gradient*** γ as a function of height or range.

When (85) is applied, it ***must*** be recalled that there is a definite possibility of analytical requirements wherein the implications of the exponential parameter a of (84) may require investigation. If the value of a varies ***within*** the interval of integration, the amplitude of the reflection will be found to vary accordingly.

Equation (85) may be modified by introduction of a symbol $h_x\equiv h_2-h_1$, which represents the range interval of integration of the original expression in a region of constant dielectric gradient. In addition, this expression may be manipulated into the form of an equivalent-gradient maximum-reflection coefficient

$$R_{g(\max)} = c\gamma_c[1 + 1/(1 + h_x/h_1)]/16\pi f \quad (86)$$

where c is the speed of light and the distance h_x extends for at least several wavelengths (perhaps at least four wavelengths). This value may be used in the same fashion as other reflection coefficients.

Comparison of Theoretical Predictions with Experimental Data

A. Sounding System Parameters

1. *The Received Signal as a Function of Apparatus, Propagation, and Reflection Parameters:* If energy is radiated uniformly in all directions, from an "isotropic radiator," the power which flows through a unit area is

$$P_t/A = P_t/4\pi d^2. \quad (87)$$

As a result of reflecting systems and radiating element arrangements, the energy may be more or less concentrated in various directions in space. Very narrow beams may be rather easily produced in microwave systems. The concentration of power density S in space at a distance d from the transmitting antenna is

$$S = P_tG_t/4\pi d^2. \quad (88)$$

If a receiving antenna of effective area A_e is placed at a distance d, the total received power at that position in space is SA_e. The effective cross section of a receiving antenna is

$$A_e = G_r\lambda^2/4\pi. \quad (89)$$

Then, from (88), the total power received, at an antenna in space, is

$$P_r = P_tG_tG_r\lambda^2/(4\pi)^2d^2. \quad (90)$$

If reflection takes place at a plane boundary which is quite extensive, so that most of the radiated energy is intercepted by the boundary, (88) remains valid when multiplied by the power reflection ratio or coefficient, or by the square of the electric field-strength coefficient $|R|$. The distance d is then the total distance from the transmitter to the reflecting surface and back to the receiver. In vertical-beam sounding of the troposphere where the height of the reflecting stratum is h, and where the transmitter and receiver are close together,

$$d = 2h. \tag{91}$$

Equation (89) therefore becomes

$$P_r = \frac{P_t G_t G_r \lambda^2 |R|^2}{64\pi^2 h^2}. \tag{92}$$

This is the complete equation for the received power which is reflected from a thin dielectric stratum of reflection coefficient $|R|$ in the atmosphere, normal to the direction of propagation of the transmitted wave.

In the case of a small discrete reflecting region of cross-sectional area A_0, the portion of the intercepted energy which is scattered back toward the receiver may spread out in space as though from a new source of radiation. If isotropic radiation is assumed, the term $(4\pi h^2)^{-1}$ enters into both the transmitted and reflected propagation fields. The amount of power captured by the isotropic reflecting region is

$$S_0 = P_t G_t A_0 / 4\pi h^2. \tag{93}$$

These two considerations enter into the case of the small reflecting region. The power density returned to the receiver location, when an electric field reflecting coefficient $|R|$ is assumed, becomes, therefore,

$$S_r = P_t G_t A_0 |R|^2 / (4\pi)^2 h^4. \tag{94}$$

This introduces the familiar attenuation according to the inverse fourth power of the range h, as in radar detection of small objects.

The returned power is intercepted by the receiving antenna system. This may be the same antenna which is used for transmission. The effective cross section was given by (89). From (89) and (94), the received pulse power is

$$P_r = \frac{P_t G_t G_r \lambda^2 A_0 |R|^2}{64\pi^3 h^4}. \tag{95}$$

If an antenna reflector is used, and if the gain of the antenna system has not been measured, it is possible to use the equation

$$G = a(4\pi A/\lambda^2) \tag{96}$$

where a is the antenna "efficiency" and A the equivalent geometrical area of the antenna system or the geometrical area of the antenna reflector or collector of energy. The MIT Radiation Laboratory[24] used $a = 0.65$ for paraboloidal reflectors, but the true value may vary from 0.50 to 0.80. The Signal Corps has employed a value of $a = 2/3$. In the case of the cylindrical paraboloid with an array of dipoles excited from a waveguide, as used in the AN/CPS-1 (MEW) radar, the value of a appears to be about 0.750.

The echoes which are returned from meteorological phenomena of relatively small dimensions, in comparison with a widespread stratum, should be of magnitudes within the limits stated by (92) and (95).

The Minimum Detectable Signal

The well-known expression for the random noise power introduced into a receiver by thermal agitation is

$$P_n = F_n k T_n B \tag{97}$$

where F_n is the noise factor of the receiver; $F_n = 1$ for a perfect receiver completely overcoupled to the dummy antenna; $F_n = 2$ for a perfect receiver completely matched to the dummy antenna. The specifications of sensitive receiving systems usually include the actual noise level in terms of db above $kT\Delta f$ or kTB.

On the occasions of a number of discussions of technical papers delivered by the author, before World War II, there were discussions relating to the relative strength of a detectable synchronized pulse in the "noise" received by and generated within a receiver. It was stated that pulses had been observed at voltage levels somewhat below the noise voltage. It is well known that the noise voltage level represented upon a cathode-ray type-A presentation is rather indefinite, and that its amplitude relative to a received pulse amplitude may be a matter of personal judgment, unless it is computed from power measurements. One is, therefore, justified in concluding that these statements were rather vague.

Since that time, efforts have been made to determine more precisely the signal-to-noise ratios which prevailed during those earlier periods of measurement. It is known that the systems which the author operated on medium and high frequencies, between 1935 and 1942, utilized a system of operation with synchronized cw signal injection, which has since been called a "homodyne" or "coherent cw" arrangement.[25] A small amount of cw output which leaked out from the rf power amplifier of the transmitter was received along with the echo pulse signals. A similar additional synchronized signal has been shown[25] to contribute approximately 10 db improvement in the detectability of pulse signals passing through the receiver. Footnote 25 cites later intentional application of the principle to radar receivers.

Haeff[26] has shown that it is possible to repeat experiments relating to the limit of optical detection of pulse signals in a thermal noise background with reasonable accuracy. He used a 205-Mc carrier frequency with a

[24] D. E. Kerr and P. J. Rubenstein, "Introduction to microwave propagation," MIT Radiation Laboratory Report No. 406, p. 2; September 16, 1943.

[25] H. D. Webb, W. S. McAfee, and E. D. Jarema, "C.w. injection as a means of decreasing the minimum detectable signal of a radar receiver," Camp Evans Signal Laboratory, CESL Technical Report No. T-32, 34 pp., December 19, 1944.

[26] A. V. Haeff, "Minimum detectable radar signal and its dependence upon parameters of radar systems," Proc. I.R.E., vol. 34, pp. 857-861; November, 1946.

detector and a video amplifier, without coherent cw injection, and recorded the observations of many individuals upon a so-called type-A indicator, with signal amplitude versus time (range). These tests indicated that the minimum observable peak pulse power was

$$P_{\min} = k_0 k T_n B F_n \cdot \frac{1}{4}\left(1 + \frac{1}{B\tau_p}\right)^2 \cdot \left(\frac{1670}{r}\right)^{1/3} \quad (98)$$

where $P_{\min}$ is the minimum detectable pulse power at the terminals of the receiver, in watts, and k_0 is the personal constant of the observer ($k_0=1$ for the *average* observer).

A recent paper by Norton and Omberg[27] states that the noise figure *with the antenna in operation* is

$$F_n' = \left[\frac{F_n}{F_d} + (F_{en} - 1)\right], \quad (99)$$

where F_n, the receiver noise figure, is equal to the apparent noise power input to the receiver when operating from a pure resistance source at $T=300°K$ divided by kTB.

The Required Peak Transmitted Power for the Minimum Detectable Signal

Consideration of reflections from a plane stratum, of extent greater than the cross section of the radiated sounding beam, leads to a combination of (92), (98), (99), with $P_{\min}=P_r$. The minimum transmitted peak pulse power $P_{t(\min)}$ which may be utilized for detecting an echo from a region of voltage reflection coefficient $|R|$ and height h is found to be given by

$$P_{t(\min)} = \frac{(4\pi)^2 k_0 k_1 k T_n B f^2 [F_n/F_d + (F_{en} - 1)](1 + k_2/B\tau_p)^2 (1670/r)^{1/3} h^2}{G_t G_r c^2 |R|^2} \quad (100)$$

where k_1 is inserted as a constant to include the effects of the type of presentation of the signal, of photographic recording and other integrating methods, of "homodyne" or "cw injection," and of any and all additional devices and method for improving the effective signal-to-noise ratio, and k_2 is a pulse shape factor which is equal to unity for a rectangular pulse. The corresponding value of minimum detectable peak pulse signal level is

$$P_{r(\min)} = \tfrac{1}{4} k_0 k_1 k T_n B [F_n/F_d + (F_{en} - 1)](1 + k_2/B\tau_p)^2 (1670/r)^{1/3}. \quad (101)$$

The units are watts, meters, and cycles per second.

The Power Required on 2.398 Mc

It may be assumed that a half-wave dipole transmitting antenna may be used on 2.398 Mc, that it may be suspended at a height of about 0.1 wavelength above a horizontal reflecting plane surface, that the receiving antenna may be of similar design or that it may be a loop antenna suspended several meters above the earth, that the noise of the receiver is $F_n=1.64$ (which corresponds with a measured value of 2.16 db above kTB), that a coherent cw signal is injected into the receiver, and that an experienced observer monitors the signal indicator. This leads to a value of power P_{KW}, in kw, required for detection of echoes of electric field reflection coefficient $|R|$, from a height of h km, which is expressed by the relation

$$P_{KW} = 2.58 \times 10^{-13}\, h^2/|R|^2 \quad (102)$$

when the pulse duration is 8.33 microseconds and the receiver bandwidth is optimum. The values computed from this expression are believed to be usually reliable within somewhat less than one order of magnitude, depending upon the external noise and propagation conditions.

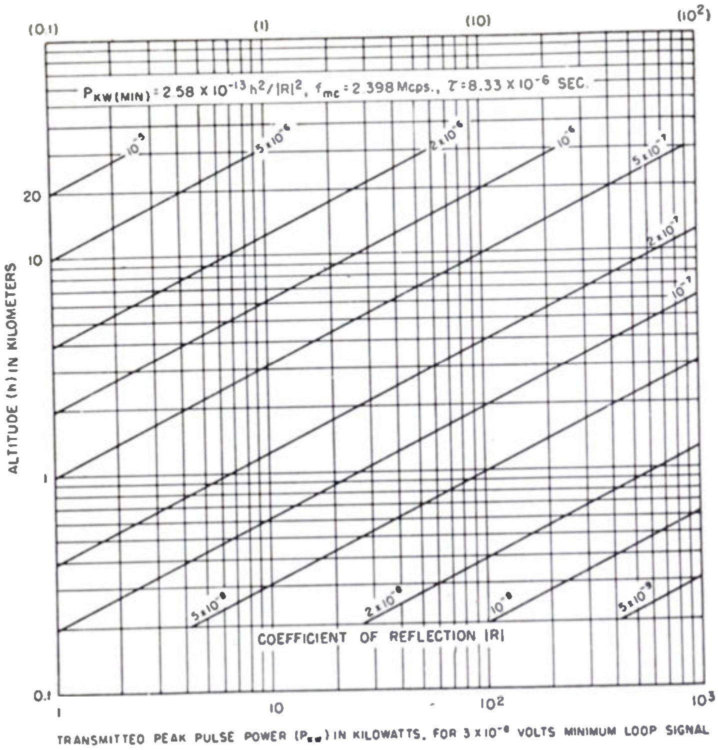

Fig. 10—Power required to produce a detectable reflection from an isolated stratum of reflection coefficient $|R|$.

The chart of Fig. 10 is plotted from (102). These data may be compared with possible meteorological conditions by converting the meteorological data to dielectric information by use of the charts of the report,[28] the key chart of which is shown in Fig. 11, or by the equation

[27] K. A. Norton and A. C. Omberg, "The maximum range of a radar set," Proc. I.R.E., vol. 35, pp. 4–24; January, 1947.

[28] A. W. Friend, "Charts of dielectric constant or refractive index of the troposphere," Cruft Laboratory, Harvard University, Cambridge, Mass., Contract N5ori-76, Task Order No. 1, Technical Report No. 34, March 10, 1948. Also, *Bul. Amer. Met. Soc.*, vol. 29, pp. 500–509; December, 1948.

$$10^6\epsilon_r' = (\epsilon_r - 1)10^6$$
$$= 158.3p\left[\frac{1}{T} + \frac{4815w}{(621 + w)T^2}\right]. \quad (103)$$

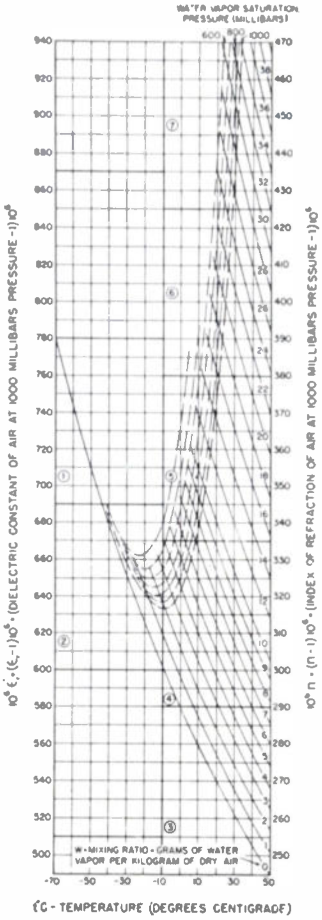

Fig. 11—Index chart of tropospheric dielectric constant or refractive index at 1,000 millibars pressure as a function of temperature and mixing ratio.

The air of the troposphere is never entirely free of water vapor, but, as a first approximation, it may be assumed that a temperature transition stratum of 1.0°C exists in perfectly dry air, that its thickness is less than one-tenth wavelength (12.5 meters), and that it is at a height of about 10.0 km. This indicates a voltage reflection coefficient of about 2.15×10^{-7}. The chart of Fig. 10 indicates a required peak pulse power of 600 kw for detection of this stratum by the 2.398-Mc system. If an echo is detected from that height when the peak transmitted pulse power is 40 kw, then a temperature transition of 2.7°C is indicated within a fraction of a wavelength, unless the presence of a water-vapor or particle-transition stratum is indicated. Thunderheads are observed by optical and microwave radar methods to heights of at least 15 km, so there must be significant amounts of water at such heights in both the crystalline and vapor forms up to at least this level. Cirrus clouds are regularly observed at 8 to 11 km altitude. The values of the precise quantities of water in the different states and the gradients which may be expected at these levels are not now available, but it seems that they could quite conceivably be responsible for the production of many of the echoes which have been observed with peak pulse powers of 40 kw, and on many occasions with even less than 10 kw.

These and other comparisons indicate that water vapor and particles of solid or liquid water, acting to increase the dielectric constant of the atmosphere on frequencies below perhaps 500 Mc, are of major importance in the production of detectable echoes from the troposphere.

A very extensive study of the United States Weather Bureau radiosonde records from Portland, Maine; Nantucket, Mass.; Lakehurst, N. J.; and Buffalo, N. Y., of special radiosonde records made in the Boston vicinity, by Harvard University, and of a series of meteorological data recorded by means of aircraft near Morgantown, W. Va., in connection with West Virginia University, was made in relation to simultaneous radio echo soundings which extended over a period from 1935 to 1942. From these data, an estimate of the approximate mean peak reflection coefficient $|R|$, on 2.398 Mc, has been plotted as a function of altitude h, in kilometers, as shown in Fig. 12.

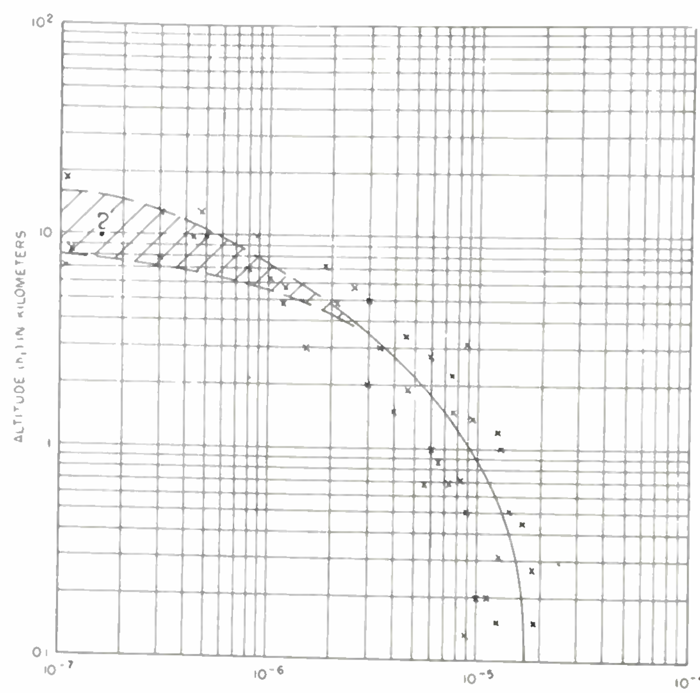

Fig. 12—Approximate mean peak reflection coefficient $|R|$ on 2.4 Mc, from northeast United States RAOB data.

The Power Required on 110 Mc

A system has been proposed for test on about 110 Mc for a number of years. There has to date been no opportunity to conduct these experiments, but this system might be arranged with a pulse length of perhaps either 2 or 8.33 microseconds, with optimum bandwidth, a pulse-repetition rate of 625 pulses per second, a homodyne receiving system, and an antenna power gain of 300 times with respect to an isotropic radiator in both transmitting and receiving. A 62-foot paraboloidal antenna reflector could be used with a dipole antenna radiator. When the 2-microsecond pulse is considered, the minimum required peak pulse power in kilowatts $P_{kw\,min}$ for detecting an echo from a stratum of voltage reflection coefficient $|R|$ from a height (range) h is computed from (99) to be

$$P_{kw(min)} = 1.08 \times 10^{-15}\, h^2/|R|^2. \qquad (104)$$

With a transmitted peak pulse power of 1 megawatt, this becomes

$$|R| = 1.04 \times 10^{-9}\, h. \qquad (105)$$

With a pulse length of 8.33 microseconds, these equations become

$$P_{kw(min)} = 2.59 \times 10^{-16}\, h^2/|R|^2, \qquad (106)$$

and, with 1-megawatt peak pulses,

$$|R| = 5.087 \times 10^{-10}\, h. \qquad (107)$$

All these values are for detection by observation on a cathode-ray indicator of range versus amplitude (type-A radar indicator) without the aid of the photographic integration, which may increase the detectability by at least 10 db. The results of (105) and (107) are plotted in Fig. 13.

In this case, with dry air at 500 millibars pressure, if there is a temperature transition of only 0.1°C within a range interval of 30 cm (about one foot), $N=\Delta\epsilon_r=1.14\times10^{-7}$, so that the stratum intercepts the entire beam at normal incidence, and the indicated detectable range is 26.5 km with the 2-microsecond pulse or 35 km with the 8.33 microsecond pulse. Similarly, a temperature change from −3°C to −5°C with a change of water-vapor mixing ratio from 2.5 to 2.0 gm/kg leads to $N=\Delta\epsilon_r=-6.75\times10^{-6}$. If this transition occurs within a range interval of 1 meter, $|R|=3.4\times10^{-7}$ for total normal interception, and the indicated possible range is 327 km with a 2-microsecond pulse. Without the use of homodyne reception, these ranges would be about one-third of the indicated values. In any event, the indicated results are exceedingly interesting to speculate upon.

The Indicated Performance with Stratum Echoes on 2500 Mc

The modified AN/CPS-1 (MEW) radar system, which was used in the experiments mentioned at the beginning of this paper, was operated with a 2-microsecond pulse at a repetition rate of 400 pulses per second with about 600 kw peak pulse power output. With type-A presentation, the indicated results are given by

$$P_{kw(min)} = 4.42 \times 10^{-13}\, h^2/|R|^2 \qquad (108)$$

or

$$|R| = 2.103 \times 10^{-8}\, h. \qquad (109)$$

The latter relation is plotted in Fig. 14.

The 0.1°C temperature change, within 30 cm, which was considered in connection with the proposed 110-Mc system, should yield $Lf_{Mc}=840$ on 2,800 Mc. Hence, for $N=\Delta\epsilon_r=1.14\times10^{-7}$, the reflection coefficient $|R|$ should be only about 5×10^{-17}, which is negligible. It seems clear that one must consider such changes within shorter range intervals of the order of magnitude of a wavelength or less. Such effects might readily occur in turbulent air, especially within large air-mass boundary regions. If only the moisture in the vapor state is con-

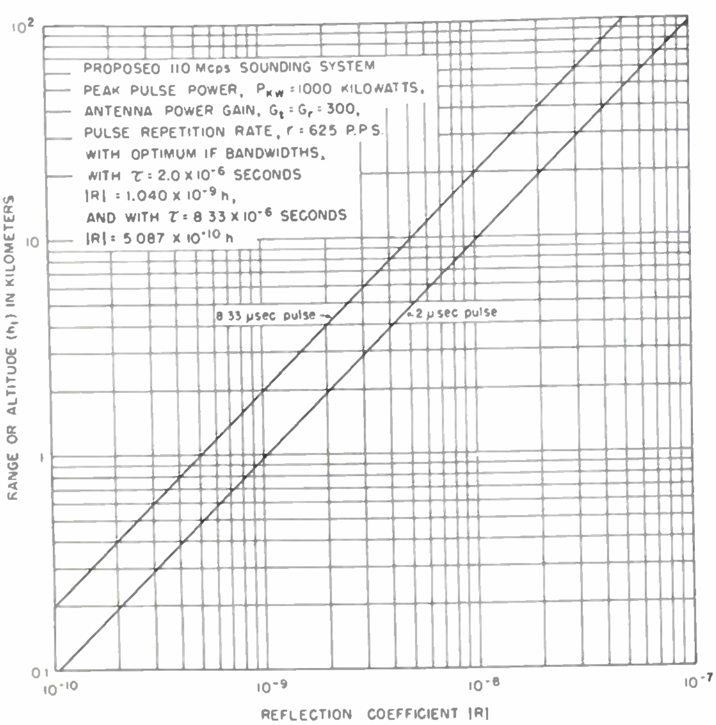

Fig. 13—Minimum voltage reflection coefficients $|R|$ detectable at height h with 10^6 watts peak pulse power, $f_{Mc}=110$ Mc, and $G_t=G_r=300$.

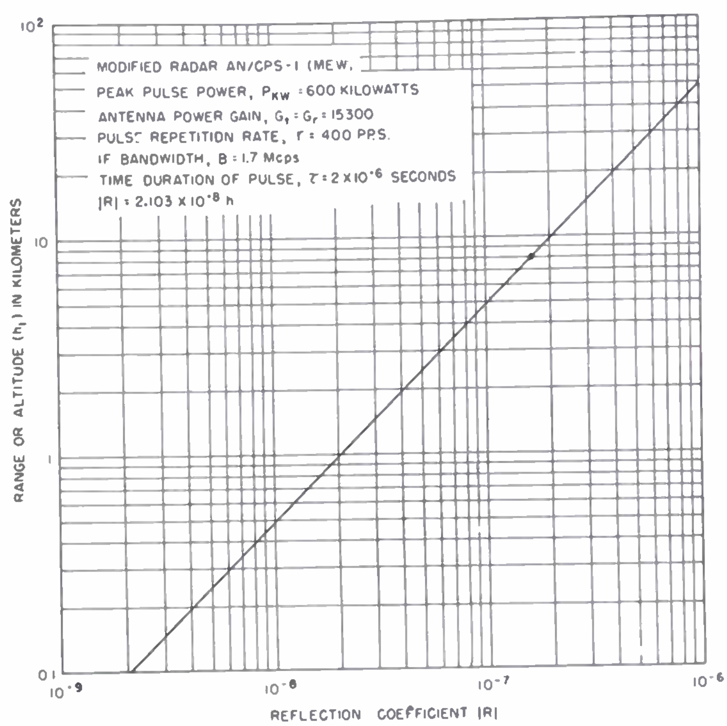

Fig. 14—Minimum voltage reflection coefficient $|R|$ detectable at height h with 6×10^5 watts peak pulse power, $f_{Mc}=2{,}800$ Mc, and $G_t=G_r=15{,}300$.

sidered, with a transition at an altitude of about 5.8 km, with $t_1=-3°C$, $w_1=2.5$gm/kg, $t_2=-5°C$ and $w_2=2.0$ gm/kg within a range interval of 2 cm, then $N=\Delta\epsilon_r=-6.75\times10^{-6}$ and $|R|=1.08\times10^{-6}$. This leads to a maximum range (for complete normal interception of the beam) of 51 km. At the same height, with $p=500$ mb, $t_1=-3°C$, $w_1=2.5$ gm/kg, $t_2=-3.3°C$, and $w_2=2.3$ gm/kg, $N=\Delta\epsilon_r=1.30\times10^{-6}$ and $|R|=2.08\times10^{-7}$. This yields an indicated maximum range of 9.8 km. *Many* of the *photographically recorded* echoes from other than particle-scattering regions appear to indicate reflection coefficients of *less than* 10^{-7}.

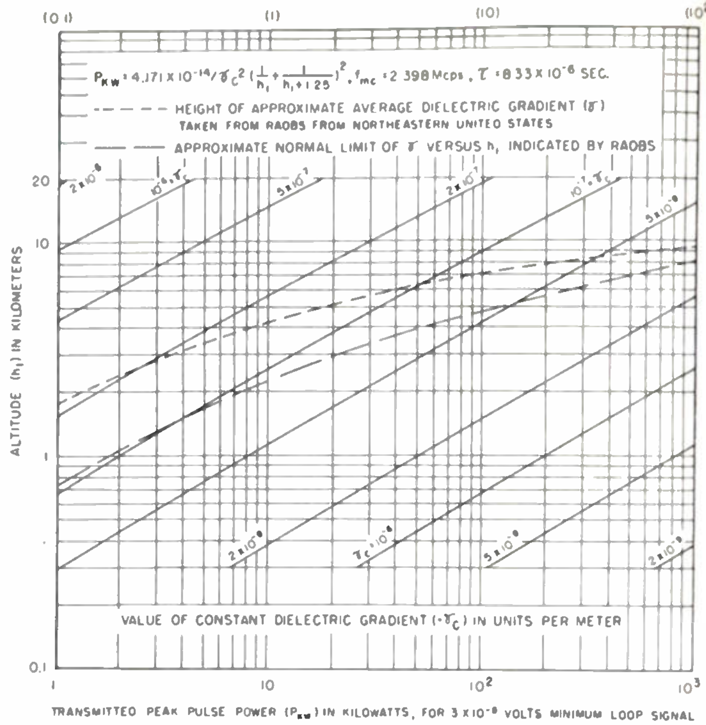

Fig. 15—Power required to produce a detectable reflection from a region of constant dielectric gradient γ_c.

The Required Peak Pulse Power for the Detection of Dielectric Gradient Echoes

The peak pulse power required for detection of dielectric gradient echoes may be determined by reference to (85) and (98). These relations may be converted to the same terms, combined and solved to yield an expression which reveals the required peak pulse output of the sounding transmitter, on 2.398 Mc. The conditions for solution of the problem when different antenna systems are used may be developed by introducing, in addition, (89) and (90). A general equation for the required minimum peak transmitted rectangular pulse power, in watts, is

$$P_{t(\min)}=\frac{16(4\pi)^4k_0kT_nBf^4[F_n/F_d+(F_{en}-1)](1+k_2/B\tau_p)^2\cdot(1670/r)^{1/3}}{G_tG_r\gamma_c^2c^4\left(\frac{1}{h_1}+\frac{1}{h_2}\right)^2}. \quad (110)$$

a. *Minimum Gradient Detectable on 2.398 Mc.* Operation on 2.398 Mc, with an 8.33-microsecond pulse, of 60-cps repetition rate, with optimum receiver bandwidth, homodyne reception, antenna power gains by reflection of four times, and the measured noise level, indicate a required peak pulse power output of

$$P_{kw}=4.17\times10^{-14}/\gamma_c^{-2}\left(\frac{1}{h_1}+\frac{1}{h_1+1.25}\right)^{-2} \quad (111)$$

in kilowatts, when the height h_1 is measured in kilometers and the average dielectric gradient γ_c is in units per meter. The chart of Fig. 15 is plotted from (111). Data from northeastern United States radiosonde records have been analyzed to determine the approximate average dielectric gradient γ_c values indicated by the lower dashed line. The upper dashed line indicates an approximate usual maximum limit of the radiosonde recorded average values of gradient as a function of altitude.

These data indicate that, with 40 kw peak pulse power output, gradient echoes may usually be detected, on 2.398 Mc, up to about 3.7 km altitude, and often to as high as 6.0 km. These figures appear to agree quite well with the experimental results. They may be a trifle conservative in many instances. This may be caused by an averaging of the maximum dielectric gradient values with the minima, which was practiced in an effort to provide a conservative analysis.

b. *Minimum Dielectric Gradient Detectable on 110 Mc:* Equation (110) may be applied with the conditions of (104) to (107) for the case of the postulated 110-Mc sounding station. When the 8.33-microsecond pulse is used, the minimum required power, in kilowatts, is

$$P_{kw}=8.83\times10^{-14}/\gamma_0^2\left(\frac{1}{h_1}+\frac{1}{h_1+1.25}\right)^2, \quad (112)$$

and the minimum detectable dielectric gradient is

$$\gamma_0=9.41\times10^{-9}(h_1+1.25)/(2+1.25/h_1) \quad (113)$$

when h_1 is in kilometers. The 2-microsecond pulse leads to the relations

$$P_{kw}=3.68\times10^{-13}/\gamma_c^2\left(\frac{1}{h_1}+\frac{1}{h_1+0.30}\right)^2 \quad (114)$$

and

$$\gamma_c=1.92\times10^{-8}(h_1+0.30)/(2+0.30/h_1). \quad (115)$$

Equations (113) and (115) are plotted in Fig. 16. Reference to the gradient curves of Fig. 15 in relation to Fig. 16 leads to the conclusion that a highly developed 110-

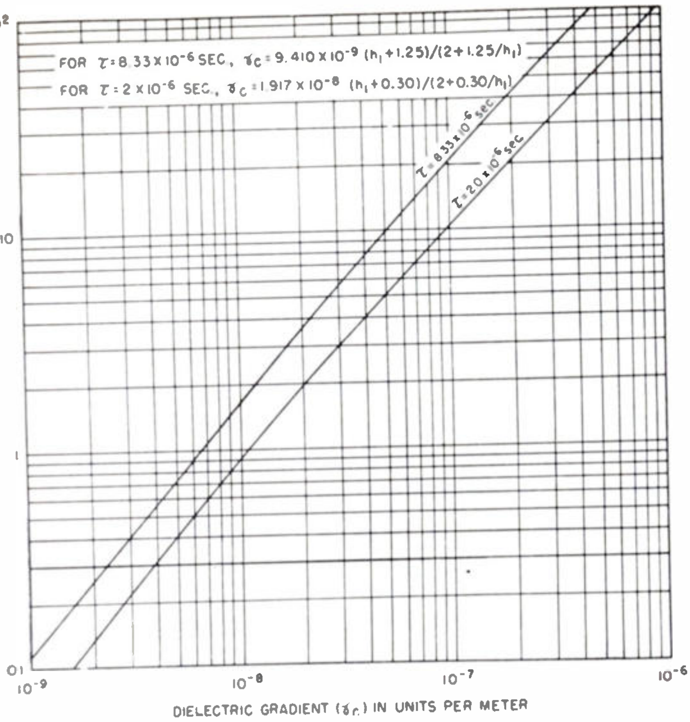

Fig. 16—Minimum detectable dielectric gradient γ_c at indicated height h_1 with 10^6 watts peak pulse, $f_{Mc}=110$ Mc, and $G_t=G_r=300$.

Mc sounding system should allow detection of continuous dielectric gradient echoes up to at least 8 km on many occasions, and on the average at heights up to 5 km.

c. *Minimum Dielectric Gradient Detectable on 2,800 Mc with Modified AN/CPS-1 Radar:* When (110) is applied to the modified 2,800-Mc type AN/CPS-1 radar system, which was outlined in connection with (108) and (123) the results are expressed by the equations

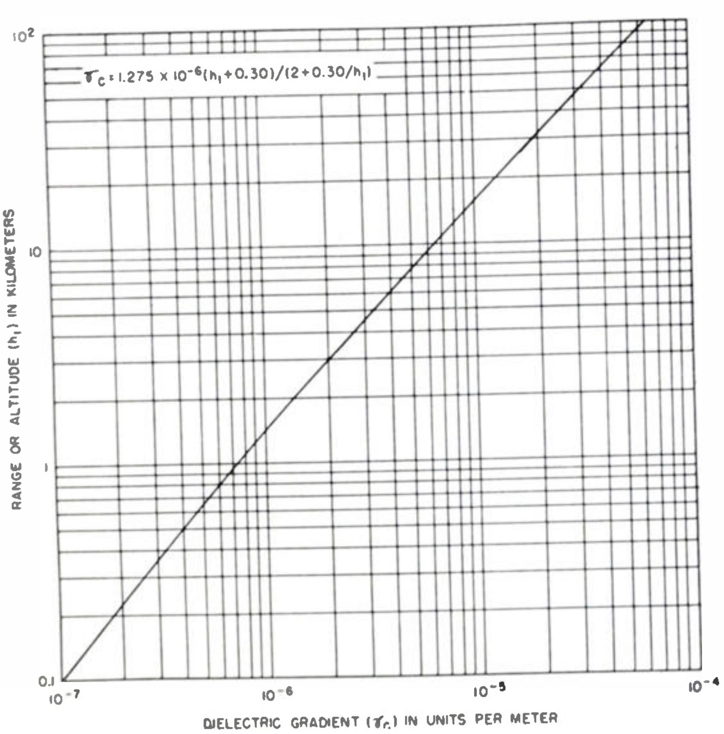

Fig. 17—Minimum detectable dielectric gradient γ_c at indicated range h_1 modified AN/CPS-1 (MEW) radar, $P_t=6\times10^6$ watts, $f_{Mc}=2{,}800$ Mc, $\tau=2\times10^{-6}$ seconds.

$$P_{kw} = 9.75 \times 10^{-9}/\gamma_c^2\left(\frac{1}{h_1} + \frac{1}{h_1 + 0.30}\right)^2 \qquad (116)$$

and

$$\gamma_c = 1.28 \times 10^{-6}(h_1 + 0.30)/(2 + 0.30/h_1), \qquad (117)$$

which are plotted in Fig. 17. This figure indicates that extended gradient reflections are probable only from ranges less than about 350 meters when the indicator is visually monitored. Photographic recording could increase this range to perhaps 1.0 km, and with homodyne detection also added, it might be increased to perhaps 3.0 km.

There is also the very probable possibility that much larger gradients which may extent for short distances, of the order of several wavelengths ($\lambda=10.7$ cm) may produce detectable echoes from much greater ranges. Equation (86) may be applied in the approximate analysis of these effects if the range interval (in meters) of the large gradient region is substituted for the value h_z, where f is in cycles per second, and where the boundaries of the gradient interval are quite sharply defined.

Relations of the Observed Reflection Effects to the Theoretical Developments

The medium-frequency echo patterns, which are typified by the patterns shown in Figs. 1 and 2, appear to be susceptible to separation into extended gradient and discrete thin stratum echoes and combinations thereof. These effects are believed to be explained in a broad sense by the theories of extended dielectric gradient and discrete dielectric transition reflection effects which have been outlined in this treatise. It is believed that these and similar theoretical developments may quite readily be extended to portray more precisely the effects which may be expected in connection with the details of the fine-structure of the atmosphere as these data become more clearly outlined and more readily available. At present, the pulse radio-wave soundings indicate that most dielectric transition strata are from one-fourth to one-half as thick as the radiosonde records lead one to believe.

The theories of the dielectric transition and deviation reflection coefficients lead to the conclusion that there are many important eddies and stratifications of the troposphere which are of the order of thickness of a very few centimeters or less. It appears further that such effects are quite often associated with air-mass boundaries of various sorts. The surfaces which separate quite moist air from very dry air appear to be particularly effective in the establishment of conditions which are favorable to the production of the "dot" type of echo. Theory and reason both seem to support these observations, since it is evident that water vapor plays a major role in the production of outstanding dielectric gradient effects of both the extended and sharply defined types.

Conclusions

It is concluded that the troposphere may be successfully sounded for various sorts of dielectric transition, deviation, and gradient effects by the use of pulses of electromagnetic waves within the frequency spectrum from perhaps 1.6 Mc to at least 30,000 Mc. The theory indicates that an optimum band of frequencies for the most effective sounding of dielectric phenomena, alone, appears to be that between 100 and 500 Mc. A peak pulse power output of perhaps 100 kw should be adequate for most purposes, when an antenna system with an area of about 3,000 square feet is used for both transmitting and receiving, and when the receiver employs the homodyne or coherent cw signal-injection system. A peak pulse power output of 1 megawatt should allow detection of all important regions of dielectric variation. In addition, the use of separate transmitting and receiving antennas should allow detection at a minimum range of only a few yards.

Optimum sounding for both dielectric variations (from air-mass boundaries and turbulence regions) and particle scattering effects (from precipitation or dust particles) may be most readily achieved in the frequency range between perhaps 2,000 and 4,000 Mc. The peak transmitted pulse power should be at least 500 kw, and the antenna power gains should be at least 15,000 times with respect to an isotropic radiator. Homodyne reception and separate transmitting and receiving antennas are desirable features. A peak pulse power of 5 megawatts, or more, should add a remarkable amount of information to the sounding record.

Frequencies in the 3,000- to 10,000-Mc interval appear to be extremely useful for the sounding of most types of precipitate regions if the system possesses the features listed above, for the 2,000- to 4,000-Mc system. These frequencies, with the very high transmitted power levels, allow the penetration of regions of high concentration of scattering particles.

Frequencies between 10,000 and 50,000 Mc appear best suited for the location of regions of very small concentration of small precipitate or dust particles. Maximum available power and sensitivity are demanded.

It is assumed that photographic recording is to be used in all cases, so that the detectability of echoes may be increased by about 10 db. An idealized sounding station might record simultaneously and continuously with vertical beam systems operating on perhaps 400, 4,000, and 40,000 Mc. In addition a light-beam operated ceilometer chart could be recorded. The data from all these devices should supplement those from the present visual and radiosonde observations, and perhaps also those from a microwave plan-position-indication (PPI) system (for indication of regions of precipitate within a radius of perhaps 100 miles), so that a really comprehensive weather analysis program might be established. The PPI system might operate on alternate pulses from one of the vertical-sector sounding systems. It could be provided with a vertical scanning function (for height finding and detailed analysis of semiremote cloud structures), and with an automatic tracking, computing, and plotting system for the almost instantaneous analysis and recording of radiosonde data.

The author's experimental work, which has already extended over a period of thirteen years, is now supported by the similar efforts of many others.[10,11,29] It is hoped that the addition of these new data, in combination with the somewhat further extended mathematical concepts, will be of assistance in the promotion of this program for providing increased data for meteorologists and aerologists, and for extending the scope of data relating to the propagation of electromagnetic waves in the troposphere.

Acknowledgment

This program of research and development has been made possible only through the aid, assistance, and continued encouragement of a large number of people, agencies, and institutions. The original impetus which led to this work was received through the interest of R. C. Colwell and from funds provided through West Virginia University by the State of West Virginia. The work herein described was supported mainly through aid extended the Cruft Laboratory of Harvard University through contract N5ori-76, Task Order 1, with the Office of Naval Research, United States Navy, and the Signal Corps, United States Army. Additional facilities of greatest importance were provided through the Cambridge Field Station of the Air Matériel Command, United States Air Force. Most of the radiosonde data were received through the courtesy of the United States Weather Bureau, with the assistance and co-operation of its chief, F. W. Reichelderfer. Other early assistance was provided through the Genradco Trust Fund and by anonymous donors through the Cruft Laboratory and the Blue Hill Meteorological Observatory of Harvard University, and by the National Defense Research Committee through the Radiation Laboratory of the Massachusetts Institute of Technology. H. R. Mimno, C. F. Brooks, and E. L. Chaffee of Harvard, and J. S. Stratton of the Massachusetts Institute of Technology, have been of major assistance in arranging and carrying forward the many details of this extended program. The tests with the modified AN/CPS-1 system were made possible by the co-operation of L. Mansur, J. A. Kelly, and W. J. Nadeau of the Air Matériel Command, Cambridge Field Station. The Radio Corporation of America contributed materially by personal arrangements which aided in continuation of the final phases of this project. The author gratefully acknowledges all of these contributions in the pursuit of this program, and the assistance and encouragement which has been provided by numerous associates and friends.

[29] E. Gherzi, "Radar work on the tropopause and the ozone layer," *Bull. Amer. Met. Soc.*, vol. 28, pp. 421–422, November, 1947.

Rectifier Networks for Multiposition Switching*

D. R. BROWN†, MEMBER, IRE, AND N. ROCHESTER‡, MEMBER, IRE

Summary—**A multiposition switch utilizing crystal rectifiers is used in electronic digital computers requiring switch times of less than 1 microsecond and in other applications requiring extreme compactness. Several varieties of this switch are analyzed, and generalized equivalent circuits are derived. Applications and practical limitations are discussed for both low-speed and high-speed operation.**

I. INTRODUCTION

A NEW TYPE of multiposition switch has been devised in order to achieve switching speeds of less than 1 microsecond. The switch has been used in an electronic digital computer to select the appropriate control or storage circuits in accordance with a binary code. It has also been used as a commutator to distribute signals successively from one source to each of a group of output terminals, or to collect signals from such a group of terminals. Because of its adaptability to high-speed operation in electron-tube circuits, this type of switch is extremely useful wherever rapid multiposition switching is required.

In order to describe the operation of this switch, let us consider the problem of translating the Murray code. The Murray code is used to represent alphabetical characters in teletype signals. Each alphabetical character is represented by a symbol which consists of five "yes" or "no" signals, and each signal is a positive or negative voltage. A rectifier network to translate the Murray code has five pairs of terminals of one class and thirty-two terminals of another class. If the code symbol representing the alphabetical character *A* is used to control the five pairs of terminals, a signal will appear on the *A* terminal of the group of thirty-two. Conversely, if a signal is applied to the *A* terminal of the group of thirty-two, the code symbol for *A* will appear at the five pairs of terminals. One of the things which has to be done in an electronic digital computer is to select the correct storage register (memory circuit) in response to the electrical code symbol which represents this register. The principal difference between this problem and the teletype problem is that there are usually more positions on the switch in the computer and faster action is needed.

The new multiposition switch contains a rectifier network which has n pairs of terminals of one class and 2^n terminals of another. These two sets of terminals are connected to each other by means of wires and rectifiers arranged in an intricate pattern. One set of terminals (the input) is actuated by switches, electron tubes, or relays, while the other set (the output) usually actuates some electron tubes or relays.

A circuit schematic of a four-position switch is shown in Fig. 1. The switch is set by means of coded electrical

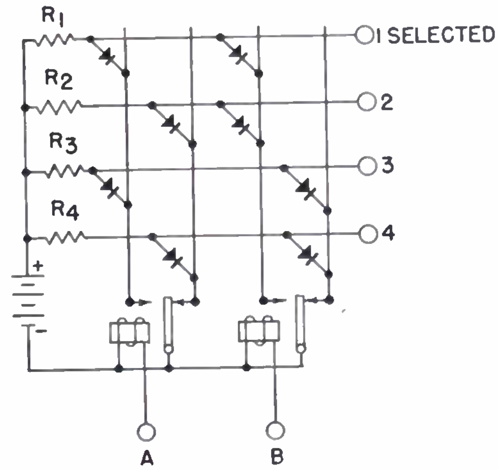

Fig. 1—A four-position switch.

pulses at terminals *A* and *B*, and the load circuits, not shown, are connected to terminals *1*, *2*, *3*, and *4*. The switch holds three of the four load circuits in an inoperative state by means of a bias voltage developed across R_2, R_3, and R_4. The load circuit connected to terminal *1* is normally operative. If, for example, the load circuit connected to terminal *3* is to be selected, a voltage must be applied to terminal *B* to operate the relay. Then current flows through resistors R_1, R_2, and

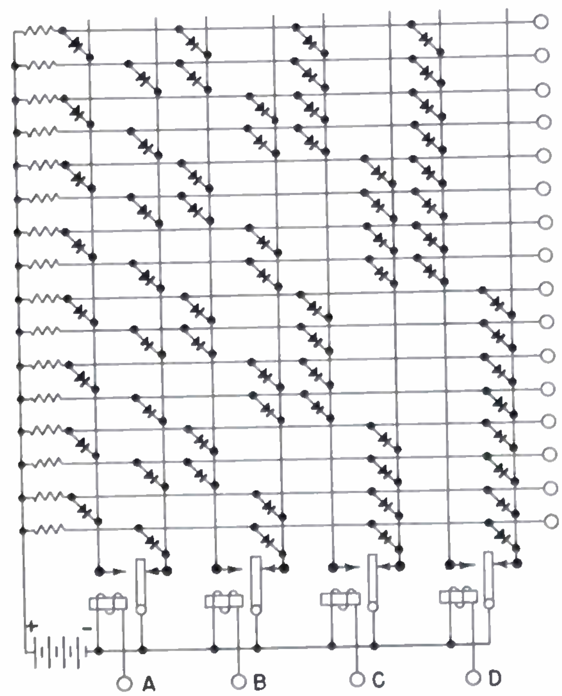

Fig. 2 A sixteen-position switch.

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‡ Formerly, Sylvania Electric Products Inc., Boston, Mass.; now, International Business Machines Corporation, Poughkeepsie, N. Y.

R_4, biasing the load circuits at terminals *1*, *2*, and *4* to an inoperative state. If a voltage had been applied to terminal *A* instead of terminal *B*, the load circuit at terminal *2* would have been selected. Application of a voltage to both terminals *A* and *B* selects the load circuit at terminal *4*. A circuit schematic of an eight-position switch is shown in Fig. 20. Note that three code terminals are required. Four code terminals are required for the sixteen-position switch shown in Fig. 2.

Each relay contacts one or the other of a pair of terminals of the rectifier network. The relays may be replaced by any element which selects one terminal or the other of a pair. In general, these "one-or-the-other" elements of the switch are oriented to select the desired load circuit by the presence or absence of a stimulus. The maximum number of load circuits is equal to the number of permutations of the orientations of the "one-or-the-other" elements. If there are *n* "one-or-the-other" elements, there may be as many as 2^n load circuits.

This type of switch does not require any special electron tubes, and, if barrier-type rectifiers are used, it does not require a large number of electron tubes. Very little voltage attenuation is suffered in the rectifier network. With the barrier-type rectifiers, the switch can be made very compact and lightweight, requiring only a small amount of power. A switch with many positions requires a few more than two crystal rectifiers for each of the 2^n terminals.

An eight-position switch has been designed and tested for use in a high-speed electronic digital computer. The switching time is less than 1 microsecond. This high speed is achieved by the use of resistance-coupled multivibrators for the "one-or-the-other" elements and type-1N34 germanium-crystal rectifiers in the rectifier network. A photograph of the switch is shown in Fig. 3.

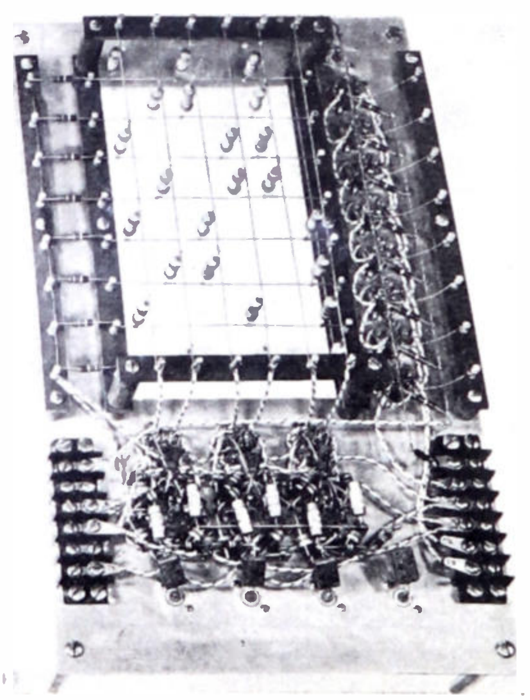

Fig. 3—Photograph of the eight-position switch.

II. Kinds of Rectifier Networks

The rectifier networks which have been shown in Figs. 1 to 3 and Fig. 20 are all termed "rectangular" networks. The name arises from the fact that the conventional method of drawing them uses a rectangular matrix of wires with rectifiers in particular places. There are other networks which accomplish nearly the same result, and some of these have important advantages. For example, if a 256-position switch were designed in the rectangular form, 2,048 rectifiers would be needed. By means of a more advanced design technique, a network of rectifiers may be designed which will produce the same result and require only 608 rectifiers. Figs. 4, 5, and 6 illustrate three of the five possible sixteen-position rectifier networks. Each of these different sixteen-position networks accomplishes the same thing that any other would. In fact, if the rectifiers were perfect

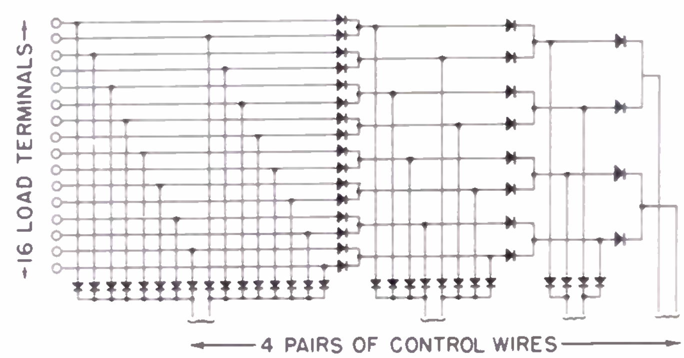

Fig. 4—Sixteen position pyramid network, 56 crystals.

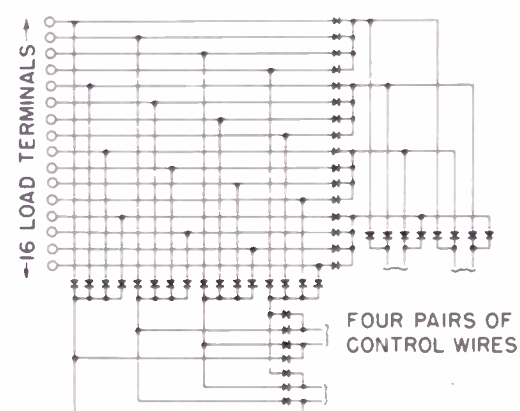

Fig. 5—Most economical sixteen-position network, 48 crystals.

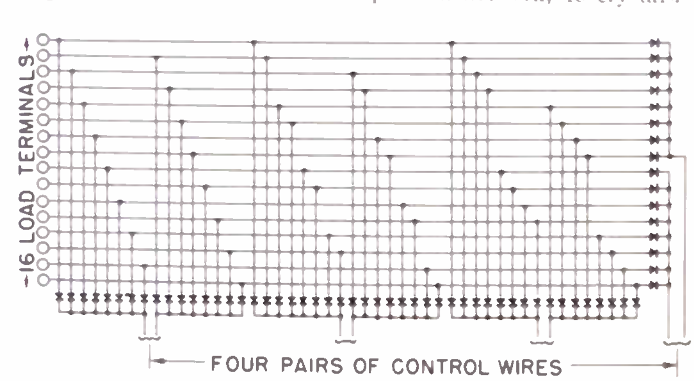

Fig. 6—Sixteen-position rectangular network, 64 crystals.

(i.e., if they had zero forward resistance and infinite back resistance) and if the network were sealed in a box with only sixteen-plus-eight terminals available, no external measurements could be made to indicate which of the five possible circuits was in the box. Note that Fig. 6 is the same circuit as Fig. 2 but redrawn to be comparable to Figs. 4 and 5. The thirty-two-position switch has twelve varieties. As the number of positions increases, the number of possible circuits increases very rapidly.

Appendix I gives a rigorous definition of the type of network considered here. Only two kinds of networks will be considered: the rectangular network, and that network which uses the fewest rectifiers (the most economical).

In order to proceed, some notation must be defined which will stress the important characteristics of the networks without any obscuring detail.

An examination of a few circuits will show what the shorthand must illustrate. Notice that in Fig. 4 two rectifiers are tied directly to each of the sixteen load terminals. In other words, proceeding into the network from the sixteen terminals, two sets of sixteen rectifiers are at once encountered. The information from one pair of control wires comes in through the bottom set, while the information from three pairs of control wires comes in from the right. The sixteen-position switch of Fig. 5, however, is different. Coming in from the sixteen terminals, two sets of sixteen rectifiers are again encountered, but the information from two pairs of control wires comes in through each set. In Fig. 6, the rectangular-network sixteen-position switch, four sets of sixteen rectifiers are encountered, and the information from each pair of control wires comes in through each set.

It would be helpful if some conventional formulas could be written to represent the situation, but this does not seem to be possible. The method of notation which has been devised employs triangles, which represent the terminals from which the selection is to be made; squares, which represent groups of rectifiers inside the network; and circles, which represent groups of rectifiers which are attached to the pairs of control terminals. For one rectifier network, one triangle is drawn. The number of terminals (2^n) is written in the triangle. The number of rectifiers in each group represented by a square or circle is written in the square or circle. The meaning and value of the distinction between squares and circles will become apparent after the rules for drawing the diagrams have been discussed. Arrows interconnecting the triangles, squares, and circles represent paths of information. One arrow represents the path of the information from one pair of control terminals.

The shorthand diagrams can be drawn by observing the following set of rules:

1. Draw a triangle, and in the triangle write the number of load terminals.
2. From the triangle, draw n arrows (the number of load terminals is 2^n).
3. Every arrow must terminate on a box or a circle.
4. If one arrow goes to an enclosure, the enclosure is a circle; no arrows leave a circle.
5. If more than one arrow goes to an enclosure, the enclosure is a box, and the number of arrows leaving must equal the number of arrows entering.
6. The number written in a box or a circle is 2^p where p is the number of arrows leaving the previous enclosure.
7. The total number of rectifiers needed to form the network is the sum of the numbers in the boxes and circles.
8. Each arrow represents part of the path of the information from one pair of control wires.

All possible diagrams drawn by the above rules represent all possible fundamental networks of the sort considered; the proof is given in Appendix II. Figs. 7, 8, and 9 show the shorthand notation for the networks shown in Figs. 4, 5, and 6, respectively. Fig. 10 shows

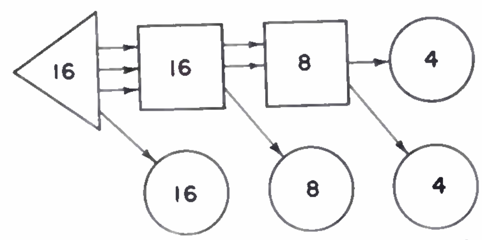

Fig. 7—Sixteen-position pyramid network (shorthand notation).

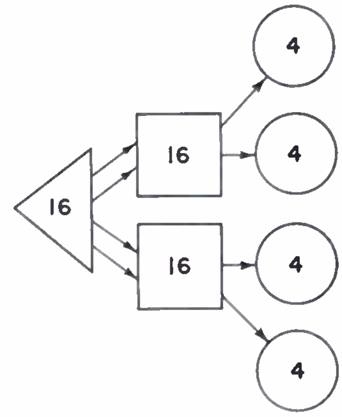

Fig. 8—Most economical sixteen-position network (shorthand notation).

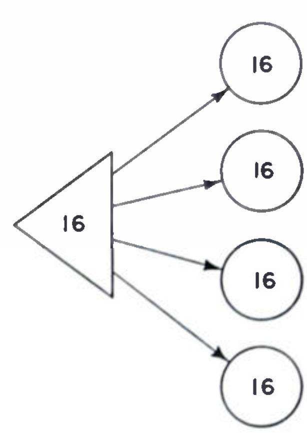

Fig. 9—Sixteen-position rectangular network (shorthand notation).

the shorthand notation for the 256-position network which has been mentioned. In order to design the most economical network, the following rules must be followed:

1. When the number of arrows leaving a box or a triangle is greater than three, the arrows go to only two boxes.

2. The arrows leaving a box or a triangle to go to boxes must be divided as evenly as possible between the two boxes toward which the arrows are pointed. If the number of arrows leaving a box or a triangle is three, both possible circuits involve the same number of rectifiers, and so the choice does not matter.

The proof that these rules lead to the most economical network is given in Appendix III.

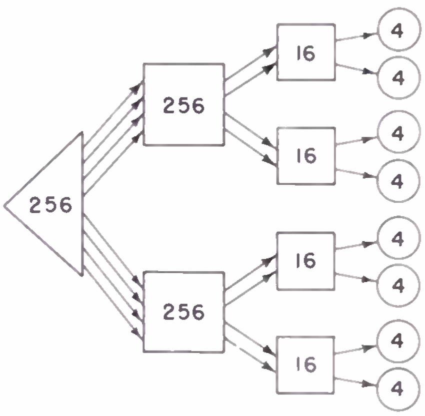

Fig. 10—Most economical 256-position network (shorthand notation).

III. Equivalent Circuits

In the previous discussion, the problem has been considered on the basis of the assumption of perfect rectifiers. These, of course, do not exist, and the discrepancies from perfection can have a profound influence on the operation of the switch.

It is easy to draw the equivalent circuit for a rectangular rectifier network. A consideration of the polarities involved will show which rectifiers are going to conduct and which are not. Fig. 11 shows an eight-position rectifier network connected into a typical circuit; Fig. 12 shows which rectifiers exhibit a low resistance, R_f, and which exhibit a high back resistance, R_b. The nature of these resistances may be understood from an examination of the resistance versus current characteristic of the rectifiers, shown for a 1N38 in Fig. 13.

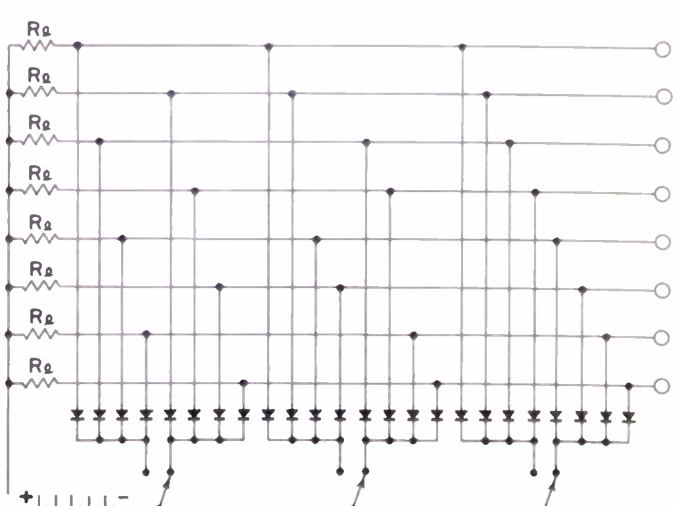

Fig. 11—Eight-position network controlled by single-pole switches.

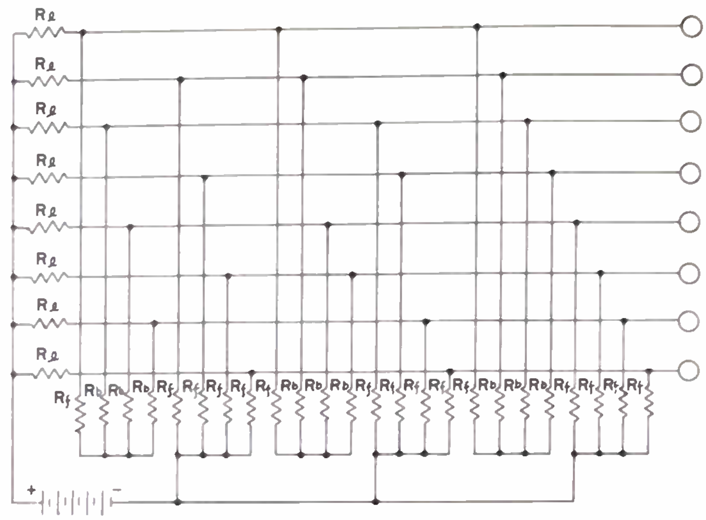

Fig. 12—Equivalent circuit for Fig. 11.

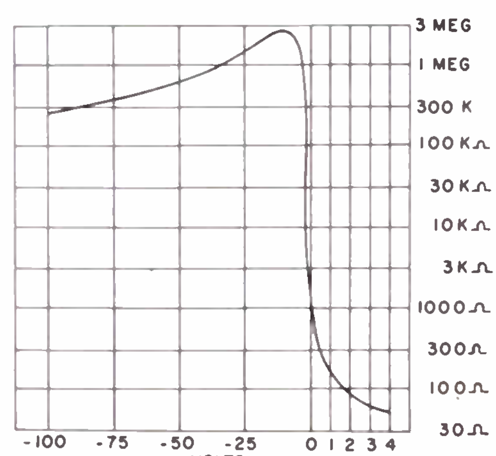

Fig. 13—Resistance of a 1N38 crystal as a function of voltage.

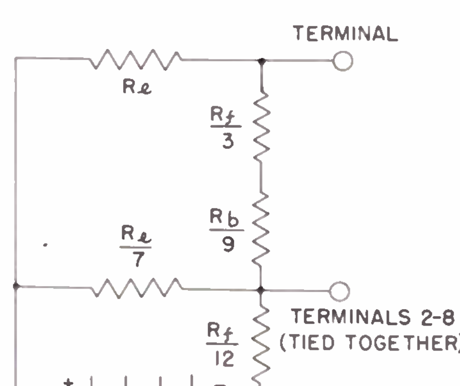

Fig. 14—Effective equivalent circuit for an eight-channel network with single-pole switches.

The values for R_b will ordinarily be about the same for all rectifiers which exhibit the high back resistance, but if the currents are not large enough there may be considerable variation in R_f. Assuming that R_f is uniformly low, and assuming that the ratio of front to back resistance is large, we can assume that terminals *2* to *8* inclusive are at the same voltage, and that terminal *1*, the selected one, is at another. The equivalent circuit is then shown in Fig. 14. Generalizing this to a 2^n terminal network is straightforward; the result is shown in Fig. 15. Networks with a very large number of

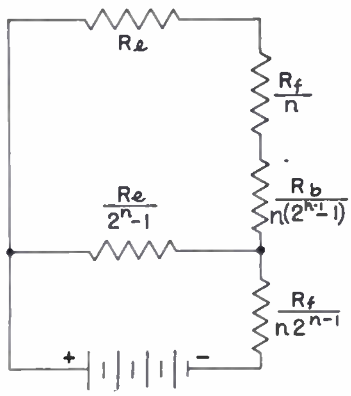

Fig. 15—Effective equivalent circuit for the general rectangular network with single-pole switches: R_l=load resistance, R_f=resistance of a conducting rectifier, R_b=resistance of a nonconducting rectifier.

terminals do not produce a very large voltage difference between the selected and nonselected terminals because

$$\frac{R_f}{n} + \frac{R_b}{n(2^{n-1} - 1)}$$

becomes so small. This situation may be greatly improved by using the network a little differently. The circuit is shown in Fig. 16, the new equivalent circuit in Fig. 17. Notice that as n gets large the performance of the network of Fig. 16 does not deteriorate as it does in Fig. 15. Except for the fact that the number of crystals

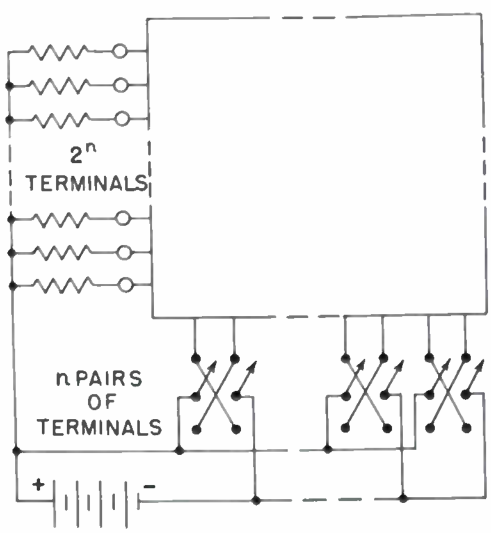

Fig. 16—$(2n+2^n)$ terminal network operated by double-pole-double-throw switches.

gets so large that the expense may get out of hand, there is no difficulty in making very large switches.

The most economical networks function better than the rectangular networks when used as in Fig. 11, but the rectangular networks function better when used in the circuit of Fig. 16. A generalized technique for deriving equivalent circuits for the most economical networks has been worked out, but will not be presented here because of its complexity.

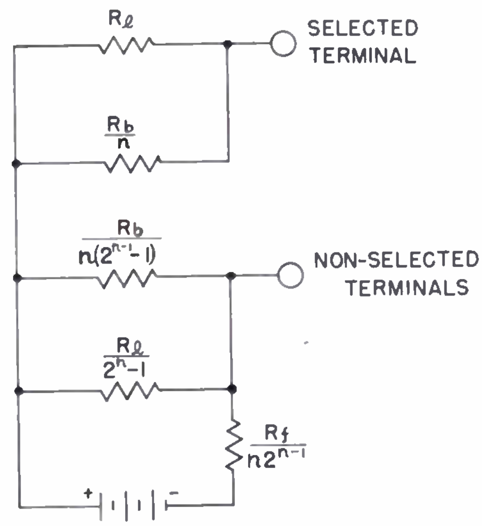

Fig. 17—Effective equivalent circuit for a general rectangular network with double-pole reversing switches.

IV. Applications

Several applications for these rectifier-network switches have arisen. In the principal computer application, an "order" which consists of the code is sent to the switch, and the switch then opens the proper electron-

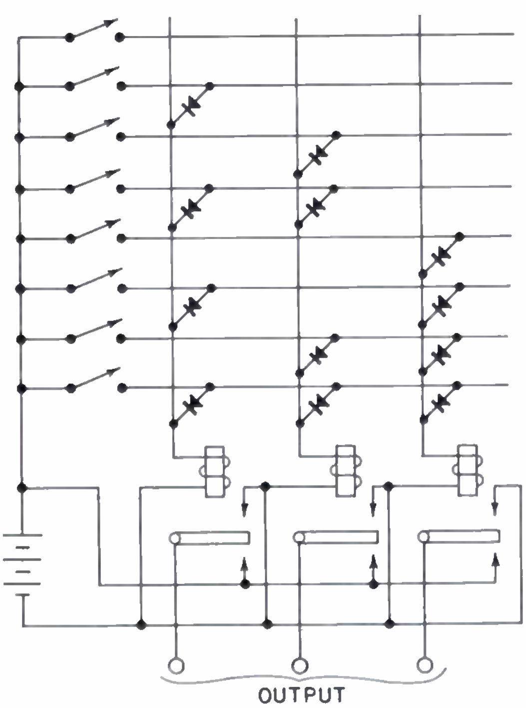

Fig. 18—Degenerate rectangular eight-position switch used to generate the binary code from a keyboard.

tube coincidence circuit or "gate." The network is really translating the binary code. Such a translator can, of course, work in reverse: a keyboard can be made with 2^n keys and connected in the circuit shown in Fig. 18. When any push button is depressed, the corresponding binary code is set up on the relays. Note that a degenerate network with half of the rectifiers omitted will serve for this purpose.

A commutator may be obtained if the multivibrators of a binary counter are used as the "one-or-the-other" elements. This will cause the various positions to be selected one after the other, as the counter counts. A commutator of this sort could be used to send pulses down various channels in a particular time sequence.

These networks can also be used to distribute or collect signals. An example of a collector is shown in Fig. 19. The signal from the selected source appears on the

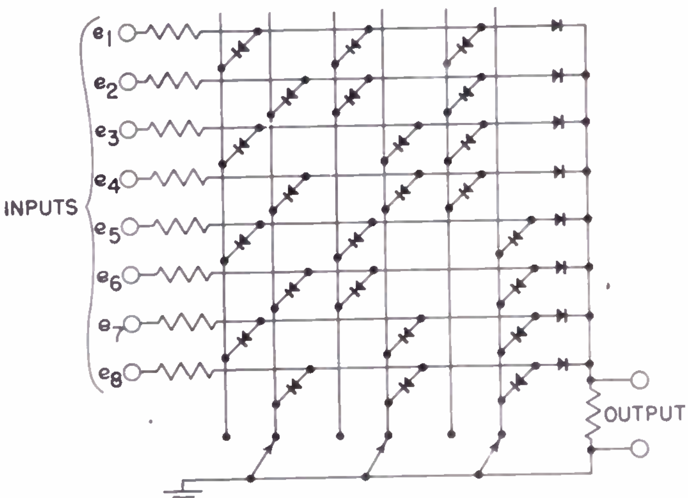

Fig. 19—Rectangular eight-position switch used to select a signal and transmit it.

output, while the signals from the nonselected sources are not transmitted. Such a network can, of course, be used in reverse to distribute a signal from one signal generator to the selected one of a group of terminals.

V. Switching Transients

Multiposition switches are required in the high-speed electronic digital computer being designed at the Massachusetts Institute of Technology. Rectifier-network switches for this computer having switching times of less than 1 microsecond have been designed, constructed, and tested. One of these, an eight-position rectangular-network switch, is shown in Figs. 3 and 20. Resistance-coupled multivibrators are used as the "one-or-the-other" elements, and isolating buffer amplifiers are placed between the multivibrators and the rectifier network. A twin triode, type 2C51, is used for each multivibrator and for each pair of buffer amplifiers; germanium-crystal rectifiers, type 1N34, are used in the rectifier network. The switch is set by a code which consists of short voltage pulses applied simultaneously to the three code terminals.

In less than 1 microsecond the desired load circuit, in this case a type 6AS6 gate tube, must be made operative, and all the gate tubes connected to other load terminals must be inoperative. The gate tube is made operative by driving the suppressor grid from below cutoff to zero or to a slightly positive voltage. A bias of 20 volts is required to change the gate tube from the inoperative to the operative state, or vice versa.

In order to study the switching transients, one or more of the code terminals was simultaneously pulsed by 0.25-microsecond negative pulses, and the reset terminal was pulsed 1 microsecond later. The process was re-

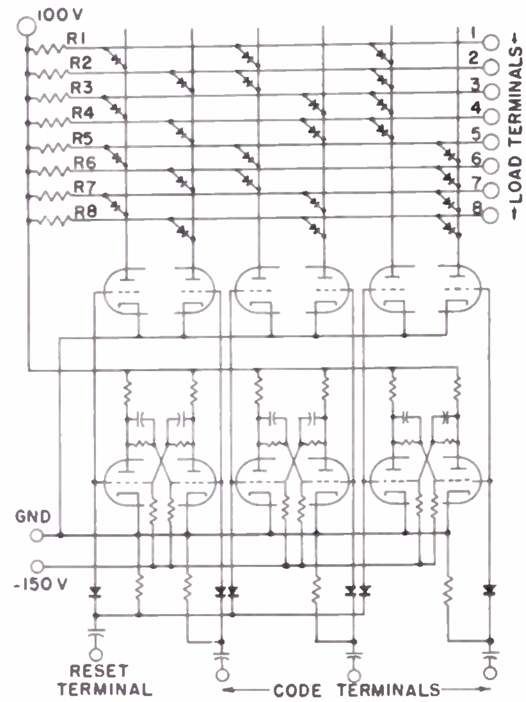

Fig. 20—Circuit schematic of eight-position switch.

peated at a 1,000-cps rate, so that the switching transients could be observed on a synchroscope. The voltage at the selected terminal increased 25 volts during the 1-microsecond period and returned to the normal nonselected voltage at the end of the period. Voltage at terminal *1*, which is the terminal selected when the switch is reset, decreased 25 volts during the 1-microsecond period and returned to its normal voltage at the end of the period. The voltage at the nonselected terminals did not change except for a brief transient pulse at the beginning and end of the 1-microsecond period.

The switching transients observed were different depending on how many multivibrators were switched by the code. For the case where all three multivibrators are switched, the wave form at terminal *1* is shown in Fig. 21(a), at a nonselected terminal in Fig. 21(b), and at the selected terminal (terminal *8*) in Fig. 21(c). The wave form at only one of the nonselected terminals is shown because all the wave forms at the nonselected terminals are the same when a given number of multivibrators are switched. The voltage scale indicated in Fig. 21 is, with respect to the supply voltage, 100 volts. These wave forms may be explained by a simple analysis. Since the longest switching time and largest transient pulses on the nonselected terminals occur when all three multivibrators are switched, this case alone will be considered. Other switching transitions may be analyzed in a similar manner.

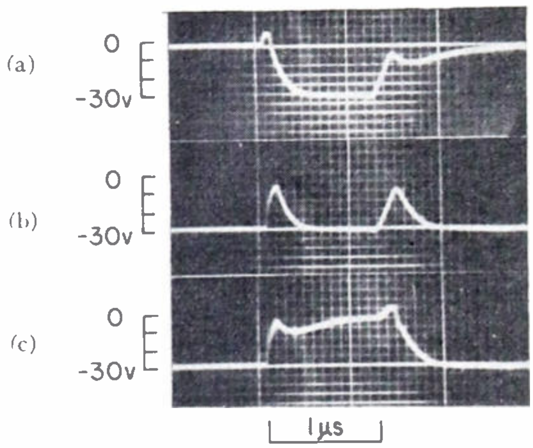

Fig. 21—Wave forms of the eight-position switch.

The nonconducting buffer amplifiers are biased well below cutoff. When the multivibrators switch, the conducting buffer amplifiers cut off before the other buffer amplifiers start to conduct because of the finite rise times of the pulses from the grids of the multivibrators. During the time that all the buffer amplifiers are cut off, the voltage of the entire network approaches the supply voltage as the shunt capacitances of the network are charged. Since the shunt capacitances are charged through resistors R_2 to R_8 in parallel, the voltage of all terminals increases very rapidly except terminal *1*, which is already at the supply voltage. When the other buffer amplifiers begin to conduct, they draw plate current through R_1 to R_7 and force the voltage at the nonselected terminals, including terminal *1*, back to the normal nonselected voltage. This accounts for the transient pulses at the nonselected terminals and the rapid rise time of the wave form at the selected terminal, terminal *8*.

Note that during the time all the buffer amplifiers are cut off, terminals *2* to *8* rise together, and after the buffer amplifiers begin to conduct, terminals *1* to *7* fall together. Soon after the buffer amplifiers begin to conduct, at the time the transient pulse at the nonselected terminal reaches its peak, the polarity of the voltage across some of the crystals changes and the selected terminal is isolated from the rest of the network. The voltage rises to its final value slowly as the shunt capacitance associated with the selected terminal is charged through the single resistor, R_8. Thus the long switching time is observed at the terminal being selected. At the end of the 1-microsecond period when the switch is reset, terminal *1* is the terminal being selected and has the same wave form as was observed at terminal *8* when it was selected. Stray coupling capacitance accounts for the small positive pulse at terminal *1* at the beginning of the 1-microsecond period and for the dip in the wave form at terminal *8* which occurs just after the initial rapid rise.

VI. Acknowledgment

Some of the work described here was carried out at the Servomechanisms Laboratory, Massachusetts Institute of Technology, under contract N5ori60 with the Special Devices Center of the Office of Naval Research; some was carried out at Sylvania Electric Products Inc., under contract W44-114-sc-88 with the Department of the Army; and some was done independently at Sylvania. The authors wish to express their thanks to Gordon S. Brown, director of the Servomechanisms Laboratory, and to J. W. Forrester, associate director, who has provided much helpful criticism. The authors are also indebted to C. H. Smith of the Naval Research Laboratory and H. C. Barlow of the Department of the Army for stimulating discussions and to J. N. Ulman for help in preparation of the paper.

Appendix I

Definition of the Rectifier-Network Switch

The switch is a network with (2^n+2n) terminals; 2^n of these terminals are of one class, and n pairs of terminals are of another. If the network is connected in the circuit of Fig. 11, current will flow in all but one of the external resistors. For each permutation of the positions of the single-pole–double-throw switches, a particular and different resistor will have no current flowing through it.

Appendix II

Validity of the Shorthand Notation

The nature of the information introduced via the two-position switches is such that every possible permutation of switch positions must be used in order to be able to select any one of the 2^n positions. Therefore, every one of the n pairs of terminals must be used.

Any circuit which does not connect two or more channels does nothing useful. Since this is so, we must connect, via rectifiers, at least two sources of information to each output wire. We may attach up to n sources of information to each output wire, but we cannot connect fewer than two or more than n in the most economical circuit.

An example of connecting two sources of information to the output wire is shown in Fig. 4. One source of information selects one of two groups of output wires, and the other selects one of eight other groups. Consequently, the result is a selection of one of $2\times8=16$ wires.

An example of connecting n sources of information to the output wires is shown in Fig. 6. Each of four sources of information selects two groups of wires, and the consequence is a selection of one of $2\times 2\times 2\times 2=16$ wires.

Each choice of whether or not a wire is to be selected, on the basis of one source of information, can be made by means of one rectifier, and one rectifier is necessary. Consequently, all possible economical circuits can be found by drawing, each time the problem of selecting between 2^n choices appears, the various circuits which have between two and n selections on each wire.

Now return to the shorthand described in Section II. A triangle represents the wires from which the selection is to be made. The number of enclosures in the column to the right of the triangle represents the number of selections to be made on each wire. If the number of enclosures is less than n, more combining circuits have to be added in order to make the selection definite. The same principles apply to the subselections which are made.

In the shorthand, the convention of using a circle when no further selection is needed and a square when further selection is needed clarifies the diagram. The arrows represent how many selections finally have to be made on each channel.

Appendix III[1]

Proof of the Rules for Deriving the Most Economical Network

Let $N(n)$ be the number of rectifiers needed to make a network which selects one from 2^n positions. Let $E(n)$ be the number of rectifiers needed to make a network by following the rules we have given for designing the most economical network. Demonstrate that

$$N(n) \geqq E(n). \tag{1}$$

Only one four position network can be drawn, and the two possible eight position networks both require the same number of rectifiers, as can be seen if all possible four-position and eight-position networks are drawn. In this paper, we have chosen to use the rectangular eight-channel network because of its symmetry.

In Table I, some values of $E(n)$ and certain other convenient quantities which will be needed in this theory have been tabulated.

Notice that, in Table I, $E(n)$ seems to approach 2^{n+1} as n gets large, but that

$$E(n) > 2^{n+1} \quad \text{when} \quad n > 2. \tag{2}$$

Inequality (2) must always apply because 2^{n+1} rectifiers are needed to attach two rectifiers to each channel, and because attaching at least two rectifiers is necessary

[1] After this paper had been prepared, it came to the attention of the authors that another proof of this same proposition had been worked out independently by C. H. Page of the Bureau of Standards. Dr. Page's proof has been extended to the case of nonbinary codes.

TABLE I

n	$E(n)$	$3\cdot 2^n$	2^n	2^{n+1}
2	8	12	4	8
3	24	24	8	16
4	48	48	16	32
5	96	96	32	64
6	176	192	64	128
7	328	384	128	256
8	608	768	256	512
9	1,168	1,536	512	1,024
10	2,240	3,072	1,024	2,048
11	4,368	6,144	2,048	4,096
12	8,544	12,288	4,096	8,192

but not adequate. It can be shown by mathematical induction that

$$E(n) \leqq 3\cdot 2^n. \tag{3}$$

As can be seen from Table I, (3) has been demonstrated experimentally when $n \leqq 12$. It will be shown that, if

$$E(p) \leqq 3\cdot 2^p \tag{4}$$

for every value of p which is less than n, then

$$E(n) \leqq 3\cdot 2^n. \tag{5}$$

This, of course, will prove (3).

If n is even,

$$E(n) = 2\cdot 2^n + 2E\left(\frac{n}{2}\right). \tag{6}$$

and so

$$E(n) \leqq 2\cdot 2^n + 2\cdot 3\cdot 2^{n/2}. \tag{7}$$

Therefore, (5) is proved if n is an even number larger than 4. If n is odd, the same process shows that (5) is proved if n is an odd number larger than 5.

In the shorthand notation, if the arrows from a box or triangle go to three or more boxes, then

$$N(n) \geqq 3\cdot 2^n. \tag{8}$$

Since this contradicts (3), the arrows which leave a box or a triangle must go to two boxes.

Next prove that the division of arrows between these boxes must be as even as possible. This must be done in two stages. The first stage will prove that the division must be within one arrow of being as even as possible, and in the second stage a division which is as even as possible will be compared with one which differs by one arrow.

First, consider the case where n is an even number. If an even division is made,

$$E(n) = 2\cdot 2^n + 2E\left(\frac{n}{2}\right) \leqq 2\cdot 2^n + 2\cdot 3\cdot 2^{n/2}. \tag{9}$$

If an uneven split is made

$$N(n) = 2\cdot 2^n + E\left(\frac{n}{2}+m\right) + E\left(\frac{n}{2}-m\right)$$

$$> 2\cdot 2^n + 2\cdot 2^{n/2+m} + 2\cdot 2 \text{[illegible]} \tag{10}$$

Equation (10) becomes larger than (9) if $m \geqq 2$ because

$$2 \cdot 2^{n/2+2} > 2 \cdot 3 \cdot 2^{n/2}. \qquad (11)$$

Next, consider the case where n is an odd number. If the most nearly even division is made,

$$E(n) = 2 \cdot 2^n + E\left(\frac{n+1}{2}\right) + E\left(\frac{n-1}{2}\right)$$

$$\leqq 2 \cdot 2^n + 2 \cdot 3 \cdot 2^{(n+1)/2}. \qquad (12)$$

If a less even division is made,

$$N(n) = 2 \cdot 2^n + E\left(\frac{n+1}{2} + m\right) + E\left(\frac{n-1}{2} - m\right)$$

$$> 2 \cdot 2^n + 2 \cdot 2^{(n+1)/2+m}. \qquad (13)$$

Equation (13) is larger than (12) when $m \geqq 2$ because

$$2 \cdot 2^{(n+1)/2+2} > 2 \cdot 3 \cdot 2^{(n+1)/2}. \qquad (14)$$

In order to proceed with the proof and demonstrate that $(m=0)$ is better than $(m=1)$, rather delicate tests must be applied. Consider four cases:

Let

$$n = 4p. \qquad (15)$$

For the even division write

$$E(4p) = 2 \cdot 2^{4p} + 2 \cdot 2 \cdot 2^{2p} + 4E(p); \qquad (16)$$

while the odd division gives

$$N(4p) = 2 \cdot 2^{4p} + E(2p+1) + E(2p-1) \qquad (17)$$

$$N(4p) = 2 \cdot 2^{4p} + 2 \cdot 2^{2p+1} + 2 \cdot 2^{2p-1} + E(p+1) + 2E(p) + E(p-1); \qquad (18)$$

Writing the appropriate inequalities for (16) and (18),

$$E(4p) \leqq 2 \cdot 2^{4p} + 2 \cdot 2 \cdot 2^{2p} + 2E(p) + 2 \cdot 3 \cdot 2^p \qquad (19)$$

$$N(4p) > 2 \cdot 2^{4p} + 2 \cdot 2 \cdot 2^{2p} + 2E(p) + 2^{2p} + 2 \cdot 2 \cdot 2^p + 2 \cdot 2^{p-1}. \qquad (20)$$

It is then obvious that

$$E(4p) < N(4p) \quad \text{when} \quad p \geqq 1. \qquad (21)$$

The same procedure may be applied when $n=4p+1$, $4p+2$, and $4p+3$, but the details will not be presented since the method has been demonstrated.

Properties of Some Wide-Band Phase-Splitting Networks*

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Summary—**Passive networks that produce polyphase output from single-phase input over wide frequency bands have great practical utility, yet published analysis of their action is incomplete. After reviewing properties of simple branch circuits useful in such networks, an expression is derived for phase difference produced between branches as a function of frequency. This expression is given a remarkably simple form, from which over-all operating properties are evident by inspection, permitting direct circuit design from required performance. General performance and design curves are presented.**

Introduction

EVERY RADIO ENGINEER has at some time produced two voltages in known phase relation, probably by use of a simple *RC* phase splitter, and many have been annoyed by the sensitivity of such devices to frequency changes. Although twenty years have gone by since Zobel[1] described some phasing circuits with constant attenuation, the possibility of using such circuits to build up wide-band phase-splitting networks has remained very little known. A recent paper by Dome[2] has therefore performed a valuable service by calling general attention to the possibility of thus making passive networks which will develop a polyphase signal from a single-phase source having complex wave form or variable frequency. He has also pointed out the great utility of such devices for single-sideband modulation, frequency-shift keying, development of circular oscilloscope sweep, and like applications.

Because of their wide applicability, the properties of these circuits deserve a more complete analysis than Dome has published. Such analysis turns out to allow the performance attainable to be displayed in a convenient and general fashion, and in addition to permit direct determination of circuit parameters from required performance. The most elementary sort of analysis will provide all the information needed, so long as a tapped signal source having negligible internal impedance can be assumed available. The properties of some simple branch circuits that may be used will first be indicated as briefly as possible, to provide a foundation for the analysis of the complete phase-splitting network which follows. The general properties of the complete circuit are then discussed and presented graphically, and the results attainable in the special case of a 90° phase splitter are indicated.

Properties of Branch Circuits

The very simple series circuit of Fig. 1(a) exhibits all essential properties, but its output must not be loaded. This circuit, with only three elements, is fully deter-

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[1] O. J. Zobel, "Distortion correction in electrical circuits," *Bell Sys. Tech. Jour.*, vol. 7, pp. 438–534; July, 1928.

[2] R. B. Dome, "Wide band phase shift networks," *Electronics*, vol. 19, pp. 112–115; December, 1946.

mined by specifying its operating characteristics of resonant impedance R, resonant frequency f_1, and selectivity Q (reactance/resistance ratio at resonance). Output voltage e is evidently related to input voltage E by

$$e = -\tfrac{1}{2}E + \frac{R}{R + j\omega L + \dfrac{1}{j\omega C}} E, \qquad (1)$$

considering the circuit as merely a voltage divider.

Introducing the operating characteristics f_1 and Q into (1), the complex voltage-transfer coefficient e/E becomes in general

$$e/E = k\frac{1 + jQ(f_1/f - f/f_1)}{1 - jQ(f_1/f - f/f_1)}, \qquad (2)$$

with the real amplitude factor k independent of frequency and having just the value $\tfrac{1}{2}$ for the particular circuit in question. Now,

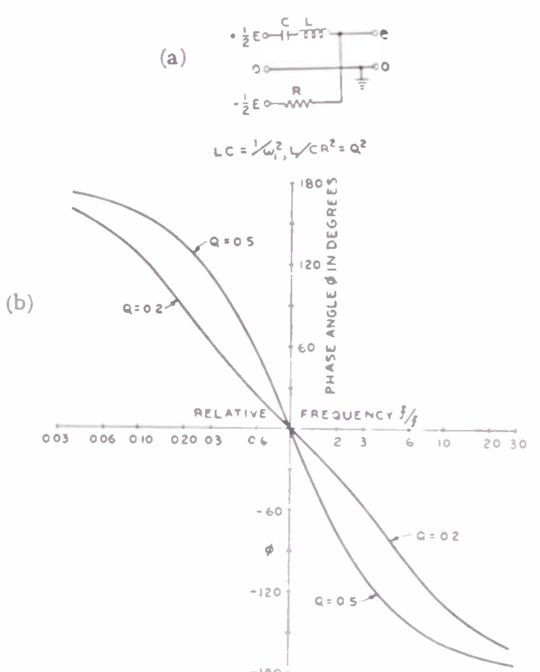


Fig. 1—Simple branch circuit and its phase characteristic.

$$Q(f_1/f - f/f_1) = \tan \tfrac{1}{2}\phi \qquad (3)$$

where $\tfrac{1}{2}\phi$ is the leading phase angle of the current in the series circuit. In terms of this angle, the voltage transfer is simply

$$e/E = k\epsilon^{j\phi}. \qquad (4)$$

The relative phase angle ϕ of the output voltage is plotted in Fig. 1(b) against f/f_1, on a logarithmic frequency scale, for two representative values of Q. At the higher value of Q shown, the curve has a simple S shape, with a single point of inflection. At the lower value of Q, the curve has a double-S shape, with three points of inflection. At an intermediate value of Q, there is a transition between these two limiting forms, and the corresponding curve is substantially straight over a considerable region. In any case, ϕ approaches $\pm 180°$ for frequencies far from resonance.

The properties of the circuit that are important to the present discussion are the constant magnitude k of input to output voltage-transfer ratio and the fairly linear region of the phase-frequency curve that is symmetrically located around its point of inflection at resonance. It is also important to remember that the shape of the phase curve is completely determined by specifying Q. On a logarithmic frequency scale, the resonant frequency f_1 merely locates the curve without altering its form, and resonant impedance R does not directly affect the transfer characteristics at all.

Several alternative circuits capable of giving the desired characteristics are shown in Fig. 2, with relations between values of circuit elements and f_1, Q, and k shown for each. Q, used here for its familiar significance, is just the reciprocal of the parameter s of Dome's notation. Since the resistance R is external to the resonant loop in Fig. 2(b), the Q in that case is the ratio of resistance to reactance of one element at resonance. Values of Q that are of present interest are less than $\tfrac{1}{2}$, so circuits using only one kind of reactance are permissible, if the penalty of sharply reduced output can be tolerated. The fourth element of Fig. 2(c) is not determined by f_1, Q, and over-all impedance required, and may be ad-

(a) $\omega_1^2 = 1/LC$; $Q^2 = L/CR^2$; $k = \tfrac{1}{2}$; $L = QR/\omega_1$; $C = 1/\omega_1 QR$

(b) $\omega_1^2 = 1/LC$; $Q^2 = CR^2/L$; $k = \tfrac{1}{2}$; $L = R/\omega_1 Q$; $C = Q/\omega_1 R$

(c) $\omega_1 = 1/R_1C_1 = 1/R_2C_2$; $Q = 1/[R_2/R_1 + 2]$; $k = 1 - 2Q$; $R_1 = R_2/[1/Q - 2]$; $C_2 = 1/\omega_1 R_2$; $C_1 = [1/Q - 2]/(\omega_1 R_2)$

(d) $\omega_1^2 = 1/L_1C_1 = 1/L_2C_2$; $Q^2 = L_1/C_1R^2 = C_2R^2/L_2$; $k = \tfrac{1}{2}$; $L_1 = QR/\omega_1$; $C_1 = 1/\omega_1 QR$; $L_2 = R/\omega_1 Q$; $C_2 = Q/\omega_1 R$

(e) $\omega_1 = 1/R_1C_1 = 1/R_2C_2 = 1/R_3C_3$; $Q = 1/[R_2/R_1 - 2] = 1/[R_3/R_2 + 2]$; $k = (1 - 2Q)/(1 + 2Q)$; $R_2 = R_3/[1/Q - 2]$; $R_1 = R_3/[1/Q^2 - 4]$; $C_3 = 1/\omega_1 R_3$; $C_2 = [1/Q - 2]/(\omega_1 R_3)$; $C_1 = [1/Q^2 - 4]/(\omega_1 R_3)$

Fig. 2—Alternative branch circuits and their properties.

justed to give maximum k for a chosen Q, a result obtained by making the two RC products equal.

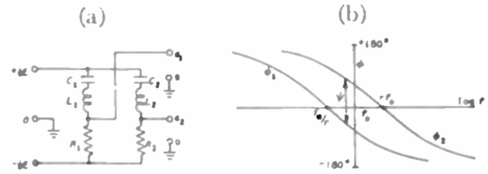
Fig. 2(d) shows one-half of the conventional symmetrical phase-compensating lattice, which is able to tolerate the resistive load $\frac{1}{2}R$ at the half-lattice output. In order to produce the desired transfer characteristics as given by (2) or (4), resonant frequencies and Q values must be made the same for the two branches shown. When these conditions are imposed, the circuit elements are fully determined and, in addition, the full lattice section exhibits a purely resistive input impedance R which is independent of frequency.

Dome[2] has also suggested the very useful circuit of Fig. 2(e), which not only avoids the use of inductances but will tolerate a load that includes shunt capacitance. This circuit is fully determined and made to follow (2) by imposing three conditions, which equalize the values of f_1 and Q for the two branches and maximize k. All circuits shown except that of Fig. 2(c) are half lattices fed in balanced fashion, and may be expanded by symmetry to the full lattice section, if a balanced output of doubled amplitude is desired.

Analysis of Complete Circuit

Any of the circuits of Fig. 2, all of which are characterized by attenuation independent of frequency and an output phase varying as in Fig. 1(b), can be used as elements of a phase splitter. If two such circuits of equal Q are tuned to different frequencies f_1 and f_2 and connected as in Fig. 3(a) to provide separate branch-output voltages e_1 and e_2, the phase relations of Fig. 3(b) will result. The mean phase of e_1 and e_2 relative to E will vary with frequency much as do the phases ϕ_1 and ϕ_2 of e_1 and e_2 separately. On the other hand, the phase difference ψ between e_1 and e_2 will take on a fairly uniform maximum value for frequencies near the mean of f_1 and f_2, where the individual phase curves are fairly linear, and will approach zero for frequencies far above f_2 or below f_1.

(a) (b)



$\omega_1^2 = \omega_0^2/r^2 = 1/L_1C_1$ $\omega_2^2 = \omega_0^2 r^2 = 1/L_2C_2$ $Q^2 = L_1/C_1R_1^2 = L_2/C_2R_2^2$

Fig. 3—Complete phase-splitter circuit and its branch-phase characteristics.

The behavior of the complete phase splitter is fully determined by the resonant frequencies f_1 and f_2, together with the common value of Q. Alternatively, the geometric mean f_0 between f_1 and f_2 and the ratio r^2 of f_2 to f_1 are convenient to use as basic parameters. In conjunction with any conditions necessary to provide the characteristic of (2), and to maximize k, the three parameters f_0, r, and Q also determine the values of all circuit elements except as regards over-all impedance level. The relation between branch-output phase difference ψ and frequency f may now be examined in detail as to its general form and its dependence on the parameters f_0, r, and Q.

Since f_1 is by definition f_0/r and f_2 is rf_0, the phase angles ϕ_1 and ϕ_2 of output voltages e_1 and e_2 relative to input voltage E are given respectively by

$$\left.\begin{aligned} \tan \tfrac{1}{2}\phi_1 &= Q\left(\frac{f_0}{rf} - \frac{rf}{f_0}\right) \\ \text{and} \qquad \tan \tfrac{1}{2}\phi_2 &= Q\left(\frac{rf_0}{f} - \frac{f}{rf_0}\right) \end{aligned}\right\}. \tag{5}$$

The phase lead ψ of e_2 over e_1 is $\phi_2 - \phi_1$, and use of (5) in the usual trigonometric expression for the tangent of the difference of two angles, gives directly

$$\tan \tfrac{1}{2}\psi = \frac{Q\left[\left(\frac{rf_0}{f} - \frac{f}{rf_0}\right) - \left(\frac{f_0}{rf} - \frac{rf}{f_0}\right)\right]}{1 + Q^2\left(\frac{rf_0}{f} - \frac{f}{rf_0}\right)\left(\frac{f_0}{rf} - \frac{rf}{f_0}\right)}. \tag{6}$$

Expanding this expression, factoring the numerator, rearranging the denominator to give perfect squares of factors in the numerator, and dividing through by Q^2 gives in turn

$$\tan \tfrac{1}{2}\psi = \frac{\frac{1}{Q}\left(r - \frac{1}{r}\right)(f/f_0 + f_0/f)}{\frac{1}{Q^2} - (r - 1/r)^2 - 4 + (f/f_0 + f_0/f)^2} \tag{7}$$

as a general expression for the frequency dependence of the phase split produced by any equal-Q pair of circuits of the type considered here.

Equation (7), with evident logarithmic symmetry about the central frequency f_0, is already quite simple and usable. Certain changes of variable render its form much simpler still, however, and in addition permit almost all numerical calculations of performance to be taken ready-made from trigonometric tables. Simplification is accomplished by use of the standard trigonometric identities

$$\left.\begin{aligned} \cot x + \tan x &= \frac{2}{\sin 2x} = 2 \operatorname{cosec} 2x \\ \text{and} \qquad \cot x - \tan x &= \frac{2}{\tan 2x} = 2 \cot 2x \end{aligned}\right\}. \tag{8}$$

Quite arbitrarily, let

$$f/f_0 = \tan \tfrac{1}{2}\eta. \tag{9}$$

$$Q\left(r - \frac{1}{r}\right) = \tan \tfrac{1}{4}\psi_0, \tag{10}$$

and

$$\frac{1}{Q^2} - \left(r - \frac{1}{r}\right)^2 = 4 \operatorname{cosec}^2 \tfrac{1}{2}\sigma \tag{11}$$

define a new independent variable η and new parameters ψ_0 and σ. Substituting these variables into (7) and using (8), there results

$$\tan \tfrac{1}{2}\psi = \tan \tfrac{1}{2}\psi_0 \frac{\operatorname{cosec}^2 \tfrac{1}{2}\sigma \operatorname{cosec} \eta}{\cot^2 \tfrac{1}{2}\sigma + \operatorname{cosec}^2 \eta}. \tag{12}$$

Now let

$$\tan \tfrac{1}{2}\sigma \operatorname{cosec} \eta = \tan \tfrac{1}{2}\theta \qquad (13)$$

define still another independent variable θ. In terms of this variable, (12) [and thereby (7)] finally takes the form

$$\frac{\tan \frac{1}{2}\psi}{\tan \frac{1}{2}\psi_0} = \frac{\sin \theta}{\sin \sigma}. \qquad (14)$$

This is the general phase-versus-frequency characteristic of the complete network. It holds for any values of f_0, r, and Q, but these have now been replaced by more convenient performance parameters ψ_0 and σ.

Properties of Phase Splitter

The physical significance of the above very compact result shows quite clearly upon examination of (13) and (14), bearing (9) also in mind, and determines the form of the curves to be given later. For an input-signal frequency equal to the network center frequency f_0, η is 90 degrees and cosecη is unity so, from (13), θ is equal to σ and, from equation (14), branch-phase difference ψ has just the value ψ_0. That is, ψ_0, determined by Q and r according to (10), turns out to be just the phase split between the two branch outputs for the center frequency f_0.

Since cosecη cannot be less than unity, (13) indicates that θ cannot be less than σ. If σ exceeds 90°, tan$\frac{1}{2}\sigma$ exceeds unity and θ remains above 90° for all frequencies. Equation (14) then indicates a single phase-difference maximum when θ, still above 90°, is a minimum. θ is a minimum when cosecη has its minimum of unity, at frequency f_0. The network with $\sigma > 90°$ thus exhibits a single maximum of phase difference, having the value ψ_0, at the center frequency f_0.

If, on the other hand, σ is less than 90°, tan$\frac{1}{2}\sigma$ is less than unity and θ will pass through 90° as cosec η takes the value cot$\frac{1}{2}\sigma$ (>1) in its variation with frequency. Since η varies from 0° to 180° as input frequency goes from 0 to ∞, cosecη will take this value twice and phase difference ψ will show in this case two maxima, of value ψ_m. These maxima will be symmetrically located with respect to f_0, since cosecη is symmetrical in log f/f_0. At f_0, θ will have its minimum value, less than 90° and equal to σ, and the phase difference ψ will now exhibit at f_0 a minimum of value ψ_0.

When there is a double maximum of phase difference, the relation

$$\operatorname{cosec} \sigma = \frac{\tan \frac{1}{2}\psi_m}{\tan \frac{1}{2}\psi_0} \qquad (15)$$

holds and, alternatively, the departure $\frac{1}{2}(\psi_m - \psi_0)$ of the extremes of phase difference from the mean $\frac{1}{2}(\psi_m + \psi_0)$ of those extremes is determined by σ as

$$\sin \tfrac{1}{2}(\psi_m - \psi_0) = \sin \tfrac{1}{2}(\psi_m + \psi_0) \tan^2 (45° - \tfrac{1}{2}\sigma). \qquad (16)$$

For values less than 90°, the physical meaning of σ is thus quite simple and clear.

Parameters f_0, ψ_0, and σ determine the relation between branch-output phase difference and signal frequency completely and in a very simple and direct way. Center frequency f_0 locates the characteristic on a logarithmic frequency scale, without affecting its form. Center-frequency phase difference ψ_0 locates the characteristic on a logarithmic-tangent phase scale without affecting its form. Form parameter σ controls the general shape of the characteristic and its spread in frequency, as well as further locating it on the phase scale. A single characteristic curve, generally useful for all values of f_0 and ψ_0, is therefore obtainable for each value of σ.

The parameters f_0, ψ_0, and σ also determine the circuit design quite directly. From (10) and (11) together,

$$Q = \tfrac{1}{2}\sqrt{1 - \tan^2 \tfrac{1}{4}\psi_0} \sin \tfrac{1}{2}\sigma. \qquad (17)$$

Letting

$$r = \cot \tfrac{1}{2}\rho, \qquad (18)$$

(8), (10), and (11) give

$$\tan \rho = \sqrt{\cot^2 \tfrac{1}{4}\psi_0 - 1} \sin \tfrac{1}{2}\sigma. \qquad (19)$$

Resonant frequencies f_1 and f_2 for the two branch networks are f_0/r and rf_0, respectively, and the values of all circuit elements may be determined from f_1, f_2, Q, and a choice of impedance level.

Fig. 4 shows the phase-frequency characteristics, as given by joint use of (9), (13), and (14), for various

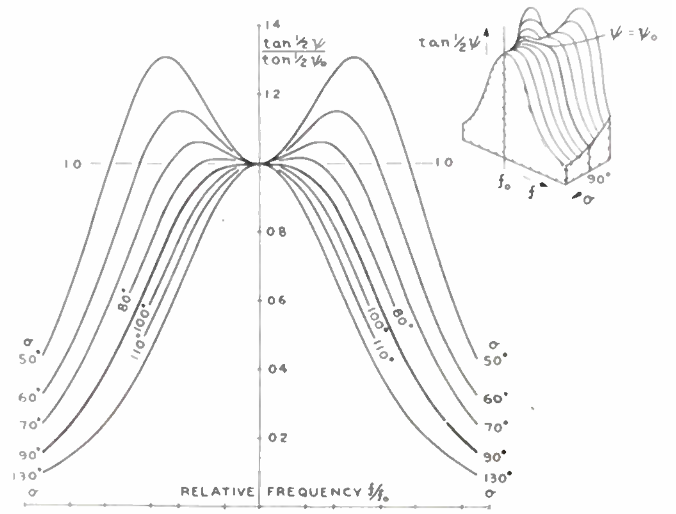


Fig. 4—General characteristics of phase splitter.

values of σ. Frequency is plotted logarithmically as f/f_0, and phase linearly as

$$\frac{\tan \frac{1}{2}\psi}{\tan \frac{1}{2}\psi_0}.$$

Curves for ψ directly would look much the same, but would change shape with changing choice of ψ_0; those shown are more general. The perspective sketch shows the over-all relationship of ψ to f and σ more clearly, though less quantitatively, than does its projection as a family of curves in the f–ψ plane. Values of σ exceeding 90° lead to rapidly varying phase difference and are not of great interest. For σ just 90°, there is a truly flat region just at f_0, where ψ_0 and ψ_m are the same. This is a useful condition for purposes which require accurately constant phase difference over a narrow frequency band, and corresponds to

$$Q = \tfrac{1}{2}\sqrt{\tfrac{1}{2}(1 - \tan^2 \tfrac{1}{4}\psi_0)}. \qquad (20)$$

For σ less than 90°, the characteristic becomes double-humped, with maxima equal to cosec σ and a minimum of unity on the relative half-angle tangent scale used. These maxima occur at frequencies for which $\sin\eta$ is just $\tan\frac{1}{2}\sigma$. Phase difference again falls to ψ_0 at the two frequencies, f_{min} and f_{max}, outside those at the phase maxima, for which $\sin\eta$ becomes $\tan^2\frac{1}{2}\sigma$. Values of less than 90° for σ lead to the conditions of most practical interest, for which phase difference remains close to a specified value over a considerable band of frequencies. When ψ_m and ψ_0 are specified, σ is given by (15) or (16) and the frequencies at which ψ_m and return to ψ_0 occur are fully determined, as are Q and r. Actual spread $\psi_m - \psi_0$ of the phase difference produced depends on the choice of both ψ_0 and σ, as do the circuit defining parameters r and Q. The ratio of maximum to minimum frequency of the working band over which ψ exceeds ψ_0 depends only on σ, however.

More detailed investigation shows that, for a given total spread of phase angle, a wider frequency band is obtained by choosing a value of σ which places ψ_m at the upper limit and ψ_0 at the lower limit of the permitted phase spread than by choosing σ to place ψ_m at the upper limit and ψ_0 at, for example, the center of the spread. If a given phase split is to be approximated over a given frequency range, requiring a definite value of σ, less phase spread is encountered when a single network section with a large value of ψ_0 is used than when several isolated or iterative sections with equal small values of ψ_0 are used in cascade. The present analysis has not been extended to such single-section networks, of more complicated structure than those shown in Fig. 2, as are necessary to give still better constancy of branch-output phase difference over the working frequency band. Extension to more than two branch outputs, considered two by two, is obvious.

Quadrature Case

The case of 90° phase difference, corresponding to production of symmetrical two-phase output, is of particular importance and will be examined further. Fig. 5 shows operating characteristics $\frac{1}{2}(\psi_m - \psi_0)$ and f_{max}/f_{min}, as well as design parameters Q and r, plotted against the form parameter σ. $\frac{1}{2}(\psi_m + \psi_0)$ is here held at 90° for σ less than 90°; ψ_0 is held at 90° for σ greater than 90°. $\frac{1}{2}(\psi_m - \psi_0)$ is the maximum departure of phase difference from 90° within the frequency range from f_{min} to f_{max}. The flat-topped phase characteristic for which $\psi_m = \psi_0 (\sigma = 90°)$ occurs in this case at a Q of 0.322, with r set at 1.835. Phase-difference characteristics with a single peak occur for all higher Q values, becoming increasingly sharp (σ larger) as Q increases. Double-

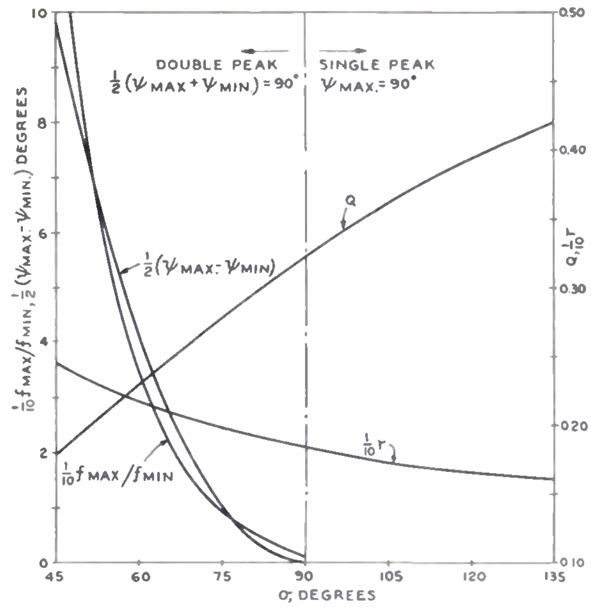


Fig. 5—Design characteristics of 90° phase splitter.

peaked characteristics occur only for Q values below 0.322, more pronouncedly as σ and Q decrease.

Fig. 6 summarizes the over-all performance of the 90° phase splitter with form parameter σ less than 90°, showing the relation between maximum error in phase difference produced in the working band and width of

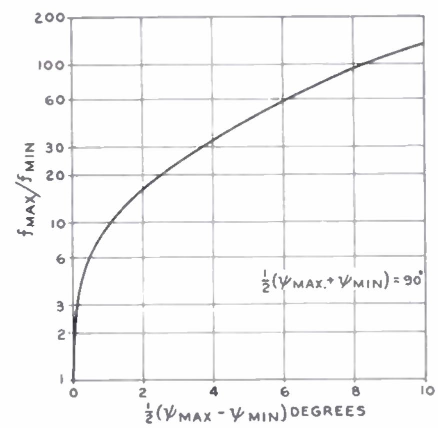


Fig. 6—Performance of 90° phase splitter.

working frequency band over which ψ remains above ψ_0. Operation over a frequency range of one full octave may in principle be had with branch-phase difference remaining within ±45 seconds of arc from 90°. The limiting frequencies are in a $9\frac{1}{2}$-to-1 ratio for operation within ±1° of 90°, while tolerance of ±5° variation of phase difference permits operation over a $43\frac{1}{2}$-to-1 band, which is as broad as the usual home-receiver audio band. Imperfections of actual circuit elements and their adjustments always degrade to some extent this ideal performance, however.

Evidently, these very simple circuits are capable ideally of very good performance, and their performance may be expressed in very convenient analytical form; this is the reason that the present study of their properties has been considered worth while.

Paralleled-Resonator Filters*

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Summary—This paper describes a class of microwave filters in which input and output waveguides are connected by a number of resonators, each coupled directly to both guides. Signal components of different frequencies can pass from the input to the output largely through different resonators. This type of filter is a realization of a lattice network. An experimental filter is described.

IN THE USUAL ladder-type microwave filter the input and output waveguides are interconnected by a series of resonators coupled in tandem, one to the next, so that a signal of any frequency must pass through all resonators in going from the input to the output.

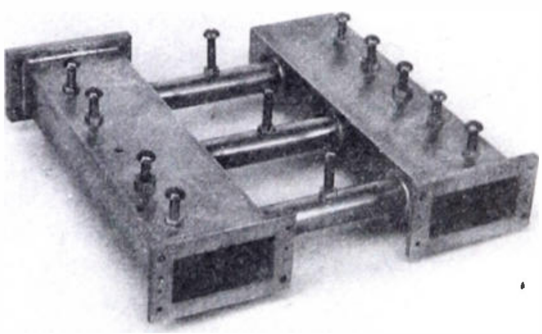

Fig. 1—An experimental paralleled-resonator filter, having three resonators.

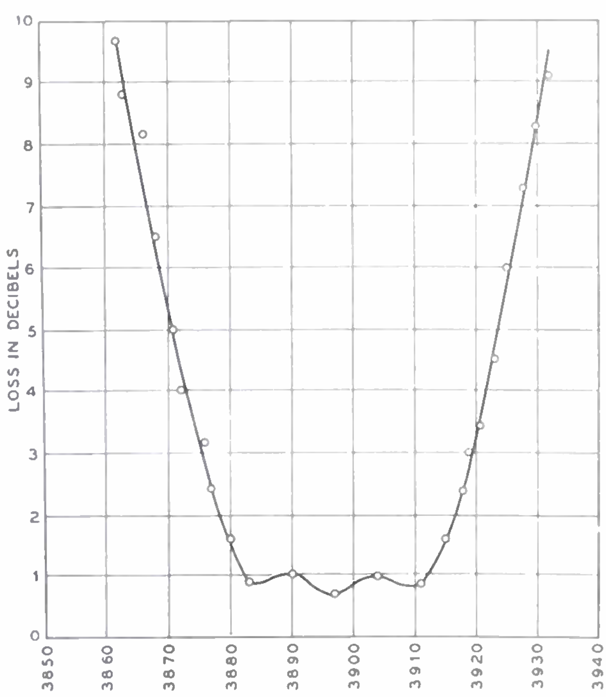

Fig. 2—Loss versus frequency for the filter in Fig. 1.

This paper describes a different class of microwave filters, which the writer has called *paralleled-resonator* filters. In such filters the input and output waveguides are connected by a number of resonators, each coupled directly to both input and output guides. Signal components of different frequencies can pass from the input to the output waveguides through different resonators.

Fig. 1 is a photograph of an experimental paralleled-resonator filter having three resonators. Fig. 2 shows loss versus frequency for the filter of Fig. 1. In the pass band there were two peaks of reflected power; at these, the reflected power was about 16 db below the incident power (a voltage-standing-wave ratio of 1.4). At three other points in the pass band the reflected power was very small. This filter was built and measured to illustrate the principles involved. A filter for a specific application might have a quite different construction.

In the filter of Fig. 1, the resonators are resonant transmission lines coupled to both input and output waveguides.[1] Fig. 3 is a cross-section view of the filter through one of the resonators. The transmission-line resonator has an inner conductor which passes through each waveguide perpendicular to the electric vector and to the direction of propagation, and through a tube between the waveguides. This line is tuned by means of a tuning screw T, which adds capacitance at the center of the line. The voltage distribution along the line is indicated below the cross-section view.

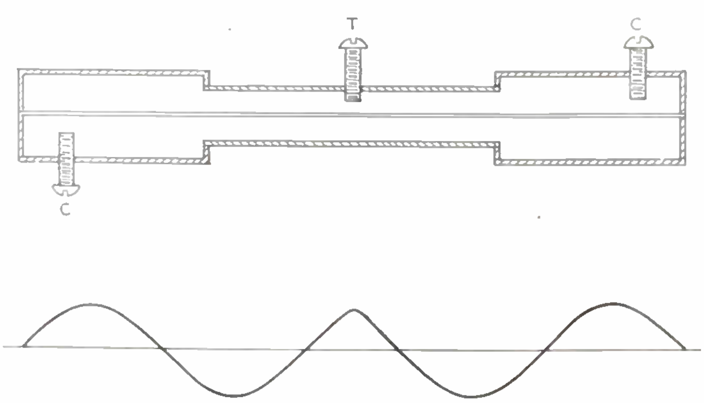

Fig. 3—Cross-section view of the filter in Fig. 1 through one of the resonators.

The line is coupled to the input and output waveguides only because two coupling screws C disturb the symmetry of the electric fields in the guides. The coupling of the line to a guide may be increased by making the coupling screw project further into the guide. The sense of the coupling may be reversed by transferring

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[1] This type of resonator and coupling is similar to that described by W. D. Lewis and L. C. Tillotson, "A constant resistance branching filter for microwaves," *Bell Sys. Tech. Jour.*, vol. 27, pp. 83–95; January, 1948.

the coupling screw from the top to the bottom of the guide.

In the filter of Fig. 1 there are three such resonators, spaced one-half wavelength apart and hence effectively in parallel. The resonators are an odd number of quarter wavelengths from the shorted ends of the input and output guides, and hence may be thought of as in parallel across the open-circuited ends of the input and output waveguides.

Three parameters of each resonator can be varied: the tuning (resonant frequency), and the coupling to each guide. Varying these parameters in a single resonator affects the filter characteristics for frequencies near the resonant frequency of that resonator only. Thus, the paralleled-resonator filter is easy to adjust. When a frequency-swept oscillator was used and the output of the filter was viewed on an oscilloscope, it was easy to obtain a flat band. The resonators could be adjusted when all were resonating, or all but one could be detuned and the active resonator could then be tuned and the coupling adjusted for appropriate broadness of response. The feature of ease of adjustment should be particularly valuable in filters with a large number of resonators.

With a lossy filter such as that of Fig. 1, it is a little difficult to obtain both a good transmission characteristic and a good input impedance unless both the reflected and the transmitted power are viewed simultaneously. This difficulty would not arise in the case of a low-loss filter.

The paralleled-resonator filter is more flexible than the ladder-type filter. The paralleled-resonator filter is a microwave realization of a lattice network, which provides the most general form of coupling between two pairs of terminals. Thus, for instance, the paralleled-resonator filter can be adjusted to approximate a linear-phase constant-amplitude characteristic in the band. Further, as the paralleled-resonator structure is equivalent to a lattice, it should be possible to apply known methods of lattice design in achieving desirable characteristics.

The properties of paralleled-resonator filters could be derived by the means usually employed in connection with lattices. A very simple analysis illustrating some of their properties is given here.

Fig. 4(a) shows a lattice network. It includes a perfect transformer N. In working with networks designed to give matched coupling between guides of equal impedance, we may let N be unity. The remaining parameters are the admittances $2Y^+$ and $2Y^-$. They provide what might be called *direct* and *contrary* coupling between input and output. If the output terminals are shorted and unit voltage is applied between input terminals, a current $Y^+ - Y^-$ will flow in the short.

Fig. 4(b) shows a network equivalent to the lattice if the admittances Y^+ and Y^- are purely imaginary. Here Y^- is represented by a number of series resonant circuits (L_1, C_1; L_2, C_2, etc.) coupled to input and output by transformers of unity ratio, and Y^+ is represented by similar circuits (L_a, C_a; L_b, C_b, etc.) with the coupling of one of the pair of transformers reversed.

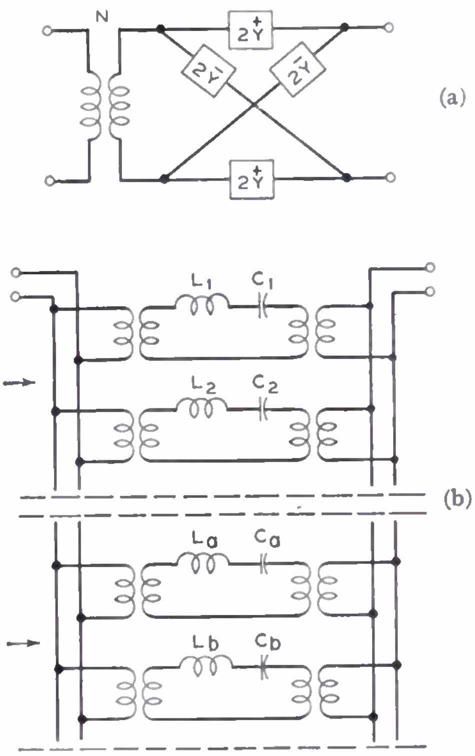

Fig. 4—Shunt representation. (a) Lattice network. (b) A network equivalent to the lattice of the admittances Y. Y are purely imaginary.

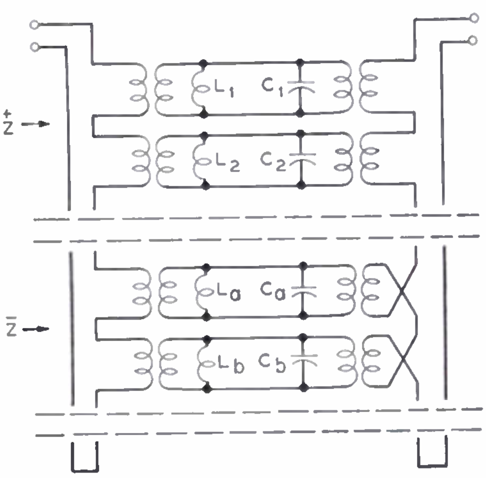

Fig. 5—Series representation in which two sets of terminals are coupled by perfect transformers connected in series to shunt resonant circuits.

Now, Fig. 4(b) is a good representation of two waveguides coupled by a number of resonators, each resonator coupled equally strongly to both guides, but some transferring power in one phase and some in the opposite phase.[2]

[2] In using the terms "direct" and "contrary" coupling, we should note that which is called direct and which contrary is arbitrary. The two types of coupling tend to excite fields in opposite directions, but which direction at what point is positive is a matter of choice.

Increasing the coupling of a resonator to the guide has the effect of decreasing L and increasing C in Fig. 4(b). Tuning a resonator has the effect of changing L or C or both.

Fig. 5 shows a similar representation in which the two sets of terminals (or, the input and output guides) are coupled by perfect transformers connected in series to shunt resonant circuits.[3] All the properties which can be deduced from the representations of Fig. 4 can be deduced equally well from that of Fig. 5, but we will work with the shunt representation.

Suppose we consider the circuit shown in Fig. 6. This represents a network connected at the input end to a source of admittance M (a pure conductance, the characteristic admittance of the waveguides) and to a load

Fig. 6—Network connect to a source of admittance M at the input end, and to a load of admittance M at the output end.

of admittance M at the output end. The energy of the source is represented by an impressed current $2I_1$. Thus, I_1 is the current which would flow in the output, were the network replaced by a direct interconnection of load and source.

We can represent the properties of the network itself without the terminations M by 4 admittances. Y_{11} is equal numerically to the current which flows in the input circuit (left side) with unit voltage across the input terminals and the output short-circuited. Y_{21} is equal numerically to the current which flows in the output circuit under the same conditions. Y_{22} and Y_{12} are similarly defined with unit voltage across the output terminals and the input short-circuited. In terms of these admittances, we may write the equations of the network with the terminations M and the impressed current $2I_1$ is

$$V_1(M + Y_{11}) + V_2Y_{12} = 2I_1 \qquad (1)$$

$$V_2(M + Y_{22}) + V_1Y_{21} = 0 \qquad (2)$$

$$V_2M = I_2 \qquad (3)$$

We may solve these for the input admittance of the network and load Y_i, and the ratio I_1/I_2. We know concerning Y_i that

$$V_1(M + Y_i) = 2I_1, \qquad (4)$$

and we find, using (1) to (3), that

$$Y_i = (M + Y_{11}) - \frac{Y_{12}Y_{21}}{M + Y_{22}} - M \qquad (5)$$

$$I_1/I_2 = + \frac{(M + Y_{22})(M + Y_{11})}{2MY_{12}} - \frac{Y_{12}}{2M}. \qquad (6)$$

As the network is passive,

$$Y_{21} = Y_{12}. \qquad (7)$$

As it is symmetrical,

$$Y_{11} = Y_{22}. \qquad (8)$$

From Fig. 4, we see that

$$Y_{11} = Y_{22} = Y^{+} + Y^{-} \qquad (9)$$

$$Y_{12} = Y_{21} = Y^{-} - Y^{+}. \qquad (10)$$

Thus we obtain

$$Y_i = \frac{M(Y^{+} + Y^{-})}{M + Y^{+} + Y^{-}} + \frac{4Y^{+}Y^{-}}{M + Y^{+} - Y^{-}} \qquad (11)$$

$$\frac{I_1}{I_2} = \frac{Y^{+} - Y^{-}}{Y^{-} - Y^{+}} + \frac{M}{2(Y^{-} - Y^{+})} + 2\frac{Y^{+}Y^{-}}{(Y^{-} - Y^{+})} \qquad (12)$$

If $I_1/I_2 = 1$, the filter gives perfect transmission.

Now, Y^{+} and Y^{-} are admittance functions having a number of poles, one located at the resonance of each resonator, and zeros between the poles, as indicated in Fig. 7.

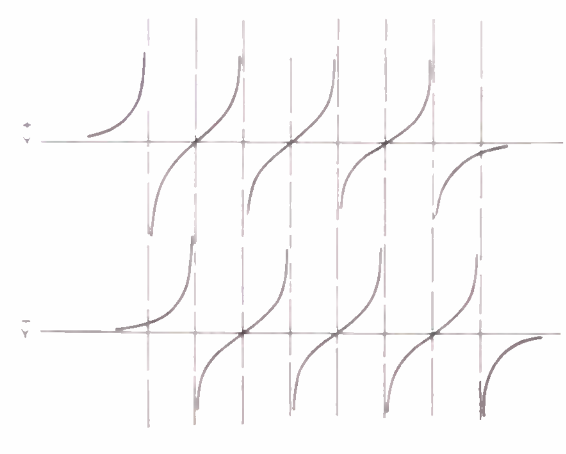

Fig. 7—Behavior of admittance functions.

From (12) we see that, if either Y^{+} or Y^{-} is infinite, the first term has a magnitude of unity and the second term is zero. If, in addition, either Y^{+} or Y^{-} is zero, the third term is zero and $I_1/I_2 = \pm 1$. Thus, if we make the poles of Y^{+} coincide with the zeros of Y^{-} and the zeros of Y^{+} coincide with the poles of Y^{-}, we will have perfect transmission at the frequencies of these pole-zero combinations.

Further, somewhere between a pole and a zero of Y^{+} we will have

$$Y^{+} = -Y^{-} = \pm jB. \qquad (13)$$

For a nearly even spacing of poles and zeros this will be about half-way between a pole and a zero. For the fre-

[3] Schelkunoff gives a very similar representation, and it was this which originally led the writer to the ideas here expressed. See S. A. Schelkunoff, "Representation of impedance functions in terms of resonant frequencies," Proc. I.R.E., vol. 32, pp. 83–90, February, 1944.

quency at which (13) holds, we have

$$\frac{I_1}{I_2} = \mp j\tfrac{1}{2}(M/2B + 2B/M). \qquad (14)$$

From this it is obvious that, at this frequency, we will have perfect transmission with $\pm 90°$ phase shift if we make

$$B = M/2. \qquad (15)$$

Thus, if (1) we tune the directly coupled (Y^+) and contrarily coupled (Y^-) resonators so that the poles of one set lie at the same frequencies as the zeros of the other set, at these frequencies we get perfect transmission with $\pm 180°$ phase shift; and if (2) we adjust the coupling of the resonators so that at frequencies where $Y^+ = -Y^-$ the absolute values of Y^+ and Y^- are equal to $M/2$, we will get perfect transmission at these frequencies.

This presents a fairly clear picture of the action of a properly adjusted paralleled-resonator filter. Suppose we consider a frequency at which a direct-coupled resonator is resonant. At this frequency the sum of transmission through contrary-coupled resonators is zero. As we raise the frequency, a contrary-coupled resonator tuned to a higher frequency begins to transmit. About halfway between the resonance of the direct-coupled resonator and that of the contrary-coupled resonator, the transmission between the two is equal and the phase shift is $-90°$. At a still higher frequency, all transmission will be through the contrary-coupled resonator, and the direct-coupled resonators will transmit nothing; the phase will then be $-180°$. As the frequency is raised, the phase will thus gradually increase, and transmission will shift from one resonator to another.

This immediately tells us that to get a substantially linear phase characteristic in the band we need merely to provide equal spacings between zeros and poles. If, however, we want to get as sharp a cutoff as possible for a given number of resonators, it would be better to tune the resonators far apart and couple them tightly near the center of the band and tune them close together, and couple them loosely near the edges of the band; always, of course, making zeros and poles coincide and fulfilling (14) at (13). This gives a slow variation of phase with frequency near the center of the band and a rapid variation near the edges.

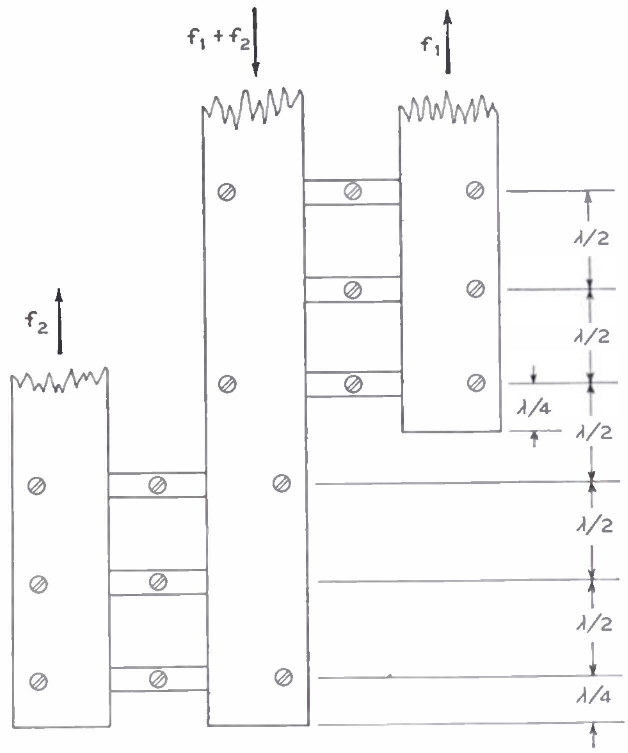

Fig. 8—Paralleled-resonator type branching filters.

There is a large number of possible variations of the paralleled-resonator filter. For instance, various resonator shapes and various coupling means can be used. Further, it should be possible to construct branching filters of the paralleled-resonator type, as indicated in Fig. 8. Here, signals of frequencies f_1 and f_2 enter the center guide, and f_1 emerges from one of the side guides, while f_2 emerges from the other. Theory indicates that it should be possible to overlap the two filters so as to give a fairly good input match over the transition region, and some rough experiments indicate at least an approach to this.

The Received Power of a Receiving Antenna and the Criteria for its Design*

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Summary—**A general formula for the received power of a receiving antenna is derived, whence two theorems are obtained. These theorems are used as criteria for receiving-antenna design under different conditions.**

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Introduction

It is well known that "the properties of an antenna, when used to abstract energy from a passing radio wave, are similar in nearly all respects to the corresponding properties of the same antenna when acting as a radiator."[1] However, this does not tell

[1] F. E. Terman, "Radio Engineers' Handbook," 1st edition, 1943; McGraw-Hill Book Co., New York, N. Y. p. 786.

how much power a receiving antenna can abstract from the incoming wave. Neiman obtained a formula establishing a relation between the receiving power and the field intensity of the incoming wave,[2] but this is only for the special case of a linearly polarized wave. Recently, Friis and Lewis gave a general transmission formula expressing the receiving power in terms of the total radiation power of the transmitting antenna, the wavelength, the effective areas of the transmitting and receiving antennas, and the distance between them.[3,4] This formula is convenient for the case of microwaves, but in general it does not help us very much regarding the receiving-antenna design. The present paper is expected to solve this problem; namely, how to design a receiving antenna to receive maximum power when the incoming wave arriving at the receiver or the transmitting antenna is known.

In this paper we shall use $\hat{E}$ and $\hat{I}$ to indicate the complex values of electric force and current, respectively, and $|\hat{E}|$ and $|\hat{I}|$, their amplitudes. The mks unit system is used.

General Formula

In an antenna, no matter how complicate it is, we take an element ds at its feeding point as reference and assume that this element is at the origin of a spherical co-ordinate system, as shown in Fig. 1. This antenna,

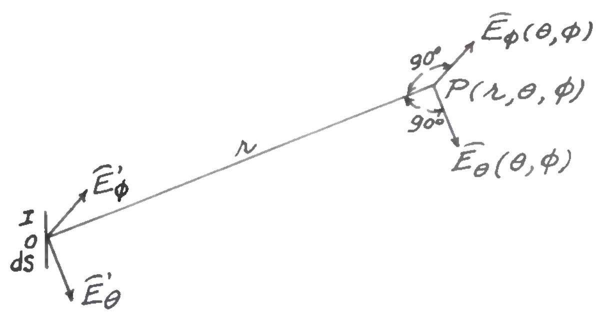

Fig. 1

if used for transmitting, will produce at an arbitrary distance point P an electric force having two components $\hat{E}_\theta(\theta, \phi)$ and $\hat{E}_\phi(\theta, \phi)$ along the increasing-θ and increasing-ϕ directions, respectively. In general, $\hat{E}_\theta$ and $\hat{E}_\phi$ are not in phase; i.e., the radiation is elliptically polarized. Now suppose this antenna is used for receiving instead of transmitting, and an incoming wave arriving at the antenna from the direction PO whose electric force also has two components $\hat{E}_\theta'(\theta, \phi)$ and $E_\phi'(\theta, \phi)$, parallel to $\hat{E}_\theta$ and $\hat{E}_\phi$, respectively. In general, $\hat{E}_\theta'$ and $\hat{E}_\phi'$ are also not in phase. Then the question arises: what power will be received by the antenna?

A current $\hat{I}$ flowing in ds, the reference point of the antenna at the origin, will produce at the distant point P an electric voltage $\hat{E}_\theta ds$ along the increasing-θ direction; reciprocally, a current $\hat{I}_1'$, in the place of $\hat{E}_\theta ds$, will produce an electric voltage $\hat{E}_1' ds$ in $\hat{I}$'s place. By the reciprocity theorem, we have

$$\frac{\hat{E}_\theta ds}{\hat{I}} = \frac{\hat{E}_1' ds}{\hat{I}_1'}.$$

Similarly, $\hat{I}$ at the origin will produce $\hat{E}_\phi ds$ at P in the increasing-ϕ direction, and reciprocally $\hat{I}_2'$ at P in $\hat{E}_\phi ds$'s place will produce $\hat{E}_2' ds$ at the origin in $\hat{I}$'s place. Thus,

$$\frac{\hat{E}_\phi ds}{\hat{I}} = \frac{\hat{E}_2' ds}{\hat{I}_2'}.$$

By the superposition theorem, the resultant voltage induced in the reference element ds at the origin is

$$(\hat{E}_1' + \hat{E}_2')ds = \frac{1}{\hat{I}}(\hat{E}_\theta \hat{I}_1' ds + \hat{E}_\phi \hat{I}_2' ds). \qquad (1)$$

Since it is reasonable to assume uniform current in ds, the available output power P_r is equal to the square of the amplitude of the induced voltage divided by 8 times the radiation resistance:

$$P_r = \frac{|(\hat{E}_1' + \hat{E}_2')ds|^2}{8R}$$

$$= \frac{1}{8R|\hat{I}|^2}|(\hat{E}_\theta \hat{I}_1' ds + \hat{E}_\phi \hat{I}_2' ds)|^2 \qquad (2)$$

where

$|(\hat{E}_1' + \hat{E}_2')ds|$ = the amplitude of the resultant induced voltage

R = the radiation resistance of the antenna with reference to the current at ds.

Because the distance PO is very large, the waves at the origin as produced by $\hat{I}_1' ds$ and $\hat{I}_2' ds$ can be considered as plane waves. Moreover, we can take it that $\hat{E}_\theta'$ is produced by the current element $\hat{I}_1' ds$, and $\hat{E}_\phi'$ by $\hat{I}_2' ds$. Then, since $\hat{I}_1' ds$ and $\hat{I}_2' ds$ are at right angles to the radius vector PO, we have

$$\hat{I}_1' ds = \frac{4\pi r}{\omega\mu}\hat{E}_\theta', \quad \text{and} \quad \hat{I}_2' ds = \frac{4\pi r}{\omega\mu}\hat{E}_\phi' \qquad (3)$$

where

ω = the angular frequency of the incoming wave

μ = the permeability of the medium

r = the distance PO.

On the other hand, the radiation power P_t of the antenna, if used for transmitting, will be

$$P_t = \frac{|\hat{I}|^2 R}{2}. \qquad (4)$$

[2] M. S. Neiman, "The principle of reciprocity in antenna theory," Proc. I.R.E., vol. 31, p. 666; December, 1943.

[3] H. T. Friis and W. D. Lewis, "Radar antennas," *Bell Sys. Tech. Jour.*, vol. 26, p. 230; April, 1947.

[4] H. T. Friis, "A note on a simple transmission formula," Proc. I.R.E., vol. 34, p. 254; May, 1946.

Substituting (3) and (4) in (2), we have

$$P_r = \frac{1}{16P_t}\left(\frac{4\pi r}{\omega\mu}\right)^2 \left| \hat{E}_\theta \hat{E}_\theta' + \hat{E}_\phi \hat{E}_\phi' \right|^2 \tag{5}$$

or, alternatively,

$$\left.\begin{aligned} P_r &= \frac{\lambda^2}{16\pi S_0}\frac{\epsilon}{\mu}\left| \hat{E}_\theta \hat{E}_\theta' + \hat{E}_\phi \hat{E}_\phi' \right|^2 \\ &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\left| \hat{E}_t \cdot \hat{E}_r \right|^2 \end{aligned}\right\} \tag{6}$$

where

λ = the wavelength of the incoming wave

$S_0 = P_t/4\pi r^2$ = the average Poynting vector radiated from the antenna if it is used for transmitting

$\mu/\epsilon = (120\pi)^2$ = the square of characteristic impedance of the medium, and ϵ is dielectric constant

$\hat{E}_t$ = the complex electric force produced by the antenna if used for transmitting

$\hat{E}_r$ = the complex electric force of the incoming wave (the dot between $\hat{E}_t$ and $\hat{E}_r$ means that it is a scalar product).

In conclusion, an antenna, if used for transmitting, radiates a certain power P_t (or average Poynting vector S_0) and an elliptically polarized wave having components of electric force $\hat{E}_\theta$ and $\hat{E}_\phi$ at a distant point P in the direction OP. Now if this antenna is used for receiving an elliptically polarized incoming wave of the angular frequency ω (or wavelength λ) having the components of electric force $\hat{E}_\theta'$ and $\hat{E}_\phi'$, the available output power as received by the antenna is given by (5) and (6).

Applications and Theorems

Let us assume that the phase differences between $\hat{E}_\phi$ and $\hat{E}_\theta$, $\hat{E}_\phi'$ and $\hat{E}_\theta'$, and $\hat{E}_\theta'$ and $\hat{E}_\theta$ are δ, δ', and γ, respectively. Then (6) can be rewritten as follows:

$$\begin{aligned} P_r &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\Big| \left|\hat{E}_\theta\right| e^{j\omega t}\left|\hat{E}_\theta'\right| e^{j(\omega t+\gamma)} \\ &\qquad + \left|\hat{E}_\phi\right| e^{j(\omega t+\delta)}\left|\hat{E}_\phi'\right| e^{j(\omega t+\delta'+\gamma)}\Big|^2 \\ &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\Big| \left|\hat{E}_\theta\right|\left|\hat{E}_\theta'\right| + \left|\hat{E}_\phi\right|\left|\hat{E}_\phi'\right| e^{j(\delta+\delta')}\Big|^2. \end{aligned} \tag{7}$$

From this equation, it is easily seen that the received power will be maximum when $\delta = -\delta'$, and maximum-maximum when, in addition, $|\hat{E}_\phi|/|\hat{E}_\theta| = |\hat{E}_\phi'|/|\hat{E}_\theta'|$, or, in other words, when the angle between $|\hat{E}_t|$ and $|\hat{E}_r|$ is zero. Therefore,

$$\begin{aligned} P_{r\max} &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\left(\left|\hat{E}_\theta\right|\left|\hat{E}_\theta'\right| + \left|\hat{E}_\phi\right|\left|\hat{E}_\phi'\right| \right)^2 \\ &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\left(\left|\hat{E}_t\right|\cdot\left|\hat{E}_r\right| \right)^2 \end{aligned} \tag{8}$$

and

$$\begin{aligned} P_{r\max\text{-}\max} &= \frac{\lambda^2}{16\pi S_0}\cdot\frac{\epsilon}{\mu}\left(\left|\hat{E}_t\right|\left|\hat{E}_r\right| \right)^2 \\ &= \frac{\lambda^2}{4\pi} G_t(\theta,\phi)\, S_r(\theta,\phi) \end{aligned} \tag{9}$$

where $G_t = \sqrt{\epsilon/\mu}\,|\hat{E}_t|^2/2S_0$ is the gain of the antenna if used for transmitting in a given direction, and $S_r = \sqrt{\epsilon/\mu}\,|\hat{E}_r|^2/2$ is the Poynting vector of the incoming wave in the same direction. Equation (8) was obtained in a different form by Neiman.[2]

In summary, the received power is maximum-maximum when the ratio of the complex components of the electric force of the antenna if used for transmitting is the conjugate of that of the incoming wave, i.e., $\hat{E}_\phi/\hat{E}_\theta$ is the conjugate of $\hat{E}_\phi'/\hat{E}_\theta'$. Now let us see how this result is applied to the design of a receiving antenna. We shall discuss four cases separately.

1. $\hat{E}_\phi'/\hat{E}_\theta'$ *of the Incoming Wave is Known*

Under this condition we can use the above result to obtain a theorem for designing the receiving antenna.

Theorem 1: For an incoming wave of given wavelength in a given direction, whose $\hat{E}_\phi'/\hat{E}_\theta'$ is known in magnitude and in phase, suppose a transmitting antenna is so designed that in the same direction it has the highest possible gain and its $\hat{E}_\phi/\hat{E}_\theta$ is the conjugate of $\hat{E}_\phi'/\hat{E}_\theta'$; then this antenna, when used as receiver, will receive maximum-maximum power from the incoming wave.

2. *For Point-to-Point Transmission, the Transmitting Antenna is Given, and the Effect of the Earth's Magnetic Field is Negligible*

The reciprocity theorem, and by consequence the general formula, holds for all conditions except short-wave transmission over long distances.[5] Hence, we obtain another theorem.

Theorem 2: For all point-to-point transmission, except short-wave transmission over long distances in the presence of a magnetic field, assume that the transmitting antenna has been designed to have the highest possible gain in the desired direction. Let a plane mirror bisect and be perpendicular to the line joining the transmitting and receiving stations. Then, to receive maximum-maximum power, the receiving antenna should be designed to be the image of the transmitting one with respect to the mirror.

The proof of this theorem is easy. Since the transmitting antenna has optimum design, it has the highest possible gain in the desired direction, and so does the receiving one, because they are exactly similar. Moreover, $\hat{E}_\phi/\hat{E}_\theta$ must be the conjugate of $\hat{E}_\phi'/\hat{E}_\theta'$, because they are images of each other. Hence, by Theorem 1, this theorem is proved.

[5] See p. 787 of footnote reference 1.

3. *For Point-to-Point Transmission, the Effect of the Earth's Magnetic Field is not Negligible*

For short-wave long-distance transmission, the effect of the earth's magnetic field is not negligible. Let us classify its effect under three cases[6] and see what antenna should be used for each.

Case A: The earth's magnetic field is parallel to the electric force of the wave, but at right angles to the direction of propagation. This corresponds to the transmission across the north or south magnetic poles of the earth. Since the magnetic field has no effect on the wave propagation, of course, Theorem 2 is applicable to the receiving-antenna design.

Case B: Propagation along the direction of the magnetic field. This corresponds to the condition of the north-south transmission. The wave is usually split into two oppositely circularly polarized rays traveling with different velocities, and consequently in different paths. By the known distance between transmitting and receiving stations, the virtual height of the ionosphere, etc., one may determine which ray can reach the antenna and carry out the design by Theorem 1.

Case C: Propagation at right angles to the magnetic field. This corresponds to the condition of east-west transmission. If the wave radiated from the transmitting antenna is elliptically polarized, its electric force has two components—one along the magnetic field, the other at a right angle to it.[7] The former is not affected by the magnetic field, but the latter is. Two components travel with different velocities and consequently in different paths. Therefore, usually only one component will arrive at the receiver. Then the design proceeds as in case B.

The above discussion under three cases is entirely theoretical; in practice, matters are not so simple. First, for very high frequencies the velocities of two polarized rays in cases B and C are practically equal; therefore, they tend to remain together if the distance is not too far. Moreover, it may happen that two rays do travel over different paths, yet both eventually arrive at the same receiver because of their different velocities. Second, the propagation of the waves may not be strictly under the three cases as discussed above; for instance, the magnetic field may make an arbitrary angle with both directions of propagation and electric force. Third, we have not taken into account the ground reflection, which usually is not negligible. Fourth, the path of wave propagation varies not only with frequency, but also with ion density, height of ionosphere, etc., and consequently varies from time to time. For these reasons, the sky wave arriving at the receiver is, in general, elliptically polarized. To have an antenna receive such a wave, it is better to design two antennas, or one antenna consisting of two parts, both of which have the highest possible gain in the given direction, one being polarized horizontally and the other vertically. From these two parts, separate transmission lines are provided to lead the energy to the receiver. Moreover, an adjustable phase shifter, which can change the phase relationship of two inputs from 0 to π positively or negatively, is inserted. The phase shifter suggested by Friis and Feldman can be used.[8] By the phase adjustment, the phase difference between the horizontal and the vertical components of the electric force of the antenna, when used as transmitter, can be adjusted to be the negative of that for the incoming wave. Then maximum power is received. Furthermore, if the horizontally or vertically polarized parts or both are adjustable so as to satisfy also the condition $|\hat{E}_\varphi|/|\hat{E}_\theta| = |\hat{E}_\varphi'|/|\hat{E}_\theta'|$, then the antenna receives maximum-maximum power; otherwise, it is sacrificed.

4. *The Transmitting Antenna and Polarization of the Incoming Wave are Completely Unknown*

If the incoming wave, though unknown, is approximately constant with time, the best design procedure is to measure E_φ'/E_θ' in magnitude and in phase. The simple method suggested by Appleton can be used, except that a heterodyne oscillator is provided instead of a ground wave, because the latter may be too weak to be useful.[9] After $\hat{E}_\varphi'/\hat{E}_\theta'$ is obtained, the antenna is designed by Theorem 1.

In case $\hat{E}_\varphi'/\hat{E}_\theta'$ of the incoming wave varies greatly with time in magnitude and in phase, an antenna consisting of two parts, together with the phase shifter described in last section, can be used.

Conclusion

The general formula gives a method of calculating the received power, and Theorems 1 and 2 give a method of designing the receiving antenna. Among the different kinds of transmission, the most troublesome one is the short-wave long-distance transmission in presence of a magnetic field. If the conditions coincide with one of the three cases as discussed above, and the polarized ray which will arrive at the receiver is known, Theorems 1 and 2 can be used to design the receiving antenna. Otherwise, we measure $\hat{E}_\varphi'/\hat{E}_\theta'$ of the incoming wave in magnitude and in phase, and use Theorem 1 to design the antenna; or we use an antenna consisting of two orthogonally polarized parts together with a phase shifter.

[6] H. W. Nichols and J. C. Schelleng, "Propagation of electric waves over the earth," *Bell Sys. Tech. Jour.*, vol. 4, p. 215; April, 1925.

[7] There is also a third component along the direction of propagation, but it is usually small and negligible.

[8] H. T. Friis and C. B. Feldman, "A multiple steerable antenna for shortwave reception," Proc. I.R.E., vol. 25, p. 841; July, 1937.

[9] E. V. Appleton, "A simple method of demonstrating the circular polarization of ionospherically reflected radio waves," *Nature* (London), vol. 151, p. 250; February, 1943.

Contributors to Proceedings of the I.R.E.

David R. Brown (S'43–A'45–M'48) was born on October 31, 1923, at Los Angeles, Calif. He received the B.S. degree in electrical engineering from the University of Washington in 1944, and the S.M. degree in electrical engineering from the Massachusetts Institute of Technology in 1947. From 1944 to 1946, he was a member of the staff of the Applied Physics Laboratory, University of Washington.

David R. Brown

In April, 1946, Mr. Brown joined the staff of the Servomechanisms Laboratory of the Massachusetts Institute of Technology, where he assisted in the development of a large-scale digital computer.

In September, 1948, he joined the staff of the Electrical Engineering Division of the University of California at Berkeley, Calif.

Mr. Brown is a member of Tau Beta Pi, Sigma Xi, and the American Association for the Advancement of Science.

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Albert W. Friend (A'34–M'39–SM'43) was born in Morgantown, W. Va. on January 24, 1910. He received the B.S.E.E. and M.S. (Physics) degrees from West Virginia University in 1932 and 1936, respectively, and the S.D. degree in communication engineering from Harvard University in 1948. Between 1929 and 1934, he was engaged in public utilities and radio interference engineering and communications consultation. From 1934 to 1937, he was an instructor, and from 1937 to 1944, an assistant professor of physics at West Virginia University, on leave after 1939. In 1939 he transferred his tropospheric-sounding research to Harvard University, where he was an instructor in physics and communication engineering and a research fellow in Cruft Laboratory until 1941, and a research fellow in the Blue Hill Meteorological Observatory until 1942.

Albert W. Friend

In January, 1941, Dr. Friend transferred to the Radiation Laboratory of the Massachusetts Institute of Technology, as a research associate and later a staff member, where he worked on the development of microwave receivers, taught in the MIT Radar School, and extended his tropospheric echo studies. From July, 1942, to August, 1944, he acted as technical director of the Heat Research Laboratory at MIT, and at the same time as a consultant on guided-missile control with Division 5 of NDRC.

In 1944 he joined the RCA Victor Division, in Camden, N. J., continuing in television receiver development until 1947. During 1946 and 1947, he was also on the research staff of the Electronic Research Laboratory at Harvard University. Since June, 1947, he has been a member of the research staff of the Radio Corporation of America, in the RCA Laboratories Division at Princeton, N. J.

Dr. Friend is a member of the AIEE, the American Meteorological Society, the American Geophysical Union, Tau Beta Pi, Sigma Pi Sigma, Sigma XI, and the Harvard Engineering Society. He is a registered professional engineer.

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David G. C. Luck (M'46) was born in Whittier, Calif., in July, 1906. He received the S.B. degree in 1927 and the Ph.D. degree in 1932, both from the Massachusetts Institute of Technology. A recipient of both fellowships and an assistantship, he remained in the department of physics at MIT during the period of 1927–1932. He joined the research division of the RCA-Victor Company, Inc., at Camden, N. Y., in 1932, and transferred in 1942 to the RCA Laboratories Division at Princeton, N. J., where he is still engaged.

David G. C. Luck

Dr. Luck's major field of interest is in electronic instrumentation. In this connection, he developed the metal detector, the omnidirectional radio range, and a cathode-ray direction finder. He has also investigated pulse-time modulation, and studied direction-finder errors. During the war he developed applications of frequency-modulated radar. At present Dr. Luck is working in the general field of radio aids to aviation. He is a member of the American Physical Society and Sigma Xi.

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For a photograph and biography of J. R. Pierce, see page 1003 of the August, 1948, issue of the Proceedings of the I.R.E.

Nathaniel Rochester (S'41–A'43–M'46) was born in Buffalo, N. Y., on January 14, 1919. He received the B.S. degree in electrical engineering in June, 1941, from the Massachusetts Institute of Technology. He received an appointment to the research staff of MIT as a research assistant in February, 1941, and, was connected with the MIT-CAA Blind Landing Project. In April, 1941, he entered the MIT Radiation Laboratory where he did research and liaison work on crystal rectifiers.

N. Rochester

He joined Sylvania Electric Products Inc. in April, 1943, where he was concerned with the development and construction of radar sets and other large devices, and where he became manager of electronic equipment engineering in the Electronics Division in Boston, Mass.

In November, 1948, Mr. Rochester joined the Engineering Department of International Business Machines Corporation, Poughkeepsie, N. Y.

Mr. Rochester is a member of the American Physical Society, Tau Beta Pi, and an associate member of Sigma Xi.

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Yung-Ching Yeh (A'35) was born in China in 1905. He received the B.S. degree in electrical engineering in 1929 from the National Chiao-Tung University, in Shanghai, and the Radio Engineer degree in 1921 at L'École Supérieure d'Électricité, in Paris, France.

Yung-Ching Yeh

At the beginning of World War II, Professor Yeh was a major in the Chinese Air Force, in charge of training radio operators and technicians. In 1939 he became a professor at the National Wu-Han University, Wu-Chang, China, and in 1943 he was in charge of the Institute of Electrical Engineering. At the end of the War he was sent by the Chinese Government to the United States, and became associated with the Radiation Laboratory and the Research Laboratory of Electronics of the Massachusetts Institute of Technology for a period of two years. Professor Yeh has now returned to the National Wu-Han University in China.

Institute News and Radio Notes

1949 IRE National Convention Program

HOTEL COMMODORE and GRAND CENTRAL PALACE—MARCH 7–10

PROGRAM

Monday, March 7, 1949

9:00 A.M.–5:30 P.M.—Registration at Hotel Commodore.
9:30 A.M.–9 P.M.—Registration at Grand Central Palace.
9:30 A.M.–9 P.M.—Radio Engineering Show, Grand Central Palace.
10:30 A.M.–12 M.—Annual Meeting. Principal Address: "Perpetual Youth in the IRE," by I. S. Coggeshall. Grand Ballroom, Hotel Commodore.
2:30–5:00 P.M.—"Systems I—Modulation Systems." "Antennas and Waveguides." Symposium: "Network Theory." "Instruments and Measurements I—Microwave." "Audio."
6:00–8:00 P.M.—Get-Together Cocktail Party, Hotel Commodore.
8:00 P.M.—Sections Committee Meeting, Hotel Commodore.

Tuesday, March 8, 1949

9:00 A.M.–5:30 P.M.—Registration.
9:30 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace.
10:00 A.M.–12:30 P.M.—"Antennas." "Passive Networks I—Synthesis." "Instruments and Measurements II—Oscillography." "Electronic Computers."
12:45 P.M.—President's Luncheon, honoring President-Elect Bailey, Grand Ballroom, Hotel Commodore. Featured speaker: Delos W. Rentzel, Civil Aeronautics Administrator; topic: "All-weather Flying." Guest of Honor: Stuart L. Bailey. Toastmaster: B. E. Shackelford.
2:30–5:30 P.M.—Symposium: "Electronic Computing Machinery." "Wave Propagation I—Television." "Passive Networks II—Analysis." "Components and Materials." "Nucleonic Instrumentation."
8:00–10:30 P.M.—Symposium: "Nuclear Science."

Wednesday, March 9, 1949

9:00 A.M.–5:30 P.M.—Registration.
9:30 A.M.–6:00 P.M.—Radio Engineering Show, Grand Central Palace.
10:00 A.M.–12:30 P.M.—"Television I." Symposium: "Radio Aids to Navigation." "Active Circuits I." "Instruments and Measurements III." "Electronics I—Tube Design and Engineering."
2:30–5:00 P.M.—"Television II." "Wave Propagation II." "Active Circuits II." "Instruments and Measurements IV." "Electronics II—Electron-Tube Cathodes."
6:45 P.M.—Annual IRE Banquet (dress optional), Hotel Commodore. Toastmaster: Raymond F. Guy. Featured speaker: Frank A. Stanton; topic: "Television and People." Awarding of the Medal of Honor, the Morris Liebmann Memorial Prize, the Browder J. Thompson Memorial Award, and Fellow Awards. Fellows' Acceptance will be delivered by Karl Spangenberg.

Thursday, March 10, 1949

9:00 A.M.–5:30 P.M.—Registration.
9:30 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace.
10:00 A.M.–12:30 P.M.—"Systems II—Relay Systems." "Navigation Aids I." Symposium: "Marketing." "Electronics III—Electron-Tube Theory."
2:30–5:00 P.M.—Symposium: "Germanium and Silicon Semiconductors." "Information Transmission and Noise." "Navigation Aids II." "Oscillators." "Electronics IV—New Forms of Tubes."

Committee Meetings

The following IRE committee meetings have been scheduled during the convention. The times and places will be announced in the Convention Program.

Board of Editors
Chairman, A. N. Goldsmith

Education Committee
Chairman, H. J. Reich

Membership Committee
Chairman, R. F. Guy

Nuclear Studies Committee
Chairman, L. R. Hafsted

Professional Groups Committee
Chairman, W. L. Everitt

Research Committee
Chairman, F. E. Terman

Sections Committee
Chairman, A. W. Graf

Standards Committee
Chairman, A. B. Chamberlain

All chairmen and vice-chairmen of technical committees are requested to be present at the Standards Committee meeting in order to present all suggestions for technical committee operations to the Standards Co-ordinator and the Chairman of the Standards Committee. The Standards Co-ordinator is looking forward to greeting the chairmen of all Technical Committees personally. This meeting will also provide the first opportunity for a get-together with the chairmen of the several new technical committees created during the past year.

ANNUAL MEETING

MONDAY, MARCH 7
10:30 A.M.

This opening meeting of the convention is for the entire membership. The meeting will feature the following address:

PERPETUAL YOUTH—AND THE IRE

IVAN S. COGGESHALL
(Western Union Telegraph Co., New York, N. Y.)

One of the significant sources of strength of the IRE, in achieving its educational objectives and in serving the engineering profession, has been the youthfulness of its membership, its ideas, and ideals. As the average age of its members increases, steps must be taken leading to the continuous renewal of corporate youth. Elements of this problem will be discussed as they are related to the 1949 Convention theme: "Radio-Electronics—Servant of Mankind."

SUMMARIES OF TECHNICAL PAPERS

NOTE

No papers are available in preprint or reprint form nor is there any assurance that any of them will be published in the PROCEEDINGS OF THE I.R.E., although it is hoped that many of them will appear in these pages in subsequent issues.

Systems I

Modulation Systems

1. DEVELOPMENT OF A HIGH-SPEED COMMUNICATION SYSTEM

DONALD S. BOND
(RCA Laboratories, Princeton, N. J.)

An experimental ultrafax system operating at transmission speeds up to 15 pages a second is described. A flying-spot scanner employing a single-line sweep scans the copy, which has been previously photographed on film. A projection kinescope of improved electron-optical focus and with a short-persistence, high-efficiency phosphor was developed for the scanner.

The receiving apparatus includes circuits similar to those required in the horizontal channel of television equipment. Clamp circuits are provided in both scanner and receiver to maintain the dc axis of the video wave in the presence of rapid changes in background brightness. A 5-inch projection-type kinescope with a P-11 screen is used. The continuous-drive recording camera uses 16-mm motion-picture film. The exposed film is developed in a hot-processing unit with an elapsed time of about 45 seconds. Tests have been conducted over a 7,000-Mc television relay system.

2. DISTORTION IN A PULSE-COUNT-MODULATION SYSTEM WITH NONUNIFORM SPACING OF LEVELS

P. F. Panter and W. Dite

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The first part of the paper deals with statistical properties of the signal. The signal power is expressed in terms of the probability density of the signal.

In the second part these properties are applied to the problem of distortion in a pulse-count-modulation system using a nonuniform distribution of the levels. It is shown that, for a signal having a probability density $P(y)$, the level spacing can be chosen to yield minimum distortion, which is given by

$$D_m^2 = \frac{2}{3N^2} \frac{\left(\int_0^V P^{1/3}(y)dy\right)^3}{\int_0^V y^2P(y)dy}$$

where N is the total number of levels in the range $-V$ to V. It is shown that the minimum distortion is significantly less than the distortion due to uniform level spacing when the crest factor of the signal exceeds 4.

Nonuniform spacing of the levels may be thought of as equivalent to compression followed by a uniform quantization of the signal. It is shown that, when the spacing of the levels is logarithmic, the resulting distortion is nearly independent of the signal.

3. CROSS-TALK CONSIDERATIONS IN TIME-DIVISION MULTIPLEX

S. Moskowitz, L. Diven, and L. Feit

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

An experimental study was made of the effects on interchannel cross talk of the bandwidth characteristics of the transmission medium in pulse-time-multiplex systems. Both pulse-amplitude and pulse-position modulation systems are considered. The effects of various types of high- and low-frequency response are discussed from both the experimental and the theoretical point of view.

4. EXPERIMENTAL VERIFICATION OF VARIOUS SYSTEMS OF MULTIPLEX TRANSMISSION

D. R. Crosby

(RCA Victor, Camden, N. J.)

Using a microwave link of 12 miles, six different modulation methods for multichannel transmission were studied. Some modulation methods used frequency division; others used time division employing pulse modulation. Some of the systems used triple modulation; one pulse system quantized the signal.

The purpose of the work was to get an acquaintance with each of the systems to see if there were important properties that the theoretical studies had not emphasized. Good agreement with the theoretical performance formulas was obtained.

For each system the signal-to-noise ratio was evaluated for four types of noise: (1) thermal noise, (2) continuous-wave interference, (3) impulse noise, and (4) co-channel system noise.

5. INTERFERENCE CHARACTERISTICS OF PULSE-TIME MODULATION

E. R. Kretzmer

(Massachusetts Institute of Technology, Cambridge, Mass.)

The susceptibility to interference of a modulation system is one of its most important characteristics. The interference characteristics of pulse-time modulation, of which pulse-duration modulation and pulse-position modulation are particular forms, are discussed with particular reference to two-station and two-path interference. Experimental investigations, as well as mathematical work largely based on N. Wiener's autocorrelation method of analysis, lead to some interesting and useful results not heretofore known. Single-channel pulse-duration modulation, in particular, is found to have some remarkable characteristics, exhibiting a strong capture effect in favor of the stronger of two interfering transmissions, and permitting relatively good reception under most conditions of severe multipath interference. The basic facts responsible for these results, as well as the practical requirements that must be met to attain them, will be briefly discussed.

6. FACTORS INVOLVED IN THE DESIGN OF AN IMPROVED FREQUENCY-SHIFT RECEIVING SYSTEM

Colin C. Rae

(Naval Research Laboratory, Washington, D. C.)

A discussion of the theoretical and practical aspects of a frequency-shift converter developed for the purpose of reducing noise and multipath interference effects on Naval facsimile circuits. Practical results, obtained over a period of six months operation on Naval radiophoto circuits, provide proof of the system's superiority over existing equipments and confirm theoretical predictions. It is expected that application of the principles involved to teletype and high-speed keying circuits will provide worth-while increases in radio circuit reliability.

Antennas and Waveguides

7. ELLIPTICALLY POLARIZED RADIATION FROM INCLINED SLOTS ON CYLINDERS

George Sinclair

(University of Toronto, Toronto, Ont.)

A method is presented for calculating the patterns of slots of finite length and arbitrary orientation on the surface of a cylinder. Previously, it has been shown that, by assuming the slot to be of infinite length and parallel to the axis of the cylinder, the pattern in a transverse plane can be calculated. It will be shown that, under certain reasonable assumptions, it is possible to calculate the 3-dimensional pattern of a slot of finite length located arbitrarily on the surface of a cylinder of circular or elliptic cross section. It is found that the radiation from an inclined slot is generally elliptically polarized.

Experimental verification of the accuracy of the calculations will be presented. Application of the theory to the design of FM antennas for linear and for circular polarization will be considered.

8. SOME PROPERTIES OF RADIATION FROM RECTANGULAR WAVEGUIDES

John T. Bolljahn

(University of California, Berkeley, Calif.)

Certain exact relationships between the radiation pattern and impedance characteristics of a radiating rectangular waveguide with vanishingly thin walls are developed. In particular, it is shown that the ratio of radiation intensities in certain preferred directions and the power gain in these directions are related in a simple manner to the reflection coefficient inside the waveguide. Although the information obtained is restricted, in the sense that it applies to only a few discrete points on the radiation pattern, it is quite general as regards the manner in which the waveguide is broken or perforated to allow radiation. The results are shown to apply equally well if conducting sheets having arbitrary shapes but lying on specified planes are present in the vicinity of the radiating waveguide.

9. THEORY OF END-FIRE HELICAL ANTENNAS

Arthur E. Marston and M. D. Adcock

(Naval Research Laboratory, Washington, D. C.)

An exact formula for the radiation field of a helix energized by a traveling wave of current along its length is obtained. This formula yields a definite condition to be satisfied by the geometry of the helix and the propagation constant of the traveling wave of current in order that the radiation field of the helix be end-fire with circular polarization on axis. In addition, the general radiation formula yields a simple, explicit expression for the ellipticity on axis of any helix excited by a traveling wave of current. Further, it is shown that the current distribution for the end-fire helices is the current distribution associated with the $n=1$ mode of the helical traveling-wave tube as analyzed recently by C. Shulman and M. S. Heagy of RCA. Combining the theory of this mode with the general results indicated above, a complete theory of the radiation fields of end-fire helices is achieved which checks well with experimental observation.

10. A BROAD-BAND TRANSITION FROM COAX TO HELIX

C. O. Lund

(RCA Laboratories, Princeton, N. J.)

A new type of coax to helix transition, consisting of a nonuniform helix in which the pitch changes continuously from infinity to the pitch of uniform helix, is described. The smooth change in pitch is accompanied by a gradual change in conductor size, so that the part with infinite pitch is essentially a cylindrical conductor with a thin longitudinal slot. The transition, designed primarily for traveling-wave tubes, can have other uses. An approximate theory for this transition is outlined and compared with measurements.

11. EQUIVALENT CIRCUITS FOR COUPLING OF WAVEGUIDES BY APERTURES

N. Marcuvitz
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

The rigorous description of the field behavior within waveguide structures may be expressed in terms of transmission lines and lumped-constant circuits. This leads to a waveguide network theory closely related to ordinary low-frequency network theory. The determination of the parameters that characterize a number of waveguide networks—in particular, those involving aperture-type discontinuities—will be considered. The Kirchhoff analysis of such waveguide networks as well as the Schwinger integral equation and variational method of evaluating the network parameters will be discussed for the case of small apertures of various shapes coupling different waveguides.

SYMPOSIUM
Network Theory

1. Modern Developments in the Topology of Networks

Ronald M. Foster
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

2. A Survey on the Status of Linear Network Theory

E. A. Guillemin
(Massachusetts Institute of Technology, Cambridge, Mass.)

3. Recent Developments in Broad-Band Active Networks

John G. Linvill
(Massachusetts Institute of Technology, Cambridge, Mass.)

4. General Review of Linear Varying-Parameter and Nonlinear Circuit Analysis

W. R. Bennett
(Bell Telephone Laboratories Inc., Murray Hill, N. J.)

The symposium on Network Theory has been planned in such a way that there will be a general review of recent work in practically all of the fields of network theory in which there is major activity. These four papers are on the geometry of networks, passive and active networks with lumped parameters, and linear varying-parameter and nonlinear circuits. Prof. Foster's paper will discuss new developments in the geometry of networks, a field in which little has been written but from which network theory has drawn many of its basic tenets. Profs. Guillemin and Linvill will review lumped-parameter network theory, with Prof. Guillemin giving a broad general review of the field and Prof. Linvill emphasizing recent work in broad-band active circuits. Dr. Bennett will give a broad general summary of linear varying-parameter and nonlinear circuit analysis.

Instruments and Measurements I
Microwave

12. MEASURING THE EFFICIENCY OF A SUPERHETERODYNE CONVERTER BY THE INPUT-IMPEDANCE CIRCLE DIAGRAM

Harold A. Wheeler and David Dettinger
(Wheeler Laboratories Inc., Great Neck, L. I., N. Y.)

There is presented a practical method of measuring conversion efficiency, input and output image impedance of crystal converters in superheterodyne receivers. The small-signal rf input impedance of the converter (with normal local oscillator excitation) is plotted on the hemisphere chart as a function of if load impedance. The largest circle is obtained by variation of a pure reactance if load from parallel resonance to short circuit. The radius of this circle is equal to the conversion efficiency if the circle is centered on the chart. The central point of an arc plotted across this circle by variation of a resistance in parallel with the if circuit at resonance marks the if resistance equal to the output image impedance. This point marks also the if input image impedance.

13. ELECTROLYTIC-TANK MEASUREMENTS FOR MICROWAVE DELAY-LENS MEDIA

Seymour B. Cohn
(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

By means of symmetry and image considerations, it is shown that the exact solution for a microwave delay lens medium consisting of a uniform array of obstacles may be obtained from the solution for a single obstacle with proper boundary conditions. It is shown that, for obstacle spacings much less than a wavelength, these boundary conditions may be easily and exactly simulated in an electrolytic tank. It is also shown how approximate frequency corrections may be applied. The results of measurements by this method are given for thin, square, and circular obstacles.

Through an application of Babinet's principle, the electrolytic tank method may also be used in measuring the magnetic polarizability of small apertures of any shape.

14. A MICHELSON-TYPE INTERFEROMETER FOR MICROWAVE MEASUREMENTS

Bela A. Lengyel
(Naval Research Laboratory, Washington, D. C.)

The optical Michelson interferometer is modified by replacing one of its branches by a directional coupler and a waveguide. The instrument serves many purposes, among which are: precision wavelength determination, the measurement of dielectric constants of materials available in sheet form, the determination of reflection from laminated sheets at normal incidence, the study of metal-loaded dielectrics and of parallel-plate metal lens media. An instrument operating at 3.2 cm is described.

15. IMPEDANCE INTRUMENTATION FOR MICROWAVE TRANSMISSION LINES

Pierre A. Portmann
(Naval Research Laboratory, Washington, D. C.)

Three measurable sets of parameters describing the condition of a load on a transmission line are considered. These are: (1) generalized voltage versus current, (2) complex standing-wave ratio, and (3) complex reflection coefficient. The utility of these parameters for measurement purposes is discussed in terms of their applicability to automatic and semiautomatic instruments. The reflection coefficient is shown to be easily transformed to the other two parameters.

The standing wave ratio is applied to a semiautomatic instrument utilizing a circular waveguide phase-shifter. The output data are presented on an oscilloscope. The reflection coefficient is applied to a semiautomatic instrument; data are presented on a meter and calibrated scale.

16. A BROAD-BAND HIGH-POWER MICROWAVE ATTENUATOR

Herbert J. Carlin
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

The principal portion of the device consists of a high attenuation probe whose broad-band characteristics are obtained by a simple yet novel arrangement of discontinuity capacitance in coaxial line. Supplementing the probe is a lossy section of coaxial line designed to function as an attenuation equalizer. The power in the attenuator is dissipated in a broad-band coaxial high-power load which can dissipate over 100 watts. A typical design, which was built and tested, has a maximum input VSWR of 1.16, and an attenuation of 50±1 db over the frequency band 1,000 to 3,000 Mc.

17. AN ABSOLUTE METHOD FOR MEASURING MICROWAVE POWER OF LOW INTENSITY

Harold Herman
(Naval Research Laboratory, Washington, D. C.)

A method of absolute power measurement is described for the microwave and millimeter wavelength region using an adiabatic microcalorimeter having a limiting sensitivity of approximately 5 microwatts. For an accuracy of about 2 per cent, the system is usable at levels above 100 microwatts. The errors involved and the application of the method to the calibration of bolometer mounts are discussed.

Audio

18. THE REPRODUCTION OF SOUND

Harry F. Olson
(RCA Laboratories, Princeton, N. J.)

All-acoustic subjective frequency-preference tests have shown that the average listener prefers the full frequency range in speech and music, as compared to a restricted frequency range. Therefore, if the listener does not prefer a full frequency range in reproduced sound, the logical conclusion must be that the reproducing system introduces objectionable distortions which are minimized by the restricted frequency range.

As sound-reproducing systems have been developed, engineers have evolved tests which depict the distortions and, as a consequence, the performance of a sound-reproducing system. The present state of the art indicates that it is possible to outline the performance characteristics of a high-fidelity reproducing system.

In order to reproduce the physical characteristics of a musical tone, the reproducing system must be free of various distortions and deviations, which will be defined.

All sound-reproducing systems introduce all of these distortions and deviations. Typical amplitude, nonlinear, phase, and transient distortions, and noise characteristics obtained in sound-reproducing systems, will be presented. Data obtained from objective and subjective measurements show that distortions in the form of nonuniform response-frequency characteristic, nonlinear distortion, noise, phase distortion, and transient response become more objectionable as the frequency range is extended.

19. NEW DEVELOPMENTS IN STUDIO DESIGN IN EUROPE

Leo L. Beranek

(Massachusetts Institute of Technology, Cambridge, Mass.)

New broadcasting houses in Oslo, Norway, and Copenhagen, Denmark, contain important advances in studio design, while preserving beauty of architecture. Use is made of perforated facings and cavity resonators as part of the wall structure to control the reverberation time versus frequency characteristic. Efficient methods of vibration insulation are used. Shaping has been designed to provide freedom from flutter echos, to increase the excitation of the normal modes of vibration in the room, and to provide some diffusion. Interesting details of doorways, windows, and wall structures have been developed to reduce sound transmission between rooms. Listening-jury judgments performed in Copenhagen have revealed a new optimum reverberation-time characteristic for orchestra studios. British judgments indicate the desirability of logarithmic decay curves free from fluctuations. European experience shows the need for a reverberant, diffusing-type enclosure around the orchestra to satisfy the players.

20. THE TECHNIQUE OF TELEVISION SOUND

Robert H. Tanner

(Northern Electric Co. Ltd., Belleville, Ont., Canada)

This paper is intended to present a general survey of the problems and possibilities of the audio portion of a complete television program. The difficulties encountered both in achieving good-quality sound reproduction under the conditions obtaining in a television studio, and in ensuring that the sound accurately matches the picture, are discussed, together with the problem of making a single studio give the wide variety of acoustic characteristics demanded by the changing scenes. Possible improvements and suggested lines of development are mentioned, and the application of binaural reproduction is proposed as one method of obviating many of the present difficulties.

21. THE MEASUREMENT OF NONLINEAR DISTORTION

Arnold P. G. Peterson

(General Radio Co., Cambridge, Mass.)

The harmonic method and intermodulation methods of measurement of nonlinear distortion are briefly reviewed. For simple, nonlinear systems that are not frequency dependent, Warren and Hewlett have computed comparative distortion values for harmonic measurements and for the modulation or carrier-analysis method of intermodulation measurement. These calculations are extended to provide comparative values for the difference-frequency method of intermodulation measurement.

A discussion of the special problems involved in determining nonlinear distortion in amplifiers, loudspeakers, recording systems, filters, noise suppressors, FM, and hearing aids leads to the conclusion that great flexibility is required in the measurement system. The most generally useful test arrangement devised so far for this purpose is a combination of a two-signal audio generator having continuously adjustable frequencies and an analyzer for measuring the various difference-frequency components. This procedure is in agreement with that recommended by the CCIF. The advantages of this method are illustrated by applying it to the measurement of distortion in amplifiers.

Antennas II

22. WIDE-ANGLE METAL-PLATE OPTICS

John Ruze

(Air matériel Command, Cambridge, Mass.)

The wide-angle scanning possibilities of metal-plate media are examined and general design equations are derived for constrained-type lenses. Pattern distortion at points other than the design points is examined with the aid of a series expansion of the phase errors.

Double-media and periodic-structure lenses are discussed. Application of the Campbell formula to the latter type indicates nearly achromatic or wide-band performance.

Experimental data are presented on both constant- and variable-refractive-index lenses. Further experimental data indicate the importance of smooth contours and compact structures in achieving low spurious radiation and large scanning angles.

23. DIFFRACTION PATTERN FROM AN ELLIPTICAL APERTURE

R. J. Adams and K. S. Kelleher

(Naval Research Laboratory, Washington, D. C.)

The radiation pattern of a microwave antenna of elliptical contour is expressed in terms of a Fourier-series approximation to the aperture illumination. The method is easily applied to a lens or paraboloidal reflector by measuring the primary patterns of the feed in two principal planes. Typical examples show that the calculated minor-lobe structure, beamwidth, and gain are in close agreement with experiment.

24. THE MEASUREMENT OF CURRENT AND CHARGE DISTRIBUTIONS ON TRANSMITTING AND RECEIVING ANTENNAS

Tetsu Morita

(Harvard University, Cambridge, Mass.)

An experimental investigation at 300 Mc was made on the current and charge distribution along a cylindrical antenna over a ground plane for both transmission and reception. With the validity of these measurements established by correlation with the King-Middleton theory, measurements were also made on coupled antennas and folded dipoles of half-length $\lambda/4$ and $\lambda/2$ for spacings of 0.04λ. Each measured distribution was resolved into its symmetrical and antisymmetrical components by the phase-measurement data, and these were identified with the radiating current and the transmission-line current. Through a slot cut along the antenna, a current or charge probe was moved over the entire length of the antenna to measure the current or charge distribution.

25. ANTENNA SYSTEMS FOR MULTICHANNEL MOBILE TELEPHONY

W. C. Babcock and H. W. Nylund

(Bell Telephone Laboratories, Inc., New York, N. Y.)

This paper describes an arrangement whereby several antennas may be mounted on a single mast at the transmitting site of a multichannel mobile system operating in the 152- to 162-Mc band. The antennas are so disposed as to minimize shadowing effects of the mounting structure, while keeping intertransmitter coupling to a tolerable minimum. Measurements of the characteristics are presented for arrangements of 6 and 12 antennas, respectively, mounted on a 62-foot steel mast. These measurements on a full-scale structure are supplemented by tests at a higher frequency on reduced-scale simplified models.

26. A LOW-DRAG AIRCRAFT ANTENNA FOR RECEPTION ON OMNIDIRECTIONAL RANGE SIGNALS IN THE 108- TO 122-Mc BAND

John P. Shanklin

(Collins Radio Co., Cedar Rapids, Iowa)

A practical antenna for the described application having 2.63 pounds drag at 250 miles per hour is made possible by the application of the unique "notch feed" principle in which the feed is across a notch cut part way through the center of the dipole. This results in an unbroken half-wave dipole. The dipole is bent into a U shape to achieve an omni-directional pattern and to reduce drag.

The "notch feed" serves as a balanced-to-unbalanced line transformer, impedance-level transformer, and broad-banding impedance corrector.

Passive Networks I

Synthesis

27. A METHOD OF SYNTHESIZING THE RESISTOR-CAPACITOR LATTICE STRUCTURE

J. L. Bower, J. T. Fleck, and P. F. Ordung

(Yale University, New Haven, Conn.)

With the advent during the past decade of devices such as servo systems and predictors, the resistor-capacitor filter has come into prominence as the one best suited to such low-frequency work. Starting from the limitations peculiar to the resistor-capacitor structure, this paper develops a general method of realizing a given transfer-admittance, transfer-impedance, or transfer-voltage ratio in a lattice network. A particular emphasis is placed on achieving the maximum gain factor possible in such a network. The method provides for the realization of a prescribed voltage-transfer ratio in a filter feeding a given resistor-capacitor load.

28. EXACT DESIGN OF BANDPASS NETWORKS USING n COUPLED FINITE-Q RESONANT CIRCUITS ($n=3$ AND 4)

M. Dishal

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

When continuously increasing high rates of cutoff are required outside the desired pass-band, triple, quadruple (etc.) tuned band-pass networks must be used in place of the very familiar double-tuned "if transformer." Unfortunately, the design equations of classical filter theory are not satisfactory when exact response shapes are desired, because they are based on infinite Q, i.e., nondissipative, coupled resonant circuits, and the so-called reflection factors and interaction factor have been neglected. Based upon a method previously presented using the coefficients of the complex determinant equation, the exact required values are given for all the coefficients of coupling and all the Q's, in triple- and quadruple tuned band pass networks in order to obtain either one of two exact symmetrical response shapes. No small percentage pass-band approximation is made. Contrary to the much used statement that dissipation in the filter elements reduces the sharpness of cutoff at the edges of the pass band, it is shown that all the resonant circuits used can have finite Q's with no degradation whatsoever of the pass-band edges.

29. NETWORK APPROXIMATION IN THE TIME DOMAIN

W. H. Huggins

(Air Matériel Command, Cambridge, Mass.)

This paper considers the direct relation between the natural frequencies of a network and the dispersion in time of its response to a unit impulse. It is shown that there exists a complex time plane, which is the inversion of the complex frequency plane, and that the network poles and zeros in this complex time plane are related in a simple, linear manner to the statistical measures such as the delay, variance, skewness, kurtosis, etc., of the impulse response in time. An application to the design of a receiver for rectangular pulses is described, and a criterion for optimum design in the sense of greatest peak-signal-to-noise ratio is suggested.

30. THE DESIGN OF FREQUENCY-COMPENSATING MATCHING SECTIONS

V. Rumsey

(Ohio State University, Columbus, Ohio)

The general problem considered is that of transforming an impedance which changes with frequency to a specified impedance which is, as nearly as possible, constant with frequency. A simple procedure for solving the general problem is established, and formulas for the design of the appropriate matching section are worked out. The formulas give the parameters of the matching section in terms of the impedance of the load at selected frequencies. The technique is mainly applicable to cases where the total bandwidths do not exceed 50 per cent of the center frequency.

31. AMPLIFIER SYNTHESIS THROUGH CONFORMAL TRANSFORMATIONS

D. L. Trautman, Jr., and J. M. Pettit

(Stanford University, Calif.)

This paper is essentially the exposition of a new development in methods of network synthesis. It considers the problem of designing a band-pass amplifier comprised of any of the conventional interstage coupling networks for any prescribed bandwidth. Maximally flat response is considered in detail, but the method is not limited to it.

The notion of complex frequency offers a method of specifying and comparing the several networks, and lays a mathematical foundation suitable for developing a general band-pass to low-pass transformation for this class of circuits. The potential analogue is exploited for its intuitive suggestions in formulating the transformation and in suitably arranging the poles (the design parameters of the over all amplifier). The general transformation is developed in detail so as to give a firm basis for using these same techniques in the solution of general network problems. Specifically, it makes possible the synthesis of staggered double-tuned circuits in the very wide-band situation.

Instruments and Measurements II

Oscillography

32. AN IMPULSE GENERATOR–ELECTRONIC SWITCH FOR VISUAL TESTING OF WIDE-BAND NETWORKS

T. R. Finch

(Bell Telephone Laboratories, Inc., New York, N. Y.)

The impulse generator–electronic switch is an electronic instrument designed primarily for testing radar components such as delay and pulse-forming networks with transmission delay or pulse length in the range 0.05 to 20 microseconds. The instrument may be used to test any network that can be arranged to store a capacitive charge.

Discharge characteristics of a standard network and the network under test are simultaneously displayed in pictorial form and instantly compared. The instrument performance is illustrated with photographs of cro traces for pulse and delay networks, sweep networks, and delay cable. Use in troubleshooting defective networks is also shown in pictorial form. Description of the operation of the instrument is supplemented with photographs of the voltage waveforms of interest.

33. A 50-MC WIDE-BAND OSCILLOSCOPE

A. M. Levine and M. Hoberman

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A wide-band oscilloscope capable of displaying waveforms having components extending from a few cps up to 50 Mc is described. This oscilloscope has been designed for general use, including high-quality television work, pulse modulation, and transient studies.

The features of this oscilloscope include a 50-Mc wide-band vertical amplifier with 50-db gain capable of full screen deflection, and a horizontal amplifier with 10-Mc bandwidth suitable for sweep-voltage co-ordination with the wide-band vertical deflection. Special sweeping and timing devices are also available. The mechanical design has been specially integrated with the general use of this type of instrument.

General theoretical considerations covering the design, a full description of the equipment, and performance results are given.

34. A TIMING-MARKER GENERATOR OF HIGH PRECISION

R. C. Palmer

(Allen B. Du Mont Laboratories, Inc., Clifton, N. J.)

A circuit is described whereby a delay line is used to control the interval between repeated pulses used as 1-microsecond timing markers. The pulse train is started by an input gate and extends for the duration of the gate. An initial pulse is introduced into a delay line, then completely reshaped so that the delayed pulse is of the same amplitude and form as the initial one. The delayed pulse is then introduced into the line, resulting in the establishment of a train of identical pulses with a spacing controlled to a large extent by the delay time of the line. Through regulation of the operating potentials, accuracy in timing of the order of ± 0.01 per cent may be achieved. Limitations of the accuracy of the markers is discussed, as is the extension of the technique to intervals of time not readily obtained by means of delay lines.

35. THE EVALUATION OF SPECIFICATIONS FOR CATHODE-RAY OSCILLOGRAPHS

P. S. Christaldi

(Allen B. Du Mont Laboratories, Inc., Clifton, N. J.)

The selection from published specifications on cathode-ray oscillographs of the equipment best suited for particular requirements is often hampered by inadequate or nonexistent information. The types of specifications usually published are reviewed and analyzed for adequacy. Some of the characteristics on which inadequate or no information is given are reviewed, and proposals are made for additional ratings.

Since the performance of a cathode ray oscillograph depends on the co-ordination of many characteristics, undue emphasis on a particular rating may be misleading. The interrelation of the ratings on characteristics such as amplifier bandwidth, time base range, maximum visual and photographic writing rates, etc., is considered in terms of self consistency of a design.

36. PHOTOGRAPHIC TECHNIQUES IN CATHODE-RAY OSCILLOGRAPHY

C. Berkley and H. Mansberg
(Allen B. Du Mont Laboratories, Inc., Clifton, N. J.)

The oscillographer may be adequately equipped to deal with the electronic phases of cathode ray tube recording but relatively unfamiliar with the basic photographic techniques. Certain problems peculiar to crt photography recur in many applications. Use of the techniques described for obviating them will result in considerable saving in development time and improvement in the quality of the recordings.

The problems discussed include:

1. Relative merits of single frame versus continuous motion recordings for recurrent, transient, and drifting phenomena.

2. Suppression of undesired screen illumination from external and internal sources.

3. Those affecting the choice of the oscillograph, the camera, the recording medium, and processing techniques.

4. Calibration and identification of recordings.

5. Image stability.

6. Interpretation of records.

Proper application of the techniques described will result in considerable improvement in the use of oscillographs as measuring instruments.

Electronic Computers

37. A DYNAMICALLY REGENERATED MEMORY TUBE

J. P. Eckert, Jr., H. Lukoff, and G. Smoliar
(Eckert-Mauchly Computer Corporation, Philadelphia, Pa.)

The necessary development work on a new high-speed memory system suitable for use in electronic digital computers and their associated apparatus has been completed. The work on this system was begun at the Moore School, University of Pennsylvania, and has been continued at the Eckert-Mauchly Computer Corporation until the present time. F. C. Williams of Manchester University has headed a similar project in England, and a certain amount of parallel work has taken place in both countries on the same subject. Both projects had as their final goal the use of a standard or nearly standard cathode-ray tube to provide storage of a large number of binary digits at relatively low cost per digit, and more particularly with access time of not more than a few microseconds. Basically, memory is provided by charged areas which lie on the inner surface of the phosphor of a standard cathode-ray tube. These areas are placed on the surface by the electron beam and a system is provided for using two differently shaped areas, one representing a 0 in the binary system and the other representing a 1. Replacement of one by the other, using the electron beam, causes a signal to be induced into a plate near the face of the tube. The shape of the induced signal can be used to distinguish between an area representing a 1 and an area representing a 0. Reading the areas at random in this fashion would cause loss of the original pattern. Areas can be made indistinguishable by the action of the beam even on a spot adjacent to those being operated on and also by leakage of the charge due to the fact that the phosphor is not a perfect insulator. These undesirable losses are prevented by a systematic reading and writing which operates faster than these deteriorating effects. The general method is similar to the regeneration in a mercury digital memory. A great advantage of this system is the ability to move from one part of the memory to any other at high speed. A second advantage is the low cost and easy maintenance allowed by the use of completely standard high production electronic components with no special tubes. At present, sufficient work and tests have been completed to make possible the detailed analysis of a computer based upon such a memory system. In parallel with this analysis, models of the memory system are being set up to further assess the reliability and practicality of such a computing system.

38. AN ELECTRONIC DIFFERENTIAL ANALYZER

Alan B. Macnee
(Massachusetts Institute of Technology, Cambridge, Mass.)

An electronic differential analyzer, capable of solving ordinary differential equations of orders through the sixth, both linear and nonlinear, will be described. This analyzer has a high speed of operation and is extremely flexible with regard to equation parameters and initial conditions. This flexibility permits rapid investigation of wide ranges of equation solutions with regard to periodicity, instability, and discontinuities. Two new computing elements, an electronic function generator and an electronic multiplier, are employed. A number of representative differential equations of the linear and nonlinear types have been solved. Comparison of observed and calculated solutions reveals an accuracy of from 1 to 5 per cent, while the repeatability of the solutions ranges from 0.002 to 0.1 per cent. An analysis of the errors introduced into the differential-equation solutions by the frequency limitations of the computing elements has been made, and the results of this analysis have been verified experimentally.

39. AN ANALOGUE COMPUTER FOR THE SOLUTION OF LINEAR SIMULTANEOUS EQUATIONS

Robert M. Walker
(Watson Scientific Computing Laboratory, New York, N. Y.)

Linear simultaneous equations occur frequently in science and in engineering. Their solution by numerical methods is straightforward, but the amount of work required increases rapidly with the number of unknowns. A device is described for the solution of systems of linear simultaneous equations with not more than twelve unknowns. It is an electrical analogue computer which accepts the problem information in digital form from a set of punched cards. This facilitates the preparation, checking, and insertion of the input data, and greatly reduces some of the usual liabilities of an analogue device. No special preparation of the problem is required, other than a simple one of scaling the coefficients. Solutions of well determined problems are easily and rapidly attained and may be refined to any desired accuracy by a simple iteration procedure.

40. THE ELECTRONIC ISOGRAPH FOR A RAPID ANALOGUE SOLUTION OF ALGEBRAIC EQUATIONS

Byron O. Marshall, Jr.
(Air Matériel Command, Cambridge, Mass.)

An electronic analogue computer is described with which the whole complex plane may be investigated for roots of up to 10th degree polynomials in a matter of minutes. The roots are given to three places, and the character of the roots is determined. In addition, the value of the polynomial is given for all values of the complex variable. Although the input is limited to real coefficients, practically any polynomial of any degree may be solved by this method. The isograph may be used alone if engineering accuracy is desired, or as an adjunct to a large-scale computing machine. It is believed that the isograph will prove extremely valuable in servomechanism and other fields where the problem may be represented as a polynomial, and that, in general, theoretical analyses of engineering problems will be furthered by its use.

41. A PARAMETRIC ELECTRONIC COMPUTER

Charles J. Hirsch
(Hazeltine Electronics Corporation, Little Neck, L. I., N. Y.)

A novel form of computer is described that operates on the principle of an alignment chart wherein data voltages are aligned in time in the same manner that data quantities are aligned in distance on a slide rule. Since the "time scales" of this electronic slide rule may be calibrated according to any function of time which can be electrically realized, a large variety of operations can be performed. If x and y are known functions of time and n is a known constant, the computer can perform the following operations: x^n; xy; x/y; $x+y$; $\sin x$; $\sin nx$; $\sin^{-1}$; $\cos x$; $\cos nx$; $\cos^{-1} x$; $\log x$; $\int x dt$; $\int x dy$; dx/dy. As the operations can be completed in a matter of milliseconds or less, they can be repetitive and performed on variable parameters. Actual circuits which perform the operations listed above are described. Just as slide rules can be made to have any desired accuracy if they are long enough, this computer can be made as accurate as desired if high enough voltages and rapid enough samplings are taken.

SYMPOSIUM
Electronic Computing Machines

1. Results of Tests on the Binac

J. W. Mauchly
(Eckert-Mauchly Computer Corporation, Philadelphia, Pa.)

2. The Mark III Computer

H. H. Aiken
(Harvard University, Cambridge, Mass.)

3. The IBM Type 604 Electronic Calculator

Ralph Palmer
(International Business Machines Corporation, New York, N. Y.)

4. Electrostatic Memory for a Binary Computer

F. C. Williams
(Manchester University, Manchester, England)

5. Counting Computers

G. R. Stibitz
(Consultant, Burlington, Vt.)

6. Programming a Computer for Playing Chess

Claude E. Shannon
(Bell Telephone Laboratories, Inc., New York, N. Y.)

Last year's symposium on "Advances Significant to Electronics" included one paper on electronic computers and two others on somewhat related subjects. The enthusiasm with which these papers were received, and the discussions provoked by them and others on the same subject presented during the regular technical sessions, demonstrated such a widespread interest in high speed computers that a symposium on recent advances in the art has been scheduled for this convention. The speakers include several of the pioneers and acknowledged leaders in this relatively new but highly important branch of electronics.

During the past year some computers have been completed and operating tests carried to the point where new evaluations of their capabilities and limitations have become possible. Three of the speakers at this symposium will discuss the construction and operating characteristics of new computers. These results should be of great value in pointing the way toward future development. As an example of the tremendous advances that have already taken place in this field, it might be pointed out that the BINAC, which forms the subject of one of the papers, achieves greater versatility and memory capacity than its logical predecessor, the ENIAC, but at the same time uses less than 5 per cent of the number of tubes.

The BINAC's operating characteristics will be discussed by Dr. Mauchly. This device is of particular interest because, except for input and output equipment, it is entirely electronic, and because it represents the practical application of elements discussed at last year's convention, such as a mercury-delay-line memory and magnetic tape input and output equipment.

Dr. Aiken of Harvard will discuss the Mark III Computer, latest in the series of large-scale computers developed at that institution. The Mark III has been designed for greater speed and reliability, more flexible memory facilities, and greater ease of preparation of input data than were found in the earlier computers.

The third in this series of papers on complete systems will be presented by Mr. Palmer and will deal with the type 604 electronic calculator built by International Business Machines Corporation. This computer combines an electronic arithmetic element, including a 13 digit electronic counter, with punched-card input and output equipment and additional mechanical storage registers with the possibility of carrying out automatically a "program" with as many as 20 arithmetic operations.

Since the basis of any successful large-scale computer is a memory having adequate capacity and means for rapid introduction and abstraction of data, the paper on an electrostatic memory for a binary computer by Dr. Williams should be of particular interest. Dr. Williams has carried out extensive research in electrostatic memories, which probably show the greatest promise of any of the basic types thus far proposed for computers, since they combine the high reading and writing speed of the delay line type of memory with a very short access time.

The paper to be presented by Dr Stibitz while not concerned directly with electronic computers, is nevertheless of great interest in view of the fact that analogue or continuous variable computers have received so little attention since the development of high speed digital or discrete variable types. Dr. Stibitz proposes a new type of computer which combines the more accurate elements of the familiar differential analyzer such as gears and differentials with a new type of "function unit" resulting in a computer having the simplicity and low cost of an analogue computer and the high accuracy of the digital type.

The remaining paper will appeal not only to those interested in computers as such but to all those attending the convention, since it deals with the application of computer techniques to a problem familiar to all. Dr Shannon, whose paper on "Information Theory" was received with much interest at last year's symposium will discuss the programming of a chess game on a computer. While the possibility of such applications was recognized early in the course of large scale computer development and has been mentioned from time to time in the press and in public discussion, this probably represents the first serious attempt to analyze the programming of such an operation.

Wave Propagation I

42. VHF TELEVISION—PROPAGATION ASPECTS

Edward W. Allen, Jr.
(Federal Communications Commission Washington, D. C.)

This paper comprises a brief report of the Engineering Conferences on this subject held by the Federal Communications Commission in November and December, 1948 and of the subsequent work of the ad hoc committee in reducing our present knowledge of vhf propagation to workable standards for the allocation of vhf television stations.

43. PROPAGATION VARIATIONS AT VHF AND UHF

Kenneth Bullington
(Bell Telephone Laboratories, Inc., New York, N. Y.)

The variations of received signal with location (shadow losses) and with time (fading) greatly affect both the usable service area and the required geographical separation between co-channel stations. An empirical method is given for estimating the magnitude of these variations at vhf and uhf. These data indicate that the required separation between co-channel stations is from 3 to 10 times the average radius of the usable coverage area, and depend on the type of service and on the degree of reliability required. The application of this method is illustrated by examples in the mobile radiotelephone field.

44. PROPAGATION TESTS AT UHF

Joseph Fisher
(Philco Corporation, Philadelphia, Pa.)

This paper reports the results of broadcast propagation over varied terrain at both 500 Mc and 3,300 Mc. While the principal interest was in television broadcast coverage in the 500 Mc band, data was also taken at 3,300 Mc, using pulse transmission in order that certain trends apparent in the 500-Mc region might be carried to sufficient extremes to permit reliable evaluation.

The 500-Mc tests were made in the Washington, D.C. area and utilized the signal transmitted by the experimental RCA television station. The mobile television receiving equipment designed by Philco was mounted in a station wagon to facilitate quick setup. There were provisions to measure field strength, and the picture quality was analyzed by competent observers.

The 3,300-Mc tests were made in the Philadelphia area. The field equipment included means for measuring field strength and photographically recording the attenuation and time delay of multipath signals.

45. A TEST OF 450-MC URBAN-AREA TRANSMISSION TO A MOBILE RECEIVER

A. J. Aikens and L. Y. Lacy
(Bell Telephone Laboratories, Inc., New York, N. Y.)

Measurements were made of transmission at 450 Mc in New York City using frequency modulation. Comparison was made with transmission at 150 Mc using the same modulation. Effective radiated powers were about equal. A-B tests (direct comparisons) were made with the receivers installed in a moving automobile. The transmitter and receiver used at 450 Mc were developed especially for the job. The receivers used at the two frequencies had substantially the same noise figures. The tests permitted estimates of the relative magnitudes of the shadow losses at the two frequencies and included measurements of rf noise and signal-to-noise ratios. Subjective tests of circuit merit comparing the two frequencies were made by a number of observers.

46. ECHOES IN TRANSMISSION AT 450 MC FROM LAND-TO-CAR RADIO UNITS

W. R. Young and L. Y. Lacy
(Bell Telephone Laboratories, Inc., New York, N. Y.)

By the use of short pulses of 450-Mc carrier, the echoes which appear in transmitting from a land station to a moving car in New York City have been investigated. The results, which show the multiple-path nature of transmission, are presented in sample and in statistical form. They are of use in considering the possibilities of systems employing a wide modulation band.

Passive Networks II

Analysis

47. IMPEDANCE CURVES FOR TWO-TERMINAL NETWORKS

E. L. Michaels
(Northwestern University, Evanston, Ill.)

This paper presents a thorough and systematic investigation of the characteristic curves of the complex impedance of all possible combinations of one-, two-, and three-element linear networks at all frequencies from zero to infinity. Use of the properties of inverse, reciprocal, and equivalent networks is made, so that a single set of curves applies to from two to six different networks.

The curves of complex impedance, plotted in the complex plane, are of particular interest because they give the resistive and reactive components of the complex impedance, as well as the impedance magnitude and phase shift, at all frequencies on just one sheet of graph paper.

Of the large number of special applications possible, the most interesting are those for the design of low- and high-frequency-compensation networks for wide-band amplifiers.

The information presented by these curves is extremely useful in determining the frequency ranges over which certain elements may be neglected in the final calculations.

48. AN ANALYSIS OF TRIPLE-TUNED COUPLED CIRCUITS

N. W. Mather
(Princeton University, Princeton, N. J.)

An analysis is given of triple-tuned coupled circuits for high-Q cases in which the couplings between circuits are kept small in both the synchronously tuned and asynchronously tuned cases, which are investigated in detail. Contour plots of gain versus band-width product are given.

The results for the transitional and triple-peak response cases are compared with the double-tuned coupled circuit. The advantages of the triple-tuned circuit are shown to be its more uniform frequency and phase characteristics in the pass band and its 50 per cent better sideband selectivity.

49. THE BRIDGED PARALLEL-TEE NETWORK FOR SUPPRESSED-CARRIER SERVO SYSEMS

C. F. White
(Naval Research Laboratory, Washington, D. C.)

The bridged parallel-tee network is the ac servo system analogue of the shunted high-pass network in the dc servo system. In suppressed-carrier servo systems a symmetrical upper- and lower-sideband transmission characteristic is highly desirable. The parallel-tee resistance-capacitance null-type network with a bridging and a retuning resistance exhibits the desired characteristic when operated between suitably selected generator and load resistances. The network output is resolved into a main component and a minimized quadrature component, and is shown by families of curves displaying amplitude and phase versus frequency. Carrier-frequency shifts are found to have only small effect. All network parameter data required by the servo system designer are summarized on a single curve.

50. TRANSIENT RESPONSE OF LINEAR NETWORKS WITH AMPLITUDE DISTORTION

M. J. Di Toro
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The transient response of linear networks having phase distortion is characterized by an excessive amount of "ringing" effects. When this is corrected so that a phase response linear with frequency prevails, ringing effect may still persist and give interference between messages adjacent in time. An example is the so-called "ideal" (and unrealizable) low-pass filter having linear phase and flat amplitude response up to the cutoff frequency and no response beyond. The impulse transient response of this circuit decays at a rate not larger than inversely with time. The resulting interference due to overlapping messages decreases the transmitting capacity of the channel. The extent of ringing is determined by the shape of the amplitude-response characteristics. A number of cases of this effect are presented. A convenient way to analyze this problem is to introduce the use of self-reciprocal transforms, such as the parabolic cylinder functions.

51. SPECTRUM ANALYSIS OF TRANSIENT-RESPONSE CURVES

H. A. Samulon
(General Electric Co., Syracuse, N. Y.)

A new method of spectrum analysis of measured transient-response curves is proposed, based on the work of W. R. Bennett, C. E. Shannon, and others who have studied the synthesis of curves with limited frequency spectra and known cutoff frequency. Under these restrictions, which are fulfilled in many technical applications, this method yields higher accuracy; at the same time, it gives a criterion for the necessary number of sample points of the curve. Comparison is made with another method previously proposed by A.V. Bedford and G.L. Fredendall.

Components and Materials

52. SUBMINIATURIZATION OF IF AMPLIFIERS

Gustave Shapiro and Robert L. Henry
(National Bureau of Standards, Washington, D. C.)

A new approach to the miniaturization of electronics equipment is described wherein some concepts and techniques new to the equipment design engineer are used. Circuitry construction where tubes occupy a large percentage of the total volume involves problems of cooling which must be met by special high-temperature materials in the components. Technical information and techniques required to use these materials is discussed along with the necessity of and methods for controlling the size and shape of the components and assemblies. These techniques are illustrated in the design of typical if amplifiers. The models described have bandwidths of 10 Mc, centered at 60 Mc, with gains of 95 db. Liberal use is made of miniature components in one assembly and printed-circuit techniques in another. A high degree of stability is achieved with physical volumes per stage approaching ½ cubic inch. They may be readily mass-produced despite their small size.

53. NEW APPLICATIONS OF A FOUR-TERMINAL TITANATE CAPACITOR

Adriano A. Pascucci
(Radio Hispano Suize, S. A., Barcelona, Spain)

Results are given for an investigation of the properties of a Ba-Sr Ti O_3 four-terminal capacitor. The capacitance and power factor observed between two terminals of the capacitor are shown to be dependent upon a perpendicular electric field applied by means of the other pair of terminals. Some results are presented for the use of these titanate capacitors as variable reactances in FM modulators, and particularly as nonlinear elements in a new type of aperiodic frequency doubler and amplifier.

54. FREQUENCY-CONTROL UNITS

August E. Miller
(August E. Miller Co., North Bergen, N. J.)

The frequency-control unit discussed in this paper comprises an oscillator and quartz crystal contained within an oven. The unit is factory-adjusted to supply a specific output and frequency, accurate and stable to within 0.0001 per cent over an ambient temperature range of $-55°$ to $+80°$ C. These units become especially advantageous in the ultra- and super-high frequencies. Because of inadequately accurate frequency control of both transmitting and receiving equipment, the present systems require bandwidths far in excess of that necessary for the transmission of intelligence. As an illustration, the bandwidths necessary to transmit a 3,000-cps voice channel now spread over several megacycles, but with more accurate frequency control the bandwidth can be reduced to a value nearer that of the intelligence. In so doing, the transmitted power necessary for operation is reduced in proportion to the bandwidth reduction.

Pictures, data, and reasons for using frequency-control units will be presented.

55. THE TYPE 5811 AND TYPE 5807 TUBES—THE SMALLEST COMMERCIAL PENTODE AMPLIFIERS

L. Grant Hector and H. R. Jacobus
(Sonotone Corporation, Elmsford, N. Y.)

The Sonotone type 5811 and type 5807

sub-subminiature pentode amplifier tubes were designed for the new small-size hearing-aid equipment. They are the smallest electronic tubes manufactured commercially. They are less than one-third of an inch in diameter and have greater gain and power than the larger subminiature tubes they replace. The mechanical design that permits the assembly of an electrode structure having high electrical gain in a T-2 envelope is discussed and compared to the subminiature and miniature tubes of similar characteristics. Circuit applications for both socket type and soldered-lead type of equipment are shown.

56. CONDUCTIVE PLASTIC MATERIALS

Myron A. Coler
(Markite Company, New York, N. Y.)
F. Robert Barnet, Albert Lightbody, and H. A. Perry
(Naval Ordnance Laboratory, Silver Spring, Md.)

The properties of a new class of plastic materials are presented by means of data secured on representative types. The outstanding characteristic of these materials resides in the fact that they have substantial and predeterminable electrical conductivities, and yet possess the general mechanical and fabrication properties of ordinary plastics. Thermosetting, thermoplastic, and elastomeric variants have been produced. Many of the materials exhibit enhanced thermal conductivity and heat stability. Data have been obtained showing the effect on resistivity of changes in major variables such as frequency, current density, temperature, and continuous loading. Practical applications in the field of electrical engineering devices are indicated.

Nucleonic Instrumentation

57. THE RF SYSTEM FOR THE UNIVERSITY OF ROCHESTER 130-INCH SYNCHROCYCLOTRON

Winfield W. Salisbury
(Collins Radio Co., Cedar Rapids, Iowa)

The design and construction of the rf system of the University of Rochester 130 inch synchrocyclotron are described. This cyclotron produces protons having energies of approximately 270 Mev. The rf system includes an oscillator, nominally rated at 160 kw. The principal frequency determining element is the cyclotron dee itself, which forms a part of the oscillating circuit. The frequency of the system is modulated between the limits of 25.5 Mc and 18.5 Mc by a rotating capacitor connected from the dee to the ground.

58. ELECTROMETER TUBES AND CIRCUITS

H. F. Starke
(Raytheon Manufacturing Co., Newton, Mass.)

Special characteristics of electrometer tubes used in radiation-detecting and measuring devices will be discussed. Sources of spurious grid current will be pointed out, while the form of the variation of grid current with operating potential and filament temperature will be indicated. Circuits to take proper advantage of tube characteristics will be described.

59. DESIGN OF A G-M COUNTER TUBE FOR HIGH COUNTING RATES

W. W. Managan
(The Victoreen Instrument Co., Cleveland, Ohio)

The problems encountered in the design of Geiger-Mueller counter tubes for high counting rates differ from those encountered at low rates near background. The major problems are presented and their solutions discussed in terms of a novel counter tube, the 1B89, capable of counting at rates greater than 10,000 counts per second. The 1B89 is used in an X-ray diffraction spectrometer.

60. HIGH-VOLTAGE SUPPLIES FOR RADIATION-MEASURING EQUIPMENT

Richard Weissman and Stewart Fox
(Nuclear Instrument and Chemical Corporation, Chicago, Ill.)

Two new high-voltage-source circuits are described which provide the stability, control range, and small size often required by radiation-measuring instruments. They use small receiving tubes operated at less than rated voltage, regardless of the voltage output of the supply. One of these circuits uses a saturable reactor, while the other, which provides control over an unusually large range, uses a specially developed dc controlled "diverted flux" transformer.

A relatively efficient battery operated portable high voltage supply, providing regulated output, is also described. This supply uses a moving coil vibrator system and high-voltage subminiature VR-type regulator tubes also designed specifically for the unit described.

61. PROPORTIONAL-COUNTER EQUIPMENT FOR BETA DETECTION

William Bernstein
(Brookhaven National Laboratory, Upton, L. I., N. Y.)

Methane-flow proportional counters are now in use for the detection of both soft and energetic beta particles. Counters of this type offer advantages over conventional end-window or thin-wall Geiger-Muller tubes with respect to window absorption, geometry, background, spurious counts, and resolution time. They are also more easily adapted to quantitative analytical chemistry techniques.

Experimental results are submitted to illustrate the performance of these counters under general operating conditions. A detailed description of the counter construction is given. The requirements of the amplifier, pulse-height selector, scaler, and power supply are discussed in detail, and some circuits in current use are described.

Some applications of proportional counters to specific beta counting problems are considered briefly.

62. INDUSTRIAL THICKNESS GAUGES EMPLOYING RADIOISOTOPES

J. R. Carlin
(Tracerlab, Inc., Boston, Mass.)

Two general types of industrial thickness gauges are described. One employs the principle of absorption of radiation by matter. The second utilizes the backscattering effect of beta particles by matter and provides a method of measuring thickness of one material superimposed on another of different atomic number. Both types can be adapted to measure continuously the thickness of moving sheets of material or individual samples, without physically contacting the specimen. Ionization chambers serve as the detecting units.

Theoretical expectations and empirical findings concerning sensitivity, accuracy, and scale linearity are presented, together with a discussion of ion-chamber design criteria.

SYMPOSIUM
Nuclear Science

1. **The Fundamental Particles**

Donald J. Hughes
(Argonne National Laboratory, Chicago, Ill.)

2. **The Detection and Measurement of Nuclear Radiation**

Howard L. Andrews
(National Institute of Health, Bethesda, Md.)

3. **The Effects of Ionizing Radiation on Tissue**

James P. Cooney
(Atomic Energy Commission, Washington, D. C.)

4. **The Application of Nuclear Radiation to Industry**

John R. Menke
(Nuclear Development Associates, Washington, D.C.)

Engineers in the electronics industry are assuming an ever-increasing responsibility in the development and production of equipments related to the science of nucleonics. Their participation is still in its early stages, and is still limited to a small fraction of the tasks which invite their attention. So that those in the electronics industry may see a broader horizon to the possibilities of work in the atomic energy field, The Institute of Radio Engineers has organized a symposium, which, many may say, is in advance of its proper time.

For engineers and scientists associated with electronics, exposure to a discussion of nuclear particles on a broad base may seem a little premature, but only if the complete pictures is presented can we hope to enlist the full co-operation of their services. Dr. Hughes, one of the foremost of our young American physicists, will discuss the fundamental particles and present the basis underlying the interrelation of nuclear particles. Following this, Dr. Andrews and Col. Cooney, both participants in five atomic bomb tests, will set forth an explanation of how these fundamental particles are measured, and what their effect is on the human organism.

Having thus described the nuclear particles as to what they are, how they are measured, and what they do, the symposium concludes with Mr. Menke's survey of the practical applications to which these can be put in industry and research.

Television I

63. A UNIDIRECTIONAL REVERSIBLE-BEAM ANTENNA FOR TWELVE-CHANNEL RECEPTION OF TELEVISION SIGNALS

O. M. Woodward, Jr.
(RCA Laboratories, Princeton, N. J.)

The present commercial channels assigned for television range from 54 to 88 Mc and from 174 to 216 Mc. This paper describes a unidirectional receiving antenna which is effective over these channels without adjustment.

The array is made up of dipole elements which themselves maintain desirable characteristics over the entire television range. These elements are united by a simple transmission-line network to yield a directive pattern.

The antenna maintains a high front-to-back ratio over all twelve channels, and is particularly useful in fringe areas where it is necessary to reduce co-channel interference.

The directional beam is reversed on any channel by a simple switch which transposes a single transmission line.

64. A METHOD OF MULTIPLE OPERATION OF TRANSMITTER TUBES PARTICULARLY ADAPTED FOR TELEVISION TRANSMISSION IN THE UHF BAND

George H. Brown, Wendell C. Morrison, W. L. Behrend, and J. G. Reddeck
(RCA Laboratories, Princeton, N. J.)

A combining network has been developed which allows two transmitter tubes to be operated simultaneously into a common load without interaction between tubes and without reduction in bandwidth. A theoretical analysis is presented which shows that the necessary balancing adjustments are not critical. A pair of tubes and a combining network may then be considered as a unit, and again combined with other units to provide a transmitter with a large number of tubes in multiple operation.

A complete television transmitter with a carrier frequency of 850 Mc has been developed, using four tubes in multiple. The physical arrangement and the operating characteristics of this transmitter are discussed.

65. TRANSIENT-RESPONSE TESTS ON THE WPTZ TELEVISION TRANSMITTER

R. C. Moore
(Philco Corporation, Philadelphia, Pa.)

In order to determine the performance of the new WPTZ transmitter, a program of transient tests was undertaken utilizing the procedures recommended by a subcommittee of the RMA Television Systems Committee. The equipment consists of a 100-kc square-wave generator with steep wavefronts, a wide-band oscilloscope, and a VSB monitor-receiver of prescribed characteristics. Following a determination of the monitor transient response, the transient response of the transmitter and monitor together is found. From these results the behavior of the transmitter is deduced. The distortions measured at WPTZ due both to the rf and vf circuits will be discussed. Compensation by predistortion with vf networks will be described.

66. TELEVISION TRANSMITTER CARRIER SYNCHRONIZATION

R. D. Kell
(RCA Laboratories, Princeton, N. J.)

Laboratory tests have indicated that a substantial reduction in co-channel interference between television stations can be obtained if the carrier frequencies are synchronized. Actual tests using television stations in New York and Washington confirm the laboratory observations.

Signals from New York and Washington are compared in a phase discriminator at the output of two receivers located in Princeton. The information regarding relationship of the two carriers is carried as frequency modulation of a 1,000-cps tone by telephone line to New York. The frequency shift of this tone is utilized to change the frequency and phase of the New York carrier to maintain a fixed phase relationship between the New York and Washington carriers as observed at Princeton.

A description of the synchronizing equipment used and the test results obtained to date will be given.

67. TELEVISION BY PULSE-CODE MODULATION

W. M. Goodall
(Bell Telephone Laboratories, Inc., Deal, N. J.)

Fundamental experiments to reveal the effects on television of time and amplitude quantization have been made. The experiment consisted in step-sampling the television signal at various rates up to ten million times per second, the coding of each level by means of a binary code, the reconversion of the code signals to quantized television, the display of the resulting signal on standard television monitoring devices, and the comparison with an unquantized signal. The relationship of the number of digits needed to the noise level, the picture quality, and other factors is discussed. Studies with this experimental system gave surprisingly good results with three or four digits. No conclusions are drawn as to the number that would be required in practice.

SYMPOSIUM

Radio Aids to Navigation

1. The Radio Technical Commission for Aeronautics—Its Program and Influence

J. H. Dellinger
(Radio Technical Commission for Aeronautics, Washington, D. C.)

2. Frequency Allocations to the Aeronautical Services above 400 Mc

V. I. Weihe
(Air Transport Association of America, Washington, D. C.)

3. Experimental Multiplexing of Functions in the 960–1660-Mc Frequency Spectrum—Its Influence on Weight and Complexity of Equipment

P. C. Sandretto and R. I. Colin
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

4. The Philosophy and Equivalence Aspects of Long-Range Radio Navigation Systems

M. K. Goldstein
(Office of Naval Research, Washington, D. C.)

5. The Future in Approach and Landing Systems

Harry Davis
(Watson Laboratories, Red Bank, N. J.)

The RTCA is an organization for the co-ordination of effort in aviation radio and electronics. It operates through the preparation of recommendations formulated by temporary technical committees with specific directives. It includes in its membership all interested bodies, both government and nongovernment. These members participate in its support, both technical and financial, and give effect to its recommendations. As an example of its work, it formulated and continues to sponsor the 15-year program of air traffic control and navigation aids whose implementation is now beginning. This work will be described by Dr. Dellinger.

Frequency allocations to the aeronautical services above 400 Mc will be discussed by Mr. Weihe. During 1944 and 1945, aviation electronic engineers, as a result of wartime advances, proposed several very noteworthy integrated systems for the navigation and traffic-control phases of aeronautical operations.

During 1946 and 1947 certain development work was initiated, and aviation was confronted with a need for evaluating the systems to find which system or system elements had the greatest merit. Early in 1947, it was realized that additional radio frequency spectrum was necessary in order to conduct the necessary development work and to provide space for installation and utilization of the adopted equipments. Frequency service allocations for these services were made by the International Telecommunications Union at Atlantic City during the summer of 1947.

Aviation is now confronted with a need for establishing a complete plan for channel allocations within the available radio frequency spectrum. Several major elements of the complete plan have already been established and considerable progress toward the ultimate plan can be anticipated for the next two or three years.

The RTCA plan contemplates multiplexing of various navigational functions on each of a series of channels. Theoretical studies have indicated that sufficient frequency space is available for this purpose, but studies are lacking on the equally important matter of methods of separating the various signals.

The present paper by Messrs Sandretto and Colin describes certain experimental equipment already built and whose initial tests substantiate the general practicality of the RTCA plan. On a 51-channel narrow-band DME there is multiplexed an azimuth service in conjunction with a ground radar, and a relay radar service for airborne pictorial displays.

Questions of interference-free mutual operation of the different functions, and duty-cycle requirements, are discussed. Novel features include a high-speed combination primary and secondary radar antenna, a two-color combination radar indicator, and altitude coding. The possibility of obtaining the full degree of multiplexing envisaged by the RTCA is discussed in the light of experience gained with this equipment.

Dr. Goldstein's paper will establish a general systems convergence or equivalence principle for all long-range radio navigation systems. It shows that there is an upper limit on the theoretical range and accuracy of such systems. This upper limit is principally determined by the parameters of:

1. Frequency. 2. Average power. 3. Azimuth magnification index (AZMI) (the ratio of electrical azimuth intelligence per degree of space azimuth). 4. Instrumental time constant (ITIC) (the permissible instrumental reading time).

It is further shown that different existing and proposed systems differ essentially in "cleverness factor"; i.e., the efficiency with which the permissible azimuth-magnification index and permissible instrumental time-constant factors are utilized. The azimuth-magnification cleverness factor gives a means for increasing the effective resolving power of the system and a means for reducing the effects of some of the ionosphere and instrumental errors. The instrumental-time-constant cleverness factor gives a means for achieving effective increases in the signal-to-noise ratio. The latter increase is reflected in increased range and accuracy performance, where white atmospheric noise would otherwise limit such navigational performance.

Present research and development activities in approach and landing systems for aircraft indicate the type of electronic landing aids which will be employed in the future common system of air navigation. Starting with an analysis of presently used systems, Mr. Davis will present a review of the newer developments with emphasis upon techniques used to eliminate present deficiencies. Included in the discussion is material on automatic ground-controlled-approach radar equipment and fixed beam systems.

Active Circuits I

68. G CURVES AS AN AID IN CIRCUIT DESIGN

KLAUS A. PULLEN
(Ballistic Research Laboratories, Aberdeen Proving Ground, Md.)

The use of G curves for design of electron-vacuum tube circuits offers several distinct advantages over present methods. These include simplification of theoretical approach and a more flexible circuit-design technique. Although basically the method gives only small-signal amplifications, use techniques will be presented which make design for large-signal conditions routine. The basic approach and considerable application data will be given.

69. A DIRECT-COUPLED AMPLIFIER EMPLOYING A CROSS-COUPLED INPUT CIRCUIT

J. N. VAN SCOYOC AND G. F. WARNKE
(Armour Research Foundation, Illinois Institute of Technology, Chicago, Ill.)

A high-gain direct-coupled amplifier employing a newly designed cross-coupled input stage is described. The amplifier has an input impedance of 100 megohms and a gain of 500 to 50,000, variable in 2-db steps between insertion of negative feedback. Either single-ended or push-pull input may be employed, while a low output impedance is obtained from a cathode-follower output stage. The upper half-power frequency limit varies from 50 kc to 150 kc, depending on gain setting. Power requirements are low and miniature tubes with ac-operated heaters are used throughout. A preamplifier is built on the same chassis and can be simply connected to the dc amplifier input through a coupling network with a large time constant. The preamplifier has a maximum gain of 50 and includes a hum-balancing control.

70. ANNULAR CIRCUITS FOR HIGH-POWER MULTIPLE-TUBE GENERATORS AT VHF AND UHF

DONALD H. PREIST
(Eitel-McCullough, Inc., San Bruno, Calif.)

A new type of vhf and uhf circuit or cavity known as the annular circuit is described, and various ways of applying it to multiple-tube generators using negative-grid tubes are discussed. The properties of annular generators are examined, and they are classified with other older types of multiple-tube generators. It is shown that there is no fundamental limit to the number of electron tubes that may be used in such a generator, and that the efficiency versus frequency characteristic is the same as that for a single tube and circuit. The influence of the annular circuit on design of the component tubes is discussed, and a survey of some possible applications in the frequency range 100 to 1,000 Mc is made, with special reference to high band television.

71. CONSIDERATIONS ON ELECTRONIC MULTICOUPLERS

(W. R. AYLWARD, AND E. G. FUBINI
Airborne Instrument Laboratory, Inc., Mineola, L. I., N. Y.)

An electronic multicoupler is, by definition, a device which, when connected to an antenna, can feed rf energy to a multiplicity of receivers without introducing undue sensitivity loss, crosstalk, spurious responses, or overload. The basic criteria which must be made a basis for the design of multicouplers will be discussed. Such factors as bandwidth, receiver noise figure, variations of antenna impedance with frequency, and effects of strong interfering signals will be considered. A new type of multicoupler design which has been used in several units built for frequency bands from 2 to 30 Mc will be briefly described.

72. IMPROVED DEGENERATIVE REGULATORS

Y. P. YU
(North Dakota Agricultural College, Fargo, N. Dak.)

Improved circuits for a direct voltage regulator, an amplitude-stabilized oscillator, and an illumination regulator are described. High stabilities are attained by the use of "constant-current" elements and memory capacitors.

Instruments and Measurements III

73. AN AM BROADCAST STATION MONITOR

H. R. SUMMERHAYES, JR.
(General Electric Co., Schenectady, N. Y.)

Standard AM broadcast stations require continuous monitoring of frequency and percentage modulation, quantities which are ordinarily measured by separate instruments.

A new AM monitor will be described which combines these measuring functions into one instrument. The new instrument features a local oscillator which is synchronized in constant phase relation with the carrier component of the modulated signal from the transmitter. The local oscillator signal, in turn, energizes a quartz-crystal discriminator which indicates deviation from assigned transmitter frequency. The local-oscillator signal is also added to the modulated wave from the transmitter to reduce the effective percentage of modulation, thereby eliminating negative peak clipping in the detector.

74. SPEED OF ELECTRONIC SWITCHING CIRCUITS

E. M. WILLIAMS AND D. F. ALDRICH
(Carnegie Institute of Technology, Pittsburgh, Pa.)

The transition or switching interval in electronic switching circuits such as multivibrators is discussed. Curves will be presented showing typical variation of the transition time with various circuit parameters. Analytic results based on the use of equivalent circuits employing mean values of transconductance and plate resistance are also presented, and means are provided for computing and minimizing the duration of the transition. Triggering delay is briefly considered.

75. A MAGNETOSTRICTIVE DELAY LINE

E. BRADBURD
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A description is given of a supersonic delay line using nickel ribbons as the sound-conducting elements; the magnetostrictive property of nickel is used for coupling into and out of the ribbons.

Smoothly varying delays of 5.3 microseconds per inch of line are obtainable, along with the added convenience of many transmitting and receiving points. The 35-db insertion loss for this delay line is substantially independent of delay.

A novel method is described for reducing the effect of echoes from the ends of the line.

76. AN ELECTROMECHANICAL STRAIN-GAGE MULTIPLIER

C. H. Woods, E. St. George, L. Isenberg, and A. C. Hall
(Massachusetts Institute of Technology, Cambridge, Mass.)

This paper describes a newly developed instrument for multiplying modulated carrier voltages. The basic multiplying element in this instrument is a wire-resistance strain-gage Wheatstone bridge. One of the modulated voltages, the multiplier, is applied to the input terminals of the bridge network and the second modulated voltage, the multiplicand, controls the degree of unbalance of the bridges. The product appears as the modulated output of the bridge network. A circuit arrangement employing a constantly excited feedback bridge effectively applies negative feedback around the active elements in the system and results in an accuracy of 0.1 per cent of full-scale output, and a multiplication time constant of less than two milliseconds.

77. RADAR-CIRCUIT-POWERED X-RAY MOVIE EQUIPMENT FOR OPERATION AT 150 FRAMES PER SECOND

D. C. Dickson, Jr., C. T. Zavales, and L. F. Ehrke
(Westinghouse Electric Corporation, Bloomfield, N. J.)

A high-powered pulse transformer has been developed which delivers 60-ampere current pulses to a special tungsten-filament X-ray tube at 150 kilovolts above ground. The pulse transformer is energized at up to 150 pulses per second by a circuit similar to a radar line-type modulator. The X-ray exposure time for each frame is 10 microseconds, which results in negligible motion blur when recording on film moving continuously at 25 feet per second. Although developed specifically for the study of small solid-fuel rockets, the principle of operation is readily adaptable to other applications. With the existing equipment, penetration is good through one inch of aluminum.

Electronics I

Tube Design and Engineering

78. MICROPHONISM INVESTIGATION

Lester Feinstein
(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

A method of investigation which indicates which elements of the tube are causing microphonism is described. The equipment consists of a vibrating system driven by an amplifier operated from a variable-frequency audio oscillator. Noise is measured on a tube voltmeter and is projected on an oscilloscope. At frequencies where peak noises are observed, a microscope and stroboscope indicate those elements that are vibrating.

An example is given in which a tube is analyzed and the information used to predict the frequencies at which microphonism will occur. Tube structural features that most frequently cause microphonism are cited, and a mathematical analysis of tube structure given which indicates how natural frequencies of the elements can be predetermined by the tube designer.

79. A CRITICAL SURVEY OF METHODS OF MAKING CERAMIC-TO-METAL SEALS AND THEIR USE FOR ELECTRON-TUBE CONSTRUCTION

Roger P. Wellinger
(University of Illinois, Urbana, Ill.)

All of the known methods of making ceramic-to-metal vacuum seals have been used in experimental klystrons. The methods have been thoroughly examined mechanically and electrically and an explanation of the physical-chemical nature of the bond is hypothecated. The methods used are: (1) the German method of metallizing the ceramic by sintering a metal powder onto it; (2) brazing onto the ceramic using a special titanium hydride flux; and (3) a new method of sealing by means of compression between the metal and the ceramic. A comparative study is presented of the advantages of each method used in respect to fabrication, as well as the respective electric properties of the seals at high frequencies.

80. RUGGED TUBES

George W. Baker
(Kip Electronics Corporation, New York, N. Y.)

An electronic tube which can take its place as a piece of mechanical equipment must be designed and tested to withstand an unlimited variety of mechanical shocks, exceeding the range of the current "W" types.

This is accomplished by dividing the shocks into three regions according to the ratio of the natural frequency of the tube element to the frequency of the shock. The most severe shocks are represented by a half-sine-wave pulse of acceleration in the frequency-ratio range of 0 to 0.5; by a sine wave, including the contribution of the starting transient, in the range 0.5 to 5; and above 5 the acceleration is considered as statically applied. In the range 0 to 0.3 the disturbing effect of the shock depends entirely upon the change in velocity during the shock. A tube that withstands these shocks consists of a very rigid structure supported by a single compliant element designed to function as an oscillating system of one degree of freedom.

81. AN IMPROVED METHOD OF TESTING FOR RESIDUAL GAS IN ELECTRON TUBES AND VACUUM SYSTEMS

E. W. Herold
(RCA Laboratories, Princeton, N. J.)

The small amount of residual gas in electron tubes is sometimes of concern to the user and, more often, of considerably greater interest in the laboratory and factory. Existing methods of test employ a negative electrode in the tube to collect ion current or, in a vacuum system, a similar test is made with an ionization gauge. The dc current to the negative electrode is used as a measure of gas pressure. Unfortunately, leakage and other stray currents are of the same nature and may even be greater in magnitude, so that ion currents well below the stray-current level cannot be determined. A new method is described in the present paper in which the desired ion current can be separated from undesired stray currents. The method is applicable to both conventional and special types of electron tubes.

82. DESIGN FACTORS, PROCESSES, AND MATERIALS FOR THE ENVELOPE OF A METAL KINESCOPE

R. D. Faulkner and J. C. Turnbull
(RCA Victor, Lancaster, Pa.)

The development of the envelope assembly of the metal kinescope was prompted by the need to achieve in large cathode-ray tubes (1) lighter weight, (2) greater manufacturing flexibility, and (3) improvement in optical quality. Such an envelope assembly presented many new problems which had to be investigated, including those of designing a suitable face plate and of making large metal-to-glass seals. It was found that if the seal shape and the stress distribution at the seal were carefully controlled, the forming of seals with considerably higher stress than previously considered possible could be performed by mass-production manufacturing methods.

The use of a metal cone permits a considerably thinner face plate. In fact, low-cost window glass 3/16-inch thick formed to a radius of 27 inches was found to be satisfactory. A chrome-iron cone is used. It has excellent sealing properties and the proper expansion characteristics. The 16-inch metal kinescope is conservatively designed with respect to pressure-strength requirements.

Television II

83. THE MEASUREMENT OF THE MODULATION DEPTH OF TELEVISION SIGNALS

R. P. Burr
(Hazeltine Electronics Corporation, Little Neck, L. I., N. Y.)

For economic reasons the tendency in television broadcasting has been towards close control of the performance of transmitters with the intent of minimizing the design complications and cost of receiving equipment.

The purpose of this paper is to discuss the measurement of the modulation depth resulting from the application of a video signal to a television transmitter. The paper considers three means for modulation-depth measurement, two of which are electronic, and the third mechanical. Any one of the methods requires relatively little apparatus and may be applied to a local source or to a distant transmitter. Several photographs are included showing the patterns resulting from various adjustments of the television transmitter, and the possible effect of these adjustments on different types of receivers.

84. DEVELOPMENT AND PERFORMANCE OF TELEVISION, CAMERA TUBES

R. B. Janes, R. E. Johnson, and R. S. Moore
(RCA Victor, Lancaster, Pa.)

Three new television camera tubes have resulted from a development program extending over several years. These are:

1. The 2P23 image orthicon, which is especially suited for remote pickups where a wide range of illumination is encountered.
2. The 5655 image orthicon, which is capable of producing better-quality pictures for studio scenes where more light is available and the range of illumination is not so great.

3. The 5769 image orthicon, which may be used for either remote or studio pickups.

The paper discusses the construction and operation of these tubes in detail. In addition, other tubes developed during the course of this work are described and their limitations brought out.

85. AN ANASTIGMATIC YOKE FOR TELEVISION DEFLECTION

KURT SCHLESINGER
(Motorola, Inc., Chicago, Ill.)

The paper describes a television deflection yoke with two coil doublets per scan and capable of reducing field distortion and spot defocussing at wide angles of deflection, if certain conditions are fulfilled. These are derived by theoretical considerations.

The performance of this yoke is shown in several tests. These include field patterns, photos of beam distortion, dot patterns on a kinescope screen, and a monoscope picture.

Three simple and economical circuits are shown for use in a television receiver employing the new yoke. They are (1) a line sweep using the field energy in the yoke to energize the focus coil, (2) a field sweep amplifier with resistive plate load, using phase correction for linearity control, and (3) a synchronous high-voltage pulse power supply. All these units operate on low supply voltage (250 volts).

86 A HIGH-EFFICIENCY SWEEP CIRCUIT

B. M. OLIVER
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

This paper describes a sweep output stage using two driver tubes in a push-pull class-B connection. The entire plate power for one of these tubes is supplied by the energy stored in the yoke by the other tube. The plate power consumption is about one-fourth that of a single-ended class-A driver stage. Except during flyback, transmission is present from the grid of a driver tube to the yoke. Negative feedback can therefore be used to provide stable sweep linearity.

High voltage for the cathode-ray tube can be obtained, either in the usual manner or in a novel fashion which prevents a loss of efficiency.

87. CURRENT DEVELOPMENTS IN UHF TELEVISION

THOMAS T. GOLDSMITH, JR.
(Allen B. Du Mont Laboratories, Inc., Clifton, N. J.)

Widespread attention is currently being devoted to the possibility of using the uhf channels from 475 to 890 Mc for expansion of television service. Consideration by the FCC, as well as great activity within the industry, makes this field of endeavor change complexion rapidly with passing weeks. This paper will be a status report of uhf television at the time of presentation.

Emphasis will be placed upon propagation characteristics, material concerning allocations distribution, availability of power, the status of transmitting and receiving equipment, and, finally, the consideration of standards as to bandwidth, black-and-white, color, and other essential factors which must be determined before commercial operation in this band is authorized.

Wave Propagation II

88. AN ANALYSIS OF DISTORTION RESULTING FROM TWO-PATH PROPAGATION

IRVIN H. GERKS
(Collins Radio Co., Cedar Rapids, Iowa)

It is shown that the distortion caused by two-path propagation for the case of amplitude modulation is principally a result of over-modulation in the resultant signal. This distortion may be very severe when the difference in path lengths becomes large and the two signals have nearly equal amplitudes, whereas it is usually negligible where the difference in path lengths is small, as in tropospheric propagation. The response of an ideal FM receiver is then investigated. It is found that the frequency of the resultant signal, when averaged over a beat-frequency cycle, is the same as that of the stronger signal, but that a very large excursion or "spike" may occur in one direction from this average value. The necessary characteristics of a receiver are determined to permit this averaging process to become effective when the two component signals are nearly equal in amplitude. It is shown that in this case the distortion can be made small, even for large differences in path lengths and an amplitude difference of less than 1 db, provided the frequency deviation is made sufficiently great.

89. ON THE ORIGIN OF SOLAR RADIO NOISE

ANDREW V. HAEFF
(Naval Research Laboratory, Washington, D. C.)

The observed anomalous radio frequency radiations from the sun are associated with sun-spot activity and are believed to be generated within intermingling streams of charged particles issuing from active areas of the sun. Such streams have the property of greatly amplifying initial fluctuations over a range of frequencies determined by the density and velocity distribution of particles in the stream. The theory of generation of radio energy resulting from space-charge interaction between streams of charged particles is reviewed and applied to the solution of the solar radio noise problem. From estimates of average density and velocity distribution of solar particles, it is possible to compute on the basis of the new theory the frequency of the most intense radiation (30 to 60 Mc) and the absolute value of radiation intensity at the surface of the earth (7 to 2)$\times 10^{-22}$ watt/(cm² cps) which agree well with the measured values. The most probable spectral distribution of the anomalous solar radiation is derived in the form:

$$\frac{E}{E_m} = \left(\frac{\lambda}{\lambda_m}\right)^2 e^{2(1-\lambda/\lambda_m)}$$

where E_m is the maximum intensity corresponding to the wavelength λ_m.

90. GEOMETRICAL REPRESENTATION OF THE POLARIZATION OF A PLANE ELECTROMAGNETIC WAVE

GEORGE A. DESCHAMPS
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A geometrical representation, introduced by H. Poincare, is briefly described. Polarization states of an electromagnetic wave, or of an antenna, are represented by points on the surface of a sphere. A simple formula is derived for the variation of the voltage induced in a given antenna when the polarization of the incident wave is changed.

Trigonometrical and graphical solutions are then provided for the following problems:

1. Determination of the polarization of a plane wave of an antenna from a minimum number of measurements.

2. Synthesis of a transducer that transforms states of polarization in a given manner.

An extension of these methods to some mode transformers and directional couplers is indicated.

91. PROPAGATION CONDITIONS AND TRANSMISSION RELIABILITY IN THE TRANSITIONAL MICROWAVE RANGE

THOMAS F. ROGERS
(Air Matériel Command, Cambridge, Mass.)

The propagation characteristics of the 5- to 1.5-cm microwave range are theoretically investigated and discussed. This range lies in a transitional region of the electromagnetic spectrum wherein absorption effects become pronounced. Quantitative sea-level values for absorption to be expected due to oxygen and water molecule resonance are found as a function of path length and atmosphere water-vapor content; these results are extended to high altitudes where atmospheric pressure, temperature, and water-vapor content differ appreciably from sea-level values. Rainfall attenuation values are also given and, on the basis of information regarding expected rainfall intensity and duration in the temperate zone, figures of percentage reliability are arrived at for operation over this range. Values are given for the system power gains necessary to maintain free-space received signal conditions under various transmission-path lengths and percentages of reliability. Other propagation effects are treated; the over-all data allows a number of conclusions to be reached regarding the value of various frequencies for specific services.

92. A FORWARD-TRANSMISSION ECHO-RANGING SYSTEM

DONALD B. HARRIS
(Collins Radio Co., Cedar Rapids, Mich.)

A new type of echo-ranging system is described in which the receiver is located at a distance from the transmitter. It is shown that with a configuration of this type the range of the target with respect to the received station is equal to the difference in path between the direct and reflected waves divided by the versine of the angle of elevation of the receiving antenna ($\rho = \delta/1 - \cos\ \theta$). By making the radial writing speed proportional to this function, a PPI presentation system can be realized which shows a profile of the propagation path viewed from the side. This system has particular application in detecting targets such as, for example, atmospheric irregularities, which have a low reflection coefficient at normal incident angles. The system, therefore, promises to be useful in connection with the study of propagation problems.

Active Circuits II

93. HIGH-POWER SAWTOOTH CURRENT SYNTHESIS FROM SQUARE WAVES

Heinz E. Kallmann
(Consulting Engineer, New York, N. Y.)

A sawtooth wave S, with rounded peaks but with a straight slope extending over nearly 90 per cent of the period, is obtained by (1) synthesis of a step wave from three or four members of a series of square waves P_n whose amplitudes and periods fall with 2^n: $S = P_1 + \frac{1}{2}P_2 + \frac{1}{4}P_4 + 1/8P_8 + \cdots + \frac{1}{2_n}P_{2_n}$ and by (2) smoothing the step wave by attenuating its higher harmonics in a simple filter network with negligible transmission at and beyond the lowest missing harmonic. Networks having good transient response are found to be most suited. The efficiency of a tube circuit for sawtooth current synthesis approaches 50 per cent when comparing power output with power supplied to the tube anodes. A modification of this circuit, using gated square waves, may have an efficiency approaching 100 per cent.

94. COMPARISON OF THE *LC* TOROIDAL FILTER WITH THE PARALLEL-TEE FEEDBACK-AMPLIFIER FILTER

A. J. Stecca
(Naval Research Laboratory, Washington, D. C.)

It is the purpose of this paper to describe the analytical study and laboratory investigation conducted with an *LC* toroidal network as a passive band-pass filter and the parallel-tee feedback amplifier as an active band-pass filter designed for a specific application. For this application the filter has a center frequency in the region of 1,000 cps and a bandwidth of 15 per cent. It was required that over the transmission band the phase shift be linear with frequency and that the attenuation be at least 30 db at a frequency ratio of 1.5. The conclusions show that the toroidal filters provided a very satisfactory operating characteristic and require less physical space, and give equally good, or better, temperature stability than can be obtained with the present art of parallel-tee construction.

95. A PEAK-SELECTOR CIRCUIT

M. J. Parker
(Naval Ordnance Laboratory, Silver Spring, Md.)

A problem required that the maximum positive peak of an asymmetrical periodically recurring wave, having a frequency in the range of from 2 to 20 cps, be used as a trigger for a sawtooth generator. The phase shift between the trigger pulse and the maximum positive peak, from which the trigger was derived, had to be less than one-half degree per cycle over the frequency range mentioned. A circuit consisting of an amplifier, a maximum peak selector, clippers, and differentiators was designed. When activated by a simulated output signal this resulted in a sharp spike which met phase-shift requirements. Due to the phase shift introduced by oscilloscopes, a novel method of measuring the performance of the circuit had to be devised. This method and associated equipment will be described.

96. A LOW-FREQUENCY SYNCHRONIZED SAWTOOTH GENERATOR PROVIDING CONSTANT AMPLITUDE SWEEP WITH APERIODIC SYNCHRONIZATION INPUT

P. Yaffee
(Naval Ordnance Laboratory, Silver Spring, Md.)

The functions of the instrument described in this paper are such as to provide a synchronized linear voltage function which sweeps between two constant reference levels, notwithstanding variation in its period in accordance with an aperiodic synchronization input. The design of this device is covered, and details concerning its circuitry and operation are given. Special circuit features, such as the method of achieving automatic control of the linear sweep between the reference levels in conjunction with the variation in the period of the sweep, are pointed out. The generator, as described, provides an output potential sweep of 0 to 150 volts with good linearity over the required range of frequency control. While the requisites of its intended use are satisfied by a synchronization frequency range of 4 to 10 cps, operation beyond this range appears feasible.

97. REGENERATIVE AMPLIFIERS

Y. P. Yu
(North Dakota Agricultural College, Fargo, N. Dak.)

This paper describes the principles and applications of regenerative amplifiers which may be used, for example, to mark the instant when two voltages become equal. A peak-voltmeter circuit based upon the switching properties of a regenerative amplifier is introduced to minimize the error encountered in measuring low-duty-cycle pulses. The use of a regenerative amplifier in forming a pulse-width discriminator circuit is also described.

98. A RECTIFIER FILTER CHART

Reuben Lee
(Westinghouse Electric Corporation, Baltimore, Md.)

All rectifiers generate ripple in the rectifier dc output. Adding a suitable filter between the rectifier and the load reduces this ripple, but it also introduces regulation, both steady and transient, and affects the rectifier peak current. A properly designed filter must keep these properties within tolerable limits. A chart has been made for choke-input filters which relates the various filter properties and enables the rectifier designer to choose quickly components such that specified performance limits are not exceeded. Use of the chart is discussed, and an example given.

Instruments and Measurements IV

99. HIGH-IMPEDANCE MILLIVOLT MEASUREMENTS ABOVE 5 MC

Walter K. Volkers
(Millivac Instruments, New Haven, Conn.)

Germanium diodes and other types of crystal rectifiers fall into the general classification of square-law detectors at input voltage levels below 25 millivolts. These rectifiers can be used as "pseudo-thermocouples" for high-impedance millivolt measurements at input power levels considerably lower than those of conventional thermocouples, and when the diode is combined with a dc vacuum-tube millivoltmeter.

The use of this technique for millivolt measurements above 5.0 Mc is described.

100. SOME ASPECTS OF THE PERFORMANCE OF MIXED CRYSTAL

P. D. Strum
(Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.)

This paper will show that the microwave performance of a mixer crystal as a circuit element can be predicted with good accuracy from its dc characteristics. Conversion loss, rf impedance, if impedance, and noise temperature of the crystal can be calculated directly for frequencies below the region where the internal reactances of the crystal become comparable with its resistance. By using the parameters as determined by the methods presented in this paper, a complete mixer-circuit analysis can be made using standard network-analysis techniques.

101. A WIDE-BAND AUDIO PHASEMETER

John R. Ragazzini and Lofti A. Zadeh
(Columbia University, New York, N. Y.)

The problem of measuring the phase angle between two sinusoidal voltages is an important one in laboratory procedure. Phasemeters which can make this measurement accurately over a substantial frequency range have heretofore been bulky and relatively stationary. This paper describes a circuit which results in a phasemeter which is accurate to about 30 minutes, has a dial adjustment which is independent of frequency, operates over a range of from 10 cps to 100 kc, and is relatively small and inexpensive. The principle used is to match the phase shift between the unknown voltages with a calibrated phase shift in the meter channel. This phase equalization is observed on a standard oscillograph, which is used only as a comparison instrument. The oscillograph amplifier phase shifts are balanced out with an approximate phase shifter cascaded in the meter channel, and do not affect the measurement.

102. A DEVICE FOR ADMITTANCE MEASUREMENTS IN THE 50-TO-500 MC RANGE

W. R. Thurston
(General Radio Co., Cambridge, Mass.)

The device to be described is based upon a null method of comparing an unknown admittance to a fixed standard of conductance and a fixed standard of susceptance. Three rotatable loops are the balancing variables, and their positions under null conditions indicate conductance, susceptance, and a multiplying factor for the first two values.

An adjustable susceptance standard is used that is set to the same value for each frequency, and loop positions are calibrated directly in millimhos and multiplying factor, independent of frequency. The model that has been constructed has scale ranges of from 0 to 20 millimhos for conductance, from −20 to +20 millimhos for susceptance, and from 1 to infinity for the multiplying factor.

103. AN IMPROVED RF CAPACITOMETER

E. F. Travis and T. M. Wilson

(General Electric Co., Schenectady, N. Y.)

Some time ago the General Electric rf capacitometer was developed to meet the need for rapid, accurate measurements of small values of capacitance and inductance. To meet the need for an instrument to measure wider ranges, the original capacitometer was redesigned to extend its range, and at the same time improve the circuit and increase the ease of operation. This paper describes some of the problems encountered in the development, manufacture, and testing of this instrument, and how these difficulties were overcome.

104. A RADIO-FREQUENCY GAS-DISCHARGE PHENOMENON AND ITS APPLICATION TO MECHANICAL MEASUREMENTS

Kurt S. Lion and John W. Sheetz

(Massachusetts Institute of Technology, Cambridge, Mass.)

A gas-filled tube with two electrodes is brought into an electric radio-frequency field of sufficient intensity to produce a gas discharge. If the rf field is slightly unsymmetrical with respect to the electrodes of the glow tube, a dc voltage can be observed between both electrodes. The magnitude and polarity of this dc voltage depends upon the asymmetry of rf field strength on both electrodes and, within wide limits, is independent of the frequency and voltage of the applied rf field. The method is suitable for the measurement of small mechanical displacements with great sensitivity and with a high degree of stability.

Electronics II

Electron-Tube Cathodes

105. THE EFFECTS OF VARIOUS BARIUM COMPOUNDS WITH RESPECT TO COLD-CATHODE BEHAVIOR AS A FUNCTION OF LIFE IN A GLOW DISCHARGE

Harold Jacobs and Armand P. LaRocque

(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

The effects of various barium compounds in glow-discharge tubes are determined and compared with those of barium metal. The various barium compounds are shown to require different activation schedules, in order to obtain the lowest initial regulation voltage. The regulating voltage, as a function of time, is determined for all the cathode surfaces. The composite surfaces are found to have greater stability than the metallic surfaces over a long period of time. The composite barium surfaces are shown to reach an asymptotic limit in their respective regulating voltages, with these values being of the same order of magnitude as the original regulating voltage of the metallic barium surface.

106. OXIDE-CATHODE PROPERTIES AND THEIR EFFECTS ON DIODE OPERATION AT SMALL SIGNALS

G. Conrad Dalman

(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

Recent experimental data obtained on commercial-type cathodes are presented which show the order of magnitude of the low-frequency cathode resistance and the high-frequency cathode impedance. These factors cause deviation from the small-signal low-frequency and high-frequency ideal-diode response given by Llewellyn, and partially explain the difficulty in obtaining diode oscillations.

A brief discussion will be given of the properties of the cathode affecting its behavior as a circuit element. Some experimental evidence that a nonlinear resistance may reside in the coating near the coating-vacuum boundary will also be presented.

107. MICROANALYSIS OF GAS IN CATHODE-COATING ASSEMBLIES

Harold Jacobs and Bernard Wolk

(Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.)

A study of gasses evolved from oxide coated cathode assemblies was made during degassing and activation conditions. It was found, first, that the volume and nature of gasses evolved from uncoated nickel cathode sleeves were practically independent of the three different cleaning methods used and, second, that hydrogen fired cathodes liberated slightly larger quantities of hydrogen when heated in vacuum. The release of hydrogen from nickel cathodes was not instantaneous but was observed to continue even after two and a half hours of continued heating at 900°C.Br.

When a similar analysis was made of gasses liberated from nickel sleeves coated with alkaline earth carbonates, the evolution of hydrogen was reduced considerably, but with a corresponding increase of CO.

The gas condition in the tubes during cathode degassing is shown to be related to the speed of exhaust.

108. EXPOSURE OF SECONDARY-ELECTRON-EMITTING SURFACES TO THE EVAPORATION FROM OXIDE CATHODES

C. W. Mueller

(RCA Laboratories, Princeton, N. J.)

Heretofore, the exposure of a secondary-electron-emitting surface to the evaporation from an oxide cathode has been considered impractical because of the rapid deterioration of the secondary emission surface. Consequently, in previous grid controlled tubes using multiplying surfaces, complicated focussing structures were necessary to bend electron beams and avoid surface contamination. Experiments discussed in this paper show that the evaporation difficulties can be overcome, and even used to enhance secondary emission in some applications. Methods will be illustrated by examples of the use of secondary-emission surfaces exposed to filamentary and indirectly heated oxide-coated cathodes in practical grid-controlled multiplier tubes.

109. THE USE OF THORIATED-TUNGSTEN FILAMENTS IN HIGH-POWER TRANSMITTING TUBES

R. B. Ayer

(RCA Victor, Lancaster, Pa.)

Thoriated-tungsten filaments have demonstrated their reliability in low-power transmitting tubes for the past quarter century. More recent experience has proved that this type of filament is equally reliable in all classes of high-power, high-voltage transmitting-tube service. This has been accomplished by improved materials and techniques.

Filament power consumption only 1/3 that required for the equivalent pure tungsten-filament type, improved mechanical strength, lower bulb temperatures, improved high frequency performance, greater emission reserve, and improved plate characteristics are a few of the advantages gained thoruch the use of thoriated-tungsten filaments in high power transmitting tubes.

Types developed for use in the broadcast and industrial field during the past year and a half are described in the paper.

Systems II

Relay Systems

110. A MICROWAVE SYSTEM FOR TELEVISION RELAYING

J. Z. Millar and W. B. Sullinger

(Western Union Telegraph Co., New York, N. Y.)

The paper describes two television relay circuits established by Western Union between New York and Philadelphia utilizing recently developed Philco 6000-Mc equipment. The transmission requirements for relaying video programs are discussed, and the methods of meeting these requirements are described. The relay circuits are independently reversible in short switching time, making possible the use of a single relay circuit for transmitting consecutive programs in opposite directions, thus reducing the amount of equipment necessary as compared to two way, or duplex, service. The equipment is described both for terminal stations, to be located at the broadcast stations of the customers, and relay stations located at Western Union sites. Photographs of the installations will be shown.

111. SYNCHRODYNE PHASE MODULATION OF KLYSTRONS

Vincent Learned

(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

The method of phase-modulating a klystron amplifier beam to obtain side-frequency output will be explained. By tuning the input resonator to the incoming microwave carrier and the output resonator to a phase modulation sideband, a klystron may be used to change the frequency of a microwave carrier by a predetermined amount. A tube utilizing this principle will be described.

112. INTERCITY TELEVISION RADIO RELAYS

William H. Forster

(Philco Corporation, Philadelphia, Pa.)

Television radio relay equipments may be divided into two classes. Systems employing second detection at each repeater are most suitable for intracity and remote pickup service, and those employing heterodyne remodulation at the repeaters are well adapted for intercity long-lines service. At the present time, television radio-relay network facilities have fidelity and signal-to-noise characteristics which exceed those provided by many other parts of the television system. As picture-generating sources are improved, better performance will be required of the network facilities.

The paper discusses specifications for television network facilities and compares them with design specifications and performance data of the Philco equipment used by Western Union in their New York-Philadelphia television circuit. Methods of improving the performance of this equipment, for a greater number of repeaters, particularly the signal-to-noise ratio and the transient response, will be described. A color film of relay installations will be shown.

113. VIDEO DESIGN CONSIDERATIONS IN A TELEVISION LINK

M. Silver, H. French, and L. Staschover

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A repeater link for the transmission of standard composite television signals must meet exacting noise and transient-response requirements. This paper is devoted to the transmitter-modulator section of a relay link consisting of 10 repeaters. Design problems encountered in conforming with specifications for frequency and transient response, and signal compression, are treated in detail. Solutions to these problems in a practical circuit design are presented.

114. SIX-CHANNEL URBAN MOBILE SYSTEM WITH 60-KC SPACING

R. C. Shaw, P. V. Dimock, W. Strack, and W. C. Hunter

(Bell Telephone Laboratories, Inc., New York, N. Y.)

This paper describes a six-channel mobile radiotelephone system in Chicago, operating in the 152- to-162 Mc band, and using 60-kc spacing of carrier frequencies, rather than the 120-kc spacing of previous practice. The measures required to achieve this frequency saving are described, including filters and special antenna arrangements at the land transmitter, "off-channel squelch" in the land receivers, agc in the mobile receivers, connection of six land receivers to a common antenna, and other special coordinating means.

Navigation Aids I

115. THE DETERMINATION OF GROUND SPEED OF AIRCRAFT USING PULSE-DOPPLER RADAR

Irving Wolff, S. W. Seeley, Earl Anderson, and W. D. Herschberger

(RCA Laboratories, Princeton, N. J.)

Apparatus is described for determining the ground speed of aircraft over both water and land, using airborne pulse-doppler radar. In this equipment pulse signals are transmitted both forward and rearward, the phases of the returning signals are compared, and the rate of phase shift is determined. Directional antennas are not required since the angle can be determined by time-gating circuits. This has the additional advantage of making the determination independent of the pitch of the aircraft. Experimental tests indicate that ground-speed determination can be made with reasonable accuracy using such equipment.

116. THE DIMEAL AIRCRAFT APPROACH AND LANDING SYSTEM

Ludlow B. Hallman, Jr.

(Wright-Patterson Air Force Base, Dayton, Ohio)

The paper suggests and briefly describes an aircraft approach and landing system which makes use of the standard radar-type distance-measuring equipment to establish the "localizer" or runway alignment pattern. The "descent" or "glide" path is established by means of altimeters of the pressure and radio types contained entirely within the aircraft. The suggested name DIMEAL is derived from the words *DI*stance *M*easuring *E*quipment and *AL*timeters.

117. THEORETICAL ASPECTS OF NONSYNCHRONOUS MULTIPLEX SYSTEMS

Warren D. White

(Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.)

In the past, considerable attention has been given to various forms of frequency-division and time-division multiplex systems in which information from a number of sources is transmitted over a common channel. Most of these systems, however, require that the information from the various sources be assembled at a common location where a systematic multiplex operation can be performed. With the advent of recent advances in the theory of communication by Shannon and others, it is now possible to examine the case where a multiplicity of geographically separated transmitters share a common channel without the benefit of various synchronizing techniques. Such nonsynchronous multiplex systems are important in the case of air navigation and other applications where the bandwidth of a channel is for some reason not consistent with the amount of intelligence to be transmitted. This paper considers the theoretical information capacity of some of the simpler forms of nonsynchronous systems.

118. BAND-PASS CIRCUIT DESIGN FOR VERY-NARROW-BAND, VERY-LONG-RANGE DIRECTION-FINDER RECEIVERS TO MINIMIZE BEARING ERROR DUE TO RECEIVER MISTUNING

Milton Dishal and Homer Morrow

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

In very-long-range navigational systems, it is desirable to use the narrowest possible receiver bandwidth to obtain maximum receiver sensitivity.

When the receiver pass band is just wide enough to accommodate the sidebands passing through the receiver, the nonlinear phase shift near the edges of the pass band produces a bearing error when the receiver is not correctly tuned.

This paper presents the analysis and design values for band-pass circuits to obtain minimum direction-finding bearing error with receiver mistuning, and points out the fact that the bandwidth of the phase-response curve of a network may be more useful than the bandwidth of the amplitude-response shape when complex waveform reproduction is of importance.

119. CRYSTAL CONTROL AT 1,000 MC FOR AERIAL NAVIGATION

S. H. Dodington

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The problems faced in the design of a crystal-controlled channelling system are described, together with the solution adopted for an airborne pulse-type distance-measuring equipment, operating on 51 channels in the 960- to 1,215-Mc band, each channel permitting simultaneous service to fifty aircraft at distances up to 100 miles with better than 1 per cent accuracy.

SYMPOSIUM

Marketing

1. Market Research

E. H. Vogel

(General Electric Co., Syracuse, N. Y.)

2. The Application of Market and Field Research in Product Planning and Design

O. H. L. Jensen

(Philco Corporation, Philadelphia, Pa.)

3. Sales Planning and Distribution

Lee McCanne

(Stromberg-Carlson Co., Rochester, N. Y.)

4. National Advertising

M. F. Mahony

(Maxon, Inc., New York, N. Y.)

5. Sales Training and Sales Promotion

W. E. Macke

(Zenith Radio Corporation, Chicago, Ill.)

Marketing television equipment is a process as significant as its manufacture, and the growing postwar trend is toward a broader use of market and product research in order to enable manufacturers to understand the buying public and forecast its needs. Estimated industry sales are based on the compilation and use of national economic data, such as extent of employment, general income, cost of living, trends of cost and prices, estimated spendable income, and dollars available for purchases other than basic necessities. Manufacturers also must consider seasonal variations in the sale of the different categories and the estimated strength of their competitors, as well as surveying their own products and customers' preferences, and tabulating such reports for use in product planning. This work will be described by Mr. Vogel.

Developing a new model in the radio or television field involves a number of factors. The manufacturer, as Mr. Jensen will point out, must first determine as far as possible the type or model of radio receiver the public wants, both with respect to performance and appearance. Cabinets and technical components must be developed simultaneously and according to definite schedules, so that over-all competition will be on time. The balance between the cost of manufacture and the selling price required in order to assure reasonable profit for the manufacturer must be reached.

Introducing a new model to the trade and to the public involves co-ordination of the various sales factors. Literature is released to all field representatives, distributors, and dealers, and eventually to the public under the classification of advertising. Timing, Mr. McCanne will state in his paper on sales planning and distribution, must be accurate, so that adequate merchandise can be provided to meet initial demands. Radio sets are generally distributed from the manufacturer to the wholesaler to the dealer. They are also, however, sold directly to the dealer and through mail-order houses, both of which methods have signal advantages over the other, principally in cutting costs, as well as disadvantages.

Mr. Mahony will describe the vital part advertising plays in merchandising radio and television sets. Advertisements must be prepared according to the medium for which they are destined: national magazines for broad registration and impression, newspapers for local point-of-sale advertising and possible dealer identification. Not only are there differences in copy appeal between newspapers and magazines, but among the various types of magazines themselves. Aside from graphic advertising, there is also tremendous value in using radio itself as an advertising medium, and indications show that television may eventually surpass even radio as an advertising power.

Sales promotion methods directly follow advertising, as Mr. Macke will show. Store backgrounds, window displays, signs, literature, broadsides—all tie in with national advertising in exploiting the advertiser's products. Salesmen and retailers themselves must be trained in the new product, and many wholesale distributors use numerous methods—charts, slides, films, and so on—in order to indoctrinate their own personnel and the dealers whom they serve.

Electronics III

Electron-Tube Theory

120. GENERAL SOLUTION OF THE TWO-BEAM ELECTRON-WAVE-TUBE EQUATION

A. V. Haeff, H. D. Arnett, and W. Stein
(Naval Research Laboratory, Washington, D. C.)

The analysis of the new mechanism of microwave energy amplification, based on space-charge wave amplification effects occurring as a result of interaction between streams of charged particles, is extended to the case of two beams of arbitrary space-charge density and velocity distribution. It is shown that, for any current distribution, amplification of space-charge waves will take place over a limited frequency range determined by the distribution. The real component of the propagation constant defining the amplifying properties of such streams is computed in terms of the "inhomogeneity factor" and the "distribution index," which are dimensionless parameters relating to velocity difference and current division between the two streams. As the ratio of currents or velocities is increased, the frequency range over which energy amplification takes place increases, but the maximum gain per unit length of such streams and the effective bandwidth decrease.

121. ASPECTS OF DOUBLE-STREAM AMPLIFIERS

J. R. Pierce
(Bell Telephone Laboratories, Inc., New York, N. Y.)
W. B. Hebenstreit
(Hughes Aircraft Co., Glendale, Calif.)
A. V. Hollenberg
(Bell Telephone Laboratories, Inc., New York, N. Y.)

In double-stream amplifiers, a signal is impressed on two near-by and hence coupled streams of electrons having different velocities, by means of a helix or a resonator. An increasing wave is set up and an amplified signal is extracted by another helix or resonator located further along the streams. An evaluation of performance should take into account the effect of physical separation of the electron streams and the initial magnitude of the increasing wave set up, as well as the rate of increase of the increasing wave. Material covering these points will be presented.

122. ON THE THEORY OF AXIAL SYMMETRIC ELECTRON BEAMS IN AN AXIAL MAGNETIC FIELD

A. L. Samuel
(University of Illinois, Urbana, Ill.)

A perplexing problem in the design of klystrons and traveling-wave tubes has been the formation of well-collimated electron beams. An axial symmetric form of a tubular electron beam is discussed in which the space-charge repulsive forces are just balanced by magnetic forces, so that the beam may be made as long as desirable without any change in its cross section. The equations governing the existence of such beams are derived, and methods of obtaining useful solutions are suggested. One of the more important consequences is that the axial velocity can be made the same for all electrons in the beam.

123. ELECTRON BEAMS IN AXIAL SYMMETRIC MAGNETIC AND ELECTRIC FIELDS

C. C. Wang
(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

This paper derives equations for the trajectories of the electrons along the outer edge of the beam for the most general case in which there are both axial and radial components of the fields. It is shown that, as a result of symmetry, the combined effects of the electric and magnetic fields can be expressed as a single generalized potential function which depends only on the axial and radial space co-ordinates. This permits the expression of the axial and radial force components as the axial and radial components of the gradient of this potential function. It is shown that there exists an equilibrium radius for which the net radial forces acting on the electrons is zero, and that the outer radius of the beam will oscillate about this equilibrium value, the amplitude being nonsymmetrical and depending upon the initial conditions, and the wavelength (distance between successive maxima) depending upon the amplitude.

124. SPACE-CHARGE EFFECTS AND FREQUENCY CHARACTERISTICS OF CW MAGNETRONS RELATIVE TO THE PROBLEM OF FREQUENCY MODULATION

H. W. Welch, Jr.
(University of Michigan, Ann Arbor, Mich.)

The purpose of this paper is to integrate and supplement the knowledge of space-charge effects and frequency characteristics of cw magnetrons. The methods of analysis and the experimental data which have bearing on the problem of cw magnetrons are discussed in their relation to the problem of frequency modulation. Major emphasis is on the quantitative interpretation of "hot-impedance" measurements made on the pre-oscillating magnetron. The results of the analysis of these measurements, which give an improved understanding of space-charge behavior, are related to the Hull and Hartree voltage boundaries and, qualitatively, to the phenomena of frequency pushing and frequency pulling.

SYMPOSIUM

Germanium and Silicon Semiconductors

1. Electrical Properties of Germanium and Silicon

K. Lark-Horovitz
(Purdue University, Lafayette, Ind.)

2. The Metallurgy of Germanium and Silicon Semiconductors

J. H. Scaff
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

3. Theory of Rectification

Frederick Seitz
(Carnegie Institute of Technology, Pittsburgh, Pa.)

4. Physics of the Transistor

Walter H. Brattain
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Within the past ten years, semiconductors have found wide and increasing uses as photosensitive, thermally sensitive, and rectifying elements in communication systems. Recent advances in theory have led to the development of a new three-electrode semiconductor amplifier. The speakers will review the theories and experiments leading to the present understanding of the electrical bulk properties and metallurgy of germanium and silicon. Surface properties of these semiconductors in combination with metallic electrodes will be related to their rectifying and amplifying behavior.

Information Transmission and Noise

125. DESIGN IN NATURE AS EXPLOITED BY THE COMMUNICATION ENGINEER

L. A. de Rosa

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

When considering the problem of extracting an optimum message from a noise background so as to enable the transmission of intelligence by a modulation system requiring the least power for a given amount of intelligence, a number of apparent difficulties in equipment design arise.

A consideration of the expedients that might theoretically be used to improve signal-to-noise ratios in communication systems leads to a number of novel and peculiar arrangements. A study of the operation of certain sensory organs indicates that apparently nature has made use of some of these same expedients.

126. EXPERIMENTAL DETERMINATION OF CORRELATION FUNCTIONS AND THE APPLICATION OF THESE FUNCTIONS IN THE STATISTICAL THEORY OF COMMUNICATIONS

T. P. Cheatham, Jr.

(Massachusetts Institute of Technology, Cambridge, Mass.)

The scope of the theoretical investigations of the statistical character of communications far outstrips its experimental and practical demonstration. A realistic and integrated research program, centered about the exploitation and extension of Wiener's general theoretical work, must devote a principal part of its early energy toward the practical problems of measuring the main statistical parameters of the theory in a rapid and efficient manner. The philosophy and results of such a program will be described. In particular, a detailed description of an electronic correlator which computes autocorrelation and cross-correlation functions of stationary time series will be given. Results, including slides, of its application to a study of random noise will be shown.

127. THE TRANSMISSION OF INFORMATION THROUGH BAND-LIMITED TRANSMISSION SYSTEMS

W. P. Boothroyd and E. M. Creamer, Jr.

(Philco Corporation, Philadelphia, Pa.)

The paper describes a few of the problems encountered in attempts to transmit a maximum of modulation through a transmission system of definite bandwidth. A technique for equalizing the amplitude and phase response of a transmission system, based on the direct analysis of the system transient response, is discussed. A transient equalizing apparatus for synthesizing a desired transient or for modifying an undesired transient response is described. Specific problems encountered in the field of multiplexed communication transmission and television broadcasting and transmission are considered in the light of this technique.

128. SIGNAL-TO-NOISE IMPROVEMENT THROUGH INTEGRATION IN A STORAGE TUBE

J. V. Harrington and T. F. Rogers

(Air Matériel Command, Cambridge, Mass.)

Random noise places a fundamental physical limitation on the precision with which a signal may be observed or measured. If signal energy arrives piecemeal in a coherent manner it may be integrated over a period of time to provide an increase in signal-to-noise ratio, thereby reducing such a limitation. Several techniques have been suggested; a short review of their characteristics and an analysis of some of their limitations are presented. Recent work with an electronic barrier-grid storage tube has shown that it exhibits many desirable features as an integrating device. An introduction to accumulator theory and an analysis of the storage tube operating as an accumulator are given; calculations are made of the expected signal-to-noise ratio improvement. Evidence of the improvement experimentally achieved is presented.

129. THEORY OF RECEIVER NOISE FIGURE

L. J. Cutrona

(Sperry Gyroscope Co., Great Neck, L. I., N. Y.)

Networks for coupling the antenna to the grid of the first tube of a receiver, which yield minimum noise figures (as defined by Friis), are derived. Induced grid noise due to transit time and shot-effect noise are considered in the coherent and incoherent cases. A chart for the rapid estimation of noise figure in one case, and design formulas for the networks, are given.

Navigation Aids II

130. VHF AIRBORNE NAVIGATIONAL RECEIVER AND ANTENNA SYSTEM

A. G. Kandoian, R. T. Adams, and R. C. Davis

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

An airborne vhf receiver has been developed to provide voice communication and navigational facilities in the range from 108 to 136 Mc. It provides for 280 channels with crystal stability for either voice communication alone, or voice with superimposed navigational signals. The accuracy of the navigational facilities may be seriously impaired by interference, and this imposes stringent requirements on the receiver and its associated antenna system. Unusual techniques have been employed to provide maximum rejection of interfering signals, high-level pulse-type interference, and vertically polarized spurious signals.

131. CERTAIN NEW PERFORMANCE CRITERIA FOR LOCALIZER AND GLIDE-SLOPE GROUND INSTALLATIONS

Paul R. Adams

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A new parameter, the "clearance fraction," F is proposed, defined as $F=(A/C)-(B/C)$, where A/C and B/C are the percentages of modulation of the 90-cps and 150-cps tones, respectively. This new parameter is applicable to both glide-slope and localizer installations. It is proposed that "side clearance" and "path width" should hereafter be defined in terms of this clearance fraction F.

For localizers only, a further function, the couse-bend stability factor S, is proposed. This stability factor indicates the degree of immunity to course bends or scalloping caused by reflecting obstacles.

132. PHASE AND OTHER CHARACTERISTICS OF 330-MC GLIDE-PATH SYSTEMS

S. Pickles

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

In the past eight years the 330-Mc equisignal glide-path system has undergone continual improvement in operating characteristics. This has been due mainly to the increased knowledge which has been gained from study and operation of the equipments. The relative phase of the signals of the upper and lower antennas has been one of the factors which did not receive as much attention as is required. Recordings showing the effects of phase, along with other recordings of other characteristics, will be shown. The most precise method of measuring phase will be described.

133. PRINCIPLES OF VOLUME SCAN

Daniel Levine

(Wright-Patterson Air Force Base, Dayton, Ohio)

The equations of motion for an improved spiral scan for a radar system are developed, based upon transmitting to each point in space a specified number of radar pulses. In addition, the optimum value of the beam crossover point is discussed, as well as the dependence of angular resolution upon position within the region searched by the radar system. Expressions are derived for the optimum wavelength for a search radar set as a function of the scanning parameters and the antenna area.

134. THE CONTROL OF RESONANCE EFFECTS ON THE RADIO BEARINGS OF AN AIRCRAFT HIGH-FREQUENCY DIRECTION FINDER

M. K. Goldstein

(Office of Naval Research, Washington, D. C.)

The metallic surface of an aircraft constitutes an environmental effect having a pronounced obscuring influence on high-frequency (1.5 to 30 Mc) bearings in aircraft direction finders. It is shown that this environmental effect is essentially an electrical resonance phenomenon whose resonance frequency can be predicted from the craft's structural dimensions. Techniques are presented for successfully combating these resonance effects which involve the use of electrical loading of the aircraft's structures. A feasible method for using these techniques is given whereby usable radio direction-finder bearings can be obtained for the entire 1.5- to 30-Mc frequency range. Without the use of these corrective measures, less than 25 per cent of this frequency range is usable for direction-finder purposes.

Oscillators

135. AN ANALYSIS OF OSCILLATOR PERFORMANCE UNDER VARYING LOAD CONDITIONS AND AN ELECTRONIC SYSTEM FOR AUTOMATIC LOAD COMPENSATION

Eugene Mittelmann

(Consulting Engineer, Chicago, Ill.)

An analysis of the class-C oscillator shows that constant power absorption in a variable external load impedance can be maintained by proper changes of the operating parameters of the oscillator. The equivalent dynamic generator impedance is primarily determined by the angle of flow and the plate voltage. A system is described in which, for low load impedances, the dc plate voltage is automatically decreased and simultaneously the angle of current flow increased. For high load impedances the dc plate voltage is increased, accompanied by a simultaneous narrowing of the angle of current flow.

Oscillators of up to 30 kw for 450 kc and 40 Mc were operated successfully.

136. LOW-POWER WIDE-TUNING-RANGE UHF OSCILLATORS

F. J. Kamphoefner and J. M. Pettit

(Stanford University, Calif.)

This paper presents a study of negative-grid triode oscillators for the uhf region. Discussion is confined to oscillators using the modified Colpitts arrangement wherein the feedback is provided by the internal tube capacitances. It can be shown that this is the most useful choice where the frequency must be easily varied over wide frequency limits. Lead inductance and transit-time effects are considered for several commercial tube types, along with the requirements of the tuned circuit.

137. REACTANCE-TUBE MODULATION OF PHASE-SHIFT OSCILLATORS

F. R. Dennis and E. P. Felch

(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Frequency modulation or electronic deviation of phase-shift oscillators may be accomplished by employing a portion of the feedback circuit itself as a quadrature network for driving a reactance tube. Advantages, in addition to simplicity, are wide deviation capability, freedom from amplitude modulation, and excellent mean frequency stability. The phase-shifting networks may be of conventional *RC* or *LC* type or may consist of distributed reactances such as sections of transmission line.

RC oscillators operating at audio frequencies, *LC* oscillators at high frequencies, and both *LC* and coaxial-line oscillators in the 100-Mc region will be demonstrated. Linear deviations up to ± 10 per cent are obtained with amplitude modulation of less than ± 1 db. Mean frequency stability compares favorably with conventional unmodulated oscillators.

138. A LOW-DISTORTION AUDIO-FREQUENCY OSCILLATOR

C. W. Clapp and C. L. Hackley

(General Electric Co., Schenectady, N. Y.)

A push-pull *RC*-type oscillator has been developed having distortion less than 0.1 per cent over the range from 35 to 15,000 cps. Distortion normally introduced by the nonlinear element used to stabilize the oscillator is largely eliminated by the push-pull mode of operation. The theory of the circuit is developed and design requirements discussed.

139. AN AUTOMATIC-FREQUENCY-CONTROL SYSTEM FOR MECHANICALLY TUNED OSCILLATORS

J. G. Stephenson

(Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.)

This paper describes a system of automatic frequency control employing mechanical tuning of an oscillator and capable of very wide pull-in range. The system operates from a 30-Mc signal, cw or pulsed, which may be generated at another frequency and heterodyned to 30 Mc by the oscillator whose frequency is controlled. For normal afc operation, a discriminator, pulse stretcher, and balanced modulator are employed to produce 60-cps output of a phase and amplitude determined by the frequency of the incoming 30-Mc signal. The 60-cps signal, amplified and applied to a servomotor, produces frequency correction by tuning the oscillator.

Electronics IV

New Forms of Tubes

140. THE GRAPHECHON—A PICTURE-STORAGE TUBE

L. Pensak

(RCA Laboratories, Princeton, N. J.)

Long-time storage of television-quality pictures is possible with a new experimental type of storage tube consisting of two independent cathode-ray guns and a target plate coated with a thin film of insulating material. A new type of bombardment-induced conduction effect provides high sensitivity and stability. The two guns can operate simultaneously, the one to "write" down any arbitrary pattern to be stored, and the other to scan repeatedly over it to both generate signals and erase the pattern at a controlled rate. The tube makes possible television pictures of oscillograms or radar patterns that can be viewed for several minutes.

141. THE PENCIL-TYPE UHF TRIODE

G. M. Rose and D. W. Power

(RCA Victor, Camden, N. J.)

A new triode construction is described which not only meets the basic requirements of minimum transit time, lead inductance, and internal capacitance, but also incorporates other desirable design features such as small size, good thermal stability, low heater wattage, and convenience of use in circuits. In addition, the construction lends itself to high-volume production methods. The double-ended type of structure is used wherein the rod-type anode and cathode connections extend outward from each side of a central grid disk. The internal elements are cylindrical and coaxial. Two sets of electrical characteristics are shown, one for a medium-μ oscillator and the other for a high-μ oscillator-amplifier. Typical circuits for the tubes are described together with performance results.

142. PRACTICAL APPLICATIONS OF THE RESNATRON IN THE HIGH-POWER TRANSMITTER FIELD

W. W. Salisbury

(Collins Radio Co., Cedar Rapids, Iowa)

Recent developments in the resnatron tube, including adaptation of the tube for operation as a high level power amplifier, means for plate-modulating the tube, operation as a class-B amplifier, and design improvements leading to quick replacement of cathode assemblies at the time of filament failure, will be presented. The advantages of continuously pumped, demountable tubes, as compared with sealed-off tubes, will be discussed. The theoretical discussion will include consideration of the structure of these tubes, which contain internal tuning cavities; the selection of dimensions and operating parameters for optimum results; transit-time effects; secondary-emission effects; improvements in the filament life and emission; and the selection and control of the operating modes of the cavities.

143. THE ELECTRON COUPLER—A DEVELOPMENTAL TUBE, UTILIZING NEW PRINCIPLES, FOR THE MODULATION AND CONTROL OF POWER AT UHF

C. L. Cuccia and J. S. Donal, Jr.

(RCA Laboratories, Princeton, N. J.)

The electron coupler is a fundamentally new type of electron tube, under development, which is installed between a uhf power generator and its load, and which permits modulation and control of the power reaching the load. In the basic tube an electron beam, passing through an input cavity which is connected to a power source, absorbs the input power in the form of rotational energy and delivers this power to an output cavity and load or to a collector. As a uhf circuit element, the electron coupler can be employed as a unilateral-circuit control impedance by using any of several methods of power-transfer control which permit the amplitude modulation of high power by use of very low modulator power without any frequency modulation of the power-generator output.

144. A WIDE-TUNING-RANGE LOW-POWER CW MAGNETRON

L. R. Bloom and W. W. Cannon

(University of Illinois, Urbana, Ill.)

A low-power cw magnetron is described which is continuously tunable over approximately a 2 to 1 frequency range from 1,500 to 2,800 Mc by changing only one parameter, the length of a coaxial resonator. The tube elements consist of a conventional magnetron interaction region with vanes electrically tied alternately to the inner and outer conductor of a coaxial resonator tuned by means of a contact-type shorting plunger. Lumped reactance of the tube is thus continuously variable by changing positions of the plunger. Hot tests indicate that this tube has a power output of 100 mw or greater in the operating frequency range. Starting currents are as low as $I = 4$ ma at $H = 2000$ gauss. Operation is in the π mode and stable over the usable range. A description of the detailed performance of the tube will be given.

Sections

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G. P. Houston 3000 Manhatten Ave. Baltimore, Md.	Baltimore	J. W. Hammond 4 Alabama Ct. Baltimore 28, Md.
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G. E. Van Spankeren San Martin 379 Buenos Aires, Arg.	Buenos Aires	A. C. Cambre San Martin 379 Buenos Aires, Arg.
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C. K. Gieringer 3016 Lischer Ave. Cincinnati, Ohio	Cincinnati February 15	F. W. King RR 9 Box 263 College Hill Cincinnati 24, Ohio
F. B. Schramm 2403 Channing Way Cleveland 18, Ohio	Cleveland February 24	J. B. Epperson Box 228 Berea, Ohio
Warren Bauer 376 Crestview Rd. Columbus 2, Ohio	Columbus March 11	George Mueller Electrical Eng. Dept. Ohio State University Columbus, Ohio
S. E. Warner Aircraft Electronics Assoc. 1031 New Britain Ave. Hartford 10, Conn.	Connecticut Valley February 17	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
J. G. Rountree 4333 South Western Blvd. Dallas 5, Texas	Dallas-Ft. Worth	J. H. Homsy Box 5238 Dallas, Texas
George Rappaport 132 East Court Harshman Homes Dayton 3, Ohio	Dayton March 3	C. J. Marshall 1 Twain Place Dayton 10, Ohio
C. F. Quentin Radio Station KRNT Des Moines 4, Iowa	Des Moines-Ames	F. E. Bartlett Radio Station KSO Old Colony Bldg. Des Moines 9, Iowa
A. Friedenthal 5396 Oregon Detroit 4, Mich.	Detroit February 18	N. C. Fisk 3005 W. Chicago Ave. Detroit 6, Mich.
R. W. Slinkman Sylvania Electric Products Emporium, Pa.	Emporium	F. M. Woodward 203 E. Fifth St. Emporium, Pa.
W. H. Carter 1309 Marshall Ave Houston 6, Texas	Houston	J. C. Robinson 1422 San Jacinto St. Houston 2, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	Indianapolis	Eugene Pulliam 931 N. Parker Ave. Indianapolis, Ind.
Karl Troeglen KCMO Broadcasting Co. Commerce Bldg. Kansas City 6, Mo.	Kansas City	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
R. W. Wilton 71 Carling St. London, Ont., Canada	London, Ontario	G. H. Hadden 35 Becher St. London, Ont., Canada
Berhard Walley Radio Corp. of America 420 So. San Pedro St. Los Angeles 13, Calif.	Los Angeles February 15	J. J. Fiske Westinghouse Electric Corp. 600 St. Paul Ave. Los Angeles 14, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	Louisville	D. C. Summerford Radio Station WKLO Henry Clay Hotel Louisville, Ky.
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K. R. Patrick RCA Victor Div. 1001 Lenoir St. Montreal, Canada	Montreal, Quebec March 9	S. F. Knights Canadian Marconi Co. P.O. Box 1690 Montreal, P. Q., Canada
L. A. Hopkins, Jr. 1711 17th Loop Sandia Base Branch Albuquerque, N. M.	New Mexico	T. S. Church 3079-Q-34th Street Sandia Base Branch Albuquerque, N. M.
J. W. McRae Bell Telephone Labs. Murray Hill, N. J.	New York March 2	R. D. Chipp DuMont Telev. Lab. 515 Madison Ave. New York, N. Y.
C. G. Brennecke Dept. of Electrical Eng. North Carolina State College Raleigh, N. C.	North Carolina-Virginia	C. M. Smith Radio Station WMIT Winston-Salem, N. C.
W. L. Haney 117 Bourque St. Hull, P. Q.	Ottawa, Ontario February 17	G. A. Davis 78 Holland Ave. Ottawa, Canada
M. W. Bullock Capital Broadcasting Co. 501 Federal Securities Bldg. Lincoln 8, Nebraska	Omaha-Lincoln	B. L. Dunbar Radio Station WOW Omaha, Nebraska
A. N. Curtiss Radio Corp. of America Camden, N. J.	Philadelphia March 3	C. A. Gunther Radio Corp. of America Front & Cooper Sts. Camden, N. J.
M. A. Schultz 635 Cascade Rd. Forest Hills Borough Pittsburgh, Pa.	Pittsburgh March 14	E. W. Marlowe Union Switch & Sig. Co. Swissvale P.O. Pittsburgh 18, Pa.
O. A. Steele 1506 S.W. Montgomery St. Portland 1, Ore.	Portland	F. E. Miller 3122 S.E. 73 Ave. Portland 6, Ore.
A. V. Bedford RCA Laboratories Princeton, N. J.	Princeton	L. J. Giacoletto 9 Villa Pl. Eatontown, N. J.
K. J. Gardner 111 East Ave. Rochester 4, N. Y.	Rochester February 17	Gerrard Mountjoy Stromberg-Carlson Co. 100 Carlton Rd. Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	Sacramento	W. F. Koch 1340 33rd St. Sacramento 14, Calif.
G. M. Cummings 7200 Delta Ave. Richmond Height 17, Mo.	St. Louis	C. E. Harrison 818 S. Kings Highway Blvd. St. Louis 10, Mo.
O. C. Haycock Dept. of Elec. Eng. University of Utah Salt Lake City, Utah	Salt Lake	M. E. Van Valkenburg Dept. of Elec. Eng. University of Utah Salt Lake City, Utah
C. L. Jeffers Radio Station WOAI 1031 Navarro St. San Antonio, Texas	San Antonio	H. G. Campbell 233 Lotus Ave. San Antonio 3, Texas
C. N. Tirrell U. S. Navy Electronics Lab. San Diego 52, Calif.	San Diego March 1	S. H. Sessions U. S. Navy Electronics Lab. San Diego 52, Calif.
F. R. Brace 955 Jones St. San Francisco 9, Calif.	San Francisco	R. A. Isberg Radio Station KRON 901 Mission St. San Francisco 19, Calif.
J. M. Patterson 2009 Nipsic Bremerton Wash.	Seattle March 10	J. E. Hogg General Electric Co. 710 Second Ave. Seattle 1, Wash.
F. M. Deerhake 600 Oakwood St. Fayetteville, N. Y.	Syracuse	S. E. Clements Dept. of Electrical Eng. Syracuse University Syracuse 10, N. Y.
A. R. Bitter 4292 Monroe St. Toledo 6, Ohio	Toledo	J. K. Beins 435 Kenilworth Ave. Toledo 10, Ohio

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T. J. Carroll National Bureau of Stand. Washington, D. C.	WASHINGTON March 14	P. DeF. Mekeel Civil Aeronautics Adm. Commerce Bldg. Washington, D. C.
J. C. Starks Box 307 Sunbury, Pa.	WILLIAMSPORT	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

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J. C. Ferguson Farnsworth Television & Radio Co. 3700 E. Pontiac St. Fort Wayne, Ind.	FORT WAYNE (Chicago Subsection)	S. J. Harris Farnsworth Television and Radio Co. 3702 E. Pontiac Fort Wayne 1, Ind.
E. Olson 162 Haddon Ave., N Hamilton, Ont., Canada	HAMILTON (Toronto Subsection)	E. Ruse 195 Ferguson Ave., S. Hamilton, Ont., Canada
A. M. Glover RCA Victor Div. Lancaster, Pa.	LANCASTER (Philadelphia Subsection)	C. E. Burnett RCA Victor Div. Lancaster, Pa.
H. A. Wheeler Wheeler Laboratories 259-09 Northern Blvd. Great Neck, L. I., N. Y.	LONG ISLAND (New York Subsection)	M. Lebenbaum Airborne Inst. Lab. 160 Old Country Rd. Box 111 Mineola, L. I., N. Y.
L. E. Hunt Bell Telephone Labs. Deal, N. J.	MONMOUTH (New York Subsection)	G. E. Reynolds, Jr. Electronics Associates, Inc. Long Branch, N. J.
J. B. Minter Box 1 Boonton, N. J.	NORTHERN N. J. (New York Subsection)	A. W. Parkes, Jr. 47 Cobb Rd. Mountain Lakes, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (Chicago Subsection) January 20	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
R. M. Wainwright Elec. Eng. Department University of Illinois Urbana, Illinois	URBANA (Chicago Subsection)	M. H. Crothers Elec. Eng. Department University of Illinois Urbana, Illinois
S. S. Stevens Trans Canada Airlines Box 2973 Winnipeg, Manit., Canada	WINNIPEG (Toronto Subsection)	S. G. L. Horner Hudson's Bay Co. Brandon Ave. Winnipeg, Manit., Canada

Calendar of COMING EVENTS

1949 IRE National Convention, New York City, March 7–10

Winter Meeting, Optical Society of America, New York City, March 10–12

1949 Chicago Production Show, Chicago, Ill., March 14–17

Annual Meeting, Armed Forces Communications Association, Washington, D. C., March 28–29

Semiannual Convention, Society of Motion Picture Engineers, New York City, April 4–8

AIEE Conference on Electron Tubes, Buffalo, N. Y., April 11–12

AIEE Southwest District Meeting, Dallas, Tex., April 19–21

IRE-RMA Spring Meeting, Philadelphia, Pa., April 25–27

TECHNICAL COMMITTEE NOTES

The **Standards Committee** met on November 18, 1948, and approved the standards on Methods of Testing Amplitude-Modulation Broadcast Receivers. The proposed definitions for Navigation Aids are still under revision by the Navigation Aids Committee and will be reviewed by the Definitions Co-ordinating Subcommittee. A subcommittee under the chairmanship of Richard F. Shea has been formed to write a standard for frequency-band designations which, if approved, may be published in the PROCEEDINGS. Standard bands in different ranges will be shown and the units to be used for each band. Two other subcommittees have been formed: the first to consider a process for numbering standards or separate items within any one standard; the second, under the chairmanship of A. G. Jensen, to begin work on a master index of all IRE definitions, the master index to be kept at Headquarters. . . . Under the leadership of L. J. Chu and J. W. Wright, subcommittees of the **Antennas Committee** are circulating to members an up-to-date list of the definitions being considered for transmission line and waveguide terms. . . . The **Audio and Video Techniques Committee** plans to review proposals for standardization, definitions, and methods of measurement and test at its February meeting. . . . The **Electron Tubes Committee** is considering radio-frequency operating tests for power-output high-vacuum tubes, methods of testing camera tubes, and methods of testing cathode-ray tubes. Work is nearly completed on grid-emission circuits. The committee will next review definitions prepared by its subcommittee on power-output high-vacuum tubes under the chairmanship of I. E. Mouromtseff, and its subcommittee on cathode-ray and television tubes under the chairmanship of L. B. Headrick. . . . The **Electronic Computers Committee** discussed tentative definitions of computers, which are being prepared by a subcommittee under the chairmanship of Robert Serrell. Activities of the other three subcommittees were discussed. Circuitry, Components, and Elements, under the chairmanship of N. Rochester; Applications, under the chairmanship of B. L. Havens; and Storage Devices, under the chairmanship of J. A. Rajchman. . . . The proposal for reference designations was reviewed and revised by the **Symbols Committee**, and a subcomittee under the chairmanship of C. Netizert was created in order to edit the proposal. . . . The **Wave Propagation Committee** met in order to make plans for gathering information to assist the JTAC in preparing its second report to the FCC. Chairman A. Gardner Fox of the subcommittee on Publications reported that the committee expected to have a list of wave propagation definitions completed shortly. These definitions will be presented to the Standards Committee and the main Wave Propagation Committee simultaneously. A number of the members of the Wave Propagation Committee have signed the petition for a professional group on antennas and wave propagation initiated by members of the Antennas Committee.

IRE People

A. Kyle Wing, Jr. (J'28–A'30–M'40–SM'43), of the Federal Telecommunication Laboratories, Inc., has been appointed chairman of the JETC's high-power electron tube committee.

Mr. Wing received the B.S. degree in electrical engineering from the Sheffield Scientific School of Yale University in 1930 and the M.S. degree from the Massachusetts Institute of Technology the following year. Mr. Wing worked with the Koester Radio Co., the Federal Telegraph Co., and the RCA Manufacturing Co. prior to joining the Federal Telecommunication Laboratory, where he now heads the electron-tube department.

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Victor P. Hessler (SM'46), who studied at Oregon and Iowa State Colleges, has joined the electrical engineering department of the University of Illinois as a full professor. Formerly, Dr. Hessler was a professor at the University of Kansas.

❖

Millard S. McVay (A'43), a graduate of Iowa State College who had been employed as electronics consultant for the U. S. Naval Research Laboratory in Washington, D. C., has been named assistant professor at the University of Illinois.

William G. Albright (A'45) has been promoted to an assistant professorship; **G. D. McMullen** (S'44–A'45), **Robert H. Wendt** (A'38–VA'39), and **M. D. Widenor** (S'46–A'47), formerly a graduate student at the University, have become instructors. **Donald E. Nelson** (S'41–A'41), a special research assistant, was advanced to the rank of special research associate.

❖

William A. Hayes (M'45) has been appointed section manager in the Westinghouse electronic tube sales department, which handles X-ray tube sales and all electronic tube contracts with U. S. government agencies.

Born in England, Mr. Hayes attended Columbia University and the University of Newark. He holds several patents on electron-tube developments and during World War II he discovered that hard-to-get steel in a radar-tube manufacturing operation could be replaced by uncooked spaghetti. In addition, he developed the "star-light" tube, a measuring device so sensitive that it can determine the feeble amount of electricity in the light of a dim star.

Mr. Hayes joined Westinghouse as a tube design engineer and later became an electron tube specialist. He is the author of technical articles on electronics, and he has lectured on the subject and conducted demonstrations before college and trade groups.

Leonard Mautner (M'46–SM'47), formerly manager and chief engineer of Allan B. DuMont's television transmitter division, has been appointed vice-president of the newly formed Television Equipment Corp., a subsidiary of the International Television Corp.

Born on October 30, 1917, in New York City, Mr. Mautner received the B.S. in electrical engineering from the Massachusetts Institute of Technology in 1939 and in that year joined the Macbeth Daylighting Corp. Leaving Macbeth in order to serve as a radio engineer with the Army Signal Corps, Mr. Mautner did graduate work at the same time for three years—at the Stevens Institute of Technology and at MIT. After serving for a short time with the National Broadcasting Company, he joined MIT's Radiation Laboratory as a member of the indicator group, later transferring to the Naval Research Laboratory in Washington, D. C. In 1945 he joined the DuMont Laboratories where he took charge of the development of television video equipment.

❖

William Brown (A'43–M'45), formerly a specialist in television transmission and reception equipment with the RCA Laboratories Industry Service Division, has joined the Television Equipment Corp. as secretary and chief engineer. A specialist in television transmission and reception equipment, he has served as a consultant to many organizations in the television field.

❖

Edwin F. Dillaby (A'29–VA'39) has been appointed chief engineer in charge of the newly formed tube division of Tracerlab Inc., Boston, Mass.

Mr. Dillaby was educated at the Lowell Institute and Northeastern University. After working in the Raytheon Production Corp.'s power tube division for nine years, he joined the staff of the Hytron Radio and Electronic Corp. in 1937 as chief design and production engineer. There he contributed substantially to pioneering design and production of the subminiature tubes important to the success of the wartime VT proximity fuze.

In his new position, Mr. Dillaby will take charge of the production and design of the latest types of radiation-sensitive tubes, as well as directing an extensive program of research and development of high-efficiency gamma-ray detectors, miniature Geiger tubes, photomultiplier tubes, and other types of particular interest to the field of radioactivity.

For the past three years Mr. Dillaby has served as a member of the Committee for Standardization of Cathode Nickel Material for Vacuum Tubes of the American Society for Testing Materials. He is also a member of the Engineering Societies of New England and the Yankee Radio Club.

Hendrik Johannes van der Bijl (M'17–F'28), author of the first book on vacuum tubes, died recently of cancer at his home in Johannesburg. A leading South African physicist, Dr. van der Bijl was responsible for the South African industrial revolution, and became the country's leading industrialist himself.

Born on November 23, 1887, at Pretoria, South Africa, Dr. van der Bijl received the B.A. degree from Victoria College, Stellenbosch; the M.A. and Ph.D. degrees from the University of Leipzig; and the honorary D.Sc. and LL.D. degrees from Stellenbosch University and Capetown University, respectively.

While serving as an instructor in physics at the Royal School of Technology, at Dresden, during 1912 and 1913, Dr. van der Bijl carried out researches which led in 1920 to his joining the Western Electric Co. as a research physicist. There he worked out of the theory of the thermionic tube.

In 1920 former premier Jan Christian Smuts induced Dr. van der Bijl to return to South Africa to become technical advisor on industrial development to the Department of Mines and Industries. In 1923 he left to help form the Electricity Supply Commission, established under his chairmanship to co-ordinate and develop South Africa's Power Supply Industry on a national basis. He also sponsored the South African Iron and Steel Industrial Corp. in 1928.

In 1937 Dr. van der Bijl founded the African Metals Corp., and two years later he initiated the Industrial Development Corp. of South Africa for the purpose of financing the building of industries on a sound economic basis. He acted as chairman for several years until the organization was established on a firm footing.

Soon after the outbreak of World War II, Dr. van der Bijl was appointed Director-General of War Supplies. In 1942 he was made Honorary Colonel of the South African Tank Corps, and the year afterward his duties were enlarged to include the organization of civilian as well as war supplies, and he became Director-General of Supplies. In 1946 the British government asked his advice on socializing the steel industry.

The author of a number of publications on scientific and engineering subjects, Dr. van der Bijl was elected Vice-President of the IRE for 1945. In 1943 he was elected Foreign Associate of the National Academy of Sciences, and in 1944 was made a Fellow of the Royal Society.

Joseph C. Ferguson

Chairman, Fort Wayne Subsection

Joseph C. Ferguson was born in Beaumont, Tex., in 1900. Entering Louisiana State University in 1926, he was graduated with the B.S.E.E. degree in 1930.

In the latter year Mr. Ferguson joined the Westinghouse Electric Company as a student and junior radio engineer. There he participated in the design of the first aircraft squadron communications equipment, the forerunner of the modern command set, and also assisted in the design and test of receivers for the Department of Commerce, now the Civilian Aeronautics Administration, and Navy shipboard transmitters.

After a year of graduate work in communications engineering at Harvard University, he joined RCA as a radio design engineer, and for the next four years was active in the development and design of high-frequency facsimile transmitters, television broadcasting equipment, a 325-Mc remote-to-studio portable television transmitter, and standard AM broadcast equipment.

In 1939 Mr. Ferguson became a senior radio engineer with the Farnsworth Television and Radio Corporation, and was placed in charge of television broadcasting equipment and design. He joined the IRE as an Associate Member in 1941. The year 1942 found him promoted to chief engineer of Farnsworth's electronic apparatus division, supervising the development and design of electronic equipment for the armed services. In 1946 he became a Senior Member of the IRE.

Mr. Ferguson is a member of the RMA Television Transmitter Committee, and served as chairman of their subcommittee on television studio facilities from 1944 to 1947.

F. Marvin Deerhake

Chairman, Syracuse Section

F. Marvin Deerhake was born on April 3, 1911, in Mankato, Minn. During the summer of 1930, while he was on vacation from the University of Illinois, he was employed by the Westinghouse Electric Company. In 1932 Mr. Deerhake was graduated from the University with the degree of Bachelor of Science in Electrical Engineering, and he joined the General Electric Company the following year. He has remained with that organization ever since.

After having completed the three-year GE advanced course in engineering, Mr. Deerhake was assigned in 1936 to the research laboratory, where he engaged in microwave development with magnetrons and velocity-modulated tubes. Two years later he joined the newly organized television section of the radio division.

In 1940 Mr. Deerhake joined the IRE as an Associate. The following year he was transferred to GE development work for government contracts; and, after the outbreak of war, he was put in charge of several radar projects. He was upgraded to the rank of IRE Member in 1944. When the war ended the following year, he returned to the television section. The year 1946 found Mr. Deerhake a Senior Member of the IRE, and project engineer in charge of General Electric's relay development.

Chairman of the organizing committee which brought the Syracuse Section of the Institute into being early in 1947, Mr. Deerhake served as Vice-Chairman of the Section during its first year of operation. He is a member of the Technology Club of Syracuse, of Tau Beta Pi, and of Eta Kappa Nu.

A Broad-Band Microwave Relay System Between New York and Boston*

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Summary—**This paper describes the principal features of a broad-band microwave relay system which has recently been installed between New York and Boston. The system operates at frequencies around 4,000 Mc and provides two two-way channels, each accommodating a signal-frequency band extending from 30 cps to 4.5 Mc. Noise and distortion characteristics are satisfactory for the transmission of several hundred simultaneous telephone conversations or a standard black-and white television program.**

Introduction

The rapid progress which has been made in microwave electronics and circuit techniques during the past few years has resulted in the development of broad-band radio relay systems capable of spanning considerable distances. In order to explore the possibilities in microwave relaying, and to evaluate some of the technical and economic factors under operating conditions, a relay system has recently been installed between New York and Boston. This system operates at frequencies in the vicinity of 4,000 Mc and provides two broad-band communication channels in each direction between the two cities. Each channel is designed to accommodate a band of signal frequencies extending from 30 cps up to 4.5 Mc, and has noise and distortion characteristics satisfactory for the transmission of several hundred simultaneous telephone conversations or a standard black-and-white television program. An important objective was to provide a high degree of reliability with unattended operation of the repeaters, so that the radio relay system could be integrated with wire and cable facilities in a nationwide network.

Selection of the Route

The map of Fig. 1 shows the route finally adopted for the system. The eight repeater sections range in length from 11 to 35 miles, with an average of 27.5 miles. The repeater locations were chosen after a careful study of topographic maps and a field inspection of a large number of prospective sites. The locations selected have elevations sufficient to provide line-of-sight transmission paths between repeater stations with tower heights not greater than 60 feet at any station. In order to guard against possible inaccuracies in the topographic maps, ground elevations at the proposed sites, and also at critical points along the transmission paths, were checked with a precision altimeter before the sites were actually purchased. As a final precaution, microwave transmission measurements were made on all except the 11-mile path, using portable radio test equipment.

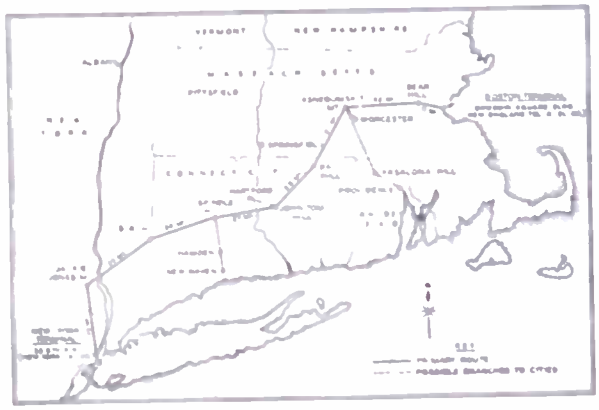

Fig. 1—Route of the New York–Boston microwave radio relay system.

In order to avoid excessive fading, it was decided to limit the path lengths in this system to a maximum of 35 miles and to require that the path clearance over intervening obstructions should correspond at least to the first Fresnel zone. The clearances were computed for a true earth radius in all cases; that is, no allowance was made for possible increases in clearance due to atmospheric refraction of the waves. These requirements were based largely on the results of microwave propagation tests[1,2,3] carried out before the New York–Boston project was undertaken. These propagation studies also influenced the decision to operate at 4,000 Mc. Broadly speaking, frequencies in the range from 1,000 to 10,000 Mc have possibilities for long-distance microwave relaying. It is advantageous to use as high a frequency as practicable, since this will result in high gain and directivity from antennas of moderate size. However, frequencies in the upper part of the range are subject to more severe fading than the lower frequencies and also are affected by rain attenuation, the severity of which increases rapidly with frequency. The choice of 4,000 Mc was considered to be a reasonable compromise between these opposing trends.

Form of the Repeater

The repeaters in the system are essentially broadband amplifiers which are capable of handling either am-

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† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] A. L. Durkee, "Results of microwave propagation tests on a 40-mile overland path," Proc. I.R.E., vol. 36, pp. 197–205; February, 1948.

[2] W. M. Sharpless, "Measurement of the angle of arrival of microwaves," Proc. I.R.E., vol. 34, pp. 837–845; November, 1946.

[3] A. B. Crawford and W. M. Sharpless, "Further observations of the angle of arrival of microwaves," Proc. I.R.E., vol. 34, pp. 845–848; November, 1946.

plitude- or frequency-modulated carriers, or various types of pulse modulation. Ideally, the repeater might consist of a number of amplifier stages operating at 4,000 Mc. While tubes are available which would provide the required gain and bandwidth at this frequency, they have not yet been developed to the point where the inherent noise level is low enough to permit their use in the early stages of a repeater. Consequently, in the New York–Boston repeaters, the incoming signal is brought down to an intermediate frequency of 65 Mc, and a good portion of the total required gain is obtained at this frequency. The signal is then translated back to the microwave range, where it is further amplified for transmission to the next repeater station.

In order to avoid interference between the output and input signals at a repeater, the incoming and outgoing microwave frequencies are made to differ by 40 Mc. This 40-Mc shift is alternately upward and downward in successive repeaters, so that two frequencies 40 Mc apart are required for transmission in one direction through the system. However, these same two frequencies are also used for transmission in the opposite direction. This is accomplished by allocating the frequencies so that, at any repeater station, one frequency is received from both directions and the other frequency is transmitted in both directions. This takes care of one of the two-way channels in the system. The other two-way channel utilizes two additional frequencies which also are 40 Mc apart, and are separated from the corresponding frequencies of the first pair by 200 Mc.

Repeater Elements

The block diagram of Fig. 2 shows the essential elements of a repeater. It also shows the gains and losses in the various elements, as well as signal and oscillator frequencies for a repeater at which the incoming and outgoing microwave frequencies are 3,930 and 3,970 Mc, respectively. These are the two frequencies which carry one of the two-way channels. The corresponding frequencies for the other channel are 4,130 and 4,170 Mc.

There are two transmitting and two receiving antennas at each repeater station. The two channels coming from New York are received on one receiving antenna and sent on toward Boston by one transmitting antenna. The other two antennas receive the two channels from Boston and transmit them toward New York. The antennas are of the shielded-lens type[4] and are constructed of stainless steel, with apertures 10 feet square. They have a measured gain of about 39 db at 4,100 Mc and provide high directional discrimination, particularly in the rear quadrants where the response is 68 db or more below maximum.

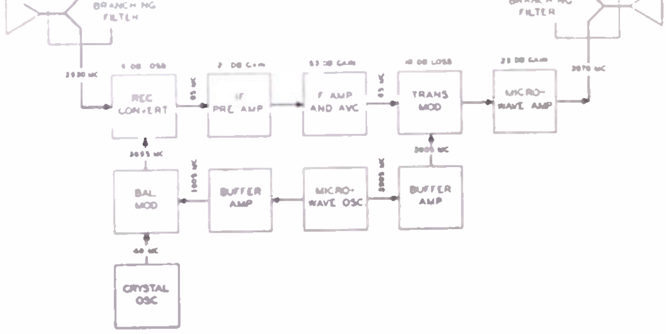

Fig. 2—Block diagram of a repeater.

The two incoming channels picked up by a receiving antenna are separated by waveguide branching filters and then delivered to individual repeaters, as indicated in Fig. 2. Translation to the 65-Mc intermediate frequency is accomplished in a balanced converter[5] using silicon crystals. The converter and first two stages of the if amplifier are mounted as a unit at the branching filter output. The noise figure[6] of the converter-preamplifier combination is about 14 db.

The if main amplifier comprises seven stages and includes an avc system of the grid-bias type which holds the output substantially constant for inputs varying over a 30-db range. This accommodates variations in received field intensity from about 10 db above to about 20 db below the free-space value. Most of the tubes in the if amplifier are of the Western Electric type 404-A. This tube is somewhat similar to the 6AK5, but has about twice the figure of merit in practical circuit arrangements.

The transmitting modulator, which follows the if amplifier, shifts the signal from 65 Mc back to the microwave range. It is similar to the receiving converter, except that it is designed to operate at higher power levels. The if input to the modulator is 50 mw and the microwave beating oscillator input is 750 mw. The desired modulation sideband is selected by a waveguide filter and delivered to the microwave amplifier at a level of about 5 mw.

The microwave amplifier is made up of four stages, each comprising a two-gap velocity-variation tube with its associated input and output cavities and focusing magnet. The amplifier is mounted in a temperature-controlled compartment. Tuning is accomplished by means of threaded studs which project into the cavities. Also extending into the cavities are small fins of resistive material which can be rotated to vary the loading. Coupling between stages is obtained with small rotatable loops connected by short coaxial lines. A photograph of the amplifier is shown in Fig. 3. The amplifier delivers approximately 1 watt of microwave power to the transmitting antenna through a waveguide filter similar to the branching filter at the repeater input.

[4] W. E. Kock, "Metal lens antennas," Proc. I.R.E., vol. 34, pp. 828–837, November, 1946.

[5] C. F. Edwards, "Microwave converters," Proc. I.R.E., vol. 35, pp. 1181–1192, November, 1947.

[6] H. T. Friis, "Noise figures of radio receivers," Proc. I.R.E. vol. 32, pp. 419–423, July, 1944.

The method of providing beating-oscillator power for the receiving converter and the transmitting modulator is indicated in Fig. 2. A stabilized microwave oscillator generates power at the frequency required for the transmitting modulator. Part of this power is delivered to the modulator through a buffer amplifier, and part is modulated with the output of a 40-Mc crystal oscillator. The appropriate sideband component produced by this auxiliary modulator is selected by a filter and supplied to the receiving converter. In this way the 40-Mc frequency shift in the repeater is accomplished. Frequency variations in the microwave oscillator are held within limits of approximately ±200 kc by an automatic control system. However, since the same oscillator is used for both modulation steps in the repeater, the outgoing microwave frequency is unaffected by frequency variations in the microwave oscillator and depends only on the stability of the 40-Mc crystal oscillator. Variations in the latter do not exceed ±2 kc. With this method of operation there is no accumulation of large frequency errors as the signals progress through the system. The stability of the outgoing frequency at a transmitting terminal is determined mainly by the stability of a microwave oscillator which is identical to those used in the repeaters.

Frequency stabilization of the microwave oscillators is accomplished by means of an electromechanical servo system.[7] A sample of the oscillator output power is fed into a waveguide discriminator, and a voltage proportional to the frequency error of the oscillator is derived. This voltage is used to drive a servomotor which is geared to a potentiometer. The potentiometer varies the repeller voltage of the reflex oscillator tube in such a way as to correct the frequency error. The frequency-determining element of the discriminator is a high-Q invar cavity in a temperature-controlled enclosure. The over-all stability achieved in the oscillator is of the order of ±200 kc.

Fig. 4 is a front view and Fig. 5 is a rear view of the repeaters in a typical installation. Each of the two

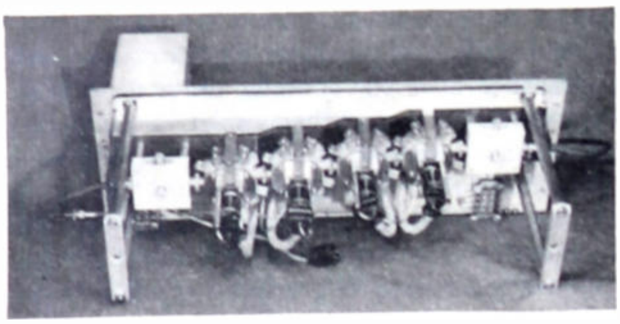

Fig. 3—Four stage microwave amplifier.

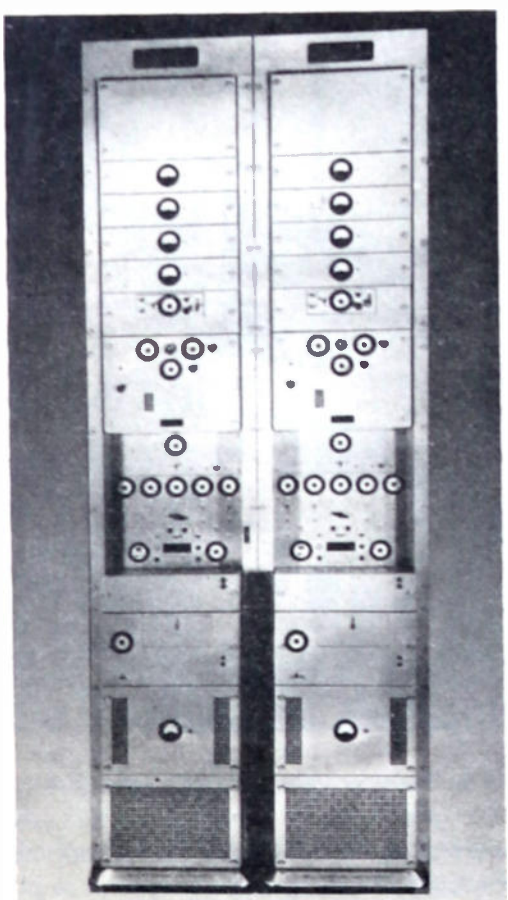

Fig. 4—Front view of microwave repeaters.

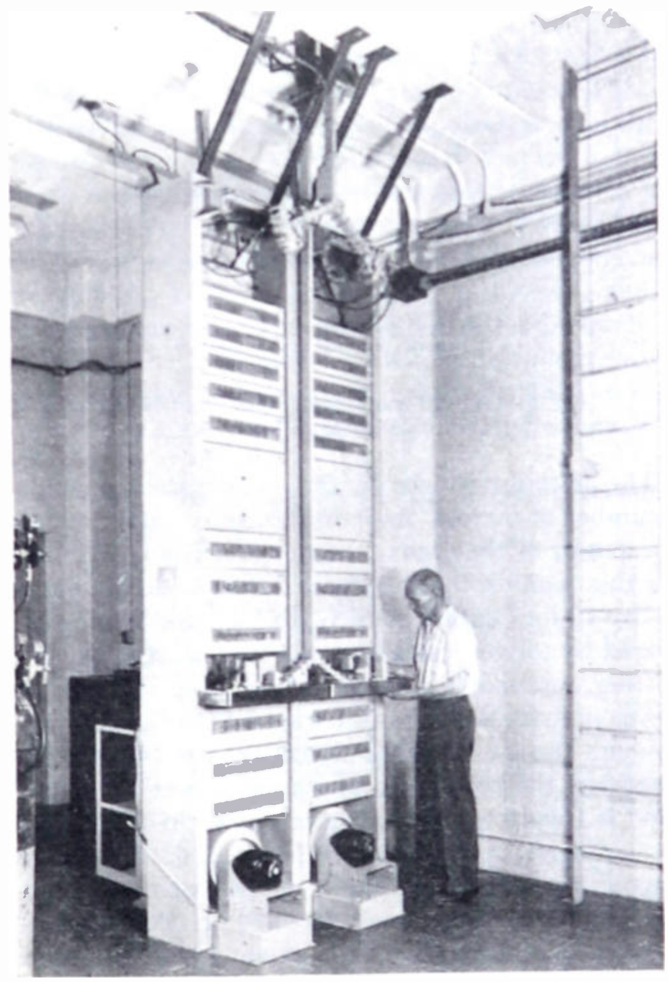

Fig. 5—Rear view of microwave repeaters.

[7] V. C. Rideout, "Automatic frequency control of microwave oscillators," Proc. I.R.E., vol. 35, pp. 767–771; August, 1947.

mounting frames contains the radio equipment and power-supply equipment for a complete one-way repeater. In Fig. 5 the receiving branching filters, with converters and preamplifiers attached, can be seen at the top of the frames. Further down are the similar filters through which the outgoing signals are fed into the waveguide leading to the transmitting antenna. The blowers at the base supply air to the various components through vertical ducts located in the sides of the frames.

Repeater Performance

Fig. 6 shows amplitude and delay distortion characteristics for a typical repeater. Amplitude variations are less than 0.1 db over a 10-Mc band. The delay distortion increases from zero at midband to an average of about 20 millimicroseconds at the edges of the 10-Mc band. The cumulative effect of this amount of delay distortion in each of seven repeaters plus the terminal stations was considered to be excessive. Consequently, an if equalizing network was designed and inserted between the if preamplifier and main amplifier at each repeater. This reduced the delay distortion over the 10-Mc band by a factor of about 10 to 1.

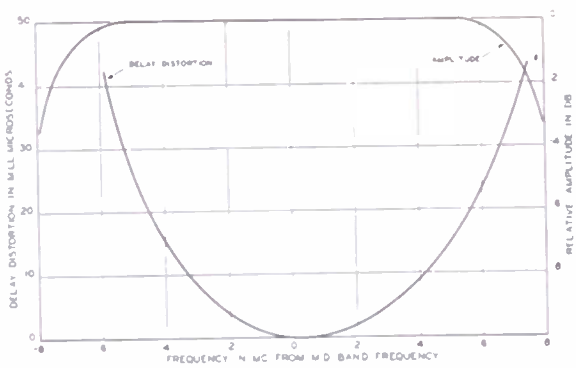

Fig. 6—Over-all transmission characteristics of a typical repeater.

The noise introduced by the relay system depends on a number of factors. As indicated previously, the average length of the eight repeater sections is 27.5 miles. For this distance the free-space loss at 4,000 Mc between 10- by 10-foot antennas is 57 db.[8] The signal power delivered by the receiving antenna for 1 watt in the transmitting antenna is 1 watt − 57 db. The thermal noise power in a 10-Mc band is 1 watt − 134 db. Taking into account the 14-db noise figure of the converter-preamplifier, the receiver input noise power for a 10-Mc band is 1 watt − 120 db. Hence, the ratio of rms carrier to rms noise at the receiver input is 63 db. The corresponding ratio at the input of the last receiver in an eight-link system becomes 54 db. For television transmission with low-index FM, as employed in the New York–Boston system, this carrier-to-noise ratio results in a ratio of peak-to-peak video signal to rms noise of 62 db. With 20-db fades simultaneously on all eight links, an unlikely condition but one which the system will tolerate, this ratio becomes 42 db.

[8] H. T. Friis, "A note on a simple transmission formula," Proc. I.R.E., vol. 34, pp. 254–256; April, 1946.

Measurements on the paths at the time the repeater sites were selected, and observations on the system since it was put into operation, have shown that the transmission losses on all sections are very close to the free-space values in the absence of fading. Continuous recording of transmission on the New York–Jackie Jones link, the longest in the system, was carried on for a period of two months during the summer season. These records showed that the received field intensity was as much as 20 db below the free-space value due to fading only about 0.04 per cent of the time.

Repeater Station

A photograph of a typical repeater station is shown in Fig. 7. The four antennas are mounted on a platform about 60 feet above the ground. This provides the necessary clearance over surrounding trees. At four of the repeater stations the additional height provided by the platform is not required, and the antennas are mounted directly on the roof. Waveguides connect the antennas to the repeaters, which are located on the second floor of the building. There are four repeaters in the station, two for each direction of transmission. The first floor of the building is occupied by heating, ventilating, and power equipment.

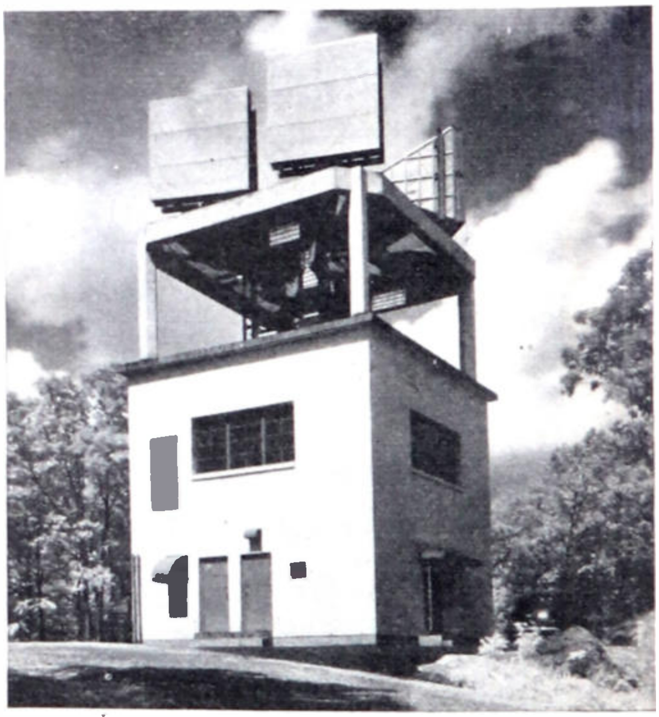

Fig. 7—Repeater station at Jackie Jones Mountain.

Commercial power at 230 volts is normally used at the repeater stations. Emergency equipment is provided in order to avoid service interruptions in case of failure of the regular supply. If the commercial power fails, the load is taken over automatically by a 230-volt ac converter driven by a 130-volt battery. The battery has capacity sufficient for six hours of operation. After two minutes of battery operation, however, a 15-kw gas-engine generator automatically starts, and after a five-minute warm-up period it takes the load if the commercial power is still off. The engine generator will carry the load until the gasoline supply is exhausted. If the gas engine should fail to start, operation would continue from the battery.

Since the repeaters operate without attendance, except for periodic maintenance visits, an alarm system is provided which automatically notifies maintenance personnel at near-by telephone offices when trouble develops at the repeater stations. The alarm circuit is combined with a telephone order-wire circuit which links the repeaters and terminal stations. The order wire is a four-wire circuit extending between New York and Boston, with four-wire bridging loops from intermediate telephone offices to the seven repeater stations. The pairs to each repeater station are simplexed to provide facilities for a ten-channel dc alarm circuit to the telephone office responsible for maintenance. The alarm system is arranged to give audible and visible indications of various trouble conditions that may occur in the radio and power equipment at the repeater stations.

Terminal Equipment

Frequency modulation with a low modulation index is the method used at present for transmission in the relay system. At New York and Boston, terminal equipment is provided for generating frequency-modulated waves for the outgoing microwave channels and receiving them from the incoming channels. The transmitting terminal equipment accepts the television or multiplex telephone input signal and frequency-modulates it on a 65-Mc carrier. This modulated carrier is then delivered to the transmitting portion of a microwave repeater which is essentially the same as the repeaters in the relay stations. The receiving terminal equipment accepts a 65-Mc frequency-modulated carrier from the receiving portion of the microwave repeater, demodulates the carrier, and recovers the video signal. The division of the standard repeater for this terminal application occurs between the if amplifier output and the transmitting modulator input.

A block diagram of the principal elements in the FM transmitting terminal is shown in Fig. 8. The video input signal may range from 0.25 to 2.5 volts, peak-to-peak, and it is first adjusted to the required level by an amplifier and attenuator in the video section of the terminal. It is then applied to the 65-Mc oscillator through two cathode-follower stages. Direct coupling is employed between the cathode followers and the oscillator to permit transmission of the dc component of a television signal. The oscillator itself is essentially a two-stage broad-band amplifier with positive feedback.

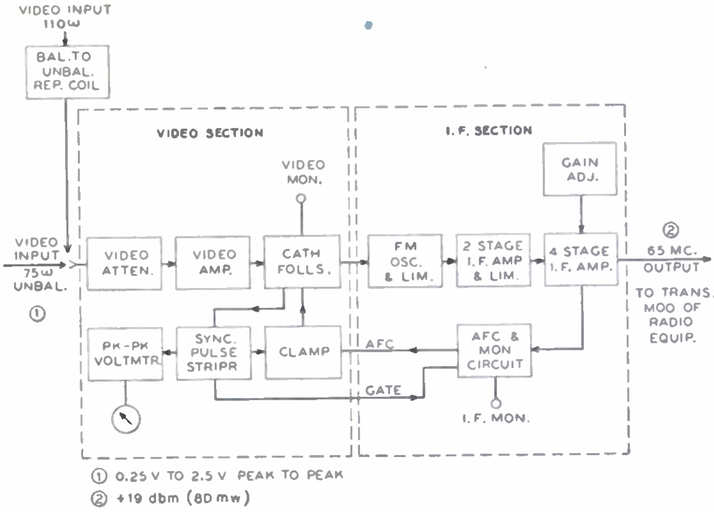

Fig. 8—Block diagram of the 65-Mc FM transmitting terminal equipment.

The phase of the feedback voltage is varied electronically by the modulating signal to produce a frequency-modulated output wave. The output is amplified, limited, and delivered to the microwave transmitter at a level of 80 mw.

For television transmission, the afc system shown in Fig. 8 is employed. A portion of the incoming video signal is diverted to the pulse stripper which supplies sync pulses to the afc gate and clamp circuits. The afc is enabled by the gate during sync pulses, and in these intervals the clamp sets the grid bias of the first cathode follower in accordance with the voltage delivered by the afc circuit. This fixes the frequency of the FM oscillator during the sync pulse intervals. The oscillator frequency corresponding to sync pulse peaks is 67 Mc, and it swings between this value and 63 Mc with full modulation. When the equipment is used for telephone transmission, the gate and clamp circuits are disabled and the afc circuit is arranged to hold the mean frequency of the oscillator at 65 Mc.

Fig. 9 is a block diagram of the receiving terminal

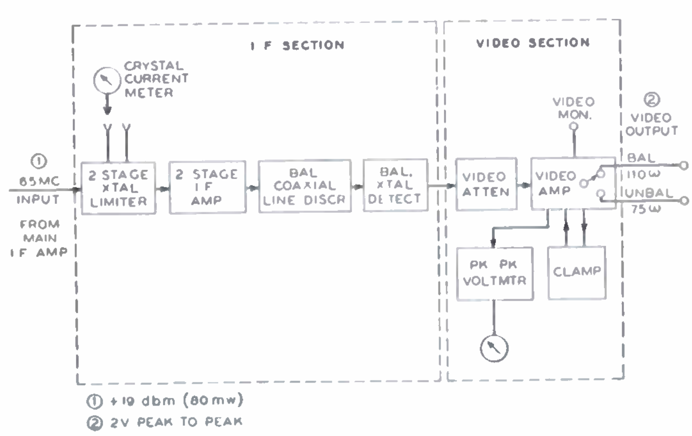

Fig. 9—Block diagram of the 65-Mc FM receiving terminal equipment.

equipment. A 65-Mc frequency-modulated carrier at a level of 80 mw is obtained from the microwave receiver. This carrier is passed through a limiter and if amplifier, and then is delivered to a balanced discriminator and crystal detector where the video signal is recovered. The detector is followed by an attenuator and video amplifier, by means of which the output is adjusted to a value of 2 volts, peak-to-peak. Associated with the video amplifier is a clamp circuit for reducing the effects of 60-cps interference and low-frequency noise during television transmission.

A photograph of the FM terminal equipment is shown in Fig. 10. The two outside frames are identical, and each contains the transmitting and receiving terminals for one two-way channel. A picture monitor is provided in each frame for viewing incoming or outgoing television pictures. The middle frame contains test equipment, including means for measuring the frequency swing of either FM transmitter. This is done by comparing the FM wave with the output of a calibrated test oscillator, using the oscilloscopes associated with the picture monitors. An electronic switch operating at a 2-kc rate alternately connects a sample of the FM transmitter output and the if test oscillator output through an FM detector to the vertical plates of the oscilloscope. The horizontal sweep is adjusted so that the wave form of the video signal is displayed. The test oscillator produces a horizontal line at a height determined by the oscillator frequency. By adjusting the frequency so that the horizontal line coincides with the appropriate portions of the video pattern, the frequency excursions of the FM transmitter, as well as the frequency of the sync pulses or other parts of the signal, can be determined from the frequency calibration of the oscillator.

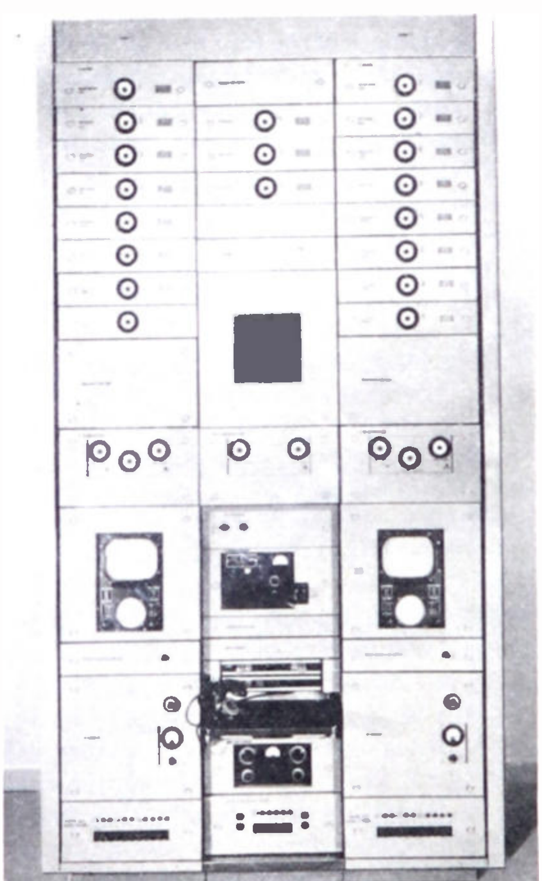

Fig. 10—Front view of the FM terminal equipment.

Conclusion

The relay system was formally opened for experimental service on November 13, 1947, and at that time its capabilities for television and multichannel telephone transmission were demonstrated. As a part of the television demonstration, the two two-way channels were connected in tandem to form a double loop, and provision was made for viewing a test pattern at New York before and after making two round trips to Boston. It was difficult to detect any impairment in the test pattern after transmission through this 880-mile system. For the telephone demonstration a standard type-K carrier system was employed, which permitted twelve conversations to be carried on simultaneously between talkers in New York and Boston. The type-K equipment was used purely as a matter of convenience, and was not indicative of the ultimate capacity of the system for multiplex telephony. Measurements of noise and intermodulation which have been made since the demonstration indicate that at least 240 channels of a type-L1 carrier system can be handled without serious degradation of the speech channels.

The operating experience now being obtained with this New York–Boston installation will be useful in the development of similar systems for other routes, and will help to determine how radio relaying can best supplement wire and cable facilities in the telephone network.

Acknowledgment

The successful outcome of this project was the result of the co-operative efforts of many individuals in the Bell Telephone Laboratories and other units of the Bell System. Fundamental research on the components for the system was done by the Laboratories groups at Holmdel and Deal, N. J.[9] Representatives of the Long Lines Department of the American Telephone and Telegraph Company and the operating companies in the New England area collaborated with the Laboratories in selecting and obtaining the repeater station sites. The Long Lines Department took care of the improvement of the sites and construction of the buildings. The project was under the general supervision of Ralph Bown, director of research of the Laboratories, and G. N. Thayer was the project engineer.

[9] An account of this work is given in the paper by H. T. Friis "Microwave repeater research," *Bell Sys. Tech. Jour.*, vol. 27, pp. 183; April, 1948.

Ionization Chambers*

JOHN A. VICTOREEN†

The field of electronic instrumentation and controls as applied to nuclear processes and the production of fissionable materials is of interest to the membership of the Institute. There are accordingly included in the PROCEEDINGS OF THE I.R.E. tutorial and comprehensive papers dealing with the atomic structure and physical particles involved therein, as well as specific devices and methods used in the nuclear field. The following instructive review is presented in this series of papers.—*The Editor.*

INTRODUCTION

IONIZATION CHAMBERS have been used for many years for the detection of ionizing radiation. The old gold-leaf electroscope was probably the first use of an ionization chamber. It was, in fact, the first complete instrument which could be used to quantitatively estimate radioactive properties. Such early instruments consisted of a pair of gold leaves suspended in a container and insulated from the walls of the container so that they might be electrically charged. The resulting electrostatic forces caused the gold leaves, which were very thin, to be deflected. and the amount of deflection was observed upon a scale. Early observers noted that it was not possible to maintain a charge upon the leaves indefinitely but that, regardless of the quality of the insulation, a uniform decrease in charge occurred. It was not until the discovery of cosmic radiation and radioactivity that the decrease in charge could be explained as resulting from ionization which occurs as a result of natural level of radiation or "background." The gold-leaf electroscope was not a very accurate instrument, but it was, nevertheless, complete. The gold leaves were fundamentally a satisfactory electrostatic voltmeter. Their electrostatic capacitance constituted a highly insulated capacitor for storing electrons. The electrostatic field within the container served to prevent recombination of ions and to collect electrons. Finally, the container enclosed a volume of air and was, indeed, an ionization chamber.

Ionization chambers as indicating devices are fundamentally simple, and it is only when critical quantitative data are required under specific conditions that difficulties become apparent. Early workers with gold-leaf electroscopes found that the charge disappeared more rapidly when the instrument was illuminated. A zinc or cadmium plate connected to the leaves and exposed to sunlight caused a very rapid discharge. We know now that these metals emitted photoelectrons, and that the kind of metal and the kind of light are so important that the one under controlled conditions may become a powerful tool for the investigation of the other.

If the term "ionization chamber" is to have a meaning or a use, these must be explicitly stated or implied in the description of the chamber. A gas phototube is certainly an ionization chamber, responsive to certain wavelengths. A Geiger-Mueller counter tube is definitely an ionization chamber, responsive to many wavelengths and particles as well. Gas phototubes and Geiger-Mueller tubes are so well known in their respective fields that they are not even popularly considered as ionization chambers.

In this paper, prime consideration is given to ionization chambers suitable for measuring radiation of X-ray wavelengths, which also includes the longer gamma wavelengths. This is the region where ionization chambers are of particular value in this atomic age.

Roentgenologists using X radiation for therapeutic purposes were the first to feel the need of an accurate radiation-measuring instrument, and after much international collaboration a unit, the roentgen, was finally evolved and accepted. As a unit, it defines an amount of ionization produced in 1 cc of air under standard conditions by a beam of X or gamma radiation. Thus, a roentgenologist may deliver to a patient 250 roentgens as a dose. Ionization chambers for indicating roentgens for therapeutic purposes are called thimble chambers because of their size and shape. Air was specified in the definition of the roentgen because it is composed of nitrogen and oxygen, which closely approach human tissue in atomic number. Thimble chambers can be made nearly independent of wavelength over the normal therapy range of wavelengths, so that measurement is independent of the particular X-ray generator being used.

The roentgen, however, was originally defined to provide a unit which could be practically standardized and practically used for the wavelengths of radiation used at the time of agreement. Since then, the development of electron accelerators and the availability of radioactive isotopes have made possible radiation therapy at such short wavelengths that it has not, as yet, been possible to develop suitable primary standards for the calibration of practical clinical measuring instruments.

On the other hand, beryllium-window X-ray tubes have made available very-long-wavelength radiation where the wall of a thimble ionization chamber must be made very thin to avoid wall absorption. This is in contrast to the short-wavelength requirement that the wall be of sufficient thickness to provide equilibrium with high-speed electrons generated in the air surrounding the chamber. There are many difficulties which have yet to be resolved.

* Decimal classification: 539.7×621.375. Original manuscript received by the Institute, October 21, 1948.

† The Victoreen Instrument Company, Cleveland, Ohio.

Medical use of energized particles, such as electrons, neutrons, etc., present special problems, both biologically and in instrumentation. Ionization chambers calibrated in roentgens deliberately exclude the simultaneous measurement of particles and electromagnetic radiation in a unit which is related to biologic reaction.

The Physics of Radiation

An ionization chamber may be defined as a volume of ionizable substance from which an ionization current may be collected when irradiated by ionizing radiation. Ionization chambers are used for quantitatively indicating gamma rays, X rays, electrons, alpha particles, or any other energized particles which produce ionization either by primary or secondary effects, and may be classified into two distinct categories: those which may be calibrated to measure a particular type of radiation in an accepted unit, and those which merely indicate radiation in arbitrary units. These two types of chambers are analogous to an ammeter and a galvanometer; the ammeter being calibrated directly in international amperes, while the galvanometer indication is an arbitrary number.

The principles involved in the measurement of ionizing radiation may be most easily understood by considering the range of wavelengths from about 0.1 to 1.0 angstroms or 1×10^{-9} to 1×10^{-8} cm. This is the range of wavelengths used in X-ray therapy and for which an internationally accepted unit, the roentgen, has been defined as a standard degree of ionization produced in air. For this reason, present consideration will be confined to this region, although the principles involved find general application.

Obviously, all of the energy collected from an ionization chamber must be obtained by absorbing it from the radiation which penetrates the ionizable volume so that the amount of current obtained at a given wavelength is determined by the absorption properties of the ionizable substance. For electromagnetic radiation of wavelengths from about 0.01 to 1.0 Å, which includes the gamma rays of radium, to long X rays, the law of absorption for thin absorbers is given by

$$\frac{E_P}{E_0} = e^{-(\mu/\rho)xd} \qquad (1)$$

where E_0 is the energy of the beam impinging the front surface of the absorber, E_P is the final energy of the beam after traversing the absorber, e is the base of the natural system of logarithms, and the term $-\mu xd/\rho$ is the specific absorption coefficient for the absorber in which x is thickness in centimeters and d is density. μ/ρ is the mass absorption coefficient for unit mass of the absorbing material.

That part of the energy which is merely scattered and not truly absorbed is, of course, not collectable as electrical energy.

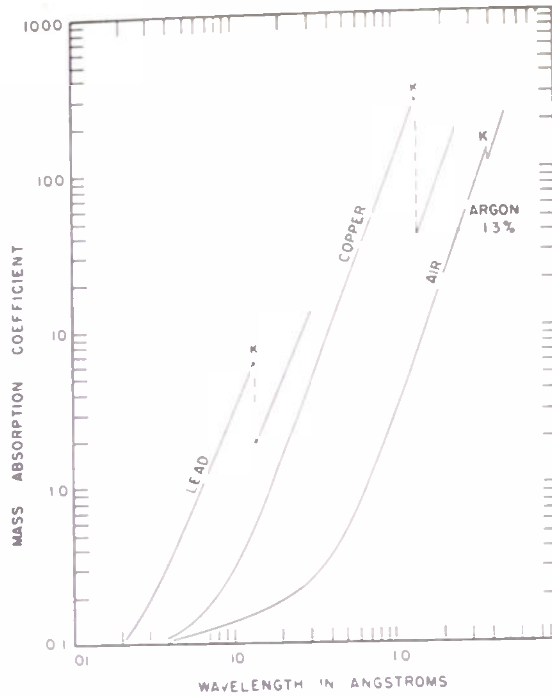

Fig. 1—Mass absorption coefficients of lead, copper, and air, showing the K critical absorption discontinuity.

Fig. 1 shows the variation of the mass absorption coefficient for air, copper, and lead for various wavelengths. The K discontinuity shown for copper and lead exist for each element, occurring systematically at shorter wavelengths as atomic number is increased. The small discontinuity shown for air results from the presence in atmospheric air of 1.3 per cent argon, which has a K discontinuity at 3.865 angstroms.

The absorption of radiation in the wavelength region shown is the result of two distinct processes, the production of (1) recoil electrons and (2) photoelectrons in the absorber. The mass absorption coefficient μ/ρ is, therefore, composed of two coefficients: the photoelectric absorption coefficient τ/ρ, and the scattering or recoil coefficient σ/ρ.

$$\frac{\mu}{\rho} = \frac{\tau}{\rho} + \frac{\sigma}{\rho}. \qquad (2)$$

At the shorter wavelengths, the absorption is almost entirely due to the production of recoil electrons; and at the long wavelengths, it is due almost entirely to the production of photoelectrons.

At wavelengths shorter than about 0.01 Å, a third type of absorption occurs as a result of pair production, and another coefficient must be added in (2). This, however, will not be included for the present. Equation (1) applies to thin absorbers and does not apply rigorously to thick absorbers. It is also important to note that, although $1-E_P/E_0$ is the fraction of the energy lost from the beam of radiation, it does not necessarily follow that all of the energy can be collected as electrical energy. Nevertheless, the mechanism of absorption is one of ionization, and is required to explain the determinants in ion-chamber design.

Electromagnetic radiation, such as X rays or gamma rays, is assumed to be propagated as discrete quanta or units of energy which, as particles, decrease in mass as the energy content decreases. Each quantum at a wavelength λ contains energy E_Q, as represented by the expression:

$$E_Q = \frac{1}{2} mv^2 = h\nu = h\frac{c}{\lambda} \tag{3}$$

where ν is the frequency, v is velocity, c is the velocity of light, m is mass, and h is Planck's constant of action.

For present purposes, an atom may be described as comprising a nucleus having a number of positive charges which determine its atomic number Z. Thus, hydrogen with one positive nuclear charge has an atomic number $Z=1$, and oxygen with eight charges has an atomic number $Z=8$. Each positive charge on she nucleus binds one electron to the atom. Oxygen in its stable state will then contain eight electrons, and these will rotate around the nucleus in orbits having definite binding energies which are a characteristic of the particular atom, the highest binding energy holding the innermost or K electron, while the least binding energy is required to hold the outermost electron, which may be in the K, L, M, N, etc., orbit.

The curve in Fig. 1 showing the mass absorption coefficient of air is typical of all elements. The discontinuity due to the presence of argon shows the importance of small impurities of higher atomic number. Air has an effective atomic number of about 7.3, while argon has an atomic number of 18. All elements have discontinuities, and so care must be exercised in ionization-chamber design to prevent the existence of a discontinuity in the wavelength region where operation is desired.

The mass absorption coefficients of each element also increase systematically as atomic number is increased. The mass photoelectric-absorption coefficient τ/ρ is given within experimental error[1] for a given element of atomic number Z by the following:

$$\left(\frac{\tau}{\rho}\right)_z = \left\{3 - \frac{\nu_3}{\nu_0}\right\} \frac{\nu_1\nu_2}{\nu_0^3} \times \frac{K}{A} \tag{4}$$

where K is a constant for all elements, A is atomic weight, ν_0 is the frequency of the incident radiation in cps, and ν_1, ν_2, and ν_3 are assumed to be critical frequencies characteristic of the particular atom. When the values of ν_1, ν_2, and ν_3 and A, the atomic weight, are given, τ/ρ may be calculated for an incident frequency ν_0.

In general, τ/ρ increases inversely as the cube of the incident frequency until the K absorption discontinuity is approached, when the term ν_3/ν_0 becomes appreciable.

Each of the critical frequencies, ν_1, ν_2, and ν_3, may be calculated for an element of atomic number Z.

$$\nu_c = R(Z - s)^2\left(\frac{1}{n_1^2} - \frac{1}{n_2^2}\right), \tag{5}$$

where R is Rydberg's constant, Z is atomic number, s is a small screening constant, and n_1 and n_2 are quantum numbers 1, 2, or 3, etc., which are constant between wavelength discontinuities.

$$\nu_1\nu_2 = R^2(Z - s)^4\left(\frac{1}{n_1^2} - \frac{1}{n_2^2}\right)\left(\frac{1}{n_3^2} - \frac{1}{n_4^2}\right), \tag{6}$$

from which it may be observed that τ/ρ varies almost as the fourth power of atomic number Z.

Actually, (5) and (6) hold only for low atomic numbers, and relativity corrections must be added for higher atomic numbers. Details of this, together with calculated values of τ/ρ, have been given elsewhere.[1]

Process of Absorption

Radiant energy is absorbed by an atom of any substance by the interaction of a quantum with the orbital electrons of the atom. The impinging quantum may give up all or part of its energy, depending upon the type of the collision with the electrons. A high-energy quantum may collide with electrons of many atoms, losing energy by the process of attrition, until finally coming to rest. Thus, a quantum may be simply scattered from atom to atom, changing its direction each time, until it may, indeed, by the laws of chance, return to its point of origin. Each time the quantum bounces off an electron, the electron must recoil, so that recoil electrons are associated with the scattering process.

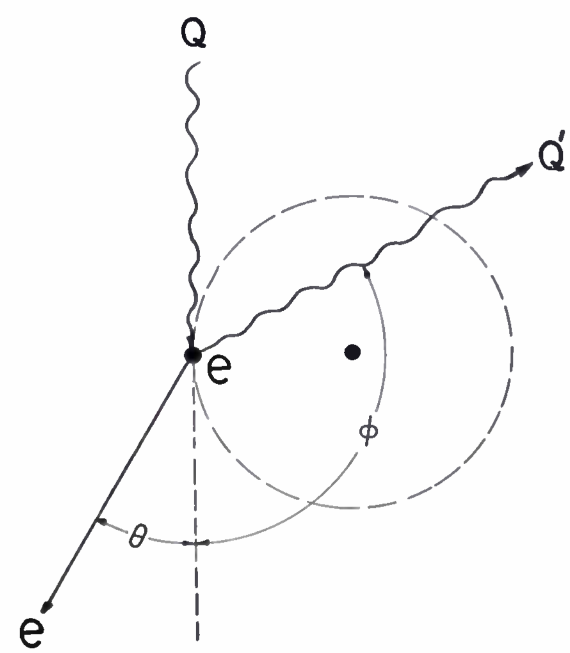

Fig. 2—Scattering of a quantum of radiation by an orbital electron *e* of an atom.

[1] J. A. Victoreen, "The absorption of incident quanta by atoms, as defined by the mass photoelectric absorption coefficient," *Jour. Appl. Phys.*, vol. 19, pp. 855–860; September, 1948.

Absorption Due to Scattering

When a quantum is scattered by an electron in an atom, all of the laws governing the conservation of energy are observed as shown in Fig. 2.

The single quantum Q, containing energy E, collides with the electron e, changes direction through an angle ϕ, and continues with reduced energy as Q'. The energy lost by the quantum is imparted to the electron as kinetic energy. The recoil electron is then propagated at an angle θ, following the classic laws of elastic collision. This process may be repeated many times, producing many recoil electrons. If collected alone, these electrons would constitute a primary ionization current. They would then liberate their kinetic energy as heat in the collecting electrode. Actually, these recoil electrons will encounter other atoms before reaching the collector, and from each encounter some energy will be consumed in removing another electron. A single recoil electron may contain enough kinetic energy to remove thousands of other electrons before finally coming to rest. Thus, by the process of attrition, all of the energy originally taken from the original quantum may be consumed in liberating a quantity of electrons, which constitutes an electric current.

Absorption due to the production of recoil electrons is truly an electronic process and is dependent upon the number of electrons per gram of absorber, so that the element lead, having a ratio of atomic number to atomic weight of 0.395, has a correspondingly lower absorption per gram due to scattering than oxygen, where $Z/A=0.500$. Recoil electron production is independent of atomic structure and, therefore, varies little with different elements.

The mass scattering coefficient σ/ρ may be calculated for a given element Z and atomic number A as follows:

$$\frac{\sigma}{\rho} = \sigma_e N_0 \frac{Z}{A} \qquad (7)$$

where values of $\sigma_e N_0$ for each wavelength are computed from the Klien-Nishina formula. A graph showing the variation of $\sigma_e N_0$ versus λ is given in Fig. 3, the values decreasing at shorter wavelengths due to the relativistic change in mass of the electron under these conditions.

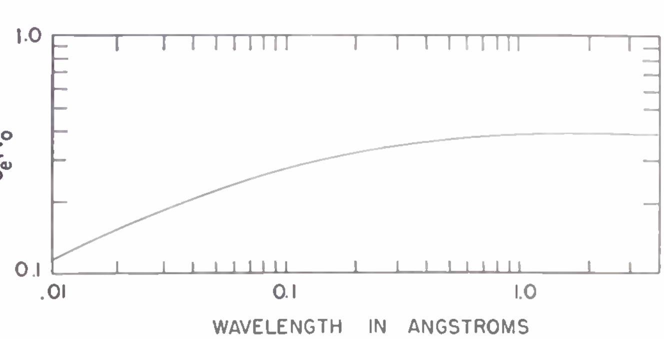

Fig. 3—Variation of $\sigma_e N_0$ with wavelength in angstroms. This factor, multiplied by Z/A for a given element, gives the mass scattering coefficient at the given wavelength.

Photoelectric Absorption

Unlike absorption due to scattering, photoelectric absorption occurs when a quantum enters an atom in such a manner that the atom emits electromagnetic radiation in the form of a new quantum which is characteristic of the energy level in the atom from which it was emitted and which may be emitted in *any direction.* This "fluorescent" radiation has a fixed value for a given energy level, and any of the original quantum energy left over is expended in the ejection of a photoelectron having proportional kinetic energy. The photoelectron may also be ejected in *any direction.* Photoelectric absorption, being entirely dependent upon the electronic structure of an atom, varies roughly as the fourth power of atomic number, and, being dependent upon the ability of the quantum to penetrate various orbits, varies roughly as the third power of wavelength, as shown by (4) and (6).

The process of photoelectric absorption is illustrated in Fig. 4, where the original quantum Q removes, for

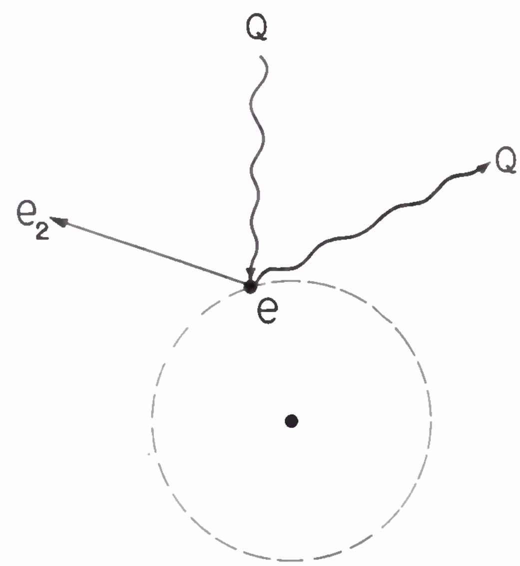

Fig. 4—Photoelectric absorption of a quantum of radiation with the resulting ejection of both a photoelectron and a new quantum of radiation.

example, a K electron from its orbit. An electron then jumps from the L orbit to the K orbit to replace it, thus liberating a new quantum Q_1 of characteristic radiation having exactly the same energy content as the difference between the binding energies of the electrons in the $K-L$ orbits. The balance of the original quantum energy appears as kinetic energy imparted to the photoelectron e, while the new quantum Q_1 may produce many more photoelectrons by the same process.

Photoelectrons constitute a part of the primary ionization current if collected, and once they are ejected, are identical to recoil electrons and may produce many secondary electrons before coming to rest.

Principles of Ionization Chambers

The prime function of an ionization chamber is the production of an electric current comprising liberated electrons. Liberated electrons will quickly recombine with the ionized atoms in the absorber, unless a strong enough electrostatic field is maintained across the absorber to prevent recombination. Choice of an absorber, therefore, is limited to those substances across which a potential gradient may be maintained, i.e., insulators. Gases are usually used in preference to liquids or solids, because of the relatively lower field potentials required for saturation.

The shape of the ionizable volume is limited electrically only to electrode construction which maintains a sufficient field gradient through all parts of the volume. This may be illustrated as in Fig. 5 (a), (b), and (c).

The simplest or parallel-plate type of ionization chamber, shown in Fig. 5 (a), must have its volume limited by

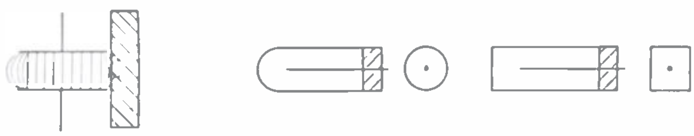

Fig. 5—Several types of ionization chamber construction showing electrostatic field distortion.

some means such as the insulating spacer *S*, or the volume from which ions will be collected will be indeterminate due to the indeterminate field as shown.

A round chamber, as shown in Fig. 5 (b), has a definite uniform field distribution throughout the ionizable volume.

A square chamber, as shown in Fig. 5 (c), is undesirable, for it may be impossible to collect all of the electrons from the corners, where the electrostatic field intensity will be seriously reduced.

Size

The size of an ionization chamber may be as small as its practical insulation can be made. Large chambers are limited by the high potentials required to prevent recombination of ions, for at excessive values corona discharge occurs, producing spurious ionization. The size may be decreased by proportionately increasing the pressure of gas within the chamber, or by using a gas having a higher absorption coefficient.

Special Types of Ionization Chambers

Ionization chambers for use with neutrons (which are uncharged particles) are designed to make use of secondary effects to produce ionization. Neutrons are not appreciably absorbed by most elements, and so secondary effects must be produced in those elements where the effects of collision or nuclear reaction with neutrons may be made to produce ionization. Boron, cadmium, and hydrogen are examples. A chamber may be filled with boron trifluoride, which is a gas, or the walls of a chamber containing any gas may simply be coated with boron, which produces alpha particles by reaction with slow neutrons. Ionization of the gas is produced by the alpha particles.

Hydrogen is used in chambers for detecting fast neutrons. Ionization in this case is produced by protons resulting from neutron collisions with hydrogen nuclei.

The response of ordinary ionization chambers to neutrons is usually dependent upon the presence of hydrogen in the composition of the walls, i.e., bakelite or other plastic.

Most ionization chambers are usually used under conditions which do not permit the indication of individual ionizing events. Geiger-Mueller tubes are designed to produce a single pulse for each event, and each pulse is counted individually. Ionization chambers, however, are usually used to integrate the number of events, and statistical fluctuations are not usually observable except for very low levels of radiation and high sensitivity. Alpha particles, because of greater ionizing ability, produce more statistical fluctuations than do high-speed electrons for an equal electrical output.

Occasionally, it is desirable to indicate individual events with an ionization chamber. Electrostatic capacitance must then be kept at a minimum and a pulse amplifier used in a manner similar to, but not identical with, a Geiger-Mueller counter system. Geiger-Mueller counters are "triggered" by an ionizing particle, and produce a pulse shape which depends upon tube design and not upon the ionizing particle, the energy of the pulse being obtained from the local source of power.

Ionization chambers produce pulses, the energy and shape of which are determined solely by the ionizing event itself and, therefore, can be used for analysis of the particles.

It is possible to use a fissionable element, such as uranium, within the chamber and to indicate radiation in terms of its ability to produce fission in the particular element.

Physical Significance of Ionization Current

When ionization chambers are used merely to indicate the presence of ionizing radiation, no great problems arise; but when any physical significance is required, even to the duplication of data, ponderous difficulties arise.

1. The law of absorption given by (1) applies only to "thin" absorbers, and does not hold for the absorption in a "volume." It is not strictly applicable to the shape of an ionization chamber.

2. The absorber in the ionization chamber is not isolated in free space, but it is surrounded by other ionizable elements which inject ionizing electrons and quanta into it. This is called wall effect, for whatever is used to limit the ionizable volume is composed of atoms, and these, too, are exposed to radiation and cannot be neglected.

3. In practice, most radiation is not monochromatic but heterogeneous. A single beam of X rays may con-

tain wavelengths producing an ionization current of which 95 per cent or more results from recoil electrons, while at the same time containing wavelengths producing an ionization current of which 95 per cent or more results from photoelectrons.

4. When the spectral composition of the radiation is not known, correction factors cannot be applied, even if they were available.

5. Critical phenomena appear as the wavelength of the radiation is changed. At long wavelengths, photoelectric absorption predominates; and at short wavelength, absorption is due almost entirely to recoil electrons. As the wavelengths are increased, the range or distance traveled by the recoil electron becomes less than the diameter of the chamber. At another wavelength, the range of the photoelectrons becomes less than the diameter of the chamber, each producing a change in ionization current. These are but several possible factors.

6. The elements comprising the chambers must be of low atomic number, so as not to have a critical absorption discontinuity in the wavelength range being measured. Such an absorption discontinuity may be observed in Fig. 1, which shows the values of μ/ρ for copper.

7. An indicated value obtained by means of an ionization chamber, to be of practical use as a "measurement," must have a fundamental relation to the utilization of the particular radiation. Because of the atomic structure of matter and the unique electronic properties of each substance, it is not usually possible to relate the effects produced in one type of matter with the effects produced in another type of matter without taking into account the differences in electronic structure. It is difficult, if not impossible, to define a unit for the measurement of radiation which would have universally fundamental significance under all possible conditions.

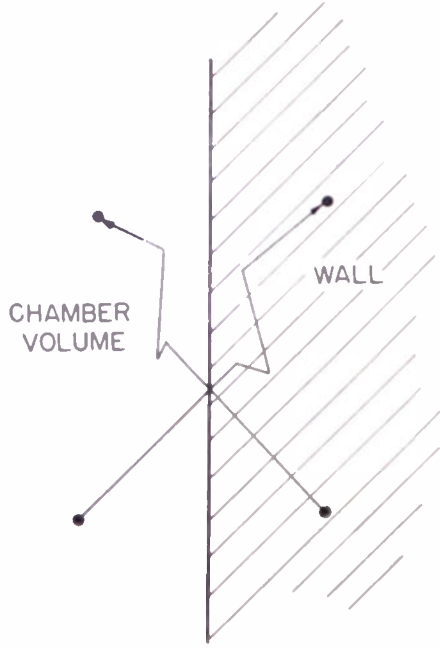

Fig. 6—"Wall effects" which exist at the field boundary of an ionization chamber.

8. The roentgen is the only unit presently available for the calibration of radiation-measuring instruments. Its use is confined to the purposes intended over a limited range of wavelengths and types of radiation, and its significance is practical rather than fundamental.

If an ionization chamber is to be made which is to be calibrated in roentgens and which has no wavelength dependence when compared to a standard air chamber, its wall effects must be similar to those in a standard air chamber. A section of the "wall" in a standard air chamber is shown in Fig. 6 where the ionizable volume and the "wall" are made of the same material, i.e., air. In this case, any ionizing particles lost through the field boundary, and which, therefore, do not contribute to the ionization current, are completely compensated by particles injected into the inside through the field boundary from outside the ionizable volume, or vice versa. Plainly, then, the effect is to collect the ions formed in 1 cc of air which exists within a larger irradiated volume of air. The roentgen measured in this way defines the degree of ionization produced by radiation which is proportional to quantity at a given wavelength.

When a given volume of air is surrounded by a substance which has a different coefficient of absorption, then the balance shown in Fig. 6 no longer holds, for there will be a loss or gain of particles by the ionizable volume which will not be compensated.

Fig. 7 shows the wavelength dependence produced in several 1-cc thimble chambers by different wall materials, as compared to the output of a standard air chamber.

Parallel-plate chambers, as shown in Fig. 5 (a), have considerable field distortion at the edges. The "ionizable

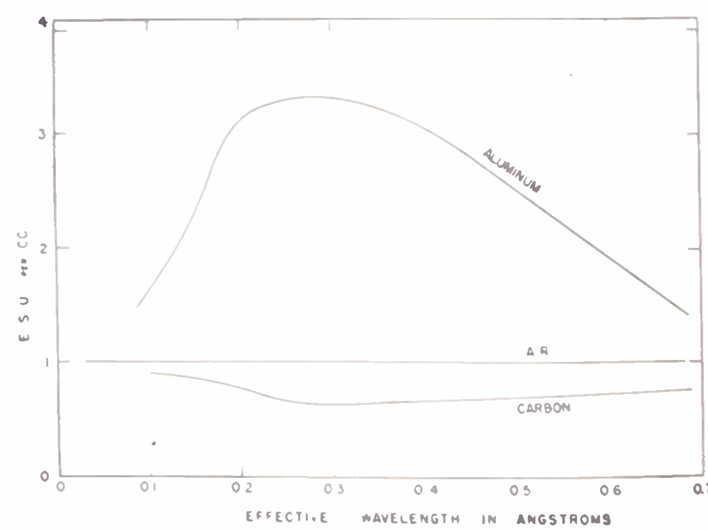

Fig. 7—Wavelength dependence or change in output which occurs at different wavelengths with different wall materials. Values were obtained with highly filtered heterogeneous radiation assumed to be equivalent to monochromatic radiation. Wavelengths are, therefore, not strictly accurate.

volume" of such a chamber is poorly defined, and is not suitable where accuracy is required. Improvement may be made by limiting the volume by an insulator, as shown, but some spurious ionization current would still be collected. This is unimportant where a mere indication is required, but where reproducibility or calibration is required, great difficulties arise.

Thimble chambers, as shown in Fig. 5 (b), have an entirely enclosed field, and by the proper choice of materials they may be made suitable for accurate calibration. Chambers of the type shown in Fig. 5 (c) should be avoided, for they are not likely to give a linear response to increase in intensity of radiation.

In general practice, ambiguities are avoided by calibrating ionization chambers in the internationally accepted unit, the roentgen. One roentgen is defined as that quantity of X or gamma radiation which produces, in 0.001293 gm of atmospheric air, a transfer of 1 esu of charge when all wall effects are eliminated. Ionization chambers to be calibrated are compared to a primary standard air chamber which is designed in accordance with the international definition of the roentgen.

Fig. 8 shows the schematic construction of a primary standard air chamber.

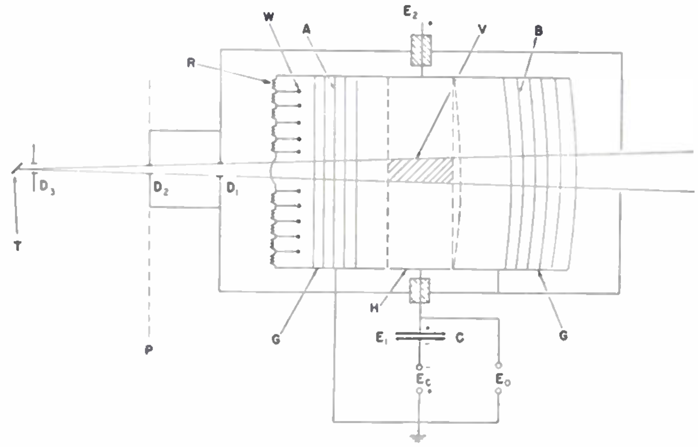

Fig. 8—Schematic construction of a primary standard air chamber as used for determining the international roentgen.

The volume of a standard air chamber is defined electrostatically to eliminate the effects of a wall, and the spacing of the collecting electrode is large enough to prevent scattered electrons or photoelectrons from impinging on the collector plates. A field potential of about 1,500 volts is maintained between the plates E_2 and GH. If the plates G are long enough, the electric field will be perpendicular to the plates, and plate H will collect ions only from an equivalent length of the beam. The "volume" is determined by the length of the collector H and the radius of the diaphragm D_2, by the relation $\pi r^2 H$. The diaphragms D_1 and D_3 serve to collimate the beam and to remove spurious radiation.

Unless the guard plates G are very much longer than the collector plate H, the electric field will be distorted or curved as shown at B, which has the effect of increasing the length of the collecting region. In standard designs this is prevented by the use of guard wires W, which are connected at spaced intervals to the potential divider R, thus providing a uniform potential gradient between plates. The electric field is then perpendicular to the collector plate H. The same construction is used at both ends of the chamber. Similar field distortion obviously will occur if the potential of the plate H differs from plates G, so that it is necessary to use a "null" type of collecting circuit to maintain the potential E_0 at zero. This is usually indicated by means of a center-zero string electrometer. Electrons accumulated in the capacitor C increases the potential E, which is continuously compensated by adding a compensating potential E_c to maintain $E_0=0$, and at the end of exposure, $E_c=E$.

The plane of measurement is shown at P where T is a point source of radiation such as the target of an X-ray tube.

Loss of radiation by absorption in the air between the front diaphragm D_2 and the collecting volume must be considered, and becomes of major importance for the longer wavelengths.

Prohibitively large plate spacing is required for very short wavelengths, so that standard air chambers are limited in use and practicability.

The ionization current obtained from any ionization chamber is directly proportional to the density of the ionizable absorber which constitutes its volume. The density of air specified in the definition of the roentgen is 0.001293 gm/cc, which corresponds to air at 0°C and 760 mm Hg pressure. When used at other temperatures and pressures, correction factors must be used. These are given by the following:

$$\text{roentgens} = \frac{CE}{300 \text{ vol.}} \times \frac{p}{760} \times \frac{273}{273 + t} \tag{8}$$

where C is the effective capacitance of the system in centimeters, E is the potential across the capacitor in volts, volume is in cc, p is the pressure in mm of mercury, and t is in degrees centigrade. One finds from (8) that an ionization chamber will produce 8 per cent less current at 72°F than at 32°F, so that a correction factor of 1.08 is required for ordinary room temperature. This factor may be eliminated by calibrating the chamber to include it originally. Such chambers indicate "roentgens corrected to ntp when used at 72°F and normal pressure." In common use, variations are then of the order of several per cent, and these may be corrected, if desired.

Ionization chambers for permanent use at a high altitude may be calibrated to include any fixed correction, but when variable pressure is encountered, such as in airborne instruments, a hermetically sealed chamber must be used.

From (8) it is evident that the current obtained from an ionization chamber may be increased in proportion to the absolute pressure within the chamber. Increase in pressure, however, requires an increase in the field potential required to prevent recombination and provide saturation current.

Field potential required is also dependent upon the degree of ionization, so that higher potentials are required as the radiation intensity is increased. A typical saturation curve for a 1-cm-diameter thimble chamber is given in Fig. 9. Chambers of larger diameter require

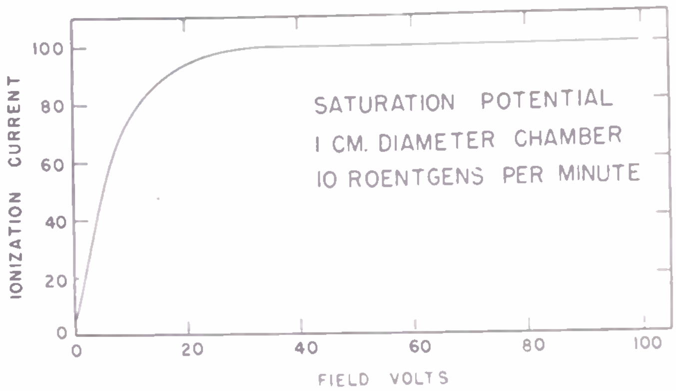

Fig. 9—Field potential required to produce saturation current in a thimble chamber 1 cm in diameter at 10 roentgens per minute.

a proportionate increase in field potential. Field potential required for saturation depends upon the radiation intensity, higher intensity requiring more field potential to prevent recombination. At very high intensities, it may even become impossible to prevent recombination.

Instrumentation

Instrumentation involving integrating ionization chambers is based on the circuit shown in Fig. 10, in which C is the total electrostatic capacitance of the insulated system in centimeters, including the potential indicating device, V is the ionizable volume of the chamber in cubic centimeters, and E is the potential change in volts occurring across the capacitor due to radiation. E_F is a field potential which, if large enough to provide saturation current, does not enter into calculated values obtained from (8).

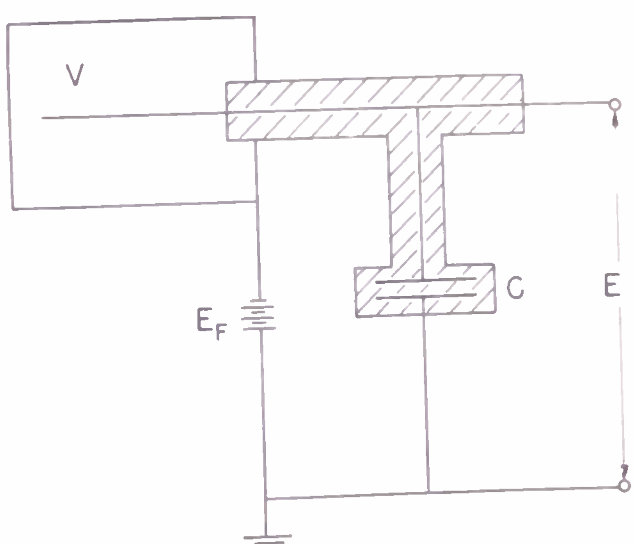

Fig. 10—Basic circuit used with practical ionization chambers.

From the definition of the roentgen, (8) provides one esu of charge for one roentgen at 0°C and 760 mm pressure. It is obvious that the potential E must be measured without taking current from the capacitance C. Electrostatic voltmeters are particularly useful for this purpose, and are extensively used.

Perhaps the most unusual feature of ionization-chamber circuits is the spurious ionization current picked up by the highly insulated circuit. This cannot be shielded from penetrating radiation, so that it must be made ionization-proof by enclosing all parts, as shown, with solid insulation of suitable quality. This, plus the desirability of a short time constant, require that this part of the circuit be kept as short as possible.

Electron tubes, when used, may be suitably mounted directly on the ionization chamber itself, so that no long leads are necessary.

The simplest practical type of radiation instrument is the capacitor type, where the ionization chamber may be considered as a combination air capacitor and ionization chamber, as shown diagramatically in Fig. 11.

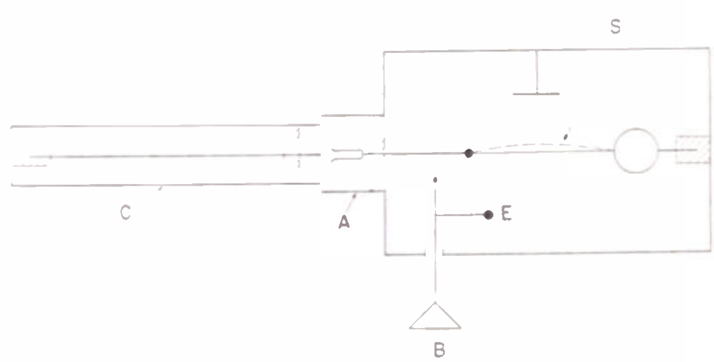

Fig. 11—Schematic construction of a capacitor-type r meter.

In this type of instrument, the capacitor-chamber unit (after being charged to a potential E by depressing the button B) is removed from the charging unit A and exposed to radiation to be measured. After exposure, the chamber unit is then reinserted in the charging unit, and the change in potential is indicated by the string electrometer S. Deflection of the string is observed with a microscope having an ocular with a scale calibrated in roentgens. In some instruments, an image of the string is projected upon a calibrated scale. Capacitor-type instruments are particularly useful in eliminating the effects of stray ionization, for it is possible to confine the ionization to the chamber proper by means of a suitable cap which normally closes the end of the capacitor unit. In one commercially available unit, full-scale deflection is obtained for 200 milliroentgens; and the chamber is shaped like a fountain pen so that it may be conveniently carried on the person during a working day for personal-health-hazard monitoring purposes, and the amount of radiation received determined at the end of the day.

Some types of pocket instruments include the complete system shown in Fig. 11, including a very small electrometer built in the shape of a fountain pen and arranged so that indication may be observed by means of a lens mounted on the end.

Similar instruments with larger capacitors and small thimble chambers are used for measuring larger quantities of radiation, such as 25, 250, or even 2,500 roentgens.

Electronic instrumentation involving ionization chambers as rate meters is severely limited in design by the low current obtainable and high insulation required. Values of 1×10^{-12} amperes may be taken as representative, while 1×10^{-16} is by no means uncommon. Electron tubes used to measure these minute currents present special problems in themselves. As a rate meter, it is common to use an electrometer tube connected as a tetrode, shown in Fig. 12; and by using a grid resistor of

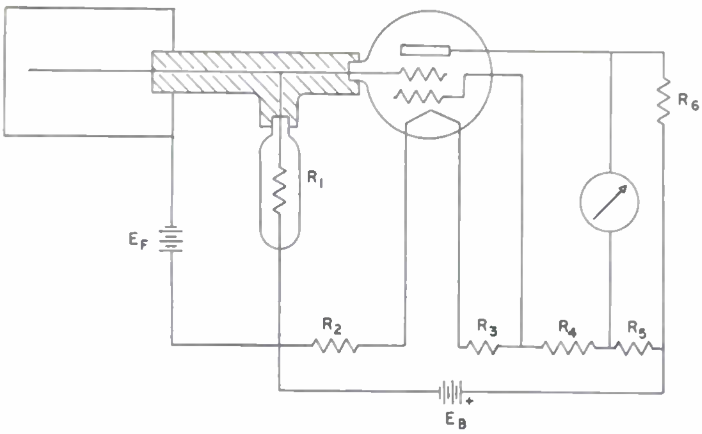

Fig. 12—Conventional tetrode electron-tube circuit as used with an ionization chamber as a rate meter.

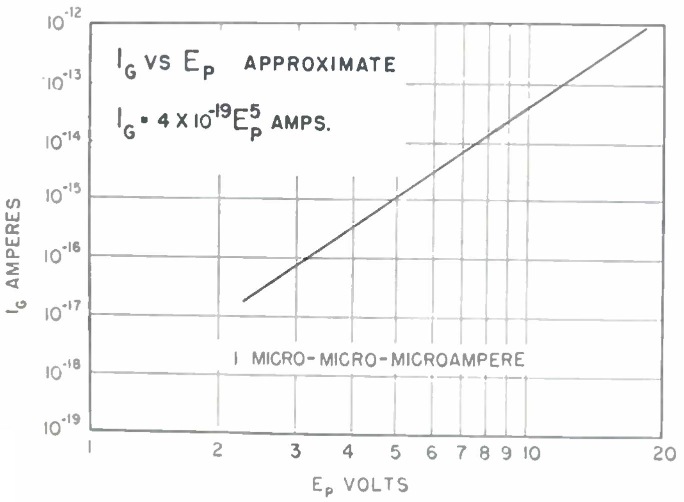

Fig. 13—Effect of plate potential on grid currents obtained with a VX-41 electrometer tube.

1×10^{12} ohms, we obtain a 1-volt grid change for a current of 1×10^{-12} amperes. If we are to maintain an accuracy of 1 per cent, the *total* leakage resistance across the circuit must be greater than 10^{14} ohms, and *all* spurious currents must be less than 1×10^{-14} ampere. Grid currents are probably the most difficult to overcome, and limit the electrometer tube to plate potentials which are of the order of 3 volts. The effect of plate potential is shown in Fig. 13, which applies to VX-41 electrometer tubes and shows that grid current varies roughly with the fifth power of plate potential. The fifth-power law presumably applies roughly to any other type of electrometer tube, with magnitude depending upon the type of tube. Low plate potentials are of the greatest importance.

High values of grid resistance may cause a prohibitive time constant in the grid circuit due to grid capacitance and the associated chamber capacitance. 10^{12} ohms produces a time constant of 1 second with a total capacitance of 1 $\mu\mu f$, so that heroic methods are sometimes required to provide both a satisfactory time constant and a measurable potential. Feedback is often employed in series with the field potential to reduce the time constant.

The accuracy of electronic ionization-chamber instruments, is dependent upon the quality of the grid resistor. Unlike grid resistors used in communications or other usual electronic devices, the grid resistor determines the calibration, and its absolute value must remain stable under all conditions to which the instrument is subjected. Accurate, stable, vacuum-sealed resistors are obtainable in values from 10^8 to 10^{14} ohms, specifically designed for this service, and are surface-treated to operate under adverse conditions of humidity where ordinary resistors would be inoperable.

Ordinary insulators are not usually satisfactory for insulation in ionization chambers or their highly insulated circuits. Most materials absorb enough moisture from the air to impair either surface or volume resistivity or both. Dielectric loss by "absorption" or charge cannot be neglected in many cases. Amber and pure polystyrene are, in general, most satisfactory and commonly used. In all cases, perfectly clean and perfectly polished surfaces are desired. Glass of certain types is sometimes permissible, but only after it has been treated to prevent surface leakage. Silicone preparations are used, and are satisfactory when properly applied. Ordinary waxes and varnishes are completely unsatisfactory, for their absorption and transmission of water vapor is high.

Capacitors present similar problems in ionization-chamber circuits. Ordinary capacitors are usually unsuited because of leakage or dielectric loss or both. Capacitors are usually located close to the chamber and must, therefore, be ionization-proof. The use of solid dielectric and the elimination of all air spaces is essential. Pure polystyrene or polyethylene film are suitable for dielectrics when proper precautions are taken. Excellent capacitors of small values can be made of thin thimbles or tubes of polystyrene coated on both sides with a conducting coating.

Some dielectrics present apparent leakage when exposed to radiation, and these must be studiously avoided both as dielectric for capacitors and as insulators.

Large chambers are often used as monitoring rate meters for controlling the output of radiation from radiation generators. Under these conditions, direct calibration of the chamber is unimportant; and wall effects, within reason, can be neglected, which permits the use of a number of thin parallel plates within the chamber to reduce electrode spacing, thus eliminating the need for high field potentials. Such a chamber is shown schematically in Fig. 14. One disadvantage of this

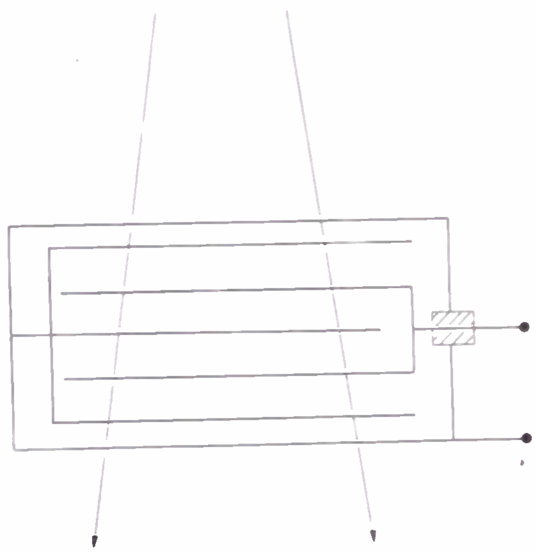

Fig. 14—Large parallel-plate chamber as used for monitoring purposes.

type is the relatively high electrostatic capacitance which results from the capacitor-like structure. The larger current obtainable, however, usually permits the

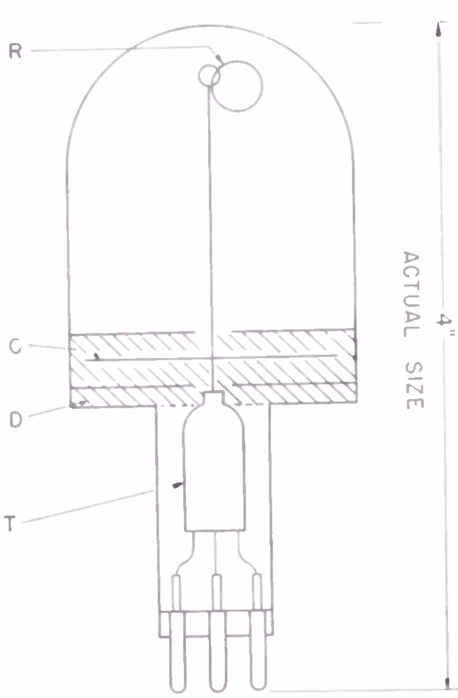

Fig. 15—Schematic construction of the sealed chamber of a practical integrating r meter. Full scale indicates 200 milliroentgens.

use of low circuit resistance, so that the time constant may not be objectionable.

A typical commercially available integrating-type radiation instrument is shown schematically in Figs. 15 and 16.

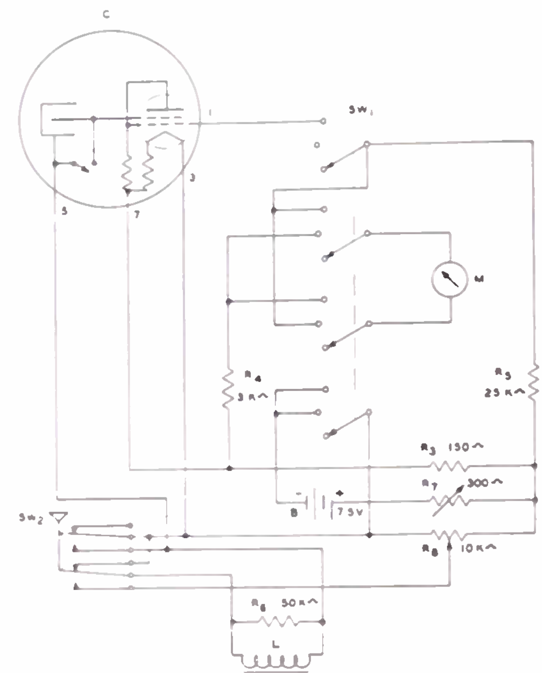

Fig. 16—Schematic circuit diagram of an instrument using the chamber shown in Fig. 15.

Fig. 15 shows schematically a plug-in ionization-chamber unit which contains a capacitor C of about 100 $\mu\mu$f, with solid dielectric D, and an electrometer tube T hermetically sealed within a plastic container, the inside of which is coated with a conducting material of suitable atomic number. The highly insulated circuit is entirely within the chamber, and only low-potential leads emerge. No resistor is used in the grid circuit, the insulation being sufficient to permit the charge to be retained for long periods of time. The loose ring R is used to discharge the chamber by inverting it, if this is desired for test purposes.

Fig. 15 shows the schematic diagram of the instrument. The electrometer tube is used here as an inverted triode, and the sum of grid, filament, and plate potentials is 6 volts, so that several hundred hours of operation may be had from five 1½-volt flashlight cells. The filament of the XV-41 tube used requires 10 ma, and becomes a part of the resistance string which supplies grid and plate potentials.

The grid is charged to a predetermined potential by means of an inductive surge provided by the inductance L when the button B is depressed. Because of the inverted triode connection, the grid may be operated at potentials which are high enough to supply saturation current in the chamber for low radiation intensities. Such an instrument is an integrating-type r meter, for it adds up all ionization received over an indefinite period of time, such as one day. Full scale reading is 200 milliroentgens, so that one-half scale corresponds to the legally accepted daily tolerance dose of radiation or 100 milliroentgens. The main use of this instrument is for the protection of personnel where exposure to radiation presents a health hazard and where monitoring is, therefore, desirable. Stability of the instrument is of the order of two divisions or less per twenty-four hours, part of which is due to normal background radiation.

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Some Additions to the Theory of Radio-Frequency High-Voltage Supplies*

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Summary—**The double-tuned overcoupled air transformer, commonly used in radio-frequency high-voltage supplies, is represented by an equivalent primary circuit. This circuit can be used very advantageously to develop the theory of operation of the oscillator, including the two possible operating frequencies, the necessary conditions for oscillation and maximum efficiency, and the variation of operating frequency and load resistance with primary tuning.**

It is shown that, at the operating frequency, the load impedance is purely resistive. The various resistances in the circuit representing the power losses and the load resistance are included in the equivalent circuit resistance, which appears as the plate load of the class-C oscillator tube. This greatly facilitates the design of a circuit to give any required power and voltage output.

There are several ways in which this design problem can be approached, and the paper outlines the method which was used and found successful by the author.

List of Symbols

R_{DC} = dc load resistance

R_L = the equivalent ac load resistance across the secondary coil

p = the total number of stages of voltage multiplication in the rectifier circuit

r_L = the equivalent series load resistance in the secondary circuit

r_S = the equivalent series loss resistance in the secondary circuit

$R_2 = r_S + r_L$

r_{Se} and R_{Se} = the equivalent primary resistances produced by r_S

r_{Le} and R_{Le} = the equivalent primary resistances produced by R_L

R_{2e} = the total equivalent secondary resistance reflected into the primary circuit

r_1 = the ac resistance of the primary coil

R_{1e} = the equivalent parallel primary-coil resistance

$G_{1e} = 1/R_{1e}$

$G_{2e} = 1/R_{2e}$

R = the equivalent oscillator load resistance

Z_e = the equivalent oscillator load impedance

E_p = rms value of the ac voltage across the oscillator tank circuit

P_T = the total ac power output of the oscillator

L_2 = the self-inductance of the secondary coil

L_1 = the self-inductance of the primary coil

$$Q_1 = \frac{2\pi f L_1}{r_1}$$

$$Q_2 = \frac{2\pi f L_2}{R_2}$$

k = the coefficient of coupling between the primary and secondary coils

C_1 = the primary tuning capacitance

C_2 = the secondary-circuit distributed capacitance

f = the operating frequency

$$f_{02} = \frac{1}{2\pi\sqrt{L_2 C_2}}$$

$$f_{01} = \frac{1}{2\pi\sqrt{L_1 C_1}}$$

$$\theta_1 = 1 - \frac{f_{01}^2}{f^2}$$

$$\theta_2 = 1 - \frac{f_{02}^2}{f^2}.$$

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Recently, considerable work has been done on the development of high-voltage, low-power, direct-current supplies, using a radio-frequency oscillator and a double-tuned, over-coupled air-core transformer for obtaining the high voltage. The usual output of these supplies is from 5,000 to 30,000 volts at from 2 to 20 watts, and they have been used extensively as a source of high voltage for the accelerating anodes of cathode-ray tubes, and the like, where most of the power is absorbed in the bleeder resistance. Several papers[1-3] on the construction and operation of these units have already been published. This report shows how an equivalent oscillator circuit can be used in the analysis and design of a high-voltage supply.

Using this equivalent oscillator circuit, the important features regarding the operation of the circuit can be determined very readily. The first part of the paper will outline the development of the equivalent circuit, and from it an analysis of the necessary conditions for the operation of the circuit and maximum efficiency will be made.

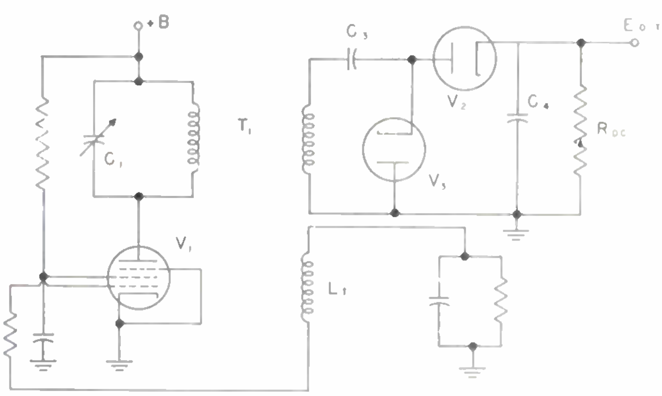

Fig. 1—Typical radio-frequency high-voltage supply circuit.

The chief advantage to be gained by the use of such an equivalent circuit, however, is in the design of a high-voltage supply unit. The source of the high-frequency power is a class-C oscillator. Because of this, and the fact that the circuit is double-parallel-tuned, the usual method of approach to double-tuned circuits was found to be impractical. Using the equivalent circuit, however, the various resistances representing power used in the circuit can be represented by equivalent resistances across the plate of the oscillator tube. This greatly facilitates the design, since the work is reduced to a simple process of substitution in the expressions for the latter. The second half of the paper outlines the method of approach to the design which was used by the author and found successful. Other slightly different methods may be used, and in some cases may be simpler.

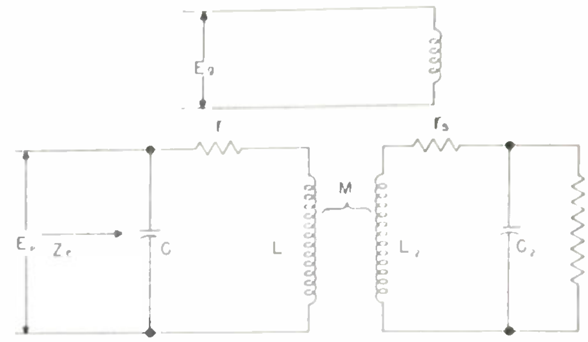

Fig. 2—Equivalent ac circuit for radio frequency high-voltage supply.

The use of the equivalent circuit and analysis is not confined to the present application. Some use has also been made of the expressions for the load resistances to facilitate the design of intermediate-frequency amplifiers, for which case the equations can be simplified considerably.

Essentially, the circuit is quite simple, as shown by the typical example in Fig. 1, and the equivalent ac circuit in Fig. 2. The equivalent ac secondary load resistance R_L is equal to $R_{DC}/2p^2$ where p is the number of stages of voltage multiplication in the rectifier circuit.

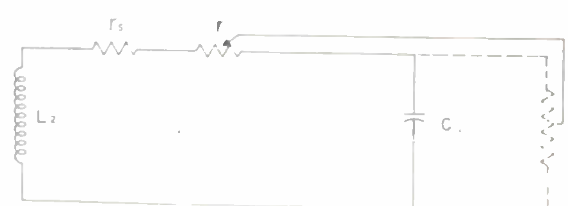

Fig. 3—Equivalent secondary circuit.

For convenience, the parallel combination of R_L and C_2 is converted into the equivalent series circuit shown in Fig. 3 as follows:

$$\frac{R_L\dfrac{1}{j2\pi fC_2}}{R_l+\dfrac{1}{j2\pi fC_2}}=\frac{R_l-j2\pi fC_2R_l^2}{1+4\pi^2f^2C_2^2R_l^2}$$

$$=\frac{1}{4\pi^2f^2C_2^2R_l}-\frac{j}{2\pi fC_2}$$

$$=r_l-j\frac{2\pi f_{02}^2L_2}{f}\ \ldots \qquad (1)$$

since in practice R_L is large and $1 \ll 4\pi^2 f^2C_2^2R_L^2$.

The equivalent primary circuit shown in Fig. 4 may be developed from that shown in Fig. 2, as follows:

[1] O. H. Schade, "Radio-frequency-operated high-voltage supplies for cathode-ray tubes," Proc. I.R.E., vol. 31, pp. 158–163; April, 1943.
[2] H. C. Baumann, "Television receiver r.f. power supply design," *Communications*, vol. 26, pp. 26–27; March, 1946.
[3] R. S. Mautner and O. H. Schade, "Television high-voltage r.f. supplies," *RCA Rev.*, vol. 8, pp. 43–81; March, 1947.

$$Z_e = \frac{\dfrac{1}{j\omega C_1}\left\{r_1 + j\omega L_1 + \dfrac{\omega^2 M^2}{R_2 + jX_2}\right\}}{\dfrac{1}{j\omega C_1} + r_1 + j\omega L_1 + \dfrac{\omega^2 M^2}{R_2 + jX_2}} \quad \cdots \quad (2)$$

where

$$R_2 = r_S + r_L$$

$$X_2 = \omega L_2\left(1 - \frac{\omega_{02}^2}{\omega^2}\right);$$

$$M = k\sqrt{L_1 L_2}.$$

Dividing by ωL_z, and rationalizing the expression

$$\frac{\omega^2 M^2}{R_2 + jX_2},$$

we obtain

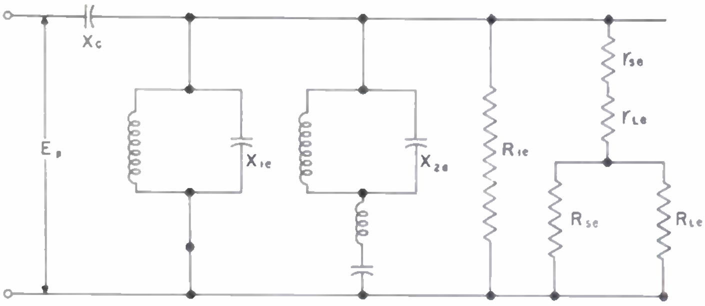

Fig. 4—Equivalent primary-circuit impedances.

From (5) and referring to Fig. 4, we have

$$R_{1e} = \frac{4\pi^2 f_{01}^4 L_1^2}{r_1 f^2} \quad (6)$$

$$X_{1e} = -j\frac{2\pi f_{01}^4 L_1}{f^3\theta_1} \quad (7)$$

$$Z_e = \frac{\dfrac{1}{j\omega C_1}\left\{\dfrac{r_1}{\omega L_1} + j + \dfrac{\omega k^2 L_2 R_2}{R_2^2 + X_2^2} - j\dfrac{\omega k^2 L_2 X_2}{R_2^2 + X_2^2}\right\}}{\dfrac{r_1}{\omega L_1} + \dfrac{\omega k^2 L_2 R_2}{R_2^2 + X_2^2} - j\left\{-1 + \dfrac{\omega_{01}^2}{\omega^2} + \dfrac{\omega k^2 L_2 X_2}{R_2^2 + X_2^2}\right\}}$$

$$= \frac{\dfrac{1}{j\omega C_1}\left\{\dfrac{r_1}{\omega L_1} + \dfrac{\omega k^2 L_2 R_2}{R_2^2 + X_2^2} - j\left(-1 + \dfrac{\omega k^2 L_2 X_2}{R_2^2 + X_2^2} + \dfrac{\omega_{01}^2}{\omega^2}\right) + j\dfrac{\omega_{01}^2}{\omega^2}\right\}}{\dfrac{r_1}{\omega L_1} + \dfrac{\omega k^2 L_2 R_2}{R_2^2 + X_2^2} - j\left\{-1 + \dfrac{\omega_{01}^2}{\omega^2} + \dfrac{\omega k^2 L_2 X_2}{R_2^2 + X_2^2}\right\}}$$

$$= \frac{1}{j\omega C_1} + \frac{\dfrac{\omega_{01}^2}{\omega^3}\cdot\dfrac{1}{C_1}}{\dfrac{r_1}{\omega L_1} + \dfrac{\omega k^2 L_2 R_2}{R_2^2 + X_2^2} + j\left\{1 - \dfrac{\omega_{01}^2}{\omega^2} - \dfrac{\omega k^2 L_2 X_2}{R_2^2 + X_2^2}\right\}}. \quad (3)$$

Putting

$$\omega = 2\pi f \qquad C_1 = \frac{1}{4\pi^2 f_{01}^2 L_1}$$

$$\omega_{01} = 2\pi f_{01} \qquad 1 - \frac{f_{01}^2}{f^2} = \theta_1$$

$$\omega_{02} = 2\pi f_{02} \qquad 1 - \frac{f_{02}^2}{f^2} = \theta_2$$

$$Z_e = -j\frac{2\pi f_{01}^2 L_1}{f} + \frac{\dfrac{2\pi f_{01}^4 L_1}{f^3}}{\dfrac{r_1}{2\pi f L_1} + \dfrac{2\pi f k^2 L_2 R_2}{R_2^2 + 4\pi^2 f^2 L_2^2\theta_2^2} + j\left(\theta_1 - \dfrac{4\pi^2 f^2 k^2 L_2^2 \theta_2}{R_2^2 + 4\pi^2 f^2 L_2^2 \theta_2^2}\right)}. \quad (4)$$

This expression is equivalent to $Z_e = Z_1 + Z_2$ where

$$\frac{1}{Z_2} = \frac{r_1 f^2}{4\pi^2 f_{01}^4 L_1^2} + \frac{f^4 k^2 L_2 R_2}{f_{01}^4 L_1(R_2^2 + 4\pi^2 f^2 L_2^2 \theta_2^2)} + j\frac{\theta_1 f^3}{2\pi f_{01}^4 L_1} - j\frac{2\pi f^5 k^2 L_2^2 \theta_2}{f_{01}^4 L_1(R_2^2 + 4\pi^2 f^2 L_2^2 \theta_2^2)} \quad \cdots \quad (5)$$

$$R_{2e} = \frac{f_{01}^4 L_1 R_2}{f^4 k^2 L_2} + \frac{4\pi^2 f_{01}^4 L_1 L_2 \theta_2^2}{f^2 k^2 R_2}.$$

Putting $R_2 = r_S + r_L = r_S + \dfrac{4\pi^2 f_{02}^4 L_2^2}{f^2 R_L}$,

we can obtain

$$R_{2e} = \frac{f_{01}^4 L_1 r_S}{f^4 k^2 L_2} + \frac{4\pi^2 f_{01}^4 f_{02}^4 L_1 L_2}{f^6 k^2 R_L} + \frac{4\pi^2 f_{01}^4 L_1 L_2 \theta_2^2}{f^2 k^2 \left(r_S + \dfrac{4\pi^2 f_{02}^4 L_2^2}{f^2 R_L}\right)}$$

$$= r_{Se} + r_{Le} + \frac{R_{Se} R_{Le}}{R_{Se} + R_{Le}}.$$

Therefore,

$$r_{Se} = \frac{f_{01}^4 L_1 r_S}{f^4 k^2 L_2} \quad (8)$$

$$r_{Le} = \frac{4\pi^2 f_{01}^4 f_{02}^4 L_1 L_2}{f^6 k^2 R_L} \quad (9)$$

$$R_{Se} = \frac{4\pi^2 f_{01}^4 L_1 L_2 \theta_2^2}{f^2 k^2 r_S} \quad (10)$$

$$R_{Le} = \frac{f_{01}^4 L_1 \theta_2^2 R_L}{k^2 f_{02}^4 L_2} \quad (11)$$

$${}_{2e} = j\left\{\frac{f_{01}^4 L_1 R_2^2}{2\pi f^5 k^2 L_2^2 \theta_2} + \frac{2\pi f_{01}^4 L_1 \theta_2}{f^3 k^2}\right\} \cdots \quad (12)$$

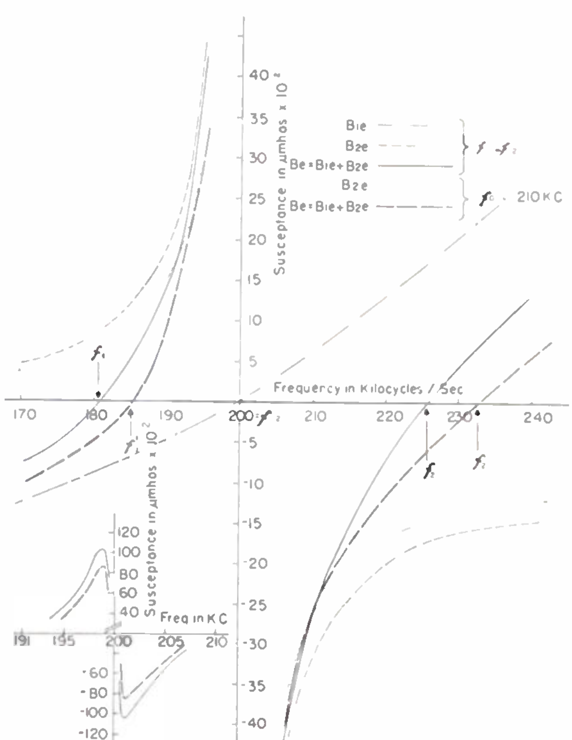

Fig. 5—Equivalent primary susceptance versus frequency for two values of f_{01}, the primary resonant frequency.

The variation of the equivalent impedances with frequency is shown in Figs. 5 and 6 for fixed values of f_{02} and f_{01}. Here the susceptances and conductances were plotted, instead of reactance and resistance, since the susceptance goes to zero at the resonant frequencies.

The equivalent secondary resistances r_{Se}, r_{Le}, R_{Se}, and R_{Le} have been combined, and are represented by the conductance G_{2e}. The graphs are plotted for the following circuit constants, which were obtained from an actual circuit:

$f_{02} = 200$ kc
$f_{01} = 200$ kc and 210 kc
$r_S + r_L = 700$ ohms
$L_2 = 43$ mh
$L_1 = 0.15$ mh
$r_1 = 2$ ohms
$k^2 = 0.05$.

The capacitive reactance X_c (Fig. 4) was found to be negligible for all values of frequency plotted, and to simplify the computations the resistances r_1, r_S, and r_L were assumed constant for all frequencies.

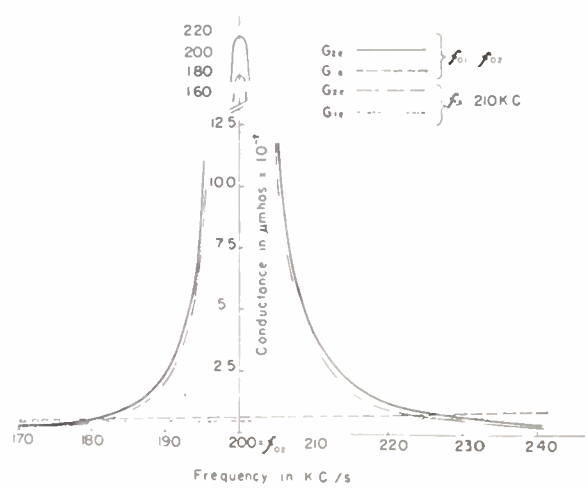

Fig. 6—Equivalent primary conductances versus frequency for two values of f_{01}.

It will be shown later that the voltage fed back to the grid of the class-C oscillator is in the correct phase for operation at the frequencies f_1 or f_2 (Fig. 5). At the other apparent resonant frequency f_{02}, the circuit is unstable if the coefficient of coupling is greater than critical. In practice this is the usual case, since the stability and efficiency are better with overcoupled circuits. The resonant frequencies can be found by putting the imaginary term in the expression for Z_e (4) equal to zero, again neglecting the effect of X_c. Thus we have

$$\theta_1 - \frac{4\pi^2 f^2 k^2 L_2^2 \theta_2}{R_2^2 + 4\pi^2 f^2 L_2^2 \theta_2^2} = 0.$$

Multiplying through by $R_2^2 + 4\pi^2 f^2 L_2^2 \theta_2^2$ and dividing by $4\pi^2 f^2 L_2^2 \theta_2$, the equation becomes

$$\frac{\theta_1}{\theta_2} \cdot \frac{1}{Q_2^2} + \theta_1 \theta_2 = k^2$$

or

$$\theta_1\theta_2 = k^2 - \frac{\theta_1}{\theta_2}\cdot\frac{1}{Q_2^2}$$
$$\doteqdot k^2 \cdots \qquad (13)$$

where

$$Q_2 = \frac{2\pi f L_2}{R_2},$$

and, in practice,

$$\frac{\theta_1}{\theta_2}\cdot\frac{1}{Q_2^2}$$

is small compared with k^2. As k is decreased the operating frequencies approach f_{02} and coincide with f_{02} for critical coupling. The secondary circuit is self-resonant, since it is desirable to have as high a secondary Q as possible, as well as to eliminate tuning elements in the high-voltage secondary circuit. In practice, therefore, the primary tuning is made adjustable, and it is found that as the capacitor C_1 is varied the secondary output will go through a maximum value.

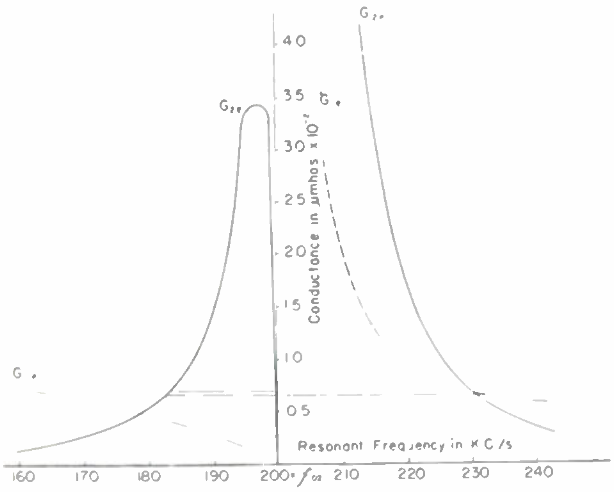

Fig. 7—Variation of equivalent conductances G_{1e} and G_{2e} with resonant frequency.

We have established that there are two possible operating frequencies and that these can be varied about f_{02} by changing the primary resonant frequency. It is, therefore, necessary to know how the load resistance will vary as the operating frequency is changed, what determines the maximum possible output with a given set of circuit constants, the point or points at which maximum efficiency can be expected, and over what range the frequency can be varied successfully. The effect of varying the operating frequency, by changing f_{01}, on the conductances G_{1e} and G_{2e}, is shown in Fig. 7. The expressions for these variables, after substituting for f_{01} from (13), are:

$$G_{1e} = \frac{1}{R_{1e}} = \frac{r_1}{4\pi^2 f^2 L_1^2\left(1 - \frac{k^2}{\theta_2}\right)^2} \cdots \qquad (14)$$

$$G_{2e} = \frac{1}{R_{2e}} = \frac{k^2 L_2}{\left(1 - \frac{k^2}{\theta_2}\right)^2 L_1\left\{R_2 + \frac{4\pi^2 f^2 L_2^2 \theta_2^2}{R_2}\right\}} \cdots \qquad (15)$$

Since the value of R_2 was again assumed constant with respect to frequency to facilitate computations, the actual value of G_{2e} should be slightly greater for $f < f_{02}$, and smaller for $f > f_{02}$, than is shown.

The power loss in the primary coil is $E_p^2 G_{1e}$, and the power delivered to the secondary circuit is $E_p^2 G_{2e}$. The ratio of the useful power output to the power loss in the secondary circuit is established primarily by the secondary-circuit design. For maximum power output, therefore, $E_p^2 G_{2e}$ should be a maximum. For a constant value of the oscillator plate voltage this requires G_{2e} to be a maximum with G_{1e} a minimum for good efficiency. Referring to Fig. 7, this condition will be realized if the circuit operates at a frequency below f_{02}. On the high-frequency side of f_{02}, a higher value of G_{2e} is obtainable, but there is also a much higher value of G_{1e}.

The operating frequency at which G_{2e} is a maximum can be obtained as follows:

Putting

$$f_{02} = f + \alpha,$$
$$\theta_2 = -\frac{2\alpha}{f}, \quad \text{since } \alpha^2 \text{ is small.}$$

Then the expression for $1/G_{2e}$ becomes

$$1/G_{2e} = 1/k^2\left(1 + \frac{k^2 f_{02}}{2\alpha} - \frac{k^2}{2}\right)\frac{L_1}{L_2}\left(R_2 + \frac{16\pi^2 L_2^2 \alpha^2}{R_2}\right). \qquad (16)$$

Differentiating with respect to α, and equating the result to zero, we obtain

$$\alpha^3 \cdot \frac{16\pi^2 L_2^2}{R_2}(2 - k^2) = k^2 f_{02} R_2,$$

from which

$$\alpha = \left(\frac{k^2 R_2^2 f_{02}}{16\pi^2 L_2^2 (2 - k^2)}\right)^{1/3}. \qquad (17)$$

From Fig. 7 it can, therefore, be seen that the efficiency is better at the lower resonant frequency because:

1. For corresponding values of G_{2e}, the value of G_{1e} is lower.
2. The secondary-circuit series load resistance r_L is higher at the lower resonant frequency. Thus the ratio r_L/r_S is higher.

It would seem advisable, then, to operate at a frequency just below f_{02} in order to obtain maximum power output and maximum efficiency. There are, however, other limiting conditions due to which it is not always possible to operate at this point. Moreover, although the variation in G_{2e} is large, it is not always possible to obtain the required output power with just any circuit constants. As stated previously, as the primary tuning is changed a point of maximum output is reached. This maximum output is determined by the factors which are discussed below.

1. The Effective Loaded Q of the Equivalent Circuit.

This should be equal to 4π for maximum efficiency.[4] If less than 4π the circuit will be unstable, and if greater than 4π the efficiency will be down. If, in the circuit shown in Fig. 2, R is equal to the total equivalent parallel load resistance across the plate of the oscillator tube at resonance, then

$$Q = R/X_{c1} \geqq 4\pi$$

$$\text{or} \quad X_{c1} \leqq R/4\pi$$

$$\leqq \frac{E_p^2}{4\pi P_l}. \tag{18}$$

To digress a moment from the main subject, we have previously neglected the reactance X_C (Fig. 4) in our discussion. This reactance is equal in magnitude to X_{c1}. It is, therefore, limited, and in most practical cases can be neglected with respect to the total equivalent load resistance R_T.

2. The Effective Load Resistance.

The effective load resistance

$$R_T = \frac{1}{G_{1e} + G_{2e}}$$

must be matched correctly to the oscillator tube, for maximum power output. How this is accomplished, using the equivalent circuit, will be shown in the second part of the report.

3. The Phase of the Grid Voltage.

So far, it has been assumed that the oscillator operates at the resonant frequency (resistive plate load) of the equivalent primary circuit. As stated previously, the actual operating frequency depends on the phase of the grid voltage. It will differ slightly, in practice, from the above resonant frequency (about 0.5 kc), although, in a properly designed circuit, not enough to make an appreciable difference in any of the equations thus far developed. If, however, the equivalent primary-circuit Q is low, or if the resonant frequency is nearly equal to f_{04} (θ_2 small and θ_1 large), the difference will be appreciable. The efficiency will, therefore, be considerably reduced, since the oscillator load will no longer be resistive and a point will be reached where the circuit will no longer operate.

The power required to feed the grid of the oscillator tube may be obtained by mutual coupling to either the primary or secondary coils. The analysis that follows will show that mutual coupling to the secondary coil is preferable since it results in better stability, for, depending on the sign of the mutual inductance between the tickler and secondary coils, the circuit can be made to operate at one, and only one, of the resonant frequencies. With primary tickler feedback the mutual inductance must be negative for oscillation, and it can operate at both frequencies.

Using the approximate equivalent circuit for the oscillator tube shown in Fig. 8, where the resistance R_1 in-

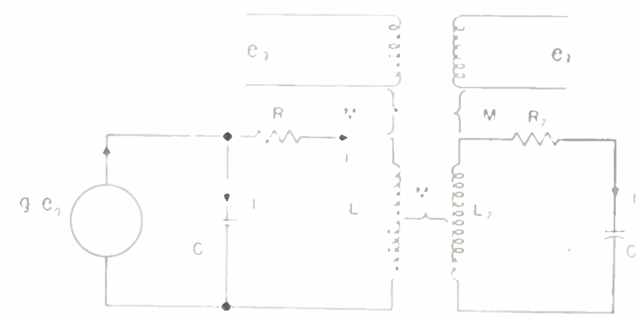

Fig. 8—Approximate equivalent oscillator circuit.

cludes the primary-coil resistance and the effective plate resistance of the oscillator tube, it will be shown that the resonant frequencies and the operating frequencies nearly coincide. The frequency at which e_g is in phase with $g_m e_g$ can be found as follows:
Putting

$$\omega = 2\pi f$$

$$g_m e_g = i_c + i_l$$

$$\frac{-j i_c}{\omega C_1} = i_l(R_1 + j\omega L_1) \cdot j\omega M i_2$$

$$\frac{-j g_m e_g}{\omega C_1} + \frac{j i_L}{\omega C_1} = i_l(R_1 + j\omega L_1) + j\omega M i_2$$

$$\frac{-j g_m e_g}{\omega C_1} = i_l\left(R_1 + j\omega L_1 - \frac{j}{\omega C_1}\right) \pm j\omega M i_2$$

$$0 = i_2(R_2 + j\omega L_2 - j/\omega C_2) \pm j\omega M i_l.$$

[4] D. C. Prince, "Vacuum tubes as power oscillators," Proc. I.R.E., vol. 11, p. 301; June, 1923.

(1) Tickler coil coupled to the secondary coil: $e_g = \pm j\omega M_1 i_2$.

Therefore,

$$i_L = -\frac{e_g / \pm j\omega M_1 (R_2 + j\omega L_2\theta_2)}{\pm j\omega M}$$

$$= \pm \frac{e_g(R_2 + j\omega L_2\theta_2)}{\omega^2 M M_1}$$

$$\frac{-jg_m e_g}{\omega C_1} = \pm \frac{e_g(R_2 + j\omega L_2\theta_2)(R_1 + j\omega L_1\theta_1)}{\omega^2 M M_1} \pm \frac{M}{M_1} e_g$$

$$\frac{-jg_m e_g}{\omega C_1} = \pm e_g \left\{ \frac{M}{M_1} + \frac{R_1R_2 - \omega^2 L_1 L_2 \theta_1\theta_2}{\omega^2 M M_1} + \frac{j(\omega L_1\theta_1 R_2 + \omega L_2\theta_2 R_1)}{\omega^2 M M_1} \right\}.$$

For e_g in phase with $g_m e_g$:

$$\frac{M}{M_1} + \frac{R_1R_2 - \omega^2 L_1 L_2 \theta_1\theta_2}{\omega^2 M M_1} = 0. \qquad (19)$$

Putting

$$M = k\sqrt{L_1L_2}$$

$$R_2/\omega L_2 = 1/Q_2, \qquad \frac{R_1}{\omega L_1} = \frac{1}{Q_{1r}}.$$

Equation (19) becomes

$$\theta_1\theta_2 = k^2 + \frac{1}{Q_{1r}Q_2} \qquad (20)$$

$$\doteqdot k^2 \text{ since } \frac{1}{Q_{1r}Q_2} \ll k^2 \text{ in practice.}$$

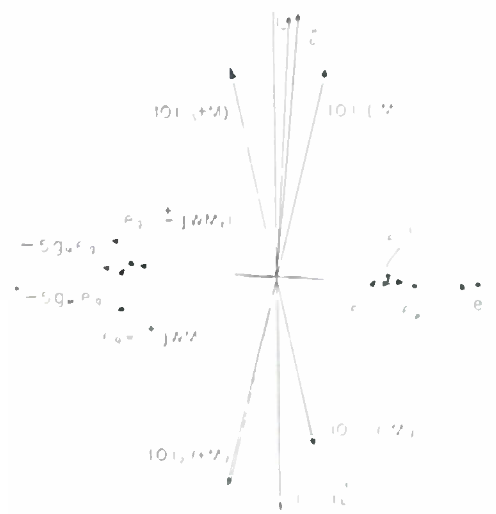

Fig. 9—Phase relations in the circuit of Fig. 8, at resonance, with the tickler coil coupled to the secondary coil. $i_{L'}$, etc., for $f<f_{02}$; and $i_{L''}$, etc., for $f<f_{02}$.

Thus the resonant frequencies determined by (13) are approximately equal to the actual operating frequencies (20), the difference in well-designed circuits being of the order of 0.5 to 1 kc. However because of the term

$$\frac{\theta_1}{\theta_2}\cdot\frac{1}{Q_2^2}$$

in (13), the difference between the operating and resonant frequencies increases as the resonant frequency approaches f_{02}; i.e., for large values of θ_1, and small θ_2. This produces a reactive load, and the efficiency is decreased. Because of this, it is not always possible for the circuit to work properly at the point of maximum G_{2e} which is fairly close to f_{02} (see Fig. 7).

The vector diagram in Fig. 9 shows how the operating frequencies can be controlled by the mutual coupling between the tickler and secondary coils. Notice that for $f<f_{02}$, M positive, e_g must equal $+j\omega M_1 i_2'$; and for $f>f_{02}$, M positive, e_g must equal $-j\omega M_1 i_2''$. (Refer to Fig. 11 for vector notation.)

It is, therefore, possible to force the oscillator to work at either the lower or higher resonant frequency by changing the coupling. This results in stable operation at the desired frequency.

The vectors as drawn in Fig. 9 are not in phase, since they were calculated using the frequencies given by the equation $\theta_1\theta_2 = k^2$. However only a slight change in frequency is necessary to bring the grid voltage into proper phase.

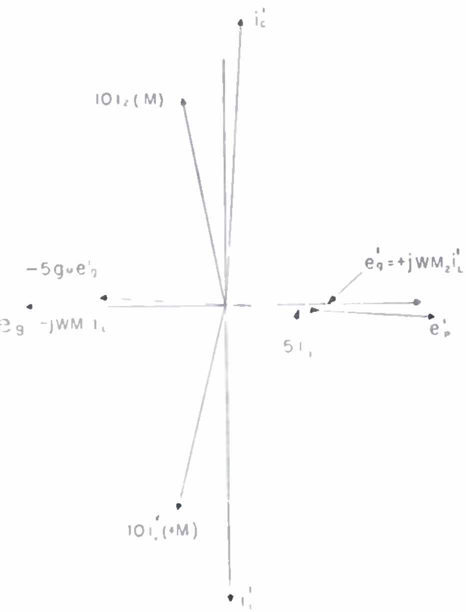

Fig. 10—Phase relations in the circuit of Fig. 8, with the tickler coil coupled to the primary coil.

(2) Tickler coil coupled to the primary coil: By a similar process to the above, it can be shown that, for e_g in phase with $g_m e_g$,

$$\theta_1\theta_2 = k^2 - \frac{\theta_1}{\theta_2}\cdot\frac{1}{Q_2^2}. \qquad (21)$$

The vector diagram for the primary-coupled tickler coil is shown in Fig. 10. Notice that the circuit will only operate if the primary-tickler coupling is positive, and also that the phase angle between $g_m e_g$ and e_g is the same for $f < f_{02}$ and $f > f_{02}$. Because of this the circuit will start oscillating at the frequency which presents the least load to the oscillator tube. If the primary circuit is then tuned to obtain maximum output, when the load reaches some value at or near maximum, depending on the absolute magnitude, the frequency will "jump" to the other resonant point which for this value of f_{01} will load the oscillator very lightly. Sudden small changes in load or voltage will also produce the same instability. (Refer to Fig. 11 for vector notation.)

To show how the preceding theory is used, the method followed in the design of a high-voltage power supply will be outlined briefly.

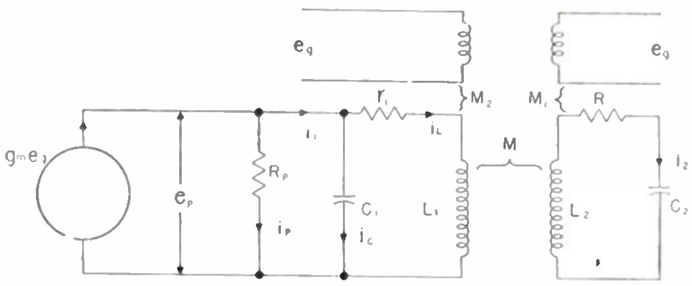

Fig. 11—Equivalent oscillator circuit used for plotting vector diagrams.

Starting from the power and voltage output required, and the dc plate voltage of the oscillator tube, a suitable rectifier circuit and value of secondary inductance are selected, using as a basis the approximate efficiency desired and the results of past experience. The self-resonant frequency of the secondary circuit is estimated, and therefore the approximate value of

$$Q_2 = \frac{2\pi f_{02} L_2}{R_2}$$

can be found. If a value of 100 is assumed for Q_1 (a figure easily obtainable), the required coefficient of coupling k can be computed. For good efficiency k should be approximately 20 k_c where

$$k_c = \frac{1}{\sqrt{Q_1 Q_2}}.$$

From the above results, the total power required for the secondary circuit can be computed and the primary circuit power loss estimated, thereby giving the total required ac power output P_T of the oscillator. Since the rest of the design depends on the value of the ac oscillator plate voltage E_p, it is advisable, at this stage, to determine a suitable value for the latter. This may be done using the usual class-C oscillator theory.

The resistances r_{Se} and r_{Le} in the equivalent circuit (Fig. 4) are negligible, in practice, compared to R_{Se} and R_{Le}. The required effective load resistance R_{Le} is therefore equal to E_p^2/P_{DC} where P_{DC} is the required power output.

There are still three factors to be determined. These are: (1) the operating frequency f; (2) the primary self-resonant frequency f_{01}; and (3) the primary inductance L_1.

These factors are related by the equations

$$\left(1 - \frac{f_{01}^2}{f^2}\right)\left(1 - \frac{f_{02}^2}{f^2}\right) = k^2 \tag{13}$$

$$R_{Le} = \frac{E_p^2}{P_{DC}} = \frac{1}{k^2}\frac{f_{01}^4}{f_{02}^4}\frac{L_1}{L_2}\left(1 - \frac{f_{02}^2}{f^2}\right)^2 \cdot R_L \tag{11}$$

$$X_{c1} = \frac{2\pi f_{01}^2 L_1}{f} \leq \frac{E_p^2}{4\pi P_T}. \tag{18}$$

These equations can be solved for f_{01}, f, and L_1. The solution of the equations, however, involves a cubic equation in the first variable selected. It has been found that a good first approximation for the solution of these expressions is to take $f = f_{02} - 4\alpha$ where α is given by (17).

Having obtained f, f_{01}, and L_1, the design of the primary tank circuit and the class-C oscillator can be completed. The various power losses in the circuit can be found by using the equivalent resistances. It may then be found necessary to change slightly some of the circuit constants previously assumed. The primary tuning and oscillator-tube screen voltage can then be adjusted to give the exact voltage output required.

The theoretical work outlined in the report has been confirmed, where possible, by actual experimental measurements on various circuits.

A more complete report of the design procedure, including all computations for inductances, etc., is being prepared and will be published in the near future as a National Research Council of Canada Departmental Report.

Acknowledgment

The author wishes to express his appreciation to J. E. Breeze, Division of Radio and Electrical Engineering, National Research Council of Canada, who suggested the problem and guided and supervised the experimental work, and to W. C. Brown of the National Research Council, who spent considerable time reviewing and constructively criticizing the report.

An Electronic Phasemeter*

EDWIN F. FLORMAN†, ASSOCIATE, IRE, AND ANDREW TAIT†

Summary—**The instrument described is designed to read and record directly on calibrated scales the phase angle between two sinusoidal voltages having a voltage range of 1 to 30 volts, over a frequency range of 100 to 5,000 cps to a sensitivity of 0.5°. The separate sinusoidal input voltages are converted to square waves through cascade amplifiers and limiters, and applied to two separate phase indicators. One indicator is ambiguous about the 180° value and measures and records the average of the algebraic sum of the square waves in the separate channels. The second type of indicator is unambiguous, and is operated by differentiating the square waves and applying the proper resultant pulses to fire a trigger circuit of the Eccles-Jordan type. The average plate current in the trigger tubes is directly proportional to the phase angle between the two input voltages to the phase meter.**

Introduction

An improved form of direct-reading audio-frequency phasemeter will be described in which the phase between two sinusoidal voltages is measured by converting them to square waves through two separate channels of amplifier limiters. A direct comparison of these square waves gives a measure of the phase between the original voltages. Two methods of comparing the square waves have been employed. One involves their direct addition in a circuit having two tubes with a common plate resistor, while in the other method the square waves are used to produce voltage "spikes" which, in turn, control a "trigger"-type unambiguous phase-indicating circuit.

Tests on this phasemeter show that it records and indicates unambiguously the phase between two input voltages from 0° to 360°, to a sensitivity of 0.5°, over a range of frequencies from 100 to 5,000 cps and for a voltage range of 1 to 30 volts. The relationship between phasemeter readings and phase is linear. A 72-hour stability test on the instrument showed that the maximum drift after the first 15 minutes of warm-up was approximately ±1.6°, while the maximum rate of drift was 0.25° per hour.

Description of Equipment

The phasemeter described in this paper embodies principles and circuits used in several other types of phasemeters.[1,2] It is believed that the circuit modifications incorporated in the instrument have increased its reliability over that of similar devices reported in the literature.

The phasemeter consists of two channels of cascaded amplifier-limiter stages, followed by two types of phase-indicating circuits, one being designated the "sum" indicator, and the other the "trigger" indicator (Fig. 1). These two indicating circuits are based on different principles, and were incorporated in the phasemeter for the purpose of making direct comparison between them.

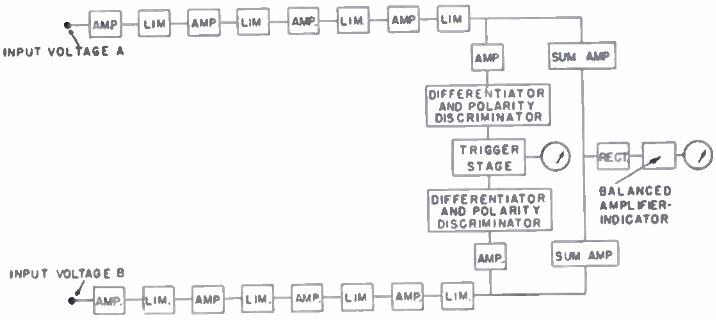

Fig. 1—Block diagram of the electronic phasemeter.

Sinusoidal input voltages are separately amplified and clipped in the four amplifier-limiter stages of each channel. The resultant square waves are then fed to the two separate indicating circuits (Fig. 2). The circuits of the first amplifier stages of each channel are designed to give symmetrical output into the first limiters for a voltage range of 1 to 30 volts. It is necessary that the output of this first stage be undistorted (or symmetrical), since otherwise there will be a spurious phase-shift indication with variation of voltage amplitude. The grid-voltage swings on the other three stages of each channel are limited to ±1.25 volts by the biased diode limiters, and hence distortion in these stages is low.

Referring to Fig. 2, the square-wave voltage at points A_1, A_2 are fed to each of the indicating circuits through separate 0.5-μf capacitors. Considering the "sum" indicator first, the square waves are applied to the grids of tubes V_9 and V_{18}, which have a common plate resistor R_{32}. The voltage appearing across resistor R_{32} is equal to the algebraic sum of the currents of the tube V_9 and V_{18}, multiplied by the value of R_{32}. Since these tubes are operated class-A, it follows that the ac component of the voltage across R_{32} is proportional to the algebraic sum of the square-wave inputs at points A_1 and A_2. The form of the ac voltage appearing across resistor R_{32} is a rectangular wave having a constant amplitude and a width which is proportional to the phase angle between the square waves applied to the grids of tubes V_9 and V_{18}. The average current obtained by rectifying the ac component of the voltage across resistor R_{32} is, therefore, a measure of the phase angle. This average value is obtained by means of the diode rectifier V_{19} and the balanced amplifier V_{20}, using a 0–1 dc milliammeter as the "sum" phase indicator. The balanced-amplifier circuit is arranged by means of proper meter shunts to give three ranges of phase-angle measurements: 180° to 135° or 225°, 180° to 90° or 270°, and 180° to 0° or 360°. It should be noted that, for each of these ranges, the

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† Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.

[1] Edward L. Ginston, "Electronic phase-angle meter," *Electronics*, vol. 15, p. 60; May, 1942.

[2] E. R. Haberland, "Direct Reading Electronic Phase Meter," Naval Ordnance Laboratory Memorandum No. 7900.

phasemeter reading is ambiguous about the 180° value.

The operation of the "trigger" indicator system is as follows: The square-wave voltages at points A_1 and A_2 are applied to the grids of amplifier tubes V_{21} and V_{22}. The amplified square waves are then differentiated in the plate circuits of V_{21} and V_{22}, and the resultant voltage spikes, which occur at the instant of rise and fall of the square waves, are applied to the diode polarity-discriminator tube V_{23}, which suppresses the positive voltage spikes and passes the negative pulses to the No. 3 grids of the "trigger" tubes V_{24} and V_{25}. The *RC* values in the plate circuits of V_{21} and V_{22} are a compromise so as to give the shortest possible pulse with sufficient amplitude to fire the trigger tubes.

The trigger-tube circuit is so connected that when a negative pulse is applied to the grid of one of the tubes it will cut off, simultaneously firing the other tube. In turn, when a negative pulse is applied to the "conducting" tube it will cut off, and, at the same time, will fire the first tube. As a result of this sequence of operation, the average plate current flowing in the plate circuit of the trigger tubes is a measure of the time interval existing between voltage spikes from the two channels of the phasemeter, and this time interval in turn corresponds to the relative phase of the sinusoidal input voltages.

The average trigger-tube plate currents are measured by means of the balanced circuit shown in Fig. 2. Both recording and indicating milliammeters are used, together with proper meter shunts, to give three ranges of phase-angle measurements: 130°–180°–230°; 80°–180°–280°; and 0°–180°–360°. A zero-centered meter is used for convenience, so that the center of the scale is the 180° position. The phase-angle readings in the trigger indicating circuit are unambiguous, but the circuit is inherently unstable for phase angles in the neighborhood of 0° and 360° because, under this condition, the order of firing of the trigger tubes alternates irregularly, with the result that the indicating meter swings from one end of its scale to the other. Fixed, regulated bias of 105 volts on the No. 1 grid of this trigger circuit was used, as recommended in footnote reference 2. The operation of this circuit over a period of several months proved it to be very stable and satisfactory.

The phasemeter indicating circuits are adjusted to read correctly for the 180° and for the 0/360° relationship between the input voltages as follows:

1) With the meter switch S_2 and the range switch S_1 in position 1, the meters are adjusted for their mechanical zeros.

2) The meter switch is moved to position 2, and the "sum" meter is set at zero by means of the "zero adjust" control on the right-hand side of the panel; this control is the balancing resistor in the plate circuits of tube V_{20}.

3) The meter switch is then advanced to position 3, and the phase of one of the input voltages is adjusted to give a minimum reading on the "sum" meter. The sum meter reading is then set at zero by means of the "sum bal." adjustment, which consists of variable resistors in the cathode circuit of tube V_{18}. This adjustment equalizes the gains of the "sum" tubes. For a more precise adjustment, the range switch is set on position 3.

4) The trigger meter is then set at zero by advancing the meter switch to position 4, and adjusting the "zero adjust" control on the left-hand side of the panel. This control is the variable resistor in the plate circuits of V_{24} and V_{25}. The above procedure determines the proper circuit conditions to give zero readings on the meters for a 180° phase relationship between the input voltages to the phase meter.

The 0/360° phase relationship between the input voltages can be indicated by the "sum" tube only, and is obtained as follows:

The range switch is set at position 1, and the phase of one of the input voltages is then adjusted to give a maximum reading on the "sum" meter. The "sum" meter

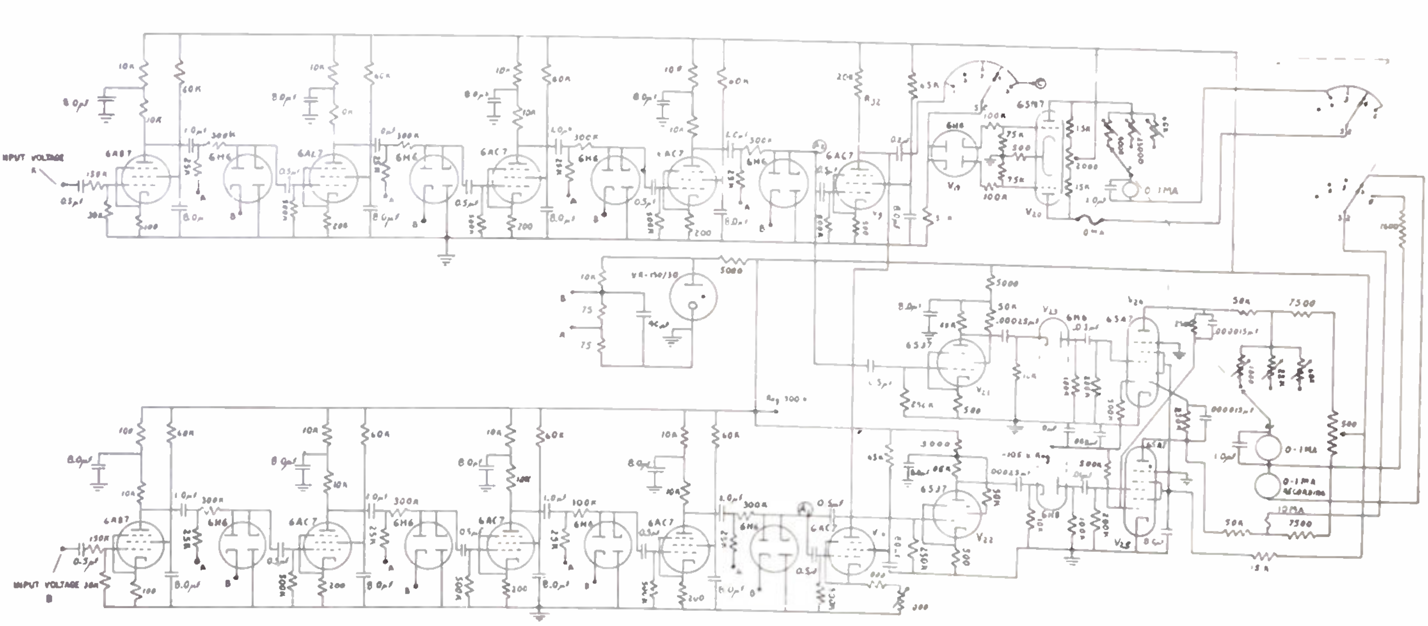

Fig. 2—Schematic wiring diagram of the electronic phasemeter.

reading is then set at 45 (360° = 180 + (4×45)) by means of the "range 1 adj." on the right-hand side of the panel.

Since the "trigger" phase-indicating circuit is unstable within ±1° of 0/360°, it is necessary to adjust the trigger meter by means of the "range 1 adj." on left-hand side of the panel, for a phase-difference between the input voltage somewhat less than 360°. This adjustment of the "trigger" phase-indicating circuit can be made in terms of the "sum" meter reading, or by means of a known phase difference between the input voltages to the phasemeter.

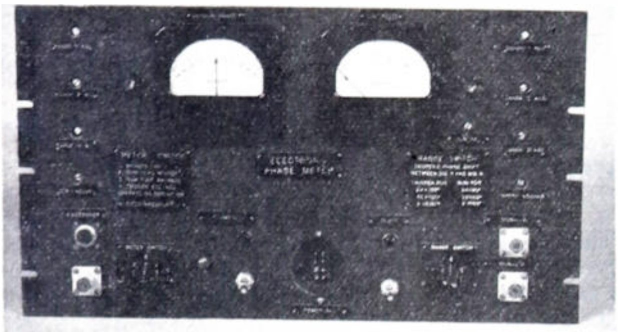

Fig. 3—Front view of the electronic phasemeter.

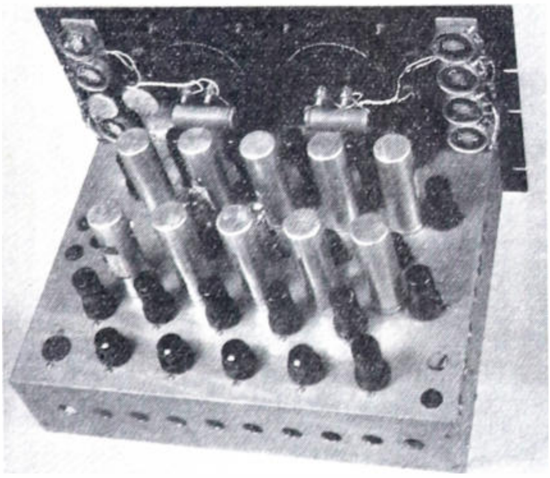

Fig. 4—Rear view of the electronic phasemeter.

Likewise, the more sensitive ranges of the phasemeter can be adjusted by means of the least-sensitive range, or by means of a known phase difference between the input voltages. In practice, these methods have been found to check each other to within ±0.5°.

Constructional details of the phasemeter are shown in Figs. 3, 4, and 5.

An independent calibration of the phasemeter was obtained by means of pairs of sinusoidal voltages having a fixed frequency ratio of 10 to 1 and a fixed relative phase. These calibration voltages were derived from the primary frequency standards of the National Bureau of Standards Radio Station WWV.

The calibration procedure consisted of feeding a sinusoidal voltage of desired frequency directly to one

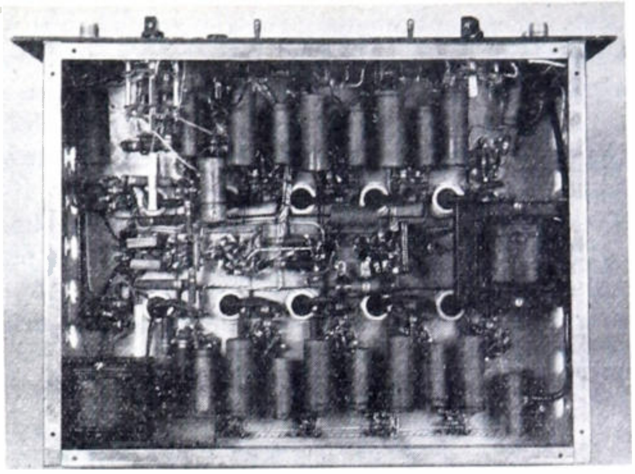

Fig. 5—Bottom view of the electronic phasemeter.

channel of the phasemeter, while simultaneously feeding this same voltage to the other channel of the phasemeter through a phase shifter. The output of this phase shifter was also connected to one set of deflection plates of an oscilloscope, while the other set of plates was connected to a reference sinusoidal voltage with a frequency 10 times the frequency of the calibrating voltage frequency and with a fixed phase relative to the calibrating voltage.

The resulting Lissajous figure on the screen of the oscilloscope indicated the phase shift between the input voltages to the phasemeter for any two positions of the phase shifter and could be read to within 0.5°, or less. Calibration points were obtained in steps of 18 electrical degrees by rotating the phase shifter until cycles were matched on the Lissajous figure.

The phase shifter, of the capacitance-goniometer type, used in this calibration procedure was capable of introducing a continuously variable phase shift of from 0° to 360°.

Performance

A series of tests on the phasemeter gave the following results:

(a) When the amplitudes of both signals were varied slowly and independently, keeping the relative phase constant, the phasemeter readings varied as shown in Table I.

TABLE I

Input voltage range, volts, rms	Indicated phase change, degrees	
	"Trigger"	"Sum"
1–10	±0.25	±0.25
1–20	±0.8	±0.35
1–30	±1.0	±0.50

(b) For very rapid random changes in the amplitudes of input voltages from 1 to 20 volts, rms, the phasemeter readings showed a phase change of ±1.0°. These rapid changes in input voltages were made by manually operating two potentiometers; the time required to vary the signal voltage from one extreme to the other was of the order of about one-quarter second.

(c) For a line-voltage variation of 100 to 120 volts, the phasemeter readings varied $\pm 0.25°$. These results were obtained for a range of signal frequencies from 100 to 5,000 cps, and for input voltages from 2 to 15 volts.

(d) After the first 15 minutes of warm-up, the phasemeter readings showed an indicated drift of $\pm 1.6°$ over a period of 72 hours, with a maximum rate of drift of 0.25° per hour.

(e) The curve of phasemeter reading versus phase was found to be linear within 1° over a range of input frequencies from 100 to 5,000 cps.

(f) The phasemeter readings were found to be independent of frequency from 100 to 5,000 cps.

Throughout all the above tests of the phasemeter the "sum" and the "trigger" indicators checked each other very closely. However, the "trigger" indicator has the advantage of being unambiguous, while, in the circuit shown in Fig. 2, the "sum" indicator circuit is about twice as sensitive as the "trigger" circuit.

The Use of "G" Curves in the Analysis of Electron-Tube Circuits*

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A SEMI-ANALYTIC method of designing circuits for electron tubes has been developed. This method makes use of small-signal amplification at a series of discrete points for calculation of large-signal characteristics. Amplification and distortion of cathode followers, degenerative amplifiers, class-A amplifiers, class-AB1, class-AB2, and class-B amplifiers, as well as cathode-coupled amplifiers, can be obtained without recourse to graphical methods.

Because of the use of small-signal amplification at a series of discrete points, determination of the linear range of operation and evaluation of distortion is simplified and rendered more accurate. The operating range is found by computing small-signal amplification over a range of grid-bias values, and comparing the results. Assuming the second harmonic to predominate and knowing that the percentage distortion is one-quarter the percentage variation of amplification, the range offering desired linearity can be chosen.

Equations in the simplest form for use with this technique are:

$$A = -g_mR_L/(1 + g_pR_L)$$
for ordinary amplifiers, (1)

$$A = g_mR_k/1 + (g_m + g_p)R_k$$
for cathode followers, and (2)

$$A = -g_mR_L/1 + (g_m + g_p)R_k + g_pR_L$$
for degenerative amplifiers. (3)

Similar equations can be determined for other circuits.

Derivations are based on the equation $di_p = g_m de_c + g_p de_p$. Use of this approach circumvents the complexity frequently found in the equivalent-circuit approach used in many texts.

The procedure for this method of circuit design follows. The load line is drawn on plate-characteristic curves carrying g_m and g_p contours (see Fig. 1). Values of g_m and g_p are determined at each intersection of the load line with a bias contour. Substitution of these data and the chosen circuit parameters into the appropriate equation gives the small-signal amplification at each bias value. Evaluation of effective amplification and distortion is then carried out by expansion of the amplification equation in a power series of grid signal. Application of a sinusoidal grid voltage gives components of fundamental and harmonics. These can be determined in terms of the change in small-signal amplification at the mean and extreme bias values. For predominant second harmonic, the formulas are:

$$A = \frac{A_1 + A_2}{2} \quad (4)$$

$$\text{per cent distortion} = 25\left(\frac{A_1 - A_2}{A_1 + A_2}\right), \quad (5)$$

where
A = effective amplification
A_1 = amplification at maximum positive excursion.
A_2 = amplification at maximum negative excursion.

All these A's are small-signal amplifications. In the case of push-pull operation, the sums of the amplifications of both tubes from instant to instant are used. The condition offering the minimum variation of this sum should be chosen. Equations similar to (4) and (5) can be derived.

Use of this method has proved to be time saving in selection of operating conditions meeting specific requirements. This results from elimination of replotting. The only graphical operation is drawing the load line.

Nomographs have been designed for these computations. Because of the similarity of (1), (2), and (3), all three can be solved through the use of two nomographs.

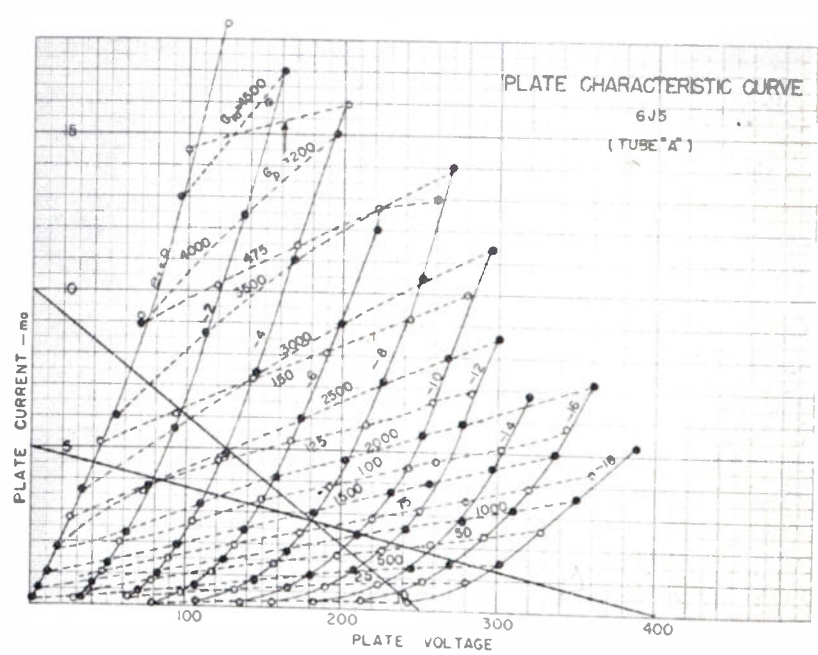

Fig. 1

* Decimal classification: R139.1. Original manuscript received by the Institute, June 7, 1948; abstract received, September 3, 1948. Presented, joint meeting URSI, American Section, and IRE, Washington Section, Washington, D. C., May, 1948.

† Ballistic Research Laboratories, Aberdeen Proving Ground, Md.

Contributors to Waves and Electrons Section

Edwin F. Florman (A'41) was born on February 16, 1904, in Venice, Ill. He worked as an electrician in a steel mill, during which time he completed his high school and undergraduate college work. In 1932 he was graduated from Washington University with the B.S. degree in electrical engineering, and in 1934 he received the M.S. degree in electrical engineering and physics from the same institution. From 1934 to 1941 he was employed as research physicist at the Western Cartridge Company of East Alton, Ill., where he was in charge of the development of electronic ballistics-testing equipment. From 1941 to 1944 he was employed as a radio engineer at the National Bureau of Standards. Here he was in charge of and completed a radio device for measuring upper-air wind velocities up to altitudes of 100,000 feet.

Edwin F. Florman

From 1944 to 1946 Mr. Florman was associated with the Philco Corporation, as a radio engineer in the research department. In 1946 he returned to the National Bureau of Standards, where he is at present conducting low-frequency radio-wave-propagation experiments in the Central Radio Propagation Laboratory. Mr. Florman is a member of Tau Beta Pi and Sigma Xi.

❖

Robert W. Friis (A'34–M'40–SM'43) was born on October 10, 1907, in Kenmare N. D. He received the B.E.E. degree from the University of Minnesota in 1930. From 1930 to 1945 he was a member of the technical staff of the radio research department of Bell Telephone Laboratories, Inc., engaged in the development and design of various types of radio and radar transmitters. Since 1945, as a member of the radio projects and television research department, he has been in charge of an equipment design group participating in the development of microwave radio relay systems. Mr. Friis is a licensed professional engineer in the State of New York.

Robert W. Friis

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For a photograph and biography of A. L. Durkee, see page 244 of the February, 1948, issue of the Proceedings of the I.R.E.

❖

George W. C. Mathers (S'45–A'47) was born on July 5, 1923, at Calgary, Alberta, Canada. He obtained the B.Sc. degree in engineering physics at the University of Alberta in 1946. While attending that University he received the Engineering Institute of Canada award in his third year, and the Henry Birks and Sons Gold Medal in his fourth year for heading the engineering class.

G. W. C. Mathers

Since graduation Mr. Mathers has worked at the National Research Council of Canada, Radio and Electrical Engineering Division, on several problems associated with radar research and development.

❖

A. A. Roetken (A'29–M'40–SM'43) was born on October 15, 1902 in Covington, Ky. He received the degree of B.E.E. in 1927 and the M.S. degree in 1929, both from the Ohio State University. He joined the Bell Telephone Laboratories, Inc., in 1929 as a technical staff member of the radio research group. His work included the development of precision frequency-measuring equipment, single-sideband radio receivers for transoceanic telephone circuits, and ultra-high-frequency radiotelephone circuits for domestic service. During World War II, Mr. Roetken's principal activity was in the development of pulse-multiplex microwave radio repeaters for the armed forces. Since then he has collaborated with others in developing the New York–Boston microwave repeater system and similar projects.

A. A. Roetken

❖

Andrew Tait was born in Bridgeport, Conn., on April 22, 1920. Upon graduation from high school, he went with the Rockefeller Institute for Medical Research, New York, N. Y., as a technical assistant in the physical chemistry laboratories. There he was engaged in the design of specialized electronic equipment. He attended evening sessions at Brooklyn Polytechnic Institute, and was also engaged in experimental work in television receivers and preamplifiers for the "Fringe" operation.

Andrew Tait

In 1942 Mr. Tait joined the National Bureau of Standards staff, where he engaged in the field of high-frequency direction finders. In 1944 he went on active duty as an officer in the Electronics Engineering Division of the U. S. Coast Guard, working with specialized military radio direction-finding equipments and associated radio communication equipment in this country and overseas.

In 1946 Mr. Tait rejoined the staff of the Central Radio Propagation Laboratory at the National Bureau of Standards, and is at present working in the field of navigation systems and the associated field of low-frequency propagation.

❖

G. N. Thayer (SM'47) was born on October 6, 1908 in Delta, Colo. Upon graduation from Stevens Institute of Technology with the M.E. degree in 1930, he joined the technical staff of Bell Telephone Laboratories, Inc., where, until the beginning of the war, he was engaged in the development of radio equipment for various mobile services. During the war he was project engineer for a number of radar systems. Since the war Mr. Thayer has been project engineer for the New York–Boston microwave system, and is now transmission development engineer in charge of the development of several radio systems for Bell System use.

G. N. Thayer

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John Austin Victoreen was born in Johnstown, Pa., on July 4, 1902. An ardent radio amateur in the early days of radio, he founded the Victoreen Radio Company in 1920, devoting his time to the development and manufacture of radio receiver components until 1930.

J. A. Victoreen

In 1925 Mr. Victoreen founded the Victoreen Instrument Company to develop and manufacture the X-ray dosage-measuring instruments used by the medical profession today. Since 1930 he has confined his efforts to the development of instruments for the measurement of X-rays and radioactivity. From 1940 to 1945 he devoted his full time to the development of radiation-measuring instruments for the atomic bomb project under the Manhattan Engineering District, setting up facilities in 1943 for the manufacture of subminiature electrometer tubes. Much of the radiation-measuring instrumentation used on the Bikini and Eniwetok bomb tests was designed and manufactured by the Victoreen Instrument Company under his supervision.

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534.213.4 **1**

On the Radiation of Sound into a Circular Tube—U. Ingård. (*Chalmers Tekn. Högsk. Handl.*, no. 70, 49 pp.; 1948. In English.) A study of the propagation of sound waves in rigid-walled tubes from sound sources of circular symmetry, in particular, a plane piston source. The radiation resistance and the pressure distribution in the tube are calculated, and verified by measurements. The equivalent circuit is discussed. The case of a cylindrical resonator is considered, and curves are given for calculated and measured resonance frequencies.

534.756+621.39 **2**

Relationship Between Rate of Transmission of Information, Frequency Bandwidth, and Signal-to-Noise Ratio—Earp.(*See* 62.)

534.78 **3**

A Playback for Visible Speech—L. O. Schott. (*Bell Lab. Rec.*, vol. 26, pp. 333–339; August, 1948.) Description of a rotatable frosted lucite drum and associated photoelectric equipment which reproduces speech from opaque patterns on the surface of the drum.

534.842:791.45 **4**

Auditorium Acoustics—J. P. Maxfield. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 169–176; August, 1948. Discussion, pp. 176–183.) A review of the factors affecting the acoustic properties of auditoria, with special reference to the cinema.

534.844.1 **5**

Measuring Reverberation Time by the Method of Exponentially Increasing Amplification—W. Tak. (*Philips Tech. Rev.*, vol. 9, no. 12, pp. 371–378; 1947 and 1948.) Periodic sound impulses are generated within the space whose reverberation time is to be measured and the voltage from a microphone within this space is applied, after exponential amplification, to a cro. The amplification of each pulse voltage begins with a low value, rises exponentially to 60 db, and then drops back to the initial low value. When the sweep period is equal to the reverberation time, the oscillogram has an approximately constant amplitude. The equipment is described and experimental results are discussed.

534.844.1:621.385.832 **6**

Reverber-O-Graph (An Electronic Reverberation Recorder)—N. B. Bhatt and D. L. Subrahmanyam. (*Electrotechnics*, pp. 47–54; March, 1948.) A portable mains-operated recorder which indicates the decay of sound intensity in rooms in db per second on the screen of a crt of long persistence. The sweep time can be 1, 2, 3, 5, or 7 seconds, so that the reverberation times of rooms of any size can be determined. A logarithmic amplifier stage enables a range of sound intensity of 75 db to be recorded either by tracing or by photographic means. Full circuit details are given, with examples of typical sound-decay curves.

534.845 **7**

Behavior of Acoustic Materials—R. K. Cook. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 192–200; August, 1948. Discussion, pp. 200–202.) The mechanism of absorption in various materials is discussed and methods for determining normal-incidence and random-incidence absorption coefficients are outlined. The latter coefficient is considered the more useful in auditorium design. Recommendations for painting acoustic materials are made and the results of such painting are illustrated.

534.862:621.396.662.3.029.3 **8**

Continuously Variable Band-Elimination Filter—Singer. (*See* 59.)

621.395.616 **9**

Circuit for Condenser Microphones with Low Noise Level—J. J. Zaalberg van Zelst. (*Philips Tech. Rev.*, vol. 9, no. 12, pp. 357–363; 1947 and 1948.) The microphone is incorporated in a bridge circuit which, as the diaphragm vibrates, supplies an AM high-frequency voltage, which is amplified and detected. The bridge circuit can be connected to the amplifier by a cable, and a high resistance in series with the microphone is not required.

621.395.616:534.76 **10**

A Condenser-Microphone for Stereophony—A. Rademakers. (*Philips Tech. Rev.*, vol. 9, no. 11, pp. 330–338; 1947 and 1948.) The requisites for such a microphone are discussed and the construction of the instrument used in Philips' laboratory is described. This instrument has a diaphragm of effective diameter about 15 mm and of thickness 15 μ, whose natural frequency is raised to 12,000 cps by using the "stiffness" of the air cushion between it and the back electrode 13 μ away. The resonance peak is eliminated by air-damping, so that the response is flat, to within 2 db, up to 14,000 cps. The threshold value is below 30 phons and the sensitivity with an applied direct voltage of 100 v is 3mV/μ bar.

621.395.625:621.396.97 **11**

Sound Recording as Applied to Broadcasting—M. J. L. Pulling. (*BBC Quart.*, vol. 3, pp. 108–121; July, 1948.) A technical discussion of the methods used by the British Broadcasting Corporation. Mechanographic methods using disks or film are compared with magnetic systems using tape or wire, with special reference to the difficulties associated with editing and continuous recording and reproduction. The requirements for both static and transportable equipment are considered and the need for international standards of recording is mentioned. No single existing system meets all requirements.

621.395.625.3 **12**

The "Edispot": A Spotting Device for Magnetic Tape-Editing—R. S. O'Brien (*Audio Eng.*, vol. 32, pp. 11–13, 46; July, 1948) Provides means for winding, timing, and playback. An auxiliary rotating drum and pickup head enable a small portion of the tape to be examined continuously, either aurally or by cro display, so that any particular section of a record can be located accurately.

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.3.09 **13**

Propagation of Electromagnetic Waves in Concentric Wide-Band Cables with Longitudinal Dielectric Spacers—H. Buchholz (*Arch Elek.* (Übertragung), vol. 1, pp 137–150; September and October, 1947.) The mathematical treatment is simplified by considering the analogous case of propagation in the rectangular channels between uniformly distributed dielectric spacers between 2 plane-parallel plates. Two solutions are obtained by combining the E and H fields. Critical frequencies and the propagation constant are calculated and shown graphically. The characteristics of the magnetic field, the low losses due to current heating, and attenuation are also investigated.

621.315.212:621.3.09 **14**

Attenuation and Phase Distortion due to Internal Irregularities and Repeater Mismatch of Coaxial Cables—G. Fuchs. (*Câbles and*

Trans. (Paris), vol. 2, pp. 233–241; July, 1948. With English summary.)

621.315.212:621.317.336 15
Study of the Impedance Irregularities of Coaxial Cables by Oscillographic Observation of Pulse Echoes—Couanault and Herreng. (*See* 142.)

621.392.029.64 16
$TM_{0,1}$ Mode in Circular Wave Guides with Two Coaxial Dielectrics—S. Frankel. (*Elec. Commun.* (London), vol. 25, pp. 152–157; June, 1948.) Reprint of 22 of 1948.

621.392.029.64:621.315.61 17
Fields in Nonmetallic Waveguides—R. M. Whitmer. (Proc. I.R.E., vol. 36, pp. 1105–1109; September, 1948.) Green's function is found for an infinite plate of dielectric material having the electric field parallel to the surfaces of the plate. The solution is found as a Fourier integral which is then replaced by a complex contour integral. The residues at the real poles, of which there are a finite number, correspond to the propagating modes in a metallic waveguide. An integral around a branch cut gives a wave radiating into space, analogous to the attenuated modes of the metallic guide.

The modal field distributions are discussed for the plate and for circular rods. The surface fields are not small, but are attenuated transversely at rates ≥ 28 db per radius. A nonmetallic waveguide can be useful wherever a cheap flexible conductor is needed and imperfect shielding can be tolerated.

621.396.67:621.317.336 18
Television Receiving Aerials: Impedance Measuring and Matching Methods—D. A. Bell. (*Jour. Brit. I.R.E.*, (London), vol. 8, pp.19–39; January and February, 1948. Discussion, pp. 39–40.) Describes impedance measurement by observation of standing-wave patterns and the use of circle diagrams, and also by the substitution method. Wide-band matching devices, the use of fractional-wavelength lines as transformers, and balance versus unbalance converters are mentioned briefly.

621.396.67:621.397.743 19
Television Antenna and RF Distribution Systems for Apartment Houses—Kallmann. (*See* 243.)

621.396.677 20
Fed-Dipole Groups as Longitudinal Radiators for Broad Frequency Bands—O. Zinke. (*Funk. und Ton*, vol. 2, pp. 435–442; September, 1948.) Arrangements of dipoles in line with separations reduced according to a certain law, are shown to give beams 30 to 45 per cent narrower than other array systems and cover a 3 to 1 frequency range. For a given half-value width of the radiation diagram, a Yagi antenna is 2 to 3 times longer than a longitudinal radiator of the new type with alternating fed and reflector dipoles. The new type corresponds in its operation to the dielectric directional arrays of Mallach and extends these for $\lambda > 50$ cm. Typical radiation diagrams are given.

621.396.677 21
A Discussion of the Maximum Directivity of an Antenna—T. T. Taylor. (Proc. I.R.E., vol. 36, p.1135; September, 1948.) Comment on 2731 of 1948 (Riblet). It is shown that as the diameter of a sphere, corresponding to the aperture of a broadside array, is reduced below 50 λ, the radiation *Q*, which is the ratio of the energy stored in the intense portion of the induction field to the radiated energy, increases at an exceptionally high rate. In consequence, critical tolerance and bandwidth as well as ohmic loss are important limitations in the design of superdirective antennas. The mathematical basis is outlined and examples are given.

621.396.677 22
The Maximum Directivity of an Antenna—D. A. Bell. (Proc. I.R.E., vol. 36, p. 1134; September, 1948. Discussion on 2731 of 1948 (Riblet).

621.396.677 23
An Experimental Investigation of the Radiation Patterns of Electromagnetic Horn Antennas—D. R. Rhodes. (Proc. I.R.E., vol. 36, pp. 1101–1105; September, 1948.) 250 radiation patterns are shown for rectangular horns, as functions of the electric- and magnetic-plane flare angles (from 0° to 50°) and of the radial length of the horn (from 0 to 50 λ). Characteristic properties are thus revealed; their effect on the design of such horns is discussed.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.018.41 24
Negative Frequency—"Cathode Ray." (*Wireless World*, vol. 54, pp. 361–364; October, 1948.) The distinction between positive and negative frequencies is discussed for 2-phase supply, and negative frequency is considered in relation to modulation and frequency changing. A 2-phase method of eliminating second-channel interference, discussed fully by Barber (2697 of 1947) is explained briefly. See also 1866 of 1948 (Madella) and 1867 of 1948 (Barber).

621.314.632:621.396.622.63 25
Applications of Crystal Rectifiers at Frequencies up to 10,000 Mc—J. H. Evans. (*Jour. Brit. I.R.E.*, (London), vol. 8, pp. 112–121; May and June, 1948. Discussion, pp. 121–125.) The construction and general mechanical details of crystal rectifiers are described. Various applications considered include frequency changing and detection in microwave receivers and low-frequency rectification for meters.

621.316.726.078.3:621.396.615:621.396.619.13 26
Simplified Automatic Stabilization of a Frequency-Modulated Oscillator—J. L. Hollis. (Proc. I.R.E., vol. 36, pp. 1164 1171; September, 1948.) A quartz-crystal discriminator is operated at the frequency of the modulated oscillator, so that heterodyning and mixing circuits are unnecessary. A special bridge circuit, operating on the modulator bias, maintains the average frequency equal to the mean frequency of the discriminator. The circuit arrangements are described; the performance and stability of a complete unit are discussed briefly.

621.316.86 27
Fixed High-Stability Carbon Resistors—T. Holmes. (*Jour. I.E.E.* (London), Part IIIA, vol. 94, no. 16, pp. 912–914; 1947.) Discussion of manufacturing problems. Ceramic rods are coated with a carbon film by pyrolitic decomposition of methane in a continuous-process furnace. The resistance is increased by cutting a spiral track on the coated surface. The ends are coated with graphite for good contact with the metal end caps which are then fitted. The whole resistor is given a lacquer finish. The furnace and its control characteristics are described. Typical stability figures for the resistors are tabulated.

621.316.89 28
Production of Frequency-Independent Purely Ohmic Resistors for Short Waves—W. Burkhardtsmaier. (*Funk und Ton*, vol. 2, pp. 381–391; August, 1948.) A device is described whose input resistance remains equal to its dc value up to very high frequencies. The limiting frequency depends on the length of the resistor used. The device resembles that of Roosenstein (137 of 1944) and has a concentric return conductor. It is particularly suitable for use in compensation circuits, whose design is discussed.

621.318.42 29
Contribution to the Calculation of Premagnetized Choke Coils—J. Kammerloher. (*Funk und Ton*, vol. 2, pp. 443–453; September, 1948.) The effective core permeability of such chokes is defined and methods for calculating it are given, with relevant formulas and numerical examples.

621.318.572 30
An Electronic Switch with Variable Commutating Frequency—E. E. Carpentier. (*Philips Tech. Rev.*, vol. 9, pp. 340–346; 1947 and 1948.) Details of an instrument suitable for the simultaneous production of 2 oscillograms, and with switching period continuously variable from 2 to 40,000 per second.

621.318.572:621.396.96 31
The Strobe Principle in Radio and Radar—L. H. Bedford. (*Jour. Brit. I.R.E.*, (London), vol. 8, pp. 62–70; March and April, 1948.) A concise theoretical explanation, with discussion of various practical circuits.

621.392 32
Network Transformations—(*Distrib. Elec.*, vol. 21, pp. 40–43; October, 1948.) A collection of formulas for (a) series-parallel transformations for resistances and for impedances, (b) star-delta and derived transformations for impedances and admittances, with simple numerical examples, some of which involve negative components.

621.392 33
Dual Circuits—W. Klein. (*Funk und Ton*, vol. 2, pp. 392–395; August, 1948.) Methods are known for deriving dual, or inverse, circuits for plane networks. With the help of 2 dual potential decoupling arrangements, which consist of ideal transformers with a 1 to 1 transfer ratio, an equivalent plane network can be obtained for any network and a dual network can then be derived.

621.392 34
Three Investigations on the Pulling of Oscillations—H. G. Möller. (*Elektrotechnik* (Berlin), vol. 1, pp. 65–71; September, 1947.) Theory is given of the pulling phenomena for 2 self-oscillating systems. The theory is based on 2 fundamental equations, 1 representing the resultant oscillation and a nonlinear 1 representing the amplitude limitation in the self-oscillating systems. Confirmation of the theory is given by the results of researches on (a) the energy transfer between 2 tube transmitters with different amplitudes, (b) the current in the coupling conductor, and (c) the transition from the heterodyne to the pulling regime.

621.392:621.385.5.001.8 35
Applications of Screen-Grid Supply Impedance in Pentodes—Sulzer. (*See* 268.)

621.392.015.3:537.311.5:517.941 36
The Determination of the Integration Constants when Calculating Transient Phenomena—B. D. H. Tellegen. (*Philips Res. Rep.*, vol. 3, pp. 24–36; February, 1948.) "A network is considered containing a voltage source *v* under the influence of which a current *i* flows in a certain branch. A method is given for calculating from the differential equation connecting *i* and *v* the discontinuities in *i* and its derivatives resulting from discontinuities in *v* and its derivatives. The method is applied to the calculation of periodic phenomena caused by periodic sources containing discontinuities."

621.392.43 37
Antenna Matching with Line Segments—J. G. Marshall. (*QST*, vol. 32, pp. 18–21, 104; September, 1948.) Simple design formulas, for series- or shunt-balanced networks using a section of transmission line as a transformer, are derived for the general case when each element of the network may have a different characteristic impedance.

621.392.5 38
Delay Networks from Low-Pass Units—J. Linke. (*Elektrotechnik* (Berlin), vol. 1, pp.

43–51; August, 1947.) The transmission characteristics of low-pass filters of various types are discussed, with special reference to their use in the construction of delay networks. Designs are given for low-pass delay networks with given damping and characteristic impedance, and the behavior of these networks is compared with that of networks constructed from phase-rotation units.

621.392.5:621.385.029.63/.64 39
Experimental Method for *a priori* Tests of Delay Lines for the Travelling-Wave Amplifier—P. Lapostolle. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 388–389; August 9, 1948.) In the calculation of the gain of the amplified wave, the part of the delay line outside the beam only involves a single coefficient *P*, which is a measure of the amplification efficiency. Details and theory are given of a method for determining *P*, in which the electron beam is replaced by a cylindrical rod of dielectric which retards the waves. The retardation can be very easily expressed in terms of *P*. See also 2325 of 1947.

621.392.5.012 40
On the Amplitude Curve Associated with a Linear Phase Characteristic in a Quadripole with Minimum Dephasage—J. Laplume. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 187–188; July 19, 1948.) The curve representing the amplitude response of such a quadripole is determined for the case where the dephasage is rigorously linear between 2 finite limits and constant outside these limits. The gain curve, has a maximum at a frequency equal to the geometric mean of the limits; it falls off more rapidly toward the high than toward the low frequencies. Between the limiting frequencies the curve is rounded, in contrast to the flat characteristic often aimed at in design.

621.394/.395[.645:621.394/.395].73 41
Line Amplifier for the Paris-Toulouse Coaxial Cable—Sueur. (*See* 212.)

621.395.665.1 42
Surgeless Volume Expansion—A. A. Tomkins. (*Wireless World*, vol. 54, p. 347, September, 1948.) Corrections to 2739 of 1948.

621.396.611.1 43
Building-Up Processes in Two Coupled Linear Oscillatory Systems with Arbitrary Damping, and Type and Magnitude of Coupling—P. Schneider (*Arch Elek* (Übertragung), vol. 1, pp. 91–100 September and October, 1947.) The essential feature of the investigation is the calculation of the biquadratic Heaviside function of the system of simultaneous homogeneous differential equations of the free oscillations, by a method previously described (2627 of 1942). Three cases are considered, depending on the value of the damping. In the lower range, 2 pairs of conjugate complex roots are found, in the middle range 1 pair of conjugate complex roots and 2 real roots, and in the upper range, 4 real roots. These correspond respectively to (a) 2 oscillatory, (b) 1 oscillatory and 1 periodic, (c) 2 aperiodic phenomena. An equivalent circuit is used in the discussion of general coupling relations. The Laplace transformation is found particularly suitable for the investigation of building-up effects. The calculations are applied to the 2 important practical cases of excitation by a damped sinusoidal voltage or by a surge voltage, for arbitrary initial conditions and different degrees of damping.

621.396.615 44
Theory of Amplitude-Stabilized Oscillators—P. Aigrain and E. M. Williams. (*Rev. Sci.* (Paris), vol. 85, pp. 839–846; August 15, 1947.) For other accounts see 980 and 1883 of 1948.

621.396.615 45
A Phase-Shift Oscillator with Wide-Range Tuning—G. Willoner and F. Tihelka. (Proc. I.R.E., vol. 36, pp. 1096–1100; September, 1948.) An oscillator covering the frequency range 100 to 10,000 cps. The phase of the feedback is controlled by the delay in a 3-terminal all-pass network. The tuning is achieved by varying a single resistance element of the network.

621.396.615 46
Positive-Grid Oscillators—A. V. J. Martin. (*Radio Tech. Dig.* (*Franç.*), vol. 2, pp. 135–143; June to August, 1948.) Simple explanation of basic principles, with formula for oscillation frequency.

621.396.615 47
On a double-Triode Oscillator and its Application to the Study of Dielectrics—R. Létienne. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 506–508; August 30, 1948.) A modified Franklin oscillator, with multivibrator locked by an oscillatory circuit and with cathode coupling and an amplitude limiter. Stability of frequency is of the same order as that of a crystal oscillator without a thermostat. With coils of good quality, the amplitude remains constant over a wide range of frequencies and the harmonic content is negligible. Calibration varies little with the age of the tubes. With this type of oscillator, the 2-oscillator method for the measurement of dielectric constants can be used for both low-loss and lossy materials

621.396.615 48
Application of the Tube 833-A as Ultra-High Frequency Oscillator—S K Chatterjee and B. V. Sreekantan (*Electrotechnics*, pp 55–63; March, 1948.) Experimental investigation of the behavior of this tube when used as a resonant line oscillator to generate frequencies from 26.5 to 210 Mc. Results are shown graphically and tabulated, efficiency falls off sharply for frequencies above 60 Mc or transit angles above 30°.

621.396.615.621.394.441 49
Stabilized Oscillator for Multiplex Voice-Frequency Telegraphy—Delvon and Manière. (*See* 209.)

621.396.615.17 50
On the Performance of the Push-Pull Relaxation Oscillator (Multivibrator)—R Feinberg. (*Phil. Mag.*, vol 39, pp 268–281 April, 1948.) A mathematical analysis of the multivibrator, based on the assumption that it acts as a 2-way electronic switch which periodically charges and discharges 2 capacitors in alternate sequence. A linear relation is introduced as an approximate equivalent for the nonlinear tube characteristics. Calculated frequencies agree within 1 to 15 per cent with experimental values.

621.396.615.17:621.319.55 51
Electrical Sawtooth Oscillations: Part 2—H. Hertwig. (*Funk und Ton*, vol 2, pp. 469–474; September, 1948.) Practical circuits are given for oscillators using gas discharge tubes, with technical data for suitable tubes Philips 4686 and EC50, AEG S1/0, 2II A and S1/0, 2II. Part 1, 3056 of 1948.

621.396.615.17/.18 52
High-Ratio Multivibrator Frequency Divider—M. Silver and A. Shadowitz (*Elec. Commun.* (London), vol. 25, pp. 160–162; June, 1948.) The capacitor between the first grid and the second anode of a multivibrator divider is removed and replaced by a high-*Q* tuned circuit followed by a 2-stage clipper. The circuit will now only oscillate in the immediate vicinity of the resonant frequency of the tuned circuit. At the output, a square wave is obtained at this frequency, which is arranged to be the required sub-harmonic of the synchronizing frequency. A circuit is shown that gives stable division ratios as high as 300 to 1. See also 1488 of 1946 (Applegarth) and 3252 of 1941 (Builder).

621.396.615.18 53
Frequency Division—J. Grosskopf. (*Fernmeldetech. Z.*, vol. 1, pp. 113–119; August, 1948.) A review of the characteristics of modern frequency dividers of the regenerative-modulation type considered by Miller (3687 of 1939) and the quasistable type discussed by Fortescue (3099 of 1939), with particular reference to the pulling-in region and division ratio. See also 3490 of 1938 (Hudec) and 3491 of 1938 (Golicke).

621.396.645+621.396.828.1 54
Reducing Heater Hum—K. G. Britton. (*Wireless World*, vol. 54, p. 360; October, 1948.) An appropriate proportion of the heater voltage is injected into each tube in antiphase to the hum, which is thus reduced below the level of the tube noise. See also 1689 of 1947 (Baxandall).

621.396.645:621.385.4 55
The FP-54 as a Stable Voltage Amplifier—N. T. Seaton *Canad. Jour. Res*, vol. 26, Sec F, pp. 302–309; July, 1948.) A simple method is given for operating the FP-54 pliotron as a stabilized amplifier with a grid current $< 10^{-17}$ ampere. The application to the construction of a compact, portable electrometer is discussed briefly.

621.396.645.371.029.3 56
Negative Feedback Calculations—E. J James. (*Wireless World*, vol. 54, pp. 326–330, September, 1948.) Simplified design formulas are given, which "enable the person with little mathematical skill or knowledge to design a feedback circuit suitable to his amplifier and his requirements."

621.396.662 57
Stub Tuners for Power Division—C. E. Smith. (*Communications*, vol. 28, pp. 22–23; August, 1948.) An arrangement of concentric-line stub tuners which provides continuously variable division of the power from a common input between 2 loads, while preserving the correct matching conditions.

621.396.662.3 58
Nonsymmetrical Matched Filters—W Herzog (*Arch Elek* Übertragung), vol 1, pp. 122–127; September and October, 1947.) Discussion of various nonsymmetrical filter circuits which allow matching within wide limits, with simple design formulas.

621.396.662.3.029.3:534.862 59
Continuously Variable Band-Elimination Filter—K Singer. (*Jour Soc. Mot Pic. Eng.*, vol 51, pp 203–210; August, 1948.) The filter is essentially a 4-stage zero gain amplifier incorporating a Wien bridge and with 26-db feedback to give extra steepness of the response curve on both sides of the rejection frequency. This frequency is adjustable from 30 to 9,000 cps, and 50 db rejection is obtained anywhere in that range which is covered in 5 overlapping bands. The filter has proved very useful for the elimination of interference frequencies, such as arc whistles, in the production of sound for motion pictures.

GENERAL PHYSICS

53.081.5:530.1 60
Some Electrical Applications of Dimensional Analysis and of the Principle of Similitude—L. Lakaye. (*Bull. Sci. Ass. Inst.* (Montefiore), vol. 61, pp. 219–257; May to August, 1948.) Discussion includes the theory of experiments on reduced-scale models. See also 2504 of 1948 (Dzung and Meldahl) and back references.

530.162+621.396.822 61
The Brownian Movement and Spontaneous Fluctuation of Electricity—D. K. C. Mac-

Donald. (*Research* (London), vol. 1, pp. 194–203; February, 1948. Bibliography, pp. 203–204.) A historical review of the development of present knowledge of the subject. The principal formulas are given.

534.756+621.39 62
Relationship Between Rate of Transmission of Information, Frequency Bandwidth, and Signal-to-Noise Ratio—C. W. Earp. (*Elec. Commun.* (London), vol. 25, pp. 178–195; June, 1948.) An integrated modern theory of communication is developed. The exact benefit to be gained by the use of the right balance between rate of transmission of information and the frequency bandwidth used is estimated for the established expanded band communication systems. The particular features of transmission systems necessary for optimum performance are considered. A new system of modulation, tentatively called "step modulation," is shown to have all known characteristics of efficiency. The exact performance of the system is examined and it is found that the process of bandwidth expansion to yield improved demodulated signal-to-noise ratio may be inverted to provide communication through reduced bandwidth at the cost of signal-to-noise ratio. The new system may define the theoretical limit of efficient use of frequency bandwidth for any transmission system, for the case when the only known characteristic of the information wave is the frequency band that contains it. See also 1057 of 1947 (Gabor).

535.317.9:621.397.5:535.88 63
The Manufacture of Correction Plates for Schmidt Optical Systems—H. Rinia and P. M. van Alphen. (*Philips Tech. Rev.*, vol. 9, pp. 349–356; 1947 and 1948.) Discussion of the optics of the Schmidt system, with a method of manufacturing the correction plate. A mould with the thickness dimensions increased in the ratio 5 to 1 is used to form a gelatin plate from a 20 per cent solution, which, on drying, shrinks to the required thickness, lateral shrinking being prevented by a glass backing plate. See also 2652 of 1948 (Friedman) and 2767 of 1948 (de Groot).

537.21:621.385.833 64
The Electrostatic Field Distribution near a Circular Aperture or Short Cylinder—G. Liebmann. (*Phil. Mag.*, vol. 39, pp. 281–296; April, 1948.) The potential is the sum of that calculated by Bertram (1215 of 1942) and a correction term. In many cases of interest in electron optics or tube design, the correction term is the more important. Both terms are tabulated.

537.228.1:621.395.625.6 65
Light Modulation by P-Type Crystals—G. D. Gotschall. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 13–20; July, 1948.) The method described uses the linear electro-optic effect in P-type crystals. "PN" crystals (dihydrogen ammonium phosphate) give good results. A parallel beam of polarized light is used. Voltages as high as 9 kv may be required, but practically no current is taken. The flat response extends well beyond the audio range. The simple construction, excellent dark-to-light ratio, high input impedance, and good transmission characteristics are particularly applicable to variable-density sound-film recording.

537.311.33:621.396.645:621.315.59 66
The Transistor, A Semi-Conductor Triode—Bardeen and Brattain. (*See* 278.)

537.311.33:621.396.645:621.315.59 67
Experimental Germanium Crystal Amplifier [the Transistor]—Wells and White. (*See* 279.)

537.312.62 68
Theory of Superconductivity—M. Born and Kai Chia Cheng. (*Nature* (London), vol. 161, pp. 968–969; June 19, 1948.) A correct theory should be based on the ionic forces. Those metals are superconductive for which the Fermi surface, supposed to be a sphere, lies very close to 1 set of the corners formed by the boundary planes of a Brillouin zone.

537.312.62:621.317.332 69
High-Frequency Resistance of Superconductors—A. B. Pippard. (*Nature* (London), vol. 162, pp. 68–69; July 10, 1948.) Discussion of measurements for Sn at frequencies of about 9,200 Mc. See also 1014 of 1948.

537.523.3 70
The Influence of Corona Formation upon the Design of High Voltage Apparatus—W. T. Thornhill and E. W. Beasley. (*Marconi Rev.*, vol. 11, pp. 87–104; July and September, 1948.)

537.562 71
On the Absorption of the Energy of High-Frequency Currents by Ionized Gases—V. P. Mihu. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 472–474; August 23, 1948.) The intensity of the current in a coil forming part of an oscillatory circuit varies with the current intensity in a Geissler tube threading the coil. The absorption of energy by the gas in the tube has a maximum value for a particular gas pressure, which for hydrogen is 0.14 mm Hg and is constant for λ 2.40 to 5.94 m.

538.221 72
On the Interpretation of the Anomalies of Ferromagnetics at Radio Frequencies—I. Épelboim. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 185–187; July 19, 1948.) The anomalies are related to the existence of strong parasitic currents induced in the surface layers by the underlying layers of higher permeability. No supplementary hypotheses concerning magnetic viscosity appear necessary.

538.567.2:621.383.5 73
The Photo-Voltaic Effect—K. Lehovec. (*Phys. Rev.*, vol. 74, pp. 463–471; August 15, 1948.) The Schottky-Mott theory of the barrier-layer rectification is extended with respect to the action of light absorbed in that layer. An "equation of state" connecting photovoltage, photocurrent, light intensity, wavelength, external resistance, etc., is derived and discussed.

538.69:538.221 74
On the Motion of Iron Particles, with Six Degrees of Freedom, in the Air round a Wire carrying a Constant Electric Current—J. A. Schedling. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 470–472; August 23, 1948.) An account of effects analogous to those described by Ehrenhaft (2755 of 1947) for uniform magnetic fields.

538.3 75
Electromagnetism. [Book Review]—J. C. Slater and N. H. Frank. McGraw-Hill, London, 1947, 240 pp., 17s. 6d. (*Beama Jour.*, vol. 55, p. 266; August, 1948.) "The book is to be regarded as a modern fundamental text on the subject" and "occupies a very high place indeed in the ranks of first-class scientific literature."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.53:621.396.96 76
The Daylight Meteor Streams of 1947 May-August—J. A. Clegg, V. A. Hughes, and A. C. B. Lovell. (*Mon. Not. R. Astr. Soc.*, vol. 107, no. 4, pp. 369–378; 1947.) Whereas the results of 4.2-m radar observations on well-known meteor showers in the autumn and winter of 1946 and 1947 conformed closely to expectation from previous visual work, a similar investigation of a shower with radiant near η Aquarii, begun May 1, 1947, revealed an unexpectedly large belt of meteoric activity observable only in daylight. Some of this activity continued throughout August, 1947. This may be a recurrent phenomenon which explains the high general level of activity found by Prentice, Lovell, and Banwell (2779 of 1948).

523.72.029.62:621.396.822 77
Shape of Pulses of Radio-Frequency Radiation from the Sun—S. E. Williams. (*Nature* (London), vol. 162, p. 108; July 17, 1948.) Analysis of the shape of the tail of 78 single 75-Mc pulses shows that 58 were probably exponential in shape, 11 less probably so, 4 probably not so and 5 definitely not. Attempts to interpret the rising portion of the pulse curve in terms of exponential functions were unsuccessful.

523.746 78
Magnetic and Electric Phenomena in the Sun's Atmosphere Associated with Sunspots—R. G. Giovanelli. (*Mon. Not. R. Astr. Soc.* vol. 107, no. 4, pp. 338–355; 1947.) A method is given for the determination of the magnetic and electric fields near a sunspot which is growing in either a nonconducting or a uniformly conducting medium. The circulation of currents in the medium is critically dependent on the presence of an external magnetic field.

523.78"1945.07.09":551.510.535 79
Solar Eclipse Observations of the Ionosphere—C. W. McLeish. (*Canad. Jour. Res.*, vol. 26, sec. A, pp. 137–144; May, 1948.) Discussion of measurements made before, during, and after the eclipse of July 9, 1945, with automatic recording equipment. Analysis of the results indicates effective recombination coefficients of 1.6×10^{-8} for the E region, 1.4×10^{-8} for the F_1 region and 1.0×10^{-9} for the F_2 region. Evidence of a variable coefficient for the E region agrees with the view that, at this level, there is a high negative-ion density.

523.854:621.396.822 80
Galactic Noise—R. v. d. R. Woolley. (*Mon. Not. R. Astr. Soc.*, vol. 107, no. 3, pp. 308–315; 1947.) Electron velocity temperatures of about 12,000° may be expected near early-type stars, where ionized hydrogen is the main constituent. In other regions of interstellar space, where metallic ions greatly outnumber hydrogen ions, the temperature is of the order of 1,000°. Galactic noise of the intensity reported can only come from the hot regions.

550.384 81
Recent Work on the Earth's Main Magnetic Field—(*Observatory*, vol. 68, pp. 144–149; August, 1948.) Report of a discussion at the Royal Astronomical Society. See 3409 of 1948.

550.385:523.745 82
Magnetic Storms and Solar Activity, 1947—H. W. Newton. (*Observatory*, vol. 68, pp. 36–37; February, 1948.) Review of sunspot records, magnetic variations, magnetic storms, and sudden commencements for the year, with a table of 19 magnetic storms recorded at Abinger.

551.508.15:551.510.41 83
An Ozone Radiosonde—Stranz. (*See* 169.)

551.510.535:523.53 84
Observations of Scatter Clouds—T. L. Eckersley. (*Nature* (London), vol. 162, pp. 24–25; July 3, 1948.) A large increase in scatter clouds, not associated with visible meteors, may be observed before a Dellinger fade. This effect is attributed to the large amount of ultraviolet radiation from the sun at the time of the fade. The main effects observed by Millman, McKinley, and Burland (2511 of 1948) are explainable by the above, together with the author's theory of meteor evaporation and the physics of scatter clouds (91 of 1948).

551.510.535:525.624:550.384.4 85
Electric Conductivity of the Ionospheric

D-Region—D. F. Martyn, T. G. Cowling, and R. Borger. (*Nature* (London), vol. 162, pp. 142–143; July 24, 1948.) Discussion on 3116 of 1948 (Cowling and Borger) of 1024 of 1948 (Martyn). Martyn considers it doubtful whether tidal motions increase with height in the ionosphere. Experimental evidence strongly suggests that the main part of the magnetic variations is produced below the E region, but this must somehow be reconciled with the relatively small theoretical conductivity likely to be found there.

Cowling and Borger emphasize that work on atmospheric tidal motions is still only exploratory, and do not think that Hall currents can set up polarization capable of counteracting the magnetic reduction in E-layer conductivity.

551.510.535:551.543 86

A Note on a New Ionospheric-Meteorological Correlation—T. G. Mihran. (Proc. I.R.E., vol. 36, pp. 1093–1095; September, 1948.) A new correlation is suggested between the time of occurrence of the maximum F_2-layer critical frequency during a single day and the mean value of ground-level barometric pressure for that day.

551.510.535:621.396.11 87

Measurements of the Interaction of Radio Waves in the Ionosphere—Huxley, Foster, and Newton. (*See* 194.)

551.510.535:621.396.11.029.58 88

The Importance of Ionosphere Conditions for Long-Distance Radio Communication—H. J. Groenewold. (*Tijdschr. ned. Radiogenoot.*, vol. 13, pp. 103–129; July, 1948. Discussion 2 pp.) A review of present knowledge of the role of the ionosphere in short-wave propagation.

551.510.535:621.396.96:523.53 89

A Study of Transient Radar Echoes from the Ionosphere—E. Eastwood and K. A. Mercer. (*Proc. Phys. Soc.*, vol. 61, pp. 122–134; August 1, 1948.) Experiments performed between January, 1945 and July, 1946, in the frequency range 20 to 40 Mc support the meteoric theory of burst formation. The small amount of solar influence is explained in terms of E-region ionization; the sun does not appear to emit radiation capable of causing bursts. Analysis of the observations suggests that the greater proportion of the bursts are created within a thin layer located at a height of 86 km; the distribution of the bursts within this layer has proved to be uniform over wide areas and no latitude effect has been detected. It is established that the rate of incidence of bursts which present echoing areas between A and $A+dA$ m² to a radio wave of frequency ν may be expressed in the form $CdA/\nu^2A^{3/2}$, with C constant.

551.594.221 90

The Mechanism of Lightning Discharge—L. B. Loeb. (*Jour. Frank. Inst.*, vol. 246, pp. 123–148, August, 1948.) A survey of recent researches and modern theories.

LOCATION AND AIDS TO NAVIGATION

621.396.93 91

The Influence of the Human Element in Direction Finding—J. D. Peat. (*Marconi Rev.*, vol. 11, pp. 69–77; July to September, 1948.) Discussion of the accuracy of direction-finding bearing observations in the 1 to 20-Mc band and of a method of eliminating human errors by means of a recording device. Results obtained by visual, aural, and recorded observations on typical transmissions are compared statistically.

621.396.93:621.396.821:551.594.6 92

Extension, to Radiogoniometers for Atmospherics, of the Definition of the Operation Threshold in Terms of a Pulse Flux—F. Carbenay. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 337–339; August 2, 1948.) See also 2902 and 3504 of 1948.

621.396.932 93

A Survey of the Problems Involved in the Provision of Radio Aids to Marine Navigation—P. G. Redgment. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 1012–1015; 1947.) Discussion of the operating frequency, type of presentation and other special factors in the design of aids required in pilotage waters, in coastal navigation, in making a land-fall or in navigation in mid-ocean. A plea is made for the design of an associated group of systems employing, if possible, a common principle and requiring as few different sets of shipboard equipment as possible. (*Note.* For international recommendations on marine radar see 3135 of 1947 and 112 of 1948.)

621.396.932 94

New Developments in Marine Radio Direction Finders—H. Busignies. (*Elec. Commun.* (London), vol. 25, pp. 196–203; June, 1948.) A general review presented at the International Merchant Marine Radio Aids to Navigation Conference, New London, Conn., 1947. Suggestions for improvement include: (a) use of other wavelengths, (b) use of pulse transmissions, (c) direct-reading indication of bearings, (d) combination of distance measurement and bearing.

621.396.932 95

A Time-Multiplex Radio-Frequency Phase-Comparison Method for Navigational Systems—H. T. Mitchell and T. Kilvington. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 977–983; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 267; 1947.) Discussion of the Post Office Position Indicator (P.O.P.I.) hyperbolic navigation system, designed to assist ships in conditions of poor visibility. The continuous-wave signals whose phases are to be compared are radiated on a common frequency at intervals and in sequence from the various antennas comprising the system. Some practical results are given for an experimental arrangement using antennas $\lambda/2$ apart, but the system is inherently capable of being used with wide antenna spacings.

621.396.933 96

Note on a Short-Range Radio Position-Finding System using Modulated Continuous Waves—R. F. Cleaver. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 984–989, 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 267; 1947.) The range of an aircraft is found by comparing the phases of the modulation envelopes of signals (a) transmitted from and (b) received at a ground station after retransmission from the aircraft on a different carrier frequency. By combining apparatus for such phase comparison with the vhf automatic direction finder and conventional communication equipment, the range and azimuth of the aircraft are displayed on a cro without ambiguity for distances up to 100 miles. Range accuracy is within about 1 mile.

621.396.933:629.13.052 97

Low-Reading Absolute Altimeters—B. A. Sharpe. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 1001–1011, 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 268, 1947.) General considerations affecting design and performance of FM radio altimeters are discussed and existing instruments operating at 400 Mc as well as projected designs for 1,600 Mc are described. At present, with suitable well-placed antennas, satisfactory operation is possible at altitudes from zero to about 5,000 feet.

621.396.933:629.139.83 98

C.W. Radio Aids to Approach and Landing—M. Birchall. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 943–952; 1947. Summary, *ibid.* part IIIA, vol. 94, no. 11, p. 267; 1947.) Discussion of the fundamental problems involved. The display of the intelligence provided is of primary importance. "Crossed-pointer" meter presentation giving elevation and azimuth on 1 instrument is preferred. Existing continuous-wave radio systems are described, the general principle of operation being the use of partially overlapping radio beams to define a particular direction in space. The most serious technical objection to existing m-λ or dm-λ systems is the distortion of the radio beams by reflections. Future trends are likely to be toward the use of highly directive cm-λ systems, to overcome siting troubles.

621.396.933.2 99

C.W. Radio Aids to Homing and Blind Approach of Naval Aircraft—D. Quinn and R. D. Holland. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 953–960; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 267; 1947.) Description of beacons at present in use in aircraft carriers and at naval air stations. These include rotating antenna beacons and those having overlapping antenna patterns. The problems peculiar to naval applications are discussed.

621.396.933.2 100

The Omni-Directional Radial-Track Guide—J. H. Ashton and A. N. Beresford. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 990–1000; 1947.) Essentially a rotating radio beacon with a cardioid type of polar diagram. When the rotation takes place at high speed, the bearing of the receiving station is given by the phase of the AM envelope of the received waves. The 2-antenna rotating systems and the 4-antenna fixed systems with mechanical or electronic modulators are described. Accuracy is limited by re-radiation and by polarization errors.

621.396.933.2:621.317.79 101

The Development of Monitoring and Remote-Control Equipment for Radio Aids to Naval Flying—Quinn and LePage. (*See* 159.)

621.396.933.23 102

Indication of Landing Courses Independent of Weather Conditions—K. F. Niessen. (*Philips Res. Rep.*, vol. 3, pp. 1–12; February, 1948.) The landing course is determined as the intersection of a vertical plane F_1 with the surface of a cone whose axis lies in F_1. See also 1769 of 1946.

621.396.933.24 103

The Consol Navigation System—A. H. Brown. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 968–976; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 267; 1947.) The principles of operation and the general arrangement and design of the equipment are described. Theoretical and operational range and accuracy characteristics are given. See also 2252 of 1948 (Jessell) and 2912 of 1946 (Clegg).

621.396.96 104

Theoretical and Practical Radar Problems—E. Istvánffy. (*Elektrotechnika* (Budapest), vol. 40, pp. 167–181; August, 1948.) A review of modern developments.

621.396.96:526.9 105

The Use of Radar in Surveying—(*See* 188.)

621.396.933 106

Loran: Long Range Navigation [Book Review]—J. A. Pierce, A. A. McKenzie, and R. H. Woodward (Eds.). McGraw-Hill, New York, N. Y., 1948, 467 pp., $6.00 (Proc. I.R.E., vol. 36, p. 1141, September, 1948.) Volume 4 of the Radiation Laboratory series. A description of the loran system, the history of its development, and the associated equipment. For another review see *Nature* (London), vol. 162, pp. 633–635; October 23, 1948.

MATERIALS AND SUBSIDIARY TECHNIQUES

534.845 107
Behavior of Acoustic Materials—Cook. (*See* 7.)

535.37 108
Excitation Time of Silver-Activated Zinc Sulfide on Electron Bombardment—J. F. Mullaney, F. Raines, and H. G. Weiss. (*Phys. Rev.* vol. 74, pp. 491–492; August 15, 1948.) The rise time of visible radiation was found to be $<10^{-8}$ second, the limit of the measurement technique.

535.37 109
Electrical Properties of Incomplete Phosphors—R. Frerichs. (*Research* (London), vol. 1, pp. 208–211; February, 1948.) An account of measurements on large single crystals of CdS grown by the reaction of Cd vapor and H_2S at high temperature.

535.37 110
The New Phenomena of Electrophotoluminescence and its Possibilities for the Investigation of Crystal Lattice—G. Destriau. (*Phil. Mag.*, vol. 38, pp. 700–739, 774–793, and 880–888; October to December, 1947.) A 4-part theoretical and experimental study of the phenomena presented by photoluminescent compounds submitted to the action of intense electric fields. Part 1: Discussion of the excitation of luminescence in certain substances by varying electric fields of high intensity. The preparation of the substances and of the cells in which they were examined are also described. Part 2: Results of a study of the momentary illumination produced by the action of an electric field in phosphorescent sulphides which have been previously stimulated. Part 3: Discussion of the extinction of visible phosphorescence by electric fields. Part 4: Discussion of the electrical acceleration of infrared extinction.

549.514.51 111
The Development of Quartz Crystal Production—C. F. Booth and J. P. Johns. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 899–910; 1947. Bibliography, pp. 910–911. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 243; 1947.) The specifications and production techniques for bulk-produced BT-cut shear-mode crystals in the frequency range above 3 Mc are described. Details are given of machine processes, the solving of the activity-aging problem and the examination for raw-material faults. The reconditioning of crystal units and future developments are also considered.

549.514.51 112
The Control and Elimination of Electrical (Dauphiné) Twinning in Quartz—W. A. Wooster, N. Wooster, J. L. Rycroft, and L. A. Thomas. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 927–938; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 243; 1947.) Experimental methods use heat treatment while the quartz plate is subjected to stresses applied by either temperature gradients, pure bending, longitudinal compression, or torque. "Piezocrescence" and the effect of crystallographic orientation are discussed. Practical application of the methods to crystals of various cuts used in telecommunication apparatus removed twinning in 60 to 70 per cent of the plates submitted.

621.3.032.53:533.5 113
Stresses in Two-Wire Glass-to-Metal Seals—O. Adams. (*Jour. Soc. Glass Tech.*, vol. 32, pp. 99–112; April, 1948.) A brief discussion of the behavior of glass in plane stress when examined in plane polarized light is followed by a description of the estimation, by Filon's graphical integration method, of the principal stresses in flat 2-wire glass-to-metal seals.

621.315.332:535.827.2 114
The Microscopic Examination of Enamelled Wire—J. H. Wredden. (*Jour. R. Micr. Soc.*, vol. 66, pp. 9–22; 1946.) Shortage of chinawood oil supplies during the war stimulated the development of synthetic resins as a substitute covering for coil wire. The 4 kinds of substitute covering were, therefore, tested: (a) standard enamel, (b) a polyvinyl-acetal covering, (c) nylon, and (d) a double covering consisting of an inner layer of standard enamel and an outer layer of nylon. Enlarged photographs show the effect of phenol on each of the coverings: nylon is superior to the others, but even its resistance to phenol is low. Enlarged photographs also show these coverings (a) as normally received, (b) after a 10 per cent elongation by stretching, (c) at the point of fracture brought about by a steady pull along the sample, (d) after bending the wire round its own diameter without previous stretching, and (e) after heating in an oven to 150°C for 4 hours, followed by treatment (d). Results of these tests are discussed and experimental methods and apparatus are described briefly.

621.315.59 115
The Variation with Vapour Pressure of the Properties of Certain Electronic Semi-Conductors—C. A. Hogarth. (*Phil. Mag.*, vol. 39, pp. 260–267; April, 1948.) Formulas are given, and compared with experimental results, for the variation with vapor pressure of the Hall effect and thermoelectric power of semiconductors having an ionic lattice structure.

621.315.591.5† 116
Thermo-Electric and Conductive Properties of Blue Titanium Dioxide—H. K. Henisch. (*Elec. Commun.* (London) vol. 25, pp. 163–176; June, 1948. Bibliography, p. 177.) Measurements were made on several samples of sintered TiO_2 over a wide range of temperatures. The results agree with expressions developed from a simple theory of thermoelectric effects, based on the electronic energy relations at a metal versus semiconductor contact.

621.315.61 117
Some Modern Developments in Electrical Insulation—R. Snadow. (*Engineer* (London), vol. 186, pp. 170–171; August 13, 1948.) A short review dealing with thermosetting varnishes, flexible enamels, improved types of fabric board, and silicone varnishes.

621.315.61.011.5.029.63 118
Dielectric Constants of Some Solid Insulating Materials at Ultra Short Waves—S. K. Chatterjee and Rajeswari. (*Indian Jour. Phys.*, vol. 22, pp. 180–188; April, 1948.) Measurements by a standing-wave method for λ 57.7 to 140 cm show that the dielectric constant of paraffin wax increases slightly with decreasing wavelength. Mica, plexiglass, ebonite, and other materials tested show the reverse effect.

621.315.61.011.5.029.63 119
Dielectric Properties of Some Solid Insulating Materials at 750 Mc/s—S. K. Chatterjee. (*Indian Jour. Phys.*, vol. 22, pp. 157–166; April, 1948.) A resonant-line method was used to measure the dielectric constant and power factor of mica, mycalex, plexiglass, and ebonite. Plexiglass was found to have the lowest loss factor.

621.315.616 120
Review of the Principal New Synthetic Insulating Materials—R. Leprêtre. (*Rev. Gén. Élec.*, vol. 57, pp. 356–360; September, 1948.) General discussion, with a table giving the mechanical and electrical properties of hard plastics, rubbers, and silicones.

621.318.32:621.317.44 121
Study of Metals by Hertzian Waves with the Help of Permeameters with Demountable Coils—Épelboim. (*See* 147.)

621.318.323.2.042.15:621.775.7 122
Nickel-Iron Alloy Dust Cores—S. E. Buckley. (*Elec. Commun.* (London) vol. 25, pp. 126–131; June, 1948.) A paper read at the symposium on powder metallurgy noted in 150 and 451 of 1948. High and constant permeability and low core losses are the most important requirements for telecommunications equipment.

621.775.7:061.3 123
Powder Metallurgy—(*Metal Ind.* (London), vol. 73, pp. 103–105 and 129-130; August 6 and 13, 1948.) A report of the proceedings at the first international symposium on powder metallurgy at Graz, July, 1948, at which more than 70 papers, covering all aspects of the theory and technology of the subject, were read.

669.018.5 124
New Alloy has Improved Electrical Resistance Properties—(*Materials and Methods*, vol. 28, pp. 62–63; August, 1948.) "Evanohm" (75 per cent Ni, 20 per cent Cr, 2.5 per cent Al, 2.5 per cent Cu) has a much higher specific resistance and tensile strength and a slightly lower temperature coefficient of resistance and coefficient of thermal expansion than existing alloys such as manganin and constantan.

669.71:620.197.2 125
Aluminium Protection—(*Metal Ind.* (London, vol. 73, p. 265; October 1, 1948.) A method developed by the Société de Produits Chimiques des Terres Rares involves the use of a solution known as Framanol. This consists of (a) Cr compounds, (b) phosphoric acid, (c) alcohol, and (d) triethanolamine. (c) and (d) remove grease, (b) and (d) dissolve the surface layer of oxide and a light deposit of Cr oxide from (a) renders the cleaned surface inert.

535.37 126
Preparation and Characteristics of Solid Luminescent Materials [Book Review]—G. R. Fonda and F. Seitz (Eds.) J. Wiley and Sons, New York, N. Y., and London, 1948, 459 pp., $5.00 (*Amer. Jour. Sci.*, vol. 246, pp. 595–596; September, 1948.) ". . . for the most part a series of papers presented at a conference at Cornell University held by the Division of Electron physics, American Physical Society, in October of 1946. These have been edited and supplemented with introductory and background material."

MATHEMATICS

512.9 127
Gabriel Kron on Tensor Analysis—S. A. Stigant. (*Beama Jour.*, vol. 55, pp. 276–284; August, 1948.) An up-to-date record of Kron's published work on the theories and applications of tensor analysis in engineering, mechanics, hydrodynamics, and physics, revealing the exceptional versatility of the methods he has pioneered. A complete list is appended of Kron's books and papers, chronological in order with names and addresses of publishers.

518.5 128
Calculating Machines—(*Nature*, (London), vol. 161, pp. 712–713; May 8, 1948.) A short account of a discussion, sponsored by the Royal Society, on automatic digital computing machines. General descriptions are given of the machines either completed or under development in the United States (relay and electronic), the University of Cambridge and the National Physical Laboratory (electronic, with mercury delay lines for storage) the University of Manchester (crt storage devices) and Birkbeck College (electronic, with parallel operation).

518.5 129

Modern Mathematical Machines: Part 1—The Bush-Caldwell Differential Anlayser—M. Berry. (*Tech. Mod.* (Paris), vol. 40, pp. 246–251, August 1 to 15, 1948.) An account of the mechanization of the different operations and of the method of use, with a short description of actual equipment. Part 2, 130 below.

518.5 130

Modern Mathematical Machines: Part 2—M. Berry. (*Tech. Mod.* (Paris) vol. 40, pp. 283–287; September 1 to 15, 1948.) Discussion of the characteristics of "universal" machines, with a short description of the Eniac. An outline is given of the chief features of a machine for the Centre national ce la Recherche scientifique. This should be completed within a year and will be much more flexible in use than the Eniac. Part 1, 129 above

518.5:512.831 131

An Electrical Network for Determining the Eigenvalues and Eigenvectors of a Real Symmetric Matrix—A. Many and S. Meiboom. (*Rev. Sci. Instr.*, vol. 18 pp. 831–836; November, 1947.) A network for the solution of a matrix of order *n* consists of *n* *LC*-circuits, each coupled to each of the others through 2 capacitors, which are adjusted according to the elements of the matrix. The solution is derived from observation of resonant frequencies and voltages when the network is excited from a variable-frequency source The theory includes a discussion of the accuracy attainable. Solution for a fifth-order matrix takes about 4 hours. An instrument for solving matrices up to the tens order is planned. See also 3562 of 1947 (Hughes and Wilson).

MEASUREMENTS AND TEST GEAR

53.087 132

A Distribution Analyzer for Lengths of Impulses—O. R. Garfield. (*Bell Lab. Rec*, vol. 26, pp. 325–330; August, 1948.) Records the number of pulses longer than each of 10 given times, on 10 groups of message registers and neon lamps marked from 0 to 9 The zero register records the pulses longer than 50 μs, while register *n* records those longer than *nt*, where *t* is adjustable in multiples of 2 from 1.56 ms to 12.8 seconds.

531.761:621.317.755 133

A Spiral Chronograph for Long Time Inter-Intervals—H D. Warshaw. (*Rev Sci. Instr.*, vol. 19, pp. 420 423, July, 1948) A chronograph differing basically from those of Emrich (3569 of 1947) and Moran (764 of 1948). It measures times up to 12 ms and develops a logarithmic spiral by means of a circuit which avoids a modulating network. Time markers on the spiral every 10 μs or 20 μs are derived from a crystal-controlled oscillator The sweep speed of the trace is 1 revolution in 2 ms

621.3.018.4 (083.74) 134

[British] Standard Frequency Transmissions—(*Wireless World*, vol 54, p 322, September, 1948) The daily schedule of experimental low-power transmissions from station GMT at Abinger, Surrey, is

958 to 1,000 G.M T, announcement of call sign. 1,000 to 1,015, radiation of unmodulated 2 Mc carrier. 1,015 to 1,025, radiation of carrier with 1,000-cps modulation. 1,025, announcement of provisional frequency correction.

Corrections are accurate to within about 2 parts in 10^8.

Arrangements are under consideration for a limited standard-frequency service, on 3 frequencies, to be operated by the General Post Office. Details will be given later. Meanwhile, an accuracy within 1 part in 10^6 is given by all BBC medium-wave transmissions, except that on 583 kc, and by those from BGR Rugby on 16 kc, Droitwich on 200 kc, GRO Skelton on 6,180 kc and GSV Daventry on 17,810 kc.

621.3.018.4(083.74):621.396.81 135

Reception at Turin of the Standard Signals of the National Bureau of Standards, Washington—C. Egidi and G. Gregoretti. (*Alta Frequenza*, vol. 17, pp. 161–167; August, 1948. In Italian, with English, French, and German summaries.) Systematic reception tests for the reception of WWV 10-Mc and 15-Mc signals, from July, 1946, to June, 1947, show that the 10-Mc signals can nearly always be received satisfactorily at Turin.

621.316.726.078.3 136

Frequency Stabilization in the Region of 10,000 Mc/s—A. V. Donnelly. (*Communications*, vol. 28, pp. 6–8, 31; July, 1948.) Frequencies in the microwave region can be stabilized (a) by reference to a frequency obtained from a quartz oscillator by successive multiplication, or (b) by a feedback system of automatic frequency control incorporating a cavity as the reference standards. Method (a) is more accurate and method. (b) more convenient.

621.317.3:621.396.611.21 137

Precision Measurement of Electrical Characteristics of Quartz-Crystal Units—W. D. George, M. C. Selby and R. Scolnik. (PROC. I.R.E., vol. 36, pp. 1122–1131; September, 1948) Discussion of: (a) the equivalent circuit of such units. (b) measurement methods and techniques in which a generator of continuously adjustable frequency and stability comparable with that of a crystal unit is combined with an rf bridge or a *Q*-meter, and (c) measurement of secondary responses. Graphs show the difference between a "normal" crystal and 1 remaining constant under limited conditions. Data based on fundamental crystal-unit characteristics are correlated with data derived from the behavior of the unit in a particular oscillator.

621.317.3:621.397.61 138

TV Transmitter Design: Part 4—G. E. Hamilton. (*Communications*, vol. 28, pp. 8–9, 33; August, 1948.) Discussion of techniques for measuring (a) transmitting power output by a calorimeter method, (b) transmitter regulation and output variation. (c) frequency response and (d) transient response. For previous parts see 2948 and 3532 of 1948.

621.317.332 139

On the Limits of Absolute Accuracy for Resistance Measurements at High Frequencies—H. H. Meinke. (*Arch. Elek* (Übertragung), vol. 1, pp. 101–107; September and October, 1947.) Discussion of methods involving measurements of nodal distances and voltage maxima and minima along a transmission line of known characteristic impedance. Methods of reducing possible errors are described.

621.317.333 140

A Statistical Method for determining the Breakdown Voltage of a Dielectric—A. S Zingerman. (*Zh. Tekh. Fiz.*, vol. 18, pp 1029–1043; August, 1948. In Russian.)

621.317.333.4:621.396.615.17 141

Measurement of Phenomena of Extremely Short Duration—W. Kroebel. (*Arch Elek.* (Übertragung), vol. 1, pp. 108–113; September and October, 1947.) A pulse generator is described which uses gas-discharge tubes, gives considerable power and can be applied for short-time measurements of high accuracy. Its use is considered for the location of faults in a cable for which the propagation velocity of electric waves is known.

621.317.336:621.315.212 142

Study of the Impedance Irregularities of Coaxial Cables by Oscillographic Observation of Pulse Echoes—G. Couanault and P. Herreng. (*Câbles et Trans.* (Paris) vol. 2, pp. 111–130 and 219 232; April and July, 1948. With English summary.) The pulses used have sensibly rectangular shape, duration 0.1 or 0.2 μs and recurrence frequency 50,000 per second. The echoes arising from cable irregularities, after amplification, are displayed on a cro. Lengths of cable up to 560 m can be tested by equipment developed by the Société Alsacienne de Constructions Mécaniques; a detailed description of this echo meter is given.

621.317.336:621.396.67 143

Television Receiving Aerials: Impedance Measuring and Matching Methods—Bell. (*See* 18).

621.317.4:538.221 144

The Measurement of Permeability and Magnetic Losses of Non-Conducting Ferromagnetic Material at High Frequencies—H. J. Lindenhovius and J. C. van der Breggen. (*Philips Res. Rep*, vol. 3, pp. 37–45; February, 1948.) A ring made of the material to be tested and a ring of the same size made of a conducting metal are successively inserted, concentrically with the inner conductor, into a coaxial cavity resonator with end-capacitance. The susceptibility of the ferromagnetic material equals the ratio of the 2 changes in resonant frequency, and the losses can be computed from the bandwidths of the cavity resonator with and without the ferromagnetic ring. This method has been used between 30 and 300 Mc. Above 300 Mc a coaxial Lecher system is substituted for the cavity resonator.

621.317.41 145

High-Frequency Permeameter—M. Cogniat. (*Câbles et Trans.* (Paris), vol. 2, pp 195–207; July, 1948. With English summary) An instrument for rapid measurements on thin tape or stampings at frequencies from 50 kc to 1.6 Mc. The value of the permeability is found from the detuning of an *LC* circuit caused by insertion of the sample into the inductor. Full construction details are given, with all coil data.

621.317.44:538.23 146

Cathode-Ray Magnetization-Curve Tracer—M. V. Scherb. (*Rev. Sci. Instr.*, vol. 19, pp. 411 419, July, 1948) An instrument which displays, upon a direct-reading crt screen, the hysteresis loop of the test specimen, whose cross section may range from 10^{-4} cm² to 0.3 cm² (one-quarter inch rod). Full circuit details and operation procedure are given. With slight modifications, the apparatus can be used for other types of magnetic measurement.

621.317.44:621.318.32 147

Study of Metals by Hertzian Waves with the Help of Permeameters with Demountable Coils—I. Epelboim (*Onde Élec.*, vol. 28, pp 322 327, August to September, 1948) Description of permeameters similar in principle to the apparatus considered in 797 of 1947, and of industrial instruments for the routine testing of annular iron-dust cores at frequencies up to 60 Mc Results of measurements on magnetic powders dispersed in rubber show that for samples in which the volume of the powder does not exceed that of the dielectric, the magnetization is proportional to the logarithm of the permeability With transformer stampings, the relaxation effect is found at km λ while in the case of iron powders dispersed in rubber it is only observed at cm λ.

621.317.6.029.3:621.317.755 148

An Automatic Audio Frequency Response Curve Tracer: Parts 1 and 2—G. L. Hamburger. (*Jour. Brit. I.R.E.*, vol. 8, pp. 154–168, July and August, 1948.) Description, with full details of the various units, of equipment for displaying on the screen of a crt the af response curve of any type of network. Logarithmic scales are used for both frequency (39 cps to 10 kc) and amplitude (−21 db to +18 db). After completing the response curve for a tone of constant voltage swept from 11 kc down to 39 cps in 1 or 2 seconds, the tone is switched off automatically and a calibration raster, in octaves and 3-db steps, appears. With a long-afterglow screen, both raster and response curve can be seen together. Part 1 was reprinted in *Proc. I.R.E.* (Australia), vol. 8, pp. 16–19; October, 1947.

621.317.725 149
A Precision High-Voltage Vacuum-Tube Voltmeter—L. C. L. Yuan. (*Rev. Sci. Instr.*, vol. 19, pp. 450–452; July, 1948.) Range, 0 to 5000 v. Input impedance $<5\times10^4$ MΩ. Linear characteristic. Accuracy within 0.05 per cent.

621.317.725:621.394/.396].44 150
Tuned Valve-Voltmeter in Carrier-Frequency Technique—W. Klein. (*Funk und Ton*, vol. 2, pp. 466–468; September, 1948.) Ordinary tube voltmeters are independent of frequency over a fairly wide range, but for carrier-current systems a tuned voltmeter is a decided advantage, as it can be used for measurements in 1 channel without interfering with communication in the other channels of a multiplex system. Suitable designs are given for direct and for heterodyne types of meter, with block diagrams.

621.317.726 151
An Improved Peak Voltmeter for Pulses—Y. P. Yu. (*Rev. Sci. Instr.*, vol. 19. pp. 447–450; July, 1948.) Comprises a gate generator, an electronic switch, a capacitor, a bridge-balanced cathode follower of very high input resistance, and a dc voltmeter. Accuracy is within 4.5 per cent for pulse widths from 0.5 μs to 50 ms and repetition frequencies from 10 to 200,000 pulses per second.

621.317.733:621.317.715 152
Location of the Galvanometer Branch for Maximum Sensitivity of the Wheatstone Bridge—F. R. Kotter. (*Jour. Res. Nat. Bur. Stand.*, vol. 40, pp. 401–404; May, 1948.) The battery and galvanometer connections to the bridge may be interchanged without altering the condition for balance, but 1 of the combinations will give higher sensitivity. A "rule-of-thumb" method is given for determining the more sensitive arrangement. Only the bridge arm resistances and the external critical damping resistance of the galvanometer need be known.

621.317.734 153
Electronic Megohmmeters—H. G. M. Spratt. (*Wireless World*, vol. 54, pp. 354–357; October, 1948.) A method of measuring resistances up to about $10^{12}\Omega$ depending upon the exponential grid-voltage versus grid-current characteristic of the normal triode.

621.317.755 154
The Principles and Practice of Panoramic Display—D. W. Thomasson. (*Jour. Brit. I.R.E.*, vol. 8, pp. 171–186; July and August, 1948. Discussion, pp. 186–189.) The technique of the production of cro displays with 1 scale a frequency scale, either linear or logarithmic, is reviewed theoretically and practically. The limitations of the apparatus used are discussed and various practical design problems are considered.

621.317.755:621.397.62.001.4 155
An Electronic Wobbulator for Television—(*Telev. Franc.*, pp. 11–15; July and August, 1948.) Details of an instrument permitting direct observation on a cro of (a) the high-frequency or medium-frequency selectivity curve for a receiver, after detection or after the video amplifier, (b) the pass-band curve at the output of the video amplifier, (c) image and synchronization signals. An instrument of this type has been in constant use for more than a year for receiver production testing.

621.317.761 156
A Direct Reading Frequency Measuring Set—F. C. F. Phillips. (*Jour. Brit. I.R.E.*, vol. 8, pp. 4–15; January and February, 1948, Discussion, p. 16.) The equipment consists of a series of decade stages terminating in a deflection type of frequency meter. The signal of unknown frequency is heterodyned successively with each of the decade stages, whose settings give the various digits of the frequency, the final figures (below, 1,000 cps) being read on the scale of the frequency meter. Circuit details are discussed and also the methods adopted to avoid any ambiguity so that the equipment can be used by relatively unskilled operators. The frequency range up to 10 Mc can be extended by the use of harmonics of an auxiliary oscillator.

621.317.79:621.396.615 157
Variable Frequency Standard Signal Generator—R. G. Stokes. (*Tele-Tech*, vol. 7, pp. 44–47; September, 1948.) An equipment designed to provide any frequency between 15 kc and 10 Mc for measuring the characteristics of piezoelectric crystals. The signal is within 1 cps of the primary standard frequency throughout the frequency range. The frequency can be varied by increments as small as 0.2 cps. The circuit used is described, the basis being a primary standard input of 100 kc, an interpolation oscillator with a range of 20 to 40 kc and 4 variable filters, any specific output frequency being obtained by a suitable combination of frequencies from the separate units.

621.317.79:621.396.615.14 158
Measurement Generators for the Decimetre Waveband—G. Megla. (*Elecktrotechnik* (Berlin), vol. 1, pp. 19–24; July, 1947.) Detailed description of various frequency-stable generators for λ 10 to 100 cm, with special reference to tuning and screening arrangements and to the design of capacitive and resistive types of potential dividers.

621.317.79:621.396.933.2 159
The Development of Monitoring and Remote-Control Equipment for Radio Aids to Naval Flying—D. Quinn and L. S. LePage. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 961–967; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 268; 1947.) Systems which provide a continuous remote indication of the performance of radio aids for homing and blind approach are discussed.

621.319.4 (083.74) 160
Absolute Capacitance Standard with a Resistive Shield—T. Slonczewski. (*Rev. Sci. Instr.*, vol. 18, pp. 848–849; November, 1947.) The addition of a high-resistance shield to the guard ring of a parallel-plate capacitor reduces the fringe effect to a negligible value so that the precise computation of capacitance is simplified.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.784 161
Piezoelectric Dynamometer for Recording Work Variations in Cutting [by machine tools].—P. Naslin. (*Rev. Gén. Élec.*, vol. 57, pp. 361–364; September, 1948.) Details of an instrument in which the sensitive element consists of 4 quartz plates with polarities reversed alternately. The charges developed on the faces are proportional to the applied pressure, which can either be measured on the scale of a calibrated microammeter or displayed on the screen of a cro.

539.16.08 162
The Discharge Spread in Geiger-Müller Counters—S. H. Liebson. (*Phys. Rev.*, vol. 74, pp. 694–695; September 15, 1948.) Reasons are given for questioning the validity of the conclusions drawn by Corson and Wilson (168 below) from comparison of results obtained by Alder and others (2748 of 1947) and by Liebson (3982 of 1947) for self-quenching counters.

539.16.08 163
Discharge Spread in Geiger Counters with Methane and Methane/Argon Filling—C. Balakrishnan, J. D. Craggs, and A. A. Jaffe. (*Phys. Rev.*, vol. 74, pp. 410–414; August 15, 1948.) See also 1103 of 1948 (Craggs and Jaffe).

539.16.08 164
A Note on Ethylene Self-Quenching G-M Counters—K. H. Morganstern, C. L. Cowan, and A. L. Hughes. (*Phys. Rev.*, vol. 74, pp. 499–500; August 15, 1948.) Tests with alcohol, ether, ethylene, or cyclopropane as the quenching gas show that ethylene gives long flat plateaux over a wider range of partial pressures than the other gases. Ethylene-filled counters give no multiple counts, are very stable with regard to counting rate and recover rapidly even after a discharge through them.

539.16.08 165
Dead-Time Reduction in Self-Quenching Counters—P. B. Smith. (*Rev. Sci. Instr.*, vol. 19, pp. 453–457; July, 1948.)

539.16.08 166
Coincidence-Counting System of High Resolution—H. L. Schultz and R. Beringer. (*Rev. Sci. Instr.*, vol. 19, pp. 424–427; July, 1948.)

539.16.08 167
High-Speed Coincidence Circuit used for Multipliers—E. Baldinger, P. Huber, and K. P. Meyer. (*Rev. Sci. Instr.*, vol. 19, pp. 473–474; July, 1948.)

539.16.08 168
Particle and Quantum Counters—D. R. Corson and R. R. Wilson. (*Rev. Sci. Instr.*, vol. 19, pp. 207–233; April, 1948.) A review of the operating mechanism and special properties of the commonly used counters.

551.508.15:551.510.41 169
An Ozone Radiosonde—D. Stranz. (*Chalmers Tekn. Högsk. Handl.*, no. 72, 49 pp.; 1948. In German, with English Summary.) Full account of experimental apparatus. Ozone content is deduced from the amount of ultraviolet radiation falling on a Cd photo cell included in the equipment.

578.087.87:621.317.35 170
The Wave Analysis of Low Frequency Potentials of the Human Body—W. E. Boyd and W. E. Benham. (*Jour. Brit. I.R.E.*, vol. 8, pp. 73–84; March and April, 1948. Discussion, pp. 84–85.) Electronic methods reveal the existence of low-frequency potential variations, originating from the human body, in the frequency range 10 to 2,500 cps. Analysis of these potentials appears to indicate that parts of this frequency range are related to the muscular contraction of the heart and that information may thus be obtained which is not given by present electrocardiograph technique.

Benham discusses Fourier analysis in connection with Boyd's work.

621.3.076.12 171
New Applications of Automatic Compensation—T. Gast. (*Arch. Elek.* (Übertragung), vol. 1, pp. 114–121; September and October, 1947.) Examples of different types of compensating apparatus are given and their principles explained. A pentode oscillator with inductive coupling between anode and grid circuits is described; this can be used for transforming mechanical into electrical quantities and is applied in the construction of a microbalance and also of an electrostatic compensator.

621.316.578.1 172
Electronic Process Timer—(*Engineer* (London), vol. 185, p. 387; April 16, 1948.) A capacitor is discharged through a resistor; the known discharge time can be varied between one quarter and 60 seconds by a single potentiometer control. Accuracy is within 5 per

cent, and is independent of supply frequency variation or voltage changes of ±15 per cent. The negatively charged capacitor plate is connected to the grid of a thyratron, the potential of which takes a positive drift until eventually the thyratron fires. Relays are thus actuated which end the timing cycle and reset the mechanism. For another account see *Engineering* (London), vol. 166, p. 139; August 6, 1948.

621.317.39:620.17 **173**

Extensometers using Electrical Resistance—R. Goethals. (*Rev. Gén. Élec.*, vol. 57, pp. 375–379; September, 1948.) The operation principles of various types of instrument are reviewed briefly and details are given of resistance extensometers suitable for static or dynamic measurements.

621.317.755 **174**

The Principles and Practice of Panoramic Display—Thomasson. (*See* 154.)

621.384.6 **175**

Experimental Work on Corrugated Waveguides and Associated Components for Linear Electron Accelerators—L. B. Mullett and B. G. Loach. (*Proc. Phys. Soc.*, vol. 61, pp. 271–284; September 1, 1948.) The general problem of feeding rf power from a conventional rectangular waveguide is considered and several successful "feeds" are described. A detailed account is given of work carried out on corrugated circular waveguides with continuously varying phase velocities, and on the associated "feeds" for the 0.5-Mev and 4-Mev sections of long accelerators. Experimental results are in close agreement with the theoretical treatment of Harvie (176 below) and Walkinshaw (177 below).

621.384.6 **176**

Travelling Wave Linear Accelerators—R. B. R.-S.-Harvie. (*Proc. Phys. Soc.*, vol. 61, pp. 255–270; September 1, 1948.) The relevant properties of accelerators of the corrugated waveguide type [see 506 of 1948 (Fry *et al.*)] and the factors influencing electron energies, are considered. Some numerical results are given for λ 10 cm. The problems of high-energy accelerators are discussed briefly and illustrated by calculations for a 500-Mev accelerator.

621.384.6 **177**

Theoretical Design of Linear Accelerator for Electrons—W. Walkinshaw. (*Proc. Phys. Soc*, vol. 61, pp. 246–254, September 1, 1948.) Theoretical analysis of the type of accelerator described by Fry *et al.* (506 of 1948). The dependence of the phase velocity on the frequency and on the dimensions of the corrugated waveguide is determined and found to be in close agreement with experiment. Tables have been computed for designing accelerator tubes for any frequency and rf power.

621.384.6 **178**

Effect of Azimuthal Inhomogeneities in the Magnetic Field of a Betatron or Synchrotron—F. K. Goward. (*Proc. Phys. Soc.*, vol 61, pp. 284–293; September 1, 1948.) Particular examples of the types of forced electron oscillations which may be encountered are considered quantitatively, with special emphasis on the problems of injecting an electron into an inhomogeneous field, and of extracting an electron by an applied inhomogeneity.

621.384.6 **179**

A 20-MeV Betatron—W. Bosley, J. D. Craggs, D. H. McEwan, and J. F. Smee. (*Jour. IEE* (London), part 1, vol. 95, pp. 352–358; August, 1948.) The construction is described. Tests made to measure the magnetic field and to locate the orbit position, etc., are fully discussed, and details are given of some new measurements on the out-of-phase fields at the instant of electron injection.

621.384.6:537.531 **180**

X-Radiation from a 20-MeV Betatron—W. Bosley, J. D. Craggs, W. F. Nash, and R. M. Payne. (*Nature* (London), vol. 161, pp. 1022–1023; June. 26, 1948.) Preliminary results on the spectrum of the betatron's radiation obtained by study of (a) electron-positron pairs produced in a thin lead sheet in a Wilson chamber, and (b) the protons liberated in the photodisintegration of deuterons.

621.385.38.001.8 **181**

Thyratrons and their Industrial Applications—Martin. (*See* 267.)

621.385.833 **182**

Recent Advances in Electron Microscopy in the United Kingdom—V. E. Cosslett. (*Research* (London), vol. 1, pp. 293–304; April, 1948.) A survey based on a paper read at the Philadelphia Conference, December 12, 1947.

621.385.833 **183**

On some Focal Properties of an Electron Optical Immersion Objective—P. A. Einstein and L. Jacob. (*Phil. Mag.*, vol. 39, pp. 20–31; January, 1948.)

621.385.833 **184**

Adjustable Aperture for Electron Microscope, RCA Type EMU—J. T. Quynn. (*Rev. Sci. Instr.*, vol. 19, pp. 472–473; July, 1948.)

621.396.9 **185**

Search Apparatus for Metal—O. Martin. (*Elektrotechnik* (Berlin), vol. 1, pp. 15–18; July, 1947.) Description of various heterodyne methods, resonance methods with indication based on phase displacement, and mutual-induction methods.

621.396.9:531.561 **186**

A Radio Method of Studying the Yaw of Shells—C. C. Gotlieb, P. E. Pashler, and M. Rubinoff (*Canad. Jour. Res.*, vol. 26, sec. A, pp. 167–198; May, 1948.) The shell fuse is replaced by a high-frequency transmitter whose signals are used to determine the angular motion of the shell in flight. Theory of the method is given and experimental results are discussed.

621.396.9:550.837.7 **187**

Prerequisites for the Use of Radio Prospecting Methods—V. Fritsch. (*Radio Tech.* (Vienna), vol. 24, pp. 429–435; September, 1948.) A historical review, with discussion of some modern methods of investigation, including high-frequency pulse technique. See also 216 and 1524 of 1947.

621.396.96:526.9 **188**

The Use of Radar in Surveying—(*Observatory*, vol. 68, pp. 100–104; June, 1948.) Short account of a discussion at the Royal Astronomical Society.

PROPAGATION OF WAVES

538.566+621.396.11 **189**

Radio-Wave Propagation Research in the Department of Scientific and Industrial Research during the Years 1937–46—R. L. Smith-Rose. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 879–891; 1947. Bibliography, pp. 891–892. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 220; 1947.) Frequencies from about 10 kc to 50,000 Mc or more are considered. Discussion of (a) transmission of radio waves along the earth's surface, and the effect of the electrical properties of the ground, earth curvature, and obstacles, (b) the nature of the ionosphere and the manner in which radiowaves are propagated through it, (c) the effect at uhf of vertical temperature and humidity gradients in the lower atmosphere (d) the electrical properties of the constituents of the atmosphere at uhf (e) atmospheric and extra-terrestrial noise, (f) the importance of the study of wave propagation for various applications of radio technique, and (g) the limitations of present knowledge.

621.396.11:551.510.535 **190**

The Accuracy of Sky-Wave Delay Measurements—K. W. G. Harrod. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 893–898; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 220; 1947.) Measurements are described of the short-period fluctuations in the time of arrival of waves of frequencies between 2 and 16 Mc, received after reflection from the ionosphere. For vertical incidence, the fluctuations were of the order of ±5 μs. At oblique incidence, for ranges up to 6,000 km, path-time differences of ±10 to ±20 μs were observed with 2 receiving antennas 400 meters apart.

621.396.11:551.510.535 **191**

The Forecasting of Maximum Usable Frequencies for Radio Links—P. Lejay. (*Onde Élec.*, vol. 28, pp. 268–274 and 328–336; July, August, and September, 1948.) American and British methods are described and the use of the world charts prepared by the Central Radio Propagation Laboratories and by the National Physical Laboratory is illustrated by the determination of the maximum usable frequency for each hour of the day for communication between Paris and Dakar. The need for more stations to carry out ionosphere research is emphasized. At present, there is only 1 in the whole continent of Africa. See also 2887 of 1948.

621.396 11:551.510.535 **192**

Anomalous Effects in Ionospheric Absorption—E. V. Appleton, W. J. G. Beynon, and W. R. Piggott. (*Nature* (London), vol 161, pp. 967–968; June 19, 1948.) The relations between any 2 of the 3 variables h' (equivalent height of reflection) i_0 (angle of incidence), and f (frequency) may be satisfactorily predicted when the third variable has a constant value, but there has been no corresponding success in deducing the absorptive behavior of obliquely incident waves from that of vertically incident waves. Martyn's conception of equivalent frequency (1727 of 1935) gives values of attenuation which are too low for first-order reflections although satisfactory for those of higher order. It is suggested that when frequencies are used which only slightly exceed the oblique-incidence critical E-layer frequency, there is scattering or partial reflection of the waves by abnormal or sporadic-E ionization which produces the additional attenuation. Measurements at Hawick on transmissions from Fraserburgh to Swansea tend to support this theory: sporadic-E echoes at Hawick and abnormal attenuation at Sawnsea were observed together.

621.396.11:551.510.535 **193**

Diffraction from the Ionosphere and the Fading of Radio Waves—J. A. Ratcliffe. (*Nature* (London), vol. 162, pp. 9–11; July 3, 1948.) A theory of fading is outlined which regards a "single" reflected wave as the sum of contributions from a large number of scattering centers in the reflecting region, moving with velocities distributed according to a Gaussian law. The resultant signal is found to be analogous to that produced when random noise is passed through a filter with a specified band-pass characteristic, and the results of an analysis of this case are applied to the present problem. It is shown that the observed fading characteristics of radio waves on various frequencies from 4 Mc to 16 kc are in accord with the theory, the rms value of the scattering-center velocity being of the same order of magnitude in each case. The theory accounts for the observed fact that the rate of fading is roughly proportional to the frequency of the wave and the distance of the transmitter. An extension of the theory to the case of scattering of a wave in transit through an ionospheric layer is discussed briefly. Simultaneous observations on the fading of single echoes from regions E and F should be useful in this respect. Various possible elaborations of the simple theory given are mentioned.

621.396.11:551.510.535 194
Measurements of the Interaction of Radio Waves in the Ionosphere—L. G. H. Huxley, H. G. Foster, and C. C. Newton. (*Proc. Phys. Soc.*, vol. 61, pp. 134–146; August 1, 1948.) Discussion of measurements of radio wave interaction in special test transmissions between the BBC stations at Droitwich and Lisnagarvey (North Ireland). The theoretical formula $M=M_0/[1+(n/G\bar{\nu})^2]^{1/2}$ accurately describes the dependence of the interaction modulation M on the modulation frequency $n/2\pi$. The electronic collisional frequency ν at the seat of interaction is deduced from $G\nu$ and the laboratory value of G. The phase of the modulation is also discussed. The theory of wave interaction and its possible use as a tool in ionospheric research are considered briefly. See also 2897 of 1947.

621.396.11.029.45 195
Very-Long-Wave Phase Differences Between Spaced Aerial Systems—P. G. Redgment and D. W. Watson. (*Jour. IEE* (London) part IIIA, vol. 94, no. 16, pp. 1016–1022; 1947.) Transmissions from Varberg (Sweden) and Tuckerton (United States) on frequencies in the neighborhood of 18 kc were received in London and Birmingham (163 km apart) and the phases compared. For the shorter distance, a distinct diurnal variation in the phase difference was observed, together with slow random fluctuations which were greater at night. At the longer distance, the diurnal variation was present during the summer but was not detectable in winter. Random fluctuations of a similar order of magnitude to those at the shorter range were observed. It is concluded that a navigational aid using similar frequencies could be constructed having an equivalent-bearing accuracy within 0.2° near the median line.

621.396.11.029.58:551.510.535 196
The Importance of Ionosphere Conditions for Long-Distance Radio Communication—Groenewold. (*See* 89.)

RECEPTION

621.396.621+621.396.69 197
New Methods of Manufacturing Radio Circuits—J. A. Sargrove. (*Onde Élec.*, vol. 28, pp. 299–307; August and September 1948.) Printed-circuit technique described in 1913 of 1947.

621.396.621 198
The Cathode-Coupled Clipper Circuit—L. A. Goldmuntz and H. L. Krauss. (PROC. I.R.E., vol. 36, pp. 1172–1177; September, 1948.) Advantages of such a circuit over conventional pentode and diode types are discussed. The proper operating conditions to give the required size of output pulse, range of input voltages, bias tubes for symmetrical clipping, etc., are determined analytically. Regenerative feedback considerably improves the clipping action. Design methods are outlined.

621.396.621:621.396.619.13 199
Is Discriminator Alignment So Difficult?—A. G. Crocker. (*Wireless World*, vol. 54, pp. 312–316; September, 1948.) Discussion of methods of alignment and the effects of misalignment. In particular, the procedure suggested by Sturley (2183 of 1944) is criticized. Provided that the discriminator is suitably designed, serious alignment difficulties need not be expected.

621.396.621:621.396.662.3 200
Crystal Filters for Radio Receivers—C. F. Floyd and R. L. Corke. (*Jour. IEE* (London), part IIIA, vol. 94, no. 16, pp. 915–926; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 243; 1947.) Discussion of: (a) a pair of single-section bridged-T filters for Army communications receivers Types R206 and R201, (b) channel and carrier filters recently designed for the British Post Office 2-channel single-sideband receiver, (c) Foster's theorem for 2-terminal reactance networks and its use to determine the properties and elements of lattice networks, (d) resistance compensation, (e) the application of the equivalent circuit of a crystal resonator to filter networks and methods of calculating L and C values, (f) the properties of certain quartz cuts, and (g) the design of a new type of 2-section half-lattice channel filter.

621.396.621:621.396.933 201
Very-High-Frequency Single-Channel Receiver—W. C. Lane and T. C. Clark. (*Elec. Commun.* (London), vol. 25, pp. 132–138; June, 1948.) A crystal-controlled ground-station superheterodyne receiver with high rejection of unwanted signals and good sensitivity, for use at major airports. The frequency range is 118 to 136 Mc.

621.396.821:621.396.93:551.594.6 202
Extension, to Radiogoniometers for Atmospherics, of the Definition of the Operation Threshold in terms of a Pulse Flux—F. Carbenay. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 337–339; August 2, 1948.) See also 2902 and 3504 of 1948.

621.396.822:621.396.619.13 203
Theory of Frequency-Modulation Noise—F. L. H. M. Stumpers. (PROC. I.R.E., vol. 36, pp. 1081–1092; September, 1948.) The energy spectrum of FM noise is computed for different signal-to-noise ratios. Numerical values are given for some simple filter amplitude characteristics. The theory is based on the Fourier concept of noise; conditions of no signal, and of signal without and with modulation are successively considered. The result is given in the form of a convergent series. The suppression of the modulation by noise is also discussed.

621.396.822.1:621.396.41:621.396.619.16 204
Study of Crosstalk in Multiplex Systems using Pulse-Position Modulation—Gardère and Oswald. (*See* 224.)

621.397.828 205
TV Circuits Cause Interference—(*Tele-Tech*, vol. 7, p. 39; September, 1948.) Summary of RMA engineering report issued by the Committee on Television Receivers. Simple precautions at the design stage of television receivers are suggested to prevent interference with radio reception.

621.396.621.029.64 206
Microwave Receivers [Book Review]—S. N. Van Voorhis (Ed.). McGraw-Hill, New York, N. Y., 1948, 611 pp., $8. (PROC. I.R.E., vol. 36, p. 1141; September, 1948.) Volume 23 of the Radiation Laboratory series. "The present-day microwave engineer . . . will find a large amount of material that he can use directly, and from the rest he will get many helpful leads. This book is recommended to him without reservation." For another review see, *Wireless Eng.*, vol. 25, p. 302; September, 1948.

STATIONS AND COMMUNICATION SYSTEMS

621.39×534.756 207
Relationship Between Rate of Transmission of Information, Frequency Bandwidth, and Signal-To-Noise Ratio—Earp. (*See* 62.)

621.394.441 208
Voice Frequency Telegraphy—S. Rodhe. (*Ericsson Rev.*, no. 2, pp. 54–64; 1948.) An outline of the broad principles of the Ericsson system, with a detailed discussion of the methods of modulation and demodulation.

621.394.441:621.396.615 209
Stabilized Oscillator for Multiplex Voice-Frequency Telegraphy—G. Delyon and M. Manière. (*Câbles et Trans.* (Paris), vol. 2, pp. 209–218; July, 1948. With English summary.) The causes of frequency instability in various common types of oscillator are examined theoretically and also experimentally. A simple feedback pentode circuit is described, which includes a compensating resistance in the grid circuit and feeds into a 500-Ω load. No thermostat is used. Frequency stability is within 1 cps for a temperature variation from 5° to 50°C and for supply voltage variations of ±5 per cent, so that full use can be made of the bandwidth of the channel filters.

621.394.5:621.396.65:621.394.33 210
International Automatic Telegraph Networks: Part 1—J. D. M. Robinson. (*Jour. Brit. I.R.E.*, vol. 7, pp. 266–283; November and December, 1947. Discussion, pp. 284–285.) War-time developments in the long-distance telegraph techniques of the fighting services of the British Commonwealth and the United States of America open up new possibilities in international communications. The problem of providing an international network of greatly enhanced scope and efficiency is analyzed, with particular reference to the "message storage" intercircuit technique developed by the United States army. This technique could form the basis of a greatly improved system, provided a single telegraph code and alphabet could be adopted universally. The various automatic systems are considered briefly, intercircuit techniques with either manual or automatic routing are compared and the main design requirements of the future international automatic telegraph network are outlined. Part 2, 211 below.

621.394.5:621.396.65:621.394.33 211
International Automatic Telegraph Networks: Part 2—E. V. D. Glazier. (*Jour. Brit. I.R.E.*, vol. 7, pp. 286–297; November and December, 1947.) Discussion of the problems associated with the design of radio links forming part of an automatic telegraph system, with particular reference to distortion, signal-to-noise ratio, reduction of misprinting on teleprinter systems, frequency-shift operation, diversity reception, and multiplex working. Part 1, 210 above.

621.394/.395].73:621.394/.395].645 212
Line Amplifier for the Paris-Toulouse Coaxial Cable—R. Sueur. (*Câbles and Trans.* (Paris), vol. 2, pp. 243–250; July, 1948. With English summary.) Discussion of the characteristics of the cable and a detailed description of the amplifier and its use as a unit in each of the 40 repeater stations. The amplifier uses a special long-life pentode and has 2 parallel amplification paths with a common negative feedback.

621.395.44:621.315.052.63 213
Power Line Carrier Communications: Part 2—R. C. Cheek. (*Communications*, vol. 28, pp. 26–27, 35; August, 1948.) Discussion of (a) single-frequency automatic simplex and multistation duplex systems, and (b) different types of calling systems. Part 1, 819 of 1948.

621.396.41:621.396.65 214
The Multiplex Radio Link between Corsica and the Mainland—P. Rivère. (*Onde Élec.*, vol. 28, pp. 259–267 and 337–344; July, August, and September, 1948.) See 3508 of 1948.

621.396.5(494):621.396.41 215
The Zurich-Geneva Multichannel Radiotelephone Link in Public Trial Service—W. Steinmann and E. Huber. (*Brown Boveri Rev.*, vol. 35, pp. 111–115; March and April, 1948.) Topographical conditions make it relatively easy to establish beam radio-telephone links in Switzerland, where the links are needed to supplement existing line and cable networks

for trunk telephony. Details are given of the Zürich-Metliberg-Chasseral-Geneva link, which is the first part of such a beam network.

621.396.619.13 216
Frequency Modulation—"Cathode Ray." (*Wireless World*, vol. 54, pp. 339–343; September, 1948.) An elementary discussion of the points that seem to cause most confusion.

621.396.619.13 217
Practical Use of Frequency Modulation on Amateur Frequencies—D. N. Corfield. (*Proc. RSGB*, no. 4, pp. 1–7; 1948.) Discussion of equipment used and results obtained using FM in the 58.5 to 60-Mc amateur band. The essential differences between FM and AM transmission and reception are also considered.

621.396.65 218
[London] Overseas Exchange (Radio Services)—R. C. Devereux. (*P.O. Elec. Eng. Jour.*, vol. 41, part 2, pp. 76–82; July, 1948.) An account of the facilities and equipment of the new exchange opened in November, 1947, to handle exclusively the large volume of traffic from Great Britain and Europe to all parts of the world. The layout of the switchroom, connections with the two radiotelephone terminals in London and at Hendon, circuit and control arrangements, and operation of the service, are described briefly.

621.396.65 219
Indirect Microwave Relay System—R. R. Wakeman. (*Tele-Tech.*, vol. 7, pp. 42–43, 106; September, 1948.) Plane reflectors which require no external power source are used as auxiliary relays when obstacles intervene between transmitter and receiver. Relay stations need not then be intervisible.

621.396.65:621.396.615.142.2 220
ST [studio/transmitter] Equipment using Klystrons—(*FM and Telev.*, vol. 8, pp. 21–23; July, 1948.) A FM radio link designed primarily for high-fidelity broadcast service. It can also be adapted for the transmission of wide-band modulation, the limits of which fall within the design of the system. Sperry Type SRL-17 reflex klystrons are used for both transmitter and superheterodyne receiver. The frequency range is 920 to 960 Mc. Automatic-frequency-control and monitoring facilities are provided. Construction and performance details are given.

621.396.65.029.64 221
Repeater Buildings for the First Radio Relay System—W. L. Tierney. (*Bell Lab. Rec.*, vol. 26, pp. 281–288; July, 1948.) The buildings, normally unattended, are identical in structure, except that the central and the 4 corner supporting columns are continued through the roof for the 3 stations requiring antennas above roof level. Standby power supplies are available; power failures are automatically indicated at a remote maintenance station. See also 1755 of 1948 (Durkee).

621.396.712(489) 222
Copenhagen Broadcasting House—F. Heegaard and E. Cohrt. (*Elec. Commun.* (London), vol. 25, pp. 106–112; June, 1948.) A general description. See also 2938 of 1947.

621.396.72:621.398 223
Unattended Low-Power Transmitting Stations with Remote Control—F. C. McLean and R. Toombs. (*BBC Quart.*, vol. 3, pp. 122–128; July, 1948.) Discussion of the transmitter building, the antenna and earth system, and the design of the remote-control system for the duplicate crystal-controlled 1,474-kc 2-kw transmitters used at each station. The combined initial and operating costs are much less for unattended stations than for attended stations of comparable size.

621.396.822.1:621.396.41:621.396.619.16 224
Study of Crosstalk in Multiplex Systems using Pulse-Position Modulation—II. Gardère and J. Oswald. (*Câbles and Trans.* (Paris), vol. 2, pp. 173–193; July, 1948. With English summary.) Crosstalk is largely due to the displacement of 1 pulse caused by elongation of the preceding pulse; it is zero for ideal square pulses, but even if pulses with an infinitely steep front could be obtained, they would be distorted in passage through the various quadripoles of the system. Examples show that the distortion arises from the limitation of the pass band and from imperfect linearity of the phase versus frequency curve, 2 effects which are closely connected. The effect of a low-pass filter on pulses of various shapes is studied and the crosstalk for each case is calculated as a function of bandwidth. The results can be extended to other quadripoles, particularly band-pass filters; the study can form a basis for a more general theory.

621.396.931 225
Technical Aspects of Experimental Public Telephone Service on Railroad Trains—N. Monk and S. B. Wright. (PROC. I.R.E., vol. 36, pp. 1146–1152; September, 1948.) 1948 IRE National Convention paper. Describes component parts of the train telephone system, results of radio coverage tests on the routes involved, and devices employed to control 2-way transmission. Special features of the installations which differ from previous mobile installations and some results of the experimental operation are pointed out.

621.396.931 226
V.H.F. Railroad Radio Link—P. B. Patton. (*Communications*, vol. 28, pp. 16, 33; July, 1948.) Details of the equipment and results obtained for a 44-mile duplex FM radio link using frequencies of about 160 Mc. Simultaneous 6-channel operation was obtained by single-sideband working. Five of the channels were for telephony; the sixth could be used for 5 pulse-signaling and 2 teletype channels with sub-channel carrier frequencies in the voice-frequency range. The 6 main channels were spaced at 4-kc intervals. The antennas were vertically polarized 16-element arrays.

621.396.931 227
Compact Universal Mobile Unit: Parts 1 and 2—L. P. Morris. (*FM and Telev.*, vol. 8, pp. 34–38 and 19–22; July and August, 1948.) Circuit details of a vehicle-dispatcher equipment, for single- or dual-frequency use, with optional selective calling.

621.396.931.029.62 228
A Multi-Carrier V.H.F. Police Radio Scheme—J. R. Brinkley. (*Jour. Brit. I.R.E.*, vol. 8, pp. 128–142; May and June, 1948. Discussion, pp. 143–147.) A full description of the system noted in 1163 of 1948.

621.396.932:621.396.5 229
The 248A Marine Radio Telephone Equipment—R. C. Newhouse. (*Bell Lab. Rec.*, vol. 26, pp. 294–297; July, 1948.) For communication with either coastal radio stations or other ships. The transmitting power is 250 watts, and any one of 30 crystal-controlled frequencies in the range 2 to 18 Mc may be used.

SUBSIDIARY APPARATUS

621.352.7 230
Fresh Progress in Dry Batteries: the Vidor "Kalium" Cell—R. W. Hallows. (*Wireless World*, vol. 54, pp. 352–354; October, 1948.) The cell supplies a constant electromotive force of 1.25 to 1.4 volts for long periods under load. The use of HgO enables depolarization to keep pace with electrolytic action, so that the objectionable sawteeth of the Leclanché cell discharge curve are absent. The "Kalium" cell costs considerably more than the Leclanché type.

621.396.68:621.316.722 231
Stabilized Power Supplies: Part 1—Practical Design Procedure for Series-Valve Types—M. G. Scroggie. (*Wireless World*, vol. 54, pp. 373–378; October, 1948.) Basic principles are discussed. Design is simplified by the use of tube characteristics and load-current versus voltage diagrams. See also 197 of 1944 (Hogg).

621-526 232
Theory of Servomechanisms [Book Review]—H. M. James, N. B. Nichols, and R. S. Phillips (Eds). McGraw-Hill, New York, N. Y., and London, 1947, 375 pp., $5. (*Nature* (London), vol. 161, p. 994; June 26, 1948.) No. 25 of the Radiation Laboratory series. Only linear systems are discussed. One of the first major works to be published on the subject "This book . . . will be welcomed universally."

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 233
Possibilities of Stratovision—M. B. Sleeper. (*FM and Telev.*, vol. 8, pp. 15–17, 45; August. 1948.) Illustrations and short discussion of Westinghouse equipment installed in a Martin-202 aircraft. In a demonstration at Zanesville, Ohio, the aircraft flew at a height of 25,000 feet, picked up signals on channel No. 2 from WMAR-TV, Baltimore, and rebroadcast them on channel No. 6. Excellent reception of the rebroadcast signals was obtained at distances of over 200 miles. Some interference was experienced, because the aircraft picked up signals from WNBT, New York, N. Y., also operating on channel No. 2. With the present limited number of channels available for allocation to ground stations in United States, the aircraft will always be liable to pick up signals from 2 or more stations using the same channel, and its transmissions may interfere with those from 1 or more ground stations. This difficulty would appear to make stratovision impracticable on the lower frequency band. The possibility of using frequencies from 475 to 890 Mc is considered briefly.

621.397.331.2 234
New Design for Medium Definition TV Camera System—J. B. Sherman. (*Tele-Tech.*, vol. 7, pp. 52–55; September, 1948.) Description of experimental iconoscope Type 5527, with full circuit diagrams. The scanning circuits provide 250-line resolution with interlace ratios of 2/1, 3/1, or 4/1 on a 7-inch viewing tube with electrostatic deflection. A small motor, with clutch to prevent over-travel, is used for optical focusing, with the controls mounted on the main chassis.

621.397.331.2:[621.395.625.6+771.3 235
Television Transcription by Motion Picture Film—T. T. Goldsmith, Jr., and H. Milholland. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 107–113; August, 1948. Discussion, pp. 113–116.) Description of the electronic and camera equipment for recording television pictures and sound on film, the pictures being recorded directly from the face of the crt. Applications to documentary recording, theater television projection, etc., are discussed. See also 236 below.

621.397.331.2:621.395.625.6 236
Television Recording Camera—J. L. Boon, W. Feldman, and J. Stoiber. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 117–126; August, 1948.) Description of the principal features of a 16-mm camera for recording at sound speed directly from a monitor receiving crt. See also 235 above.

621.397.5:535.88 237
Developments in Large-Screen Television—R. V. Little, Jr. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 37–46; July, 1948. Discussion, pp. 47–51.) Description of light source, optical

system, screen, and high-voltage power units of experimental equipment.

621.397.5:535.88 238
Optical Problems in Large-Screen Television—I. G. Maloff. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 30–36; July, 1948. Discussion, pp. 47–51.) General discussion, with details of 1 prewar and 2 postwar RCA large-screen projectors.

621.397.5:535.88:535.317.9 239
The Manufacture of Correction Plates for Schmidt Optical Systems—Rinia and van Alphen. (*See* 63.)

621.397.5:535.88:791 240
Development of Theater Television in England—A. G. D. West. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 127–167; August, 1948. Discussion, pp. 167–168.) An historical review of progress in large-screen television projection, both before and after the war, and a description of the design and performance of equipment developed for the distribution and projection of television programs. Proposals now being made for setting up a theater television service, first in London, and then throughout Great Britain, are discussed.

621.397.61:621.317.3 241
TV Transmitter Design: Part 4—Hamilton. (*See* 138.)

621.397.7 242
Salute to WPIX: Blueprint for Engineering a Television Station—(*Tele-Tech*, vol. 7, pp. 33–56; July, 1948.) An account of its construction, operation, and facilities. The station was built as extra floors to the Daily News Building, New York, N. Y., the antenna and tower being mounted on top of the 37th floor. The transmitter delivers 5 kw video power and 2.5 kw FM sound signal to the antenna transmission line. Output is approximately 18 kw. The studio control room, located at one end and above the floor of the studio, is divided into 2 levels, concerned respectively with technical and program control. The master control for the station provides complete monitoring of sight and sound signals. Problems associated with transmission from films are also discussed. The acoustic treatment of the studio involved the construction of a "room within a room." Transmissions from mobile stations are also relayed.

621.397.743:621.396.67 243
Television Antenna and RF Distribution Systems for Apartment Houses—H. E. Kallmann. (Proc. I.R.E., vol. 36, pp. 1153–1160; September, 1948.) Discussion of requirements, and description of amplifiers and distribution networks which supply boosted signals of all available stations on their original carriers to all receivers in an apartment house.

TRANSMISSION

621.316.726.078.3 244
Frequency Stabilization in the Region of 10,000 Mc/s—Donnelly. (*See* 136.)

621.394.61 245
10 kW Short-Wave Telegraphy Transmitter—M. G. Favre. (*Brown Boveri Rev.*, vol. 35, pp. 85–90; March and April, 1948.) Over-all efficiency is 45 per cent so that operating costs are low. Installation and operation are simple. The high voltage is only 6 kv and CuO rectifiers are used. The keying speed is 750 words per minute. See also 1615 of 1947 (Guyer and Favre).

621.396.619.23 246
A Broad-Band High-Level Modulator—R. J. Rockwell. (Proc. I.R.E., vol. 36, pp. 1160–1164; September, 1948.) The class-B modulator has a broad pass-band with uniform gain and low distortion and noise level. Broad-band feedback and a cathode follower are used; there is no modulation transformer.

VACUUM TUBES AND THERMIONICS

621.383.5:538.567.2 247
The Photo-Voltaic Effect—Lehovec. (*See* 73.)

621.385+621.314.6 248
Transmitting and Rectifier Tubes—R. Kesselring. (*Brown Boveri Rev.*, vol. 35, pp. 99–103; March and April, 1948.) Details of various Brown Boveri types. See also 1634 of 1947 and 2967 of 1948 (Jenny).

621.385:621.396.822 249
Induced Grid Noise and Total-Emission Noise—A. van der Ziel and A. Versnel. (*Philips Res. Rep.*, vol. 3, pp. 13–23; February, 1948.) Measurements are given of the noise resonance curve of the input circuit of pentodes for λ 7.25 meters. The asymmetry of this curve is due to the phase relationship between the induced grid noise and the normal shot-effect noise. For diodes in the 7.25-meter cutoff region, the total-emission noise can be calculated by assuming that the equivalent noise temperature of the total-emission conductance is equal to the cathode temperature.

621.385.029.63/.64 250
Nonlinear Phenomena in Traveling-Wave Valves—O. Doehler and W. Kleen. (*Ann. Radioélec.*, vol. 3, pp. 124–143; April, 1948.) Continuation of 602 of 1948. The limit of output and useful power is calculated, taking account of the nonlinearity of the gain. A method of development in series and successive approximation is used, similar to that employed by Warnecke, Bernier, and Guénard (2215 of 1945) for calculating the effect of space charge in the drift space of a klystron. Only the energy corresponding to the excess of electron velocity over wave velocity can be transferred from the electron beam to the wave and transformed into uhf energy. The electrons are retarded by the field of the traveling wave, and once this retardation has reduced the electron velocity to the wave velocity, the power can increase no further. Space-charge effects and the action of radial electric fields on the electrons also tend to reduce useful power and output.

621.385.029.63/.64:537.525.92 251
On the Effect of the Space Charge in the Traveling-Wave Valve—O. Doehler and W. Kleen. (*Ann. Radioélec.*, vol. 3, pp. 184–188; July, 1948.) The method of successive approximations applied by Warnecke, Bernier, and Guénard (2215 of 1945), to study the effects of space charge in the drift space of a klystron, is here applied to the traveling-wave tube. The treatment is limited to the case of small signals. The electron beam is assumed to be (a) narrow with respect to the dimensions of the cross section of the retarding line, (b) infinitely wide. In the traveling-wave tube, the space charge effects can either increase or decrease the tube gain, according to the conditions of operation. See also 602 of 1948 and 250 above.

621.385.029.63/.64:621.392.5 252
Experimental Method for *a priori* Tests of Delay Lines for the Travelling-Wave Amplifier—Lapostolle. (*See* 39.)

621.385.032.216:537.533 253
The Mechanism of the Thermionic Emission from Oxide Coated Cathodes—H. Friedenstein, S. L. Martin, and G. L. Munday. (*Rep. Progr. Phys.*, vol. 11, pp. 298–339; 1946 and 1947. Bibliography, pp. 339–341. An historical account is given of the main experimental data so far obtained on the general behavior of oxide cathodes. The mechanism of normal emission is considered both experimentally and theoretically, with special reference to the Wilson and de Boer models of the process. Neither is as yet completely established, but recent work tends to favor the de Boer model. The fundamental difficulty appears to be the lack of knowledge of the way in which the excess alkaline-earth metal is built into the oxide. Recent research on emission under pulsed conditions has led to a revival of Riemann's conception of a potential barrier at the core coating interface.

621.385.1 254
The Consequences of an Electron-Inertia Effect in Valves—M. J. O. Strutt and A. van der Ziel. (*Physica, 's Grav.*, vol. 8, pp. 81–108; January, 1941. In German, with English summary.) Discussion of the currents flowing to the various electrodes of a tube when a small bunch of electrons starts from the cathode, and of the case of periodic pulses from the cathode. In certain cases, only dc flows to the anode as a result of such pulses. With an alternating voltage between the input grid and the cathode, the pulse shape may differ widely from the voltage curve. Different paths in a tube for different electrons may result in curious effects such as rectification or -frequency multiplication. A single sinusoidal voltage pulse on the input grid is considered and the minimum values of input resistance and amplification are deduced. The lower limit of amplification is 0.5 for multi-electrode tubes and is >1 for secondary-emission tubes at any frequency, however high. Induction effects are caused by the passage of electrons near to conductors inside tubes; simple formulas are derived for the displacement currents. For the rapid calculation of phase angles, a "center of gravity theorem" is enunciated. Simple deductions are made for the case of transit-time modulation.

621.385.1:621.396.822 255
Fluctuations and Electron Inertia—C. J. Bakker. (*Physica, 's Grav.*, vol. 8, pp. 23–43; January, 1941. In English.) The shot effect in tubes gives rise, by induction, to fluctuating displacement currents toward the tube grids, whose negative potential prevents electrons from reaching them. The fluctuating currents in the control-grid/cathode lead have been investigated experimentally for two types of tubes. The results obtained are in satisfactory agreement with theory.

621.385.1.012 256
Dynamic Methods for the Determination of Valve Characteristics—G. Gregoretti. (*Alta Frequenza*, vol. 17, pp. 110–127; June, 1948. In Italian, with English, French, and German summaries.) Static methods cannot be used with safety for transmitting tubes on account of the risk of damage if normal operating voltages are applied. Various practical pulse methods are described. Differences between the results obtained by the different methods are analyzed.

621.385.38.001.8 257
Thyratrons and their Industrial Applications—J. Martin. (*Bull. Soc. Franç Élec.*, vol. 8, pp. 371–384; July, 1948.) Detailed description of the construction and performance of a thyratron (VHC 3/1000) made by the Société Française Radioélectrique, and also circuit diagrams and discussion of numerous applications.

621.385.5.001.8:621.392 258
Applications of Screen-Grid Supply Impedance in Pentodes—P. G. Sulzer. (*Communications*, vol. 28, pp. 10–11, 39; August, 1948.) The anode-current versus grid-voltage characteristic of a pentode with negative suppressor-grid voltage and an impedance in the screen-

grid circuit can have a negative slope. This property is used in circuits suggested for phase inverters, dc and ac coupled trigger circuits and negative-resistance oscillators.

621.396.615.141.2 259

The Frequency Stability of Magnetron Oscillations of the Dynatron Type—N. S. Zinchenko. (*Radiotekhnika* (Moscow), vol. 3, pp. 40–53; March and April, 1948. In Russian.) An experimental investigation of the "dynatron" oscillations of magnetrons for long, short, and decimeter waves is described. Curves are plotted showing the frequency variations with the magnetic field, anode voltage, and filament current and a formula (p. 42) determining the relative frequency variation in terms of these factors is derived. Conditions necessary for frequency stability are established. The stabilities of magnetron and tube oscillators operating as dynatrons are compared. Curves showing the effect of the grid voltage on the stability of split-anode magnetrons are also plotted.

621.396.615.141.2 260

The Rotating Space Charge in a Whole-Anode Magnetron—I. I. Vasserman (Wassermann). (*Zh. Tekh. Fiz.*, vol. 18, pp. 785–792; June, 1948. In Russian.) The critical current corresponding to the rotating space charge of a magnetron is calculated and compared with experimental results. The rotating current is proportional to the product of the charge density and the tangential velocity of the electrons. The experimental results seem to support the theory of Brillouin that the radial velocity of electrons is zero at all points of the interelectrode space and that the electrons move in circular orbits whose centers are on the cathode, rather than the theory of Möller, Braude, Page, and Adams that the radial velocity is zero at only one point on the surface of the anode and that the electrons describe cardioid orbits.

621.396.615.141.2 261

On the Properties of Valves using a Constant Magnetic Field: Parts 1 and 2—O. Doehler. (*Ann. Radioélec.*, vol. 3, pp. 29–39 and 169–183; January and July, 1948.) Part 1 discusses the static and dynamic characteristics of the magnetron. From Brillouin's work (1892 of 1943), formulas are derived which define the space charge, the distribution of potential, and the characteristics for magnetic fields below the critical value. The oscillation frequencies of multicavity magnetrons are calculated; the results are in good agreement with measured values. Corrections are given at the end of part 2.

In part 2 the treatment is extended to high-power resonance oscillations excited in the region above the critical point. The differences between these oscillations and electron oscillations are examined and a quantitative relation is established which gives the optimum operating conditions. The input impedances and the output are calculated and an empirical relation between the anode current and the migration current is given. The dynamic electron trajectories in a plane magnetron without space charge are determined, by way of introduction to the study of the traveling-wave tube. To be continued.

621.396.615.142.2 262

On Some Typical Models of Velocity-Modulation Valves—R. Warnecke. (*Ann. Radioélec.*, vol. 3, pp. 71–106; April, 1948.) A detailed account of the construction and characteristics of (a) reflex klystrons for low-power oscillators for receivers or transmitters operating in the 3-, 10- and 20-cm bands; (b) a frequency-multiplier klystron which enables the frequency stability of a quartz oscillator to be transferred to the 10-cm band; (c) a 2-cavity klystron with a high-frequency output in the 10-cm band of more than 1 kw. Reprinted in *Onde Élec.*, vol. 28, pp. 175–185, 243–256, and 287–294; May to July, 1948.

621.396.615.142.2 263

Multifrequency Bunching in Reflex Klystrons—W. H. Huggins. (Proc. I.R.E., vol. 36, p. 1145; September, 1948.) Corrections to 2988 of 1948.

621.396.645:537.311.33:621.315.59 264

The Transistor, A Semi-Conductor Triode—J. Bardeen and W. H. Brattain. (*Phys. Rev.*, vol. 74, pp. 230–231; July 15, 1948.) A short account of the basic principles of the triode device discussed in detail in 265 below. Such units have been operated as amplifiers at frequencies up to 10 Mc. Relevant theory is also given in articles noted in 3438 of 1948 (Shockley and Pearson) and 3439 of 1948 (Brattein and Bardeen). For another account of the transistor see *Electronics*, vol. 21, pp. 68–71; September, 1948. (D.G.F. and F.H.R.).

621.396.645:537.311.33:621.315.59 265

Experimental Germanium Crystal Amplifier [the Transistor]—W. Wells and S. Y. White. (*Audio Eng.*, vol. 32, pp. 6, 8 and 28–29, 39; July and August, 1948.) The crystals used for the "transistor" developed by the Bell Laboratories belong to the class whose unidirectional conductivity characteristics are derived from imperfections in the crystal lattice introduced by impurities in the material. In this particular case, the Ge crystal contains small amounts of Sn, interspersed between the Ge atoms in the crystal lattice. Between the inner and outer electron shells of the Ge atom, there exists a zone whose force fields do not permit the existence of orbital electrons under normal circumstances. The proximity of the Sn atom so distorts these force fields that the 4 electrons from the "O" shell of the Sn atom may circulate in this "forbidden shell" of the Ge atom. These borrowed electrons have a higher binding energy than the 4 electrons in the "N" shell of the Ge atom and, consequently, require a greater potential difference between the constant and the crystal for their release. When the contact point is made sufficiently positive in relation to the crystal, the orbits of these borrowed electrons expand, allowing them to come within the outer zone where conductivity may take place. Conversely, as the potential of the contact is made negative relative to the crystal, the orbits of the borrowed electrons shrink placing them in a zone where they are no longer available for participation in conductivity. A fairly high potential will, therefore, be required to maintain a current flow between a negatively charged contact point and the crystal, since the borrowed electrons will be pushed down below the zone of activity and only the outer electrons of the Ge atom will take part in the conductivity. If, however, a second contact point is introduced, in close proximity to the first and charged positively, the borrowed electrons will again be elevated to the active region, releasing about twice as many for participation in conductivity. These newly released electrons may take part in the conductivity of both the "anode" circuit and the "grid" circuit, so that their effect is that of modulating the current flow of the anode circuit. Energy considerations indicate that a grid input signal of the order of 3 mw would yield an output signal of 60 mw.

Amplification can only take place if $E_g I_g < E_a I_a$. Hence, circuit efficiency must be gained (a) by making the electronic paths of the anode and grid contacts through the crystal as nearly coincident as possible, consistent with low cross current flow between the contacts themselves, and (b) by using the optimum step-down impedance ratio between the anode circuit of one stage and the grid circuit of the following stage.

Details are given for the construction of experimental amplifiers using Type 1N34 Sylvania Ge crystals and tungsten contact points from Type 1N21 Si crystals.

For other accounts of the Transistor see *Bell Lab. Rec.*, vol. 26, pp. 321–324; August, 1948, and 264 above.

621.396.645:621.385.4 266

The FP-54 as a Stable Voltage Amplifier—Seaton. (*See* 55.)

MISCELLANEOUS

016:[6(43)+6(52) 267

Reports on German and Japanese Industry [up to 31st March 1948] Classified List No. 18. [Book Notice]—H. M. Stationery Office, London, 1948, 114 pp., 1s. Supersedes all previous consolidated lists and supplements. Section D includes varnishes and plastics. Section F: electrical engineering industry. Section G: glass and ceramics. Section K: metal industries. Section R: rubber industry.

025.45 268

Universal Decimal Classification—The British Standard publications dealing with this system and mentioned in 1521 of 1948 can be obtained in the United States from the American Standards Association, Grand Central Terminal Building, 70 East 45th St., New York, N. Y.

5+6]:05(43) 269

German Scientific and Technical Publications Today—W. T. Cooper. (*Research* (London), vol. 1, pp. 316–319; April, 1948.) An account of the journals now being published in Germany and of methods by which they may be obtained.

621.3 270

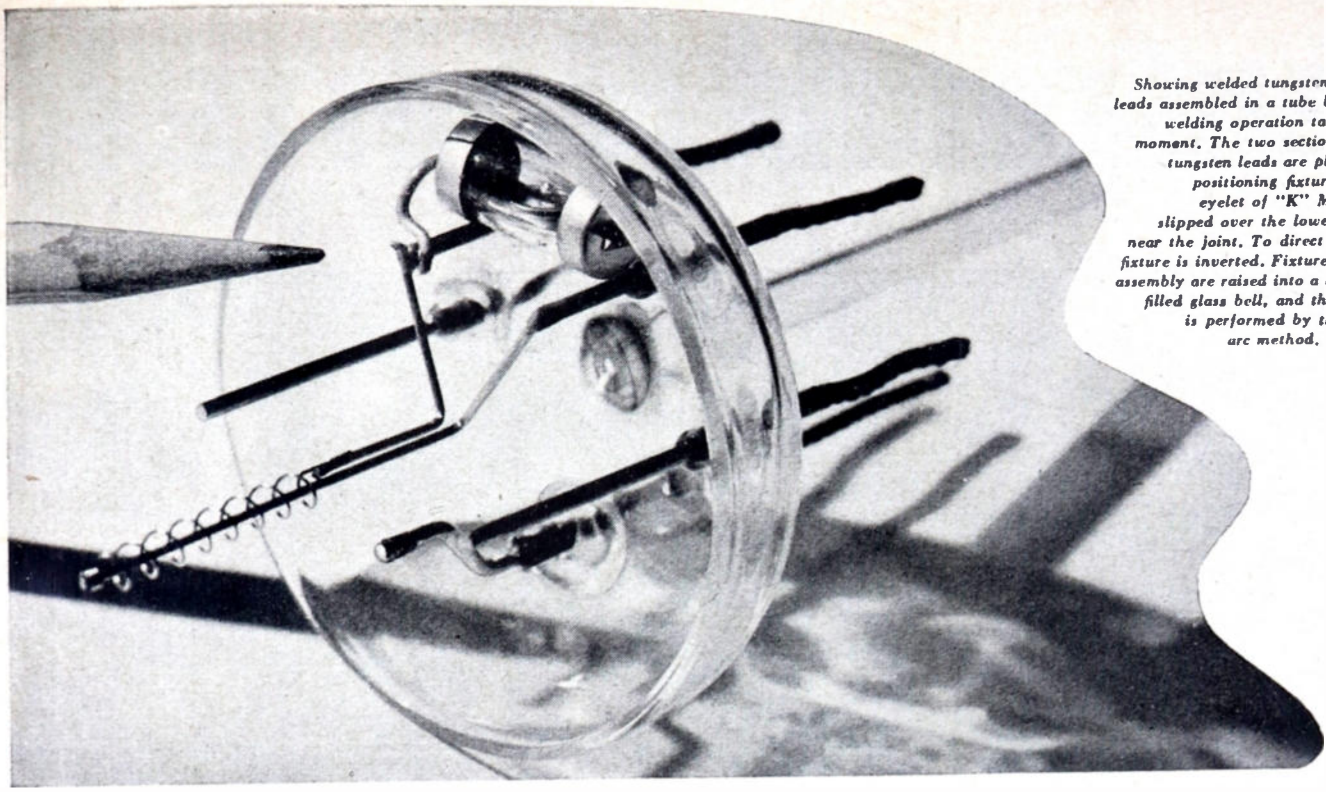
New Productions. Recent Manufactures of the Société Indépendante de T.S.F.—(*Ann. Radioélec.*, vol. 3, pp. 146–150; April, 1948.) These include (a) 2-kw and 500-w generators of compact design, for industrial high-frequency heating; (b) telegraphy telephony transmitter E.H.K. 8 for aerodromes, with frequencies in the range 265 to 415 kc; (c) high-precision absolute frequency meter covering the wide range from 14 kc to 25 Mc. This is a large rack-mounted equipment including a 100-kc quartz crystal, generators of frequencies 0.1, 1, 10, 1,000, and 10,000 kc, harmonic generators, receivers with loop antennas, and suitable apparatus for measuring the beats between the source to be measured and a convenient selected harmonic.

654.1 271

Sixth Plenary Meeting of the C.C.I.T., Brussels, May 1948—F.E.N. (*P.O. Elec. Eng., Jour.*, vol. 41, part 2, pp. 108–109; July, 1948.) A short account of the work of the various committees, particularly of the Technical Committee.

621.396 272

Essentials of Radio [Book Review]—M. S. Slurzberg and W. Osterheld. McGraw-Hill, New York, N. Y., 1948. 806 pp., $5.00. (*Electronics*, vol. 21, pp. 226–227; August, 1948.) "The authors . . . have presented here at an intermediate level, the principles of operation of the basic circuits and circuit elements used in conventional radio receivers, as essential background knowledge for understanding electronic circuits."



Showing welded tungsten filament leads assembled in a tube base. The welding operation takes but a moment. The two sections of the tungsten leads are placed in a positioning fixture. A tiny eyelet of "K" MONEL is slipped over the lower section, near the joint. To direct flow, the fixture is inverted. Fixture and lead assembly are raised into a hydrogen-filled glass bell, and the welding is performed by the carbon arc method.

How a problem in welding tungsten was solved

While improving the design of their VHF beam tetrodes, the United Electronics Company ran into a difficult technical problem.

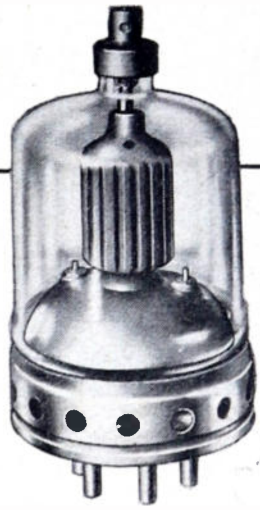
In their tube types 5D22 and 4D21, tungsten filament leads are brought out to conventional base prongs. However, to locate the filament at the center of the structure, the two internal filament leads had to be sharply offset. It was necessary, also, that the leads be accurately aligned with the base outlet holes, to eliminate stresses which might crack the glass envelope when the tube was put in service.

Bending the tungsten leads to shape proved too inaccurate a method. So it was decided to make the leads in two sections — one straight, and one bent — welding them together in precision positioning fixtures.

This method of assembly proved satisfactory, but difficulty was immediately encountered in finding a suitable joining metal.

Several metals were tried without success. Either they failed to "wet" the tungsten, or caused it to embrittle.

VHF beam tetrode tube, manufactured by the United Electronics Co., Newark, N.J. ▶



Finally, United Electronics Company engineers tried "K"* MONEL—*and it proved to be the answer to their problem.*

"K" MONEL "wet" the tungsten satisfactorily; flowed well; made strong, smooth joints; was resistant to oxidation and corrosion. In addition, "K" MONEL's melting point was safely above both exhausting and tube operating temperatures.

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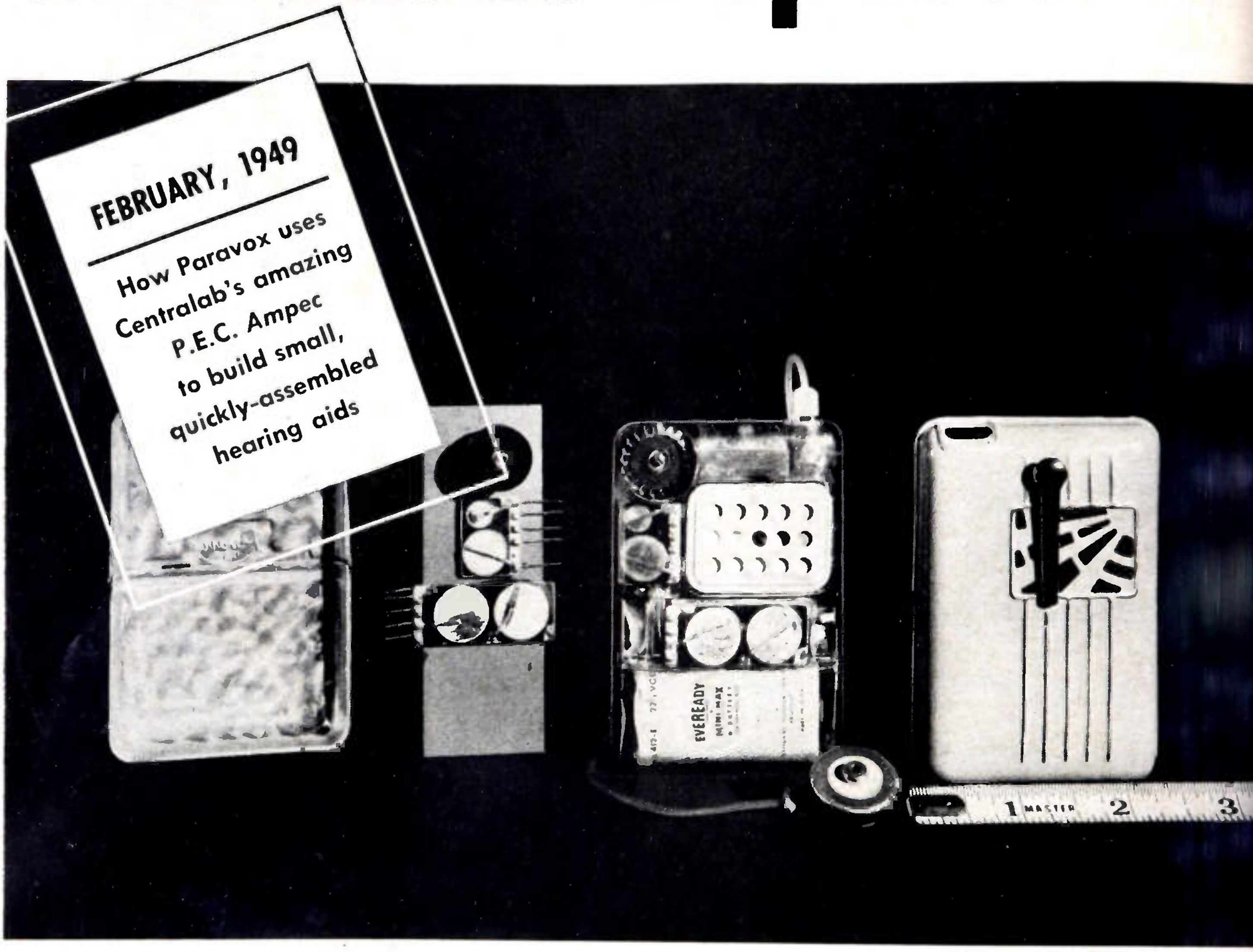
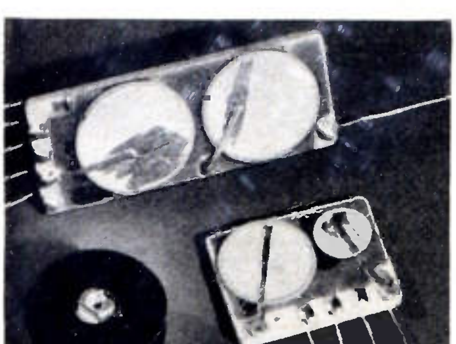
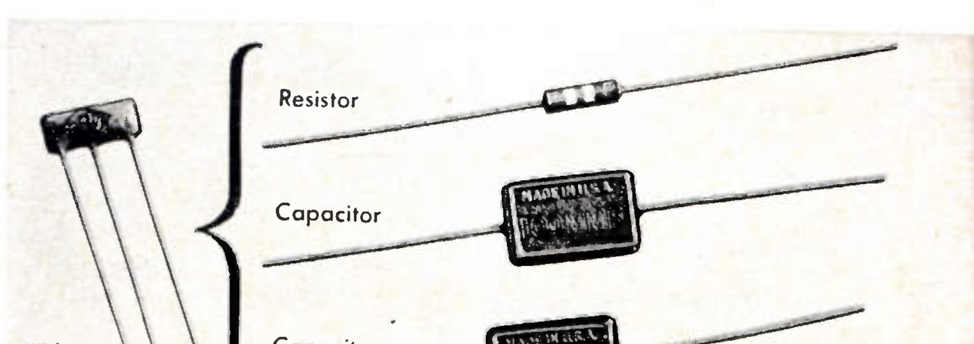


Models courtesy of Paravox, Inc.

1 Time, space and material savers! That's how Paravox, Inc., Cleveland hearing aid manufacturer, describes Centralab's revolutionary P.E.C. *Ampecs*. These tiny audio-amplifying units save time for Paravox by eliminating many assembling operations. They save space and material by reducing the number of components needed. What's more — like all CRL *Printed Electronic Circuits* — they are rugged, dependable, resistant to temperature and humidity. For Ampec facts, order Bulletin 973.

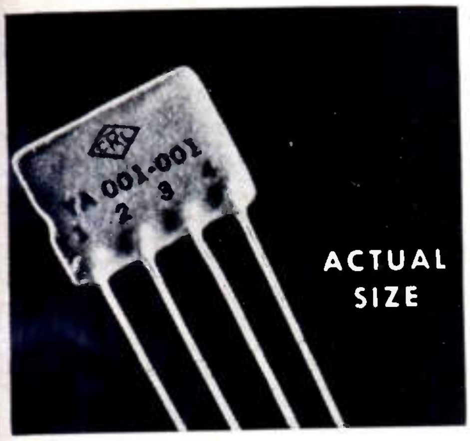


2 Two-piece *Ampec* and Model 1 *Radiohm* are the CRL units Paravox uses in its 4½-ounce hearing aid. *Ampec* is a complete 3-stage audio amplifier.

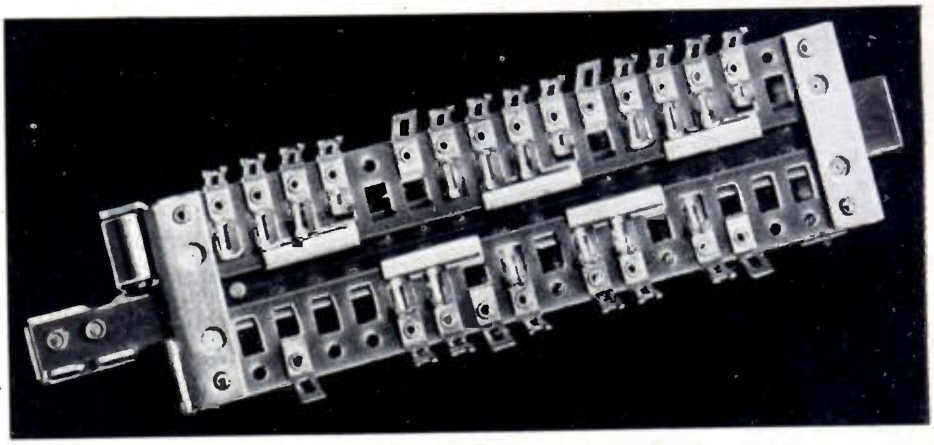


3 Centralab's *Filpec* is designed for use as a balanced diode load filter, combines up to three major components into one tiny unit, lighter and smaller than one ordinary capacitor. Capacitor values available from 50 to 200 mf. Resistor values from 5 ohms to 5 megohms. For complete information, write for Bulletin 42-9.

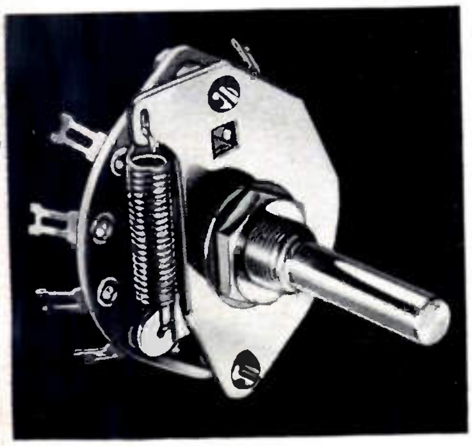
Electronic Industry

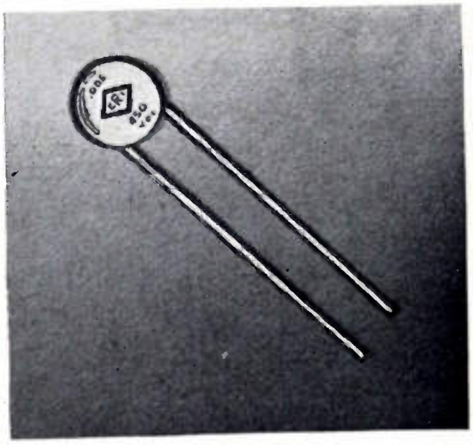


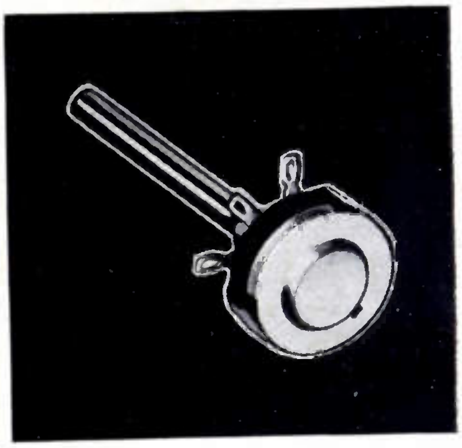
4 CRL's *Couplate* consists of a plate lead resistor, grid resistor, plate by pass capacitor and coupling capacitor. Write for Bulletin 42-6.



5 Centralab's development of a revolutionary, new *Slide Switch* promises improved AM and FM performance! Flat, horizontal design saves valuable space, allows short leads, convenient location to coils, reduced lead inductances for increased efficiency in low and high frequencies. Rugged, efficient. Write for Bulletin 953.







6 Great step forward in switching is CRL's new *Rotary Coil and Cam Index Switch*. Its coil spring gives you smoother action, longer life.

7 For by-pass or coupling applications, check CRL's original line of ceramic disc and tubular *Hi-Kaps*. For full facts, order Bulletins 42-3 and 42-4.

8 Wide range of variations in CRL's Model "M" Radiohm simplifies production and inventory. Bulletin 697-A illustrates convenience, versatility!

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SECTION MEETINGS

Baltimore

"'Symetron' 50-Kilowatt FM Broadcast Amplifier," by D. L. Balphia, Westinghouse Electric Corporation· November 30, 1948.

Beaumont-Port Arthur

"Some Traffic Control Problems and their Solution," by J. D. Southwell, City of Beaumont; September 22, 1948.

"Some Problems of Television," by B. E. Shackelford, 1948 President, The Institute of Radio Engineers; October 2, 1948.

"Construction Details of New KFDM Transmitting Plant," by S. H. Jones, KFDM; November 16, 1948.

"The Stratovision System," by C. E. Nobles Westinghouse Electric Corporation; December 8, 1948.

Buffalo-Niagara

"The Radio Engineers looks at Industrial Electronics," by E. D. Cook, General Electric Company; November 17, 1948.

"Industrial Applications of the Cathode-Ray Oscilloscope," by P. Ebbenhouse, Allen B. Dumont Laboratories, Inc.; December 15, 1948.

Cedar Rapids

"Radio—Radar—Microwaves," by J. O. Perrine, American Telephone and Telegraph Company; November 9, 1948.

Cleveland

"Servo-Mechanisms," by W. Buhl, Jack and Heintz Precision Industries; November 16, 1948.

Columbus

"A New Microwave Triode," by J. A. Morton, Bell Telephone Laboratories; November 17, 1948.

"A New Semiconductor Amplifier," by J. A. Morton, Bell Telephone Laboratories; November 17, 1948.

Connecticut Valley

"AC and DC Electronic Voltage Regulators," by L. Helterline, Sorenson and Company, Inc.; November 18, 1948.

"The Handling of Ideas Submitted to the General Electric Patent Department," by M. W. Morse, General Electric Company; December 16, 1948.

Dallas-Fort Worth

"Technical Considerations in Television Receivers," by J. E. Palmer, West Texas Appliance Company; November 10, 1948.

Dayton

"Phase Distortion In Audio Systems," by L. A. DeRosa, Federal Telecommunications Laboratories, Inc ; December 9, 1948.

Des Moines

"The Electron Microscope," by P. H. Carr, Iowa State College; November 17, 1948.

Emporium

"Modern Quality Control," by S. Collier, Johns Manville Corporation; December 7, 1948.

Election of Officers; December 7, 1948.

Houston

"Ultrasonics—Industrial Applications," by J. Woodburn, Rice Institute; October 15, 1948.

"Stratovision," by C. E. Nobles, Westinghouse Electric Corporation; December 7, 1948.

Los Angeles

"The Transistor," by J. A. Bardeen and W. Shockley, Bell Telephone Laboratories; November 18, 1948.

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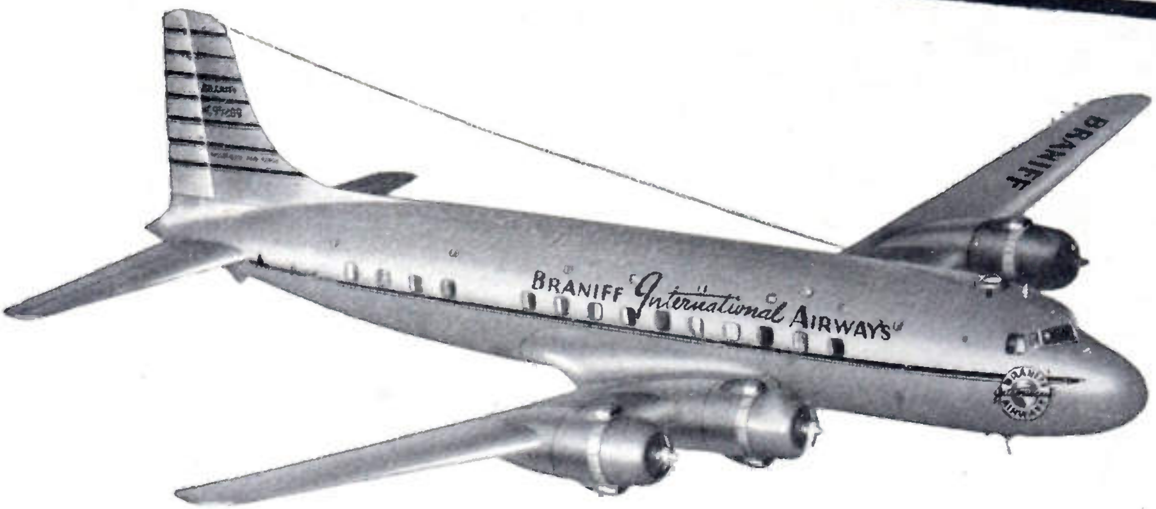

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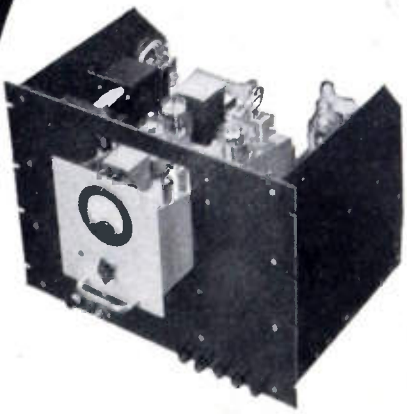

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(Continued from page 36A)

LOUISVILLE

"Application of Induction Heating in the Manufacture of Tractors," by J. Graham, International Harvester Company; November 12, 1948.

"The Design and Manufacturing of Phasing Equipment for Directional Antenna," by J. S. Brown, Andrews Corporation; December 10, 1948.

MILWAUKEE

"Common Misconceptions in the Engineering Field," by W. E. Richter, Allis-Chalmers Manufacturing Company; November 10, 1948.

"X-Ray Sources for Roentgen Therapy," by D. Trout, General Electric X-Ray Corporation; November 17, 1948.

"Roentgen Irradiation of Tumors," by M. Friedman, New York University; November 17, 1948.

"Sound and Video Separation Circuits," by B. Parmet, Motorola Corporation; December 8, 1948.

NEW MEXICO

"Some Topics in FM Theory and Application," by C. W. Carnahan, Sandia Laboratory; November 19, 1948.

"Measurements at Ultra-High Frequencies," by W. R. Hewlett, Hewlett-Packard Company; December 17, 1948.

NEW YORK

"Magnetic Materials with a Rectangular Hysteresis Loop," by E. Both, Fort Monmouth Signal Corps Engineering Laboratories; October 20, 1948.

"Fundamental Operation and Application of Magnetic Amplifiers," by R. E. Morgan, General Electric Company; October 20, 1948.

"The Transistor—A Crystal Triode," by W. Shockley and J. A. Becker, Bell Telephone Laboratories; November 10, 1948.

OTTAWA

"Pulse Time Modulation," by A. R. St. Louis, Federal Electric Manufacturing Company; October 28, 1948.

"Television," by H. S. Dawson, Canadian General Electric Company; November 18, 1948.

"GM Counters and their Applications," by E. Lever, Electronic Associates Ltd.; December 9, 1948.

PHILADELPHIA

"Recent Advances in the Theory of Communications," by C. E. Shannon, Bell Telephone Laboratories; December 2, 1948.

PITTSBURGH

"Particle Accelerators and Accelerated Particles," by E. Creutz, Carnegie Institute of Technology" November 8, 1948.

PORTLAND

"The Transistor, its Characteristics and Properties," by W. H. Brattain, The Bell Telephone Laboratories; November 18, 1948.

"Some Design Problems of a High-Gain, Wide-Band Direct-Coupled Oscilloscope," by L. Belleville, Tektronix Inc.; November 22, 1948.

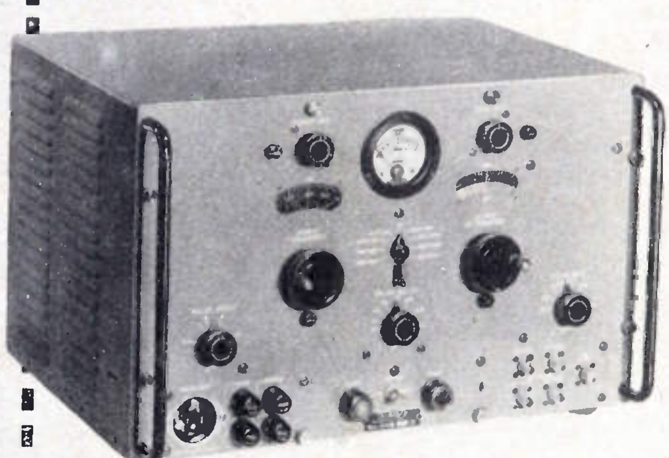
PRINCETON

"Microwave Radio Relay Systems," by J. W. McRae, Bell Telephone Laboratories, December 2, 1948.

ROCHESTER

"Teleran—Air Navigation and Traffic Control through Pictorial Display," by D. Ewing, Radio Corporation of America, December 2, 1948.

(Continued on page 40A)



SECTION MEETINGS

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Sacramento

"Electrical and Mechanical Measurements Using Electronic Instruments," by N. Porter, Hewlett-Packard Company; November 17, 1948.

"X-Ray Sources for Industrial and Therapy Applications," by E. D. Trout, General Electric Company; December 13, 1948.

"Medical Applications of X-Ray and Demonstration of New Million Volt X-Ray Unit Recently Installed at Sutter Hospital," by C. E. Grayson, Sutter Hospital; December 13, 1948.

San Diego

"The Transcontinental Coaxial Cable System," by C. T. Koerner, Pacific Telephone and Telegraph Company; November 10, 1948.

San Francisco

"Microwaves in Molecular Spectroscopy," by W. Gwinn, University of California; November 10, 1948.

"Status of Television in the Bay Area," by R. A. Isberg, KRON-TV, H. N. Jacobs, KGO-TV, and A. F. Towne, KPIX-KSFO; December 8, 1948.

Seattle

"The Transistor," by W. Brattain, Bell Telephone Laboratories; November 12, 1948.

Election of Officers; December 14, 1948.

Syracuse

"Microwave Astronomy," by C. H. Burrows, Cornell University; November 12, 1948.

"Indoor and Outdoor Experiments in Rain and Snow," by V. J. Shaefer, General Electric Company; November 29, 1948.

Toledo

"Film Projection for Television," by L. C. Downes, General Electric Company; November 22, 1948.

Toronto

"Modern Electronic Instrument Techniques," by R. Cox, National Research Council Atomic Energy Project; November 15, 1948.

"Techniques of Television Sound," by H. H. Tanner, Northern Electric Company; December 6, 1948.

Twin Cities

"Electronic Properties of Semiconductors and the Transistor," by W. H. Brattain, Bell Telephone Laboratories; November 29, 1948.

Washington

"The Electron Wave Tube—A Novel Method of Amplification and Generation of Microwave Energy," by A. V. Haeff, Naval Research Laboratory; December 13, 1948.

Election of Officers; December 13, 1948.

Subsections

Hamilton

"Twin 50-Kw B.C. Transmitters CBL-CJBC," by J. C. R. Punchard, Northern Electric Company; November 29, 1948.

Lancaster

"High Resolution Nuclear Particle Counting Circuits," by F. W. Van Name, Jr., Franklin and Marshall College; November 10, 1948.

Long Island

"The Complex Frequency Plane as a Tool for the Engineer," by W. H. Huggins, Air Materiel Command; November 17, 1948.

(Continued on page 42A)

SECTION MEETINGS

(Continued from page 40A)

MONMOUTH

"Magnetic Materials with a Rectangular Hysteresis Loop," by E. Both, Signal Corps Engineering Laboratories; September 15, 1948.

"Fundamental Operation and Application of Magnetic Amplifiers," by R. E. Morgan, General Engineering and Consulting Laboratory; September 15, 1948.

"Low Noise IF Input Circuit," by R. T. Adams, Federal Telecommunications Laboratories; October 20, 1948.

"The Transistor," by J. W. McRae, Bell Telephone Laboratories; November 18, 1948.

NORTHERN NEW JERSEY

"Analysis and Performance of Waveguide Hybrid Rings for Microwave," by H. T. Budenbom, Bell Telephone Laboratories; October 20, 1948.

"A Signal Generator for 900–2100 Mc," by G. P. McCouch, Aircraft Radio Corporation; October 20, 1948.

"A Signal Generator for 23,500–24,500 Mc," by P. B. King, Aircraft Radio Corporation; October 20, 1948.

"A Medium-Powered Travelling Wave Amplifier for Microwave," by G. C. Dewey, Federal Telecommunications Laboratories; November 17, 1948.

"The Surge Testing of High Vacuum Tubes," by H. J. Dailey, Westinghouse Electric Corporation; December 8, 1948.

"Application of Radar Techniques to a System for High-Speed X-Ray Movies," by D. C. Dickson, Jr., C. T. Zavales and L. F. Ehrke, Westinghouse Electric Corporation; December 8, 1948.

STUDENT BRANCH MEETINGS

ALABAMA POLYTECHNIC INSTITUTE—IRE BRANCH

"Audio Equipment," by D. K. Briggs, Western Electric Company; December 6, 1948.

UNIVERSITY OF ALBERTA—IRE BRANCH

"Ice Conditions on the Bow River and Iceberg Carrier of World War II," by L. H. Nichols, University of Alberta; November 2, 1948.

"Physics in 1947," by D. B. Scott, University of Alberta; November 23, 1948.

UNIVERSITY OF ARKANSAS—IRE BRANCH

"The Aims of the IRE," by C. E. Cook, Student, University of Arkansas; September 23, 1948.

"The RCA Dynamic Demonstrator," by V. Papoulias, Student, University of Arkansas; October 6, 1948.

CALIFORNIA INSTITUTE OF TECHNOLOGY—IRE

"Telemetering of Guided Missiles," by F. Lehan, Jet Propulsion Laboratory; December 6, 1948.

COLUMBIA UNIVERSITY—IRE-AIEE BRANCH

"The Mathematical Representation of Human Responses," by J. R. Ragazzini and L. J. O'Neill, Columbia University; October 29, 1948.

(Continued on page 44A)

STUDENT BRANCH
MEETINGS

(Continued from page 42A)

UNIVERSITY OF FLORIDA—IRE-AIEE BRANCH

"The Electric Eel," by J. Payne, Student, University of Florida; November 11, 1948.

"An Electronic Vapor Detector," by H. S Nielsen, Student, University of Florida; November 11, 1948.

"Rain and Radar in Ohio," by H. A. Owen, University of Florida; December 2, 1948.

IOWA STATE COLLEGE—IRE-AIEE BRANCH

"New Horizons in Communications," by J. O. Perrine, American Telephone and Telegraph Company; November 3, 1948.

"Horizons in Illuminations," by H. L. Logan, Holophane Company; November 17, 1948.

"Digital Computers," by J. V. Atenasoff, Bureau of Naval Ordnance; December 2, 1948.

"Photography and Science," (film) by A. H. Hausrath, Office of Naval Research; December 2, 1948.

STATE UNIVERSITY OF IOWA—IRE BRANCH

"Professional Engineering," by C. Stanley, Stanley Engineering Company; October 27, 1948.

" A New Horizon in Communications," by J. O. Perrine, American Telephone and Telegraph Company; November 8, 1948.

"Status of the IRE and Status of Television," by T. Hunter, Hunter Manufacturing Company; December 1, 1948.

"Use of the Cathode-Ray Oscilloscope by Unskilled Labor," by H. Talmadge, Student, State University of Iowa; December 8, 1948.

"Industrial Applications of the Photoelectric Cell," by Mr. Wilkinson, Student, State University of Iowa; December 8, 1948.

"Cadmium-Nickel Batteries," by J. E. King, Student, State University of Iowa; December 8, 1948.

"Miller Effect on Frequency Modulation," by M. B. Jones, Student, State University of Iowa; December 8, 1948.

"What the Patent Office Means to the Engineer," by D. McMannis, Student, State University of Iowa; December 8, 1949.

KANSAS STATE COLLEGE—IRE-AIEE BRANCH

"Recent Developments in Vacuum Tubes," by W. James, Radio Corporation of America; October 28, 1948.

"The Magnetron and its Applications," by H. Stout, Midwest Research Institute; November 18, 1948.

MICHIGAN STATE COLLEGE—IRE-AIEE BRANCH

"Applications of Electronics in Industry," by G. Chute, General Electric Company; November 17, 1948.

UNIVERSITY OF MINNESOTA—IRE-AIEE BRANCH

"AC Distribution Problems in the Twin Cities Area," by E. Ewold, Northern States Power Company; October 19, 1948.

NEWARK COLLEGE OF ENGINEERING—IRE-AIEE BRANCH

"Television Receivers," by I. D. Olin, Student, Newark College of Engineering; December 3, 1948.

"Television Receivers," Part II, by I. D. Olin, Student, Newark College of Engineering; December 17, 1948.

THE COLLEGE OF THE CITY OF NEW YORK—IRE BRANCH

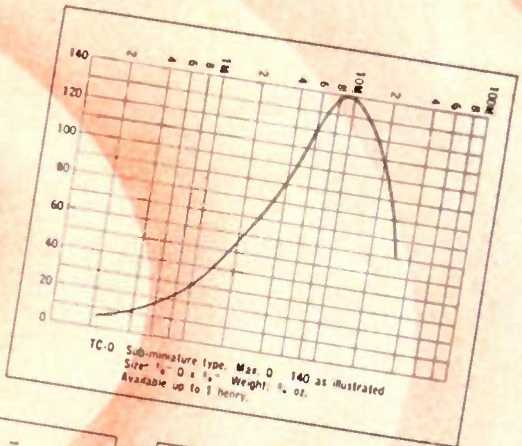
"Nuclear Accelerators," by S. Schneider, Student, College of the City of New York; October 26, 1948.

(Continued on page 46A)

BURNELL & CO., A LEADER IN THE DEVELOPMENT OF MINIATURE AND SUB-MINIATURE COILS AND FILTERS.. INTRODUCES.. THE 'WEDDING RINGS' WORLD'S SMALLEST HIGH 'Q'

TOROIDAL COILS

ACTUAL SIZE

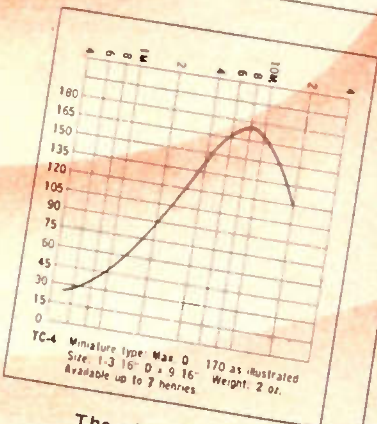


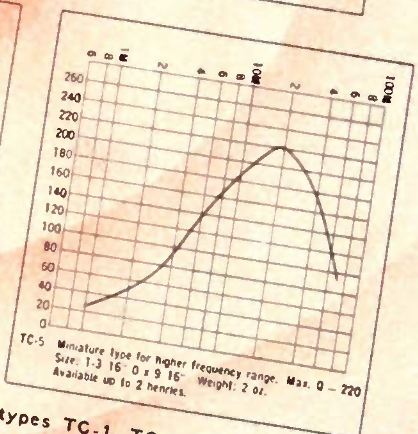
Since discovering that toroidal coils are the solution to problems in compactness of communication and control equipment, design engineers have been confronted with the ever pressing problem of miniaturization.

A major step towards a solution has now been found and we take pleasure in presenting to the electronics field, the penultimate in the design of miniature high Q coils, the types TC-4 and TC-5 and the ultimate, sub-miniature TC-0 which is not much larger than a thumb nail.

For many applications, design engineers will benefit from our specialized experience in the manufacture of miniature components and filters by utilizing the performance capabilities, degressively, of their larger associates (Types TC-1, TC-2, TC-3) but compressed into midget proportions, permitting an elegant solution to the importunate problem of miniaturization.

All of the TC series are toroidally wound on molybdenum permalloy cores providing high Q with a stability unattainable by any other material.





The characteristics of the types TC-1, TC-2 and TC-3 are available upon request.

EXCLUSIVE MANUFACTURERS OF COMMUNICATIONS NETWORK COMPONENTS

ALL INQUIRIES WILL BE PROMPTLY HANDLED

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YONKERS 2, NEW YORK
CABLE ADDRESS "BURNELL"

WRITE FOR TECHNICAL INFORMATION

STUDENT BRANCH
MEETINGS

(Continued from page 44A)

"Magnetic Tape Recording for the Home," by N. M. Haynes, Amplifier Corporation of America; November 16, 1948.

"Intermodulation Measuring Equipment," by C. R. Keith, Western Electric Company; November 30, 1948.

New York University—IRE Branch

"The Geiger Counter and its Applications," by M. Youdin, Amperex Corporation of America; November 9, 1948.

"An Electron Tube for Viewing Magnetic Fields," by S. G. Lutz, New York University; December 3, 1948.

North Carolina State College—IRE Branch

"Ionospheric Wave Propagation," by A. G. McNish, National Bureau of Standards; October 27, 1948.

"Radio Receivers, Panadapters, Wire and Tape Recorders," by S. H. Kahn, Southeastern Radio Company; November 10, 1948.

"Recent Developments in FM Broadcasting," by J. D. Bivens, Radio Engineering Laboratories, Inc., December 8, 1948.

University of North Dakota—IRE-AIEE Branch

"The Role Lignite Can Play in the Development of the E.E.," by A. C. Burr, United States Bureau of Mines; December 8, 1948.

Northwestern University—IRE-AIEE Branch

"Scientific Research in the Field of Neuropsychiatry," by W. D. McCulloch, University of Illinois; November 16, 1948.

University of Notre Dame—IRE-AIEE Branch

"The Heat Pump," by E. A. Ambrose, American Gas and Electric Service Corporation; November 16, 1948.

"The Scope of the Engineering Problem," by G. E. Messer, General Electric Company; December 9, 1948.

"Television Program," by E. L. Clark, Radio Corporation of America; December 14, 1948.

Oregon State College—IRE Branch

"Some Design Problems of a High Gain, Wide-Band, Direct-Coupled Oscilloscope," by L. Belleville, Tektronix Laboratories; November 22, 1948.

"Intertoll Dialing," by H. Beckendorf, Pacific Telephone and Telegraph Company; November 30, 1948.

Pratt Institute—IRE Branch

"Facsimile Transmission," by K. Woloschok Associated Press; November 18, 1948.

"Cold Cathode Tubes and Applications," by W. Holden, Bell Telephone Laboratories, Inc.; December 16, 1948.

Purdue University—IRE Branch

"Popular Misconceptions of Electrical Engineering Principles," by W. Richter, Allis Chalmers Manufacturing Company; December 1, 1948.

Rutgers University—IRE-AIEE Branch

"Research at Rutgers University," by Dr. Dunnington, Rutgers University; October 12, 1948.

University of Texas—IRE-AIEE Branch

"A Glimpse of the Industrial Future—Now 45,000 Kw are Generated Completely Out-of-Doors," by C. W. Geue, Texas Electric Company; November 1, 1948.

"Distributing the Power," by O. S. Hockaday, Texas Electric Company; November 1, 1948.

(Continued on page 48A)

Student Branch

(Continued from page 46A)

"Engineering Research," by A. W. Stralton University of Texas; November 15, 1948.

Utah State Agricultural College—IRE Branch

"Properties of Dielectrics at Ultra-High Frequencies," by W. L. Emery, Utah University; October 27, 1948.

"Practical Television and Programming Problems," by J. Baldwin, Station KDYL; December 1, 1948.

University of Virginia—IRE-AIEE Branch

"Two-Meter Transceiver Operation," by R. F. Stone, Student, University of Virginia; December 14, 1948.

Wayne University—IRE-AIEE Branch

"The Organization and Functioning of an FM Station," by W. Fahringer, Station WJR; November 4, 1948.

Worcester Polytechnic Institute—IRE-AIEE Branch

"The Dynamic Noise Suppressor Amplifier," by H. H. Scott, H. H. Scott Company; October 23, 1948.

"Engineering Experiences in the Orient," by W. Leavitt, Worcester County Electric Company; November 23, 1948.

MEMBERSHIP

The following transfers and admissions were approved and will be effective as of February 1, 1949:

Transfer to Senior Member

Anderson, R. S., 392 Colebourne Rd., Rochester 9, N. Y.
Auerbacher, W. F., 444 E. 21 St., Brooklyn 26, N. Y.
Avins, J., 30 Daniel Low Ter., Staten Island 1, N. Y.
Barlow, D. S., Bell Telephone Laboratories, Inc., 463 West St., New York 14, N. Y.
Battison, J. H., B52 Hudson View Gardens, Pinehurst Ave. & 183 St., New York 33, N. Y.
Borgeson, C. A., 82 Garden Rd., Red Bank, N. J.
Brown, J. S., 1658 W. 101 Pl., Chicago 43, Ill.
Bruene, W. B., 1239 First Ave., S.E., Cedar Rapids, Iowa
Chittick, K. A., 518 Maple Ave., Haddonfield, N. J.
Collins, A. A., Collins Radio Company, Cedar Rapids, Iowa
Creamer, W. J., 331 Center St., Bangor, Me.
Dolnick, A. L., Sylvania Electric Products, 40-22 Lawrence St., Flushing, L. I., N. Y.
Glaser, M., 226-07—137 Ave., Laurelton 13, L. I., N. Y.
Green, J. W., Jr., Professor of Electricity, Unitnt States Military Academy, West Poied, N. Y.
Jacobson, R. I., 120 Bennett Ave., New York 33, N. Y.
Kidd, W. E., 585 Turner Ave., Glen Ellyn, Ill.
Kozanowski, H. N., 435 Washington Ter., Audubon, N. J.
Mair, A. D., c/o Arabian American Oil Company, Dhahran, Saudi Arabia
Manley, J. M., Bell Telephone Laboratories, Murray Hill, N. J.
MacKenzie, A. A., 245 Poplar Ave., Hackensack, N. J.
Morrison, W. C., Random Rd., R.D. 1, Princeton, N. J.

(Continued on page 62A)

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

ELECTRICAL ENGINEER

Nationally known Chicago company is in need of high grade, experienced electrical engineer. Must be a college graduate with either a B.S. or E.E. degree, with high scholastic record. Should have from 2 to 5 years experience in electronics, preferably with a minimum of two years in the design of audio amplifiers. In reply give all particulars and state expected salary. Address Box 549.

DESIGN AND DEVELOPMENT ENGINEER

Design and development engineer to take care of engineering and development of receiving antennae and associated equipment. U.H.F. experience desirable. Upstate New York manufacturer. Reply giving age and qualifications to Box 450.

SALES ENGINEER

Several territories open east of Rocky Mountains, experienced sales engineer representative capable of selling and installing FM two-way radiotelephone systems for mobile operations. Nationally advertised product. Exceptional opportunity for right man. Send detailed qualifications, education, past experience and territory desired. Radiotelephone operators license, 1st or 2nd class preferred. Must have had previous experience in FM radiotelephone. Reply Box 451.

SCIENTISTS AND ENGINEERS

Wanted for interesting and professionally challenging research and advanced development in the fields of microwaves, radar, gyroscopes, servomechanisms, instrumentation, computers, and general electronics. Scientific or engineering degrees required. Salary commensurate with experience and ability. Direct inquiry to Manager, Engineering Personnel, Bell Aircraft Corporation, P.O. Box 1, Buffalo 5, N.Y.

AERO DYNAMICIST ENGINEERS

Aero dynamicist engineers wanted to work on the design of Analog computers, to simulate the flight characteristics of specific aeroplanes. 3 years experience in stability and control essential. Knowledge of servomechanisms dynamics of free flight and applied mathematics desirable. Apply in person, or submit resume to Personnel Dept. Curtis Wright Corp., Propellor Division, Route 6, Caldwell Township, New Jersey.

ELECTRICAL ENGINEER

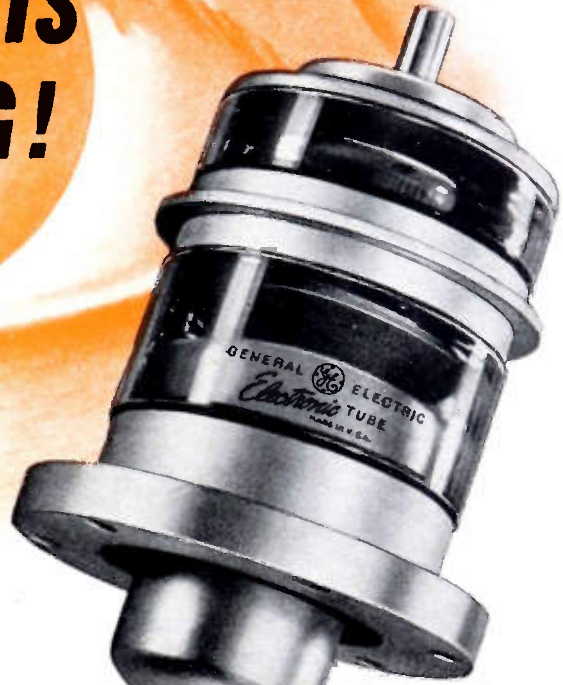
Opening for a man who has the ability to teach advanced electronic circuits and theory, frequency modulation and television. Should eventually teach course in ultra-high frequency techniques and electric wave phenomena and pulse systems. Reply to Director of Academic Administration, 1020 North Broadway, Milwaukee 2, Wisconsin.

(Continued on page 52A)

For driving stage

GL-5513

THE EYE IS EXACTING!



To assure high video-signal quality, specify G. E.'s pace-setting 220-mc triodes in your new TV transmitter design!

For output stage

GL-9C24

THE BEST, most up-to-date . . . these words must describe every unit of your television circuit IF you mean to get off on the right foot competitively. *So start by choosing General Electric ring-seal tubes.* Designed directly for grounded-grid circuits, they plug in quickly, firmly, with wide contact areas. Lead inductance is extremely low. The tubes need minimum neutralization.

All external parts are silver-plated to reduce r-f losses. Fernico metal-to-glass seals are used throughout. Sturdy, compact, built to true precision standards, Types GL-5513 and GL-9C24 are acknowledged performance leaders in the TV and FM fields.

Study the ratings of these modern yet service-proved v-h-f triodes; then phone your nearby G-E electronics office for further facts, plus (if desired) the application counsel of an experienced G-E tube engineer. Or, wire or write *Electronics Department, General Electric Company, Schenectady 5, New York.*

RATINGS AND ELECTRICAL CHARACTERISTICS

	GL-5513	GL-9C24
Filament voltage	6.3 v	6.3 v
Filament current	32 amp	240 amp
Interelectrode capacitances:		
Grid-filament	21.1 mmfd	24 mmfd
Grid-plate	8.7 mmfd	15.7 mmfd
Plate-filament	.11 mmfd	0.47 mmfd
Type of cooling	forced-air	water and forced-air
Plate ratings per tube, Class B r-f power amplifier (video service, synchronizing peak conditions):		
Max voltage	3,000 v	5,000 v
Max current	1.2 amp	2 amp
Max input	3,300 w	10 kw
Max dissipation	1,200 w	5 kw
*Power output, typical operation	1,160 w	3.4 kw
Plate ratings per tube, Class C r-f power amplifier (key-down conditions without modulation):		
Max voltage	4,000 v	6,500 v
Max current	1 amp	2 amp
Max input	3,600 w	12 kw
Max dissipation	1,200 w	5 kw
*Power output, typical operation	2.45 kw	9 kw

*Includes power transferred from driver to output of grounded-grid amplifier.

GENERAL ELECTRIC

180-H2-8850

FIRST AND GREATEST NAME IN ELECTRONICS

POSITIONS OPEN

(Continued from page 50A)

SOUND ENGINEER

For large radio and television manufacturer in Chicago. Experience in loud speaker design, audio circuits, and acoustics necessary. Fine opportunity for advancement. Please write giving full particulars. Box 552.

PROFESSOR

Professor of communications engineering needed for fall 1949 by southeastern university. Will be in charge of graduate work and research activities. $6,000.00 for nine months with extra income for summer teaching. Must have Ph.D. or D.Sc. degree. Write Box 553

ENGINEERS—PHYSICISTS

Southwestern state college engaged in research, development and operation in guided missile field has openings for graduate electrical engineers or physicists with experience in electro-mechanical devices, telemetering, audio, RF, VHF and antenna design. Salary depends on education and experience which should be fully described in first letter. Box 554.

COMMUNICATIONS—ELECTRONICS ENGINEER

Work will involve aeronautical communications and aircraft electronic projects. Some development work but primarily application engineering. Box 555.

ENGINEERS AND PHYSICISTS

An expanded program of research has created opportunities in the fields of acoustics, hydrodynamics, electronics, mechanics and mathematics. Positions flexible enough to permit part-time college teaching if desired. Opportunity for graduate study. Salary commensurate with experience. Appointments carry academic rank ranging from Research Assistant to Professor. Inquiry invited. P.O. Box 30, State College, Pa.

ELECTRICAL AND ELECTRONICS ENGINEERS

The Engineer Research and Development Laboratories of the Corps of Engineers, located at Fort Belvoir, Virginia, has a continuing need for electrical and electronics engineers to fill positions involving design and development and paying entrance salaries, depending on grade, of from $3727.20 to $6235.00 per annum. Applications should be mailed to the Civilian Personnel Office, Bldg. 211, Fort Belvoir, Virginia.

ELECTRICAL ENGINEERS

1 experienced engineer wanted for each of the following: radar pulse circuit; circuit development; radar modulators; radar devices; radar receivers; radar indicator circuit design and development. Send resume in confidence to Box 556.

ELECTRONICS ENGINEER MANAGER

Openings in equipment Development Department of Electronics Division, Boston, Mass. for graduate engineer with ability to administer and direct an engineering activity. An electrical engineer specialized in communications or electronics preferred, with industrial or equivalent in radar or communications systems design. Responsible for development and manufacture of microwave X

(Continued on page 54A)

WANTED PHYSICISTS ENGINEERS

Engineering laboratory of precision instrument manufacturer has interesting opportunities for graduate engineers with research, design and/or development experience on radio communication systems, Servomechanisms (closed loop), electronic & mechanical aeronautical navigation instruments and ultra-high frequency & microwave technique.

WRITE FULL DETAILS TO *EMPLOYMENT SECTION*

SPERRY GYROSCOPE COMPANY

DIVISION OF SPERRY CORP.
Marcus Ave. & Lakeville Rd.
Lake Success, L.I.

Electronic Engineers

BENDIX RADIO DIVISION
Baltimore, Maryland
manufacturer of
RADIO AND RADAR EQUIPMENT
requires:

PROJECT ENGINEERS

Five or more years experience in the design and development, for production, of major components in radio and radar equipment.

ASSISTANT PROJECT ENGINEERS

Two or more years experience in the development, for production, of components in radio and radar equipment. Capable of designing components under supervision of project engineer.

Well equipped laboratories in modern radio plant . . . Excellent opportunity . . . advancement on individual merit.

Baltimore Has Adequate Housing

Arrangements will be made to contact personally all applicants who submit satisfactory resumes. Send resume to Mr. John Siena:

BENDIX RADIO DIVISION
BENDIX AVIATION CORPORATION
Baltimore 4, Maryland

POSITIONS OPEN

(Continued from page 52A)

& K band radar systems, FM test equipment, industrial and laboratory test equipment, and electronic computers. Boston or New York interviews will be arranged for qualified applicants submitting complete resume including salary requirements to Supervisor of Employment, Sylvania Electric Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

MAGNETRONS DEVELOPMENT ENGINEER

An opening at our Electronics Division, Boston, Mass., for a Senior Engineer, to work in the tube development engineering group as a design and process engineer on magnetrons. Will be in charge of a project and have one or more engineers and technicians reporting to him. Will be responsible for manufacturing the tube after it has been designed in small quantities. B.S. in E.E. or physics required, and two to three years experience in the use of vacuum systems, and considerable experience on microwave type metal tube assembly. Furnish full particulars in writing regarding age, education, experience, present salary and salary requirements to: Supervisor of Employment, Sylvania Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

(Continued on page 56A)

Radio and Radar Development and Design Engineers

Openings for experienced men at

HAZELTINE ELECTRONICS CORPORATION
Little Neck, L.I., N.Y.

Please furnish complete resume of experience with salary expected to:
Director of Engineering Personnel
(All inquiries treated confidentially)

ENGINEERS – ELECTRONIC

Senior and Junior, outstanding opportunity, progressive company. Forward complete résumés giving education, experience and salary requirements to

Personnel Department
MELPAR, INC.
452 Swann Avenue
Alexandria, Virginia

ZENITH RADIO CORPORATION needs Research and Development Engineers and Physicists

Project, senior and junior engineers and physicists are required for prosecution of several very interesting developments. Openings exist for men experienced in various aspects of radio and television receiver and transmitter development, in radar, and in applications of electronics to ordnance problems. *Salaries commensurate with experience and ability.*

WRITE: **Director of Research**
Zenith Radio Corp.
6001 W. Dickens Ave.
Chicago 39, Illinois

One Relay or 100,000 . . .

CLARE RELAYS

Are *Each* Carefully Tested Against Your Specifications

Clare's Chief Inspector reports only to the president. He is responsible for the most important thing in our business . . . that the Clare Relay you order is *exactly* the Clare Relay you receive.

Whether your order is for one Clare Relay or 100,000 . . . every single relay is 100% inspected and tested against your specifications.

Mechanical adjustment, electrical characteristics, physical appearance and construction . . . every detail is gone over by experts specially trained in Clare's precise requirements. Not a single relay is sealed for packing until it conforms in every way with the highest standards of relay performance.

Such infinite capacity for taking pains is a basic reason for Clare's leadership in the industrial relay field. It accentuates the value of Clare's superior design, precise manufacture and unusual understanding of difficult relay design problems.

Clare sales engineers, fully experienced in every type of relay requirement, are located in principal cities for your convenience. If you have a relay problem that seems really tough . . . look to Clare. Enjoy the services of this organization whose entire business is devoted to making sure that you have the relay which best meets *your* needs.

Look in your classified telephone directory . . . or write: C. P. Clare & Co., 4719 West Sunnyside Ave., Chicago 30, Illinois. In Canada: Canadian Line Materials, Ltd., Toronto 13. Cable Address: CLARELAY.

Every Single CLARE Relay Must Pass These Tests

For Mechanical Adjustment

1. Contact Pressures (Make or Break)
2. Contact Follow, or Wipe
3. Sequence of Make and Break Contacts
4. Correct Airline
5. Residual Setting
6. Spring Straightness

Physical Inspection

1. Plating (For Marks or Scratches)
2. Proper Insulation
3. Condition of Insulation (No cracks, etc.)
4. Tapping of Screw Holes
5. Spring Thickness
6. Coil Data on Label and Condition of Label

Electrical Inspection

1. Coil Resistance
2. Coil Breakdown
3. Pileup Breakdown
4. Operation (as specified)
5. Direction of Winding
6. Test for Shorted Turns

All Clare Relays Are Packed Immediately Following Inspection

CLARE RELAYS

First in the Industrial Field

Positions Open

(Continued from page 54A)

ELECTRONICS ENGINEER SECTION HEAD

Our Electronics Division, Boston, Mass., is seeking a graduate electrical engineer (S.B. or S.M. electronics option preferred) to supervise a group of 17 engineers and several technicians working on X & K band radar and other electronic equipment development, including computers. Early interviews will be granted qualified applicants with actual experience in design of large scale electronics equipment and administration of engineers. Mail complete resume to Supervisor of Employment, Sylvania Electric Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

ELECTRONICS ENGINEERS

Top flight engineers. Must have 10 years design and development experience on servomechanisms and amplifiers, circuits and equipment layout. Apply in person or submit complete resume to Personnel Dept., Curtis Wright Corp., Propeller Division, Route 6, Caldwell Township, New Jersey.

ELECTRICAL ENGINEER

Nationally known Chicago company is in need of a high grade, experienced electrical engineer. He must be a college graduate with a B.S. in E.E. and have a high scholastic record. Should have from two to five years experience in electronics, preferably with a minimum of two years in the design of audio amplifiers. In reply give all particulars and state expected salary. Box 558.

SENIOR RADIO ENGINEERS

Must have good fundamental training, experience in circuit design and test of government-type portable and mobile transmitters and receivers. Write fully to Government Contract Dept., Pilot Radio Corporation, 3706-36th Street, Long Island, City, N.Y.

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcemen twithout assignment of reason.

PHYSICIST

Twenty-seven year old graduate physicist and mathematician with experience in control circuits desires either foreign or domestic employment. Box 197 W.

(Continued on page 58A)

Postions Wanted

(Continued from page 56A)

BROADCAST ENGINEER

Nine years experience. Now employed as assistant chief engineer 5 KW AM 50 KW FM, major network station, directional antennas. Desires chief engineer position or transmitter supervisor in east. Best references. All offers considered. Box 199 W.

ENGINEERING—ADVERTISING

Hard-hitting advertising executive who can talk an engineer's language.

A.B., M.A., plus 3 years electrical engineering. Last 4 years with top electrical corporation creating sales campaigns, advertisements, displays, sales aids, technical literature. Desires connection as advertising manager or assistant sales manager for medium sized New York City area electrical concern. Box 200 W.

JUNIOR ENGINEER

R.C.A. Institute graduate desires position in production, design or allied fields in New York City vicinity. Age 27. Married. 1 year laboratory and 2 years flight (radio) experience. First class radiotelephone license. Call Da. 6-7203 or write Box 201 W.

TELEVISION ENGINEER

B.S.T.E. Age. 25. Married. First class radiotelephone license. Desires position in television station or development work. Trained in operation and maintenance of R.C.A. Image Orthicon, DuMont equipment and very high frequency techniques. Box 205 W.

JUNIOR ENGINEER

Graduate 2 year course television engineering. Married. Age 25. First class FCC license. Trained in all phases of television studio work. Desires position in television broadcasting field. Box 206 W.

JUNIOR ENGINEER

Syracuse University. B.E.E. June 1948. Age 26. Married with no children. Desires work with power company or motor manufacturing company located in the east. Prefer training program if possible. Interested in transmission and mathematical design. Box 207 W.

ENGINEER

University of Minnesota, communications major. B.E.E. with distinction, August 1948. 2½ years electronics experience in U.S. Army. Desires position in production or electronic development. Will work anywhere in U.S. Box 208 W.

ELECTRONICS ENGINEER

Will graduate March 1949 Iowa State College B.S.E.E. in communications. Married. Age 25. First class Radio Telephone license. Some servicing experience. Desires position in radio or electronics anywhere in U.S. Box 209 W.

ENGINEER

B.S.E.E. Northeastern University, Boston 1947. Two years experience as Navy radio technician with Navy radar and communication equipments. Valuable experience in UHF antenna and radiation research and development at Naval Air Test Center, Patuxent River, Maryland. Desires position in research and design of antennae. Box 210 W.

(Continued on page 60A)

MICROWAVE PLUMBING

10 CENTIMETER

WAVEGUIDE DIR. COUPLER, 27 db. Navy type CABV1-47AAN, with 4 in. slotted section $42.50

ADAPTER, SQ. FLANGE TO RD CHOKE, 18 in. long OA 1½ in x 3 in guide, type "N" output and samp. probe as shown .$32.00

"S" BAND CRYSTAL MOUNT, gold plated, with 2 type "N" connectors $12.50

MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated $45.00

10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or dc arranged switching motor. Mfg. Raytheon. New and complete $150.00

721-A TR CAVITY WITH TUBE. Complete with tuning plungers $12.50

POWER SPLITTER: 726 Klystron input, dual "N" output $5.00

10 CM. McNALLY CAVITY Type SG $3.50

WAVEGUIDE SECTION, MC 445A, rt. angle bend, 5½" ft. OA, 8" slotted section $21.00

10 CM OSC. PICKUP LOOP, with male Homedell output $2.00

TS115/APS-2F 10 CM ANTENNA in lucite ball, with type "N" fitting $4.50

OAJ NAVY TYPE CYT66ADL, ANTENNA, in lucite ball, with Sperry fitting $4.50

10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola $8.00

10 CM END FIRE POLYRODS ....$1.75 ea.

"S" BAND Mixer Assembly, with crystal mount, pick-up loop, tunable output $3.00

⅞" RIGID COAX—⅜" I.C.

SLOTTED SECTION, 10 ft. L. 4" Slot $8.50

RIGHT ANGLE BEND, with flexible coax output pick-up loop $8.00

SHORT RIGHT ANGLE bend, with pressurizing nipple $4.00

RIGID COAX to flex coax connector $3.50

STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length $7.00

RT. ANGLES for above $3.75

⅞" COAX, ROTARY JOINT $8.00

RT. ANGLE BEND 15" L. OA $4.50

FLEXIBLE SECTION, 15" L. Male to female $4.25

MAGNETRON COUPLING to ⅞" rigid coax, with TR pickup loop, gold plated $7.50

⅞" RIGID COAX, Bead Supported ....$1.20 per ft.

⅞" RIGID COAX—¼" I.C.

SHORT RIGHT ANGLE BEND $3.00

Rotating joint, with deck mounting $6.00

RIGID COAX slotted section CU-60/AP $5.00

3 CM. PLUMBING

(STD. 1" x ½" GUIDE UNLESS OTHERWISE SPECIIED)

"X" BAND PREAMPLIFIER, consisting of 2 723A/B local oscillator-beacon feeding waveguide and TR/ATR Duplexer section, including 60 mc. IF amp. $47.50

WAVEGUIDE RUN, 1⅛" x ½" guide, consisting of 4 ft. section with rt. angle bend on one end and 2". 45 deg. bend other end $8.00

WAVEGUIDE SECTION, 1⅛" x ½" choke to choke, 4 ft. long $6.75

DUMMY LOAD, TS 332/UP $22.50

"X" BAND calibrated attenuator $85.00

SHIELDED KLYSTRON tube mounts with rough attenuator outputs $90.00

CU 105/APS 31 Directional coupler 25 db. $15.00

CU 106 APS—33 Directional coupler, 25 db. . $15.00

CG 176/AP Directional coupler, 20 db. $18.00

Flexible waveguide ....$4.00 per ft.

THERMISTOR: D 164699 —for mtg in "X" Band Guide $2.50

45 DEG. TWIST, 6" Long $8.50

12" SECTION, 45 deg. twist, 90 deg. bend $6.00

11" STRAIGHT WAVEGUIDE section choke to cover. Special heavy construction silver plated $4.50

15 DEG. BEND, 10" choke to cover $4.50

5 FT. SECTIONS, choke to cover, silver plated $11.50

18" FLEXIBLE SECTION $17.50

"E" or "H" PLANE BEND $12.50

BULKHEAD FEED THRU $15.00

WAVEGUIDE SECTION, CG 251/APS 15A, 26" long choke to cover, with 180 deg. bend of 2¼" rad at one end $6.00

ROTARY JOINT with slotted section and type "N" output pickup $6.50

WAVEGUIDE SECTION, 12" long choke to cover, 45 deg twist & 2½" radius, 90 deg bend $5.00

SLUG, TUNER/ATTENUATOR, W.E. guide, gold plated $6.50

TR/ATR DUPLEXER section with iris flange $8.00

TWIST, 5" 90 deg choke to cover $5.00

WAVEGUIDE SECTIONS 2½ ft long, silver plated, with choke flange $5.75

WAVEGUIDE, 90 deg bend E plane, 18" long $5.00

ROTARY JOINT, choke to choke $6.00

ROTARY JOINT, choke to choke, with deck mounting $6.00

S-CURVE WAVEGUIDE, 8" long cover to choke $3.50

3" FLEX SECTION, Sq flange to circ. flange adapter $7.50

"X" BAND WAVEGUIDE, 1¼" x ⅝" OD, 1/16" wall aluminum .... per ft. $ .75

TR CAVITY for 724-A TR tube $3.50

724-A TR tube (1I TR 1) $2.50

APS-15 DUPLEXER SECTION using 1B24 $10.00

3 CM WAVEGUIDE, 1" x ½" I.D per ft $1.50

CIRCULAR CHOKE FLANGES, solid brass .. $.55

SQ. FLANGES, flat brass ....$.55

"T" SECTION (TR-ATR) choke to choke, supplied with circ or sq flanges $4.50

"X" BAND PRESS. GAUGE SECTIONS, with 15 lbs gauge and press. nipple $18.50

TRANSMISSION LINE PRESS GAUGE $3.50

2" FLEXIBLE SECTION, cover to cover $5.00

MOUNTING SECTION for Absorp Wavemeter and Thermistor Mount, Silver Plated, Sq Flanges $6.50

1.25 CENTIMETER

MITRED ELBOW cover to cover $4.00

TR/ATR SECTION choke to cover $4.00

FLEXIBLE SECTION 1" choke to choke $5.00

KBAND Rotary joint $45.00

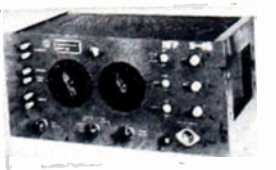
ADAPTER, rd cover to sq. cover $5.00

MITRED ELBOW and 8 sections choke to cover $4.50

"K" BAND FEEDBACK TO PARABOLA HORN, with pressurized window $30.00

PULSE EQUIPMENT

MODULATOR UNIT BC 1203-B



Provides 200-4,000 PPS. Sweep time: 100 to 2,500 microsec. in 4 steps. fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse, sweep voltages, calibration voltages, fil. voltages. Operates 115 vac, 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio #804B or #804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New as shown $125.00

MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 144 KW (12 KV at 12 amp). Duty Ratio: .001 max. Pulse duration: 5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-72's, 1-73. New ..$110.00

APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro Sec. Rep. rate 624 to 1348 Pps. Pk. pwr. out 35 KW. Energy 0.018 Joules $49.00

TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50

APS-10 MODULATOR DECK. Complete, less tubes $75.00

APS-10 Low voltage power supply, less tubes $18.50

PULSE NETWORKS

G.E. #25E5-1-350-50P2T, 25 KV, 5 sections, "E" circuit, 1 microsecond pulse length, 350 PPS, 50 ohms impedance $45.00

G.E. #6E3-5-2000-50P2T, 6KV, "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance $6.50

G.E. #3E (3-.84-810; 8-2.24-405) 50P4T: 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec, 810 PPS, 50 ohms imp.; Unit 2, 8 Sections, 2.24 Microsec, 405 PPS, 50 ohms imp. $6.50

7.5E3-1-200-67P, 7.5 KV, "E" Circuit, 1 microsec, 200 PPS, 67 ohms impedance, 3 sections $7.50

7.5E4-16-60-67P, 7.5 KV, "E" circuit, 4 sections, 16 microsec, 60 PPS, 67 ohms impedance $15.00

7.5E3-3-200-6PT, 7.5 KV, "E" Circuit, 3 microsec, 200 PPS, 67 ohms imp., 3 sections $12.50

DELAY LINES

D 168184: .5 microsec, up to 2000 PPS, 1800 ohm term ..$4.00

D-170499: .25/.50/.75/. microsec, 8 KV, 50 ohms imp. ..$16.50

D 165997: 1¼ microsec. $7.50

PULSE TRANSFORMERS

G.E.K.-2745 ....$39,50

G.E.K.-2744-A $39.50

W.E. #D166173 HI-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms, Freq. range: 10 kc to 2 mc, 2 sections parallel connected, potted in oil $12.00

W.E. KS 9800 Input transformer, Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1, Frequency range: 380-520 c.p.s. Permalloy core $2.00

G.E. #K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms, Sec. Imp: 450 Ohms, Pulse Width: 1 Microsec, Pri. Input: 9.5 KV PK, Sec. Output: 28 KV PK, Peak Output: 800 KW, Bifilar 2.75 Amp. $19.50

W.E. #D169271 Hi Volt input pulse Transformer $9.95

G.E. K2450A Will receive 13KV, 4 micro second pulse on pri., secondary delivers 14KV, Peak power out 100KW GE..$15.00

G.E. #K2748A Pulse Input, line to magnetron $12.00

#9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy 3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1¼" 19/32 ....$1.50

INVERTERS

PE 218: Input: 25-28 vdc, 92 amps. Output: 115 v. 280-500 cy. 1500 va. New, export packed (as shown) $49.95
Slightly used, ex. cond. $25.00

PE 206: Input: 28 vdc: Output: 80 v. 800 cy. 500 VA New, export packed $12,50

GE MOD 5D21NJ3A: Input 27 vdc, 35 amp: Output 115 v. 400 cy, 485 VA. New $49.95

MICROWAVE TEST EQUIPMENT

THERMISTOR BRIDGE: Power meter I-203-A. 10 cm mfg W.E. Complete with meter, interpolation chart, portable carrying case $72.50

Bell Labs Dual Mount mixer beacon assemblies, 2 complete mixer beacon mounts on gold-plated waveguide section $50.00

Slotted Line, Bell Labs, 1⅛" x 5/16" guide, gold plated $150.00

TS-238 GP. 10 cm. Echo box with resonance indicator and micrometer adjust cavity. 2700 to 2900 Mcs calibrated $85.00

TS 108 AP dummy load $65.00

W. E. I 138. Signal generator. 2700 to 2900 Mc range Lighthouse tube oscillator with attenuator & output meter. 115 VAC input, reg. Pwr. supply. With circuit diagram $50.00

3 cm. Wavemeter: 9200 to 11,000 mc transmission type with square flanges $15.00

3 cm. stabilizer cavity, transmission type $20.00

3 cm. Wavemeter, Micrometer head mounted on X-Band guide Freq. range approx. 7900 to 10,000 Mc $75.00

COAX CABLE

RG 17/U 52 ohms imp. ....$.48/Ft.

RG 57/U Twin Cond. 95 ohms ....$.55/Ft.

RG 18/U, 52 ohm im. armored ....$.51/Ft.

RG 23/U, twin coax, 125 ohm imp armored..$.50/Ft.

RG 28/U 50 ohm imp. pulse cable Corona min. starting voltage 17 KV ....$.50/Ft.

RG 35/U 70 ohm imp. armored ....$.50/Ft.

SETS IN STOCK

- **SE** (new)
- **SF** (new)
- **SG** (new)
- **SN** (new)
- **SO-1** (used)
- **SO-13** (used)
- **SQ** (used)
- **CPN-6** (unused)
- **APS-3** (used)
- **APS-4** (used & new)
- **APS-15** (near comp.)
- **QBG-1** (new)
- **TBM** (used)
- **TDE** (used)
- **RAK-7** (new)
- **TBK-19** (new)

Send for add'l info.

MICROWAVE ANTENNAS

AN MPG-1 Antenna. Rotary feed type high speed scanner antenna assembly, including horn parabolic reflector. Less internal mechanisms, 10 deg. sector scan. Approx. 12'L x 4'W x 3'H. Unused. (Gov't Cost—$4500.00) $250.00

APS-4 3 cm. antenna. Complete. 14½" dish. Cutler feed dipole directional coupler, all standard 1" x ½" waveguide. Drive motor and gear mechanisms for horizontal and vertical scan. New, complete $65.00

AN/TPS-3. Parabolic dish type reflector approx. 10' diam. Extremely lightweight construction. New, in 3 carrying cases $89.50

RELAY SYSTEM PARABOLIC REFLECTORS: approx. range: 2000 to 6000 mc. Dimensions: 4¼' x 3', rectangle, new $85.00

TDY "JAM" RADAR ROTATING ANTENNA, 10 cm. 30 deg. beam, 115 v.a.c. drive. New $100.00

SO-13 ANTENNA, 24" dish with feedback dipole 360 deg. rotation, complete with drive motor and selsyn New $120.00
Used $45.00

DBM ANTENNA. Dual, back-to-back parabolas with dipoles. Freq. coverage 1,000-4,500 mc. No drive mechanism $65.00

AS125/APR Cone type receiving antenna. 1000 to 3200 megacycles. New $4.50

140-600 MC. CONE type antenna, complete with 25' sectional steel mast, guys, cables, carrying case, etc. New $49.50

ASD 3 cm. antenna, used, ex. cond. $49.50

MAGNETRONS

TUBE	FREQ. RANGE	PK. PWR. OUT.	PRICE
2J31	2820-2860 mc.	265 KW.	$25.00
2J21-A	9345-9405 mc.	50 KW.	$25.00
2J22	3267-3333 mc.	265 KW.	$25.00
2J26	2992-3019 mc.	275 KW.	$25.00
2J27	2965-2992 mc.	275 KW.	$25.00
2J32	2780-2820 mc.	285 KW.	$25.00
2J38 Pkg.	3249-3263 mc.	5 KW.	$35.00
2J39 Pkg.	3267-3333 mc.	87 KW.	$35.00
2J40	9305-9325 mc.	10 KW.	$65.00
2J49	9000-9160 mc.	58 KW.	$85.00
2J55 Pkg.	9345-9405 mc.	50 KW.	$35.00
2J61	3000-3100 mc.	35 KW.	$65.00
2J62	2914-3010 mc.	35 KW.	$65.00
3J31	24,000 mc.	50 KW.	$55.00
5J30			$39.50
714AY			$25.00
718DY			$25.00
720BY	2800 mc.	1000 KW.	$50.00
720CY			$50.00
725-A	9345-9405 mc.	50 KW.	$25.00
730-A	9345-9405 mc.	50 KW.	$25.00

Klystrons: 723A/B $12.50; 707B W/Cavity $20.00
417A $25.00; 2 K.41 $65.00

MAGNETRON MAGNETS

GAUSS	POLE DIAM.	SPACING	PRICE
4850	⅝ in.	⅝ in.	$12.50
2500	1⅝ in.	1 11/16 in.	$12.50
1500	1⅝ in.	1½ in.	$12.50
D161302*	1⅝ in.	1 5/16 in.	$12.50

* Mfr's Number.

TUNABLE PKGD. "CW" MAGNETRONS

QK 61 2975-3200 mc. QK 62 3150-3375 mc.
QK 60 2800-3025 mc. QK 59 2675-2900 mc.
New, Guaranteed .... Each $65.00

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COMMUNICATIONS EQUIPMENT CO.

131-R Liberty St., New York, N.Y. Digby 9-4124

Positions Wanted

(Continued from page 58A)

JUNIOR ENGINEER

B.S. Television Engineering, American Television Institute of Technology, January 1949. Age 25. Married, no children. Three years experience on Navy radar. Desires position in microwave research. Anywhere in U.S. Box 211 W.

JUNIOR ENGINEER

R.C.A. Institute's graduate seeks position in research and development, or production field in New York area. Age 29. Married. Studying for engineering degree at night. Air Force officer, 2½ years experience as instructor in bomb-sight and auto-pilot theory and operation. Box 212 W.

TEACHER

Candidate for M.A. in physics August 1949. 10 years teaching. High school and Army radio school. Also Army and broadcast radio experience. Age 38 with family. Prefer college teaching in southwest. Box 220 W.

TELEVISION ENGINEER

B.E.E., Age 25, Married. 3 years radar design and maintenance, including 1½ year Navy ETM instructor; 2½ years television development and production. Available June. Prefer California. Box 221 W.

ELECTRONIC ENGINEER

B.E.E. 1943; 3 semesters postgraduate work. 5 years experience in electronic instrumentation in connection with rockets and guided missiles. Also remote guidance and control mechanisms, servos and digital computers. Interested in position involving design and development along similar lines. Box 222 W.

ELECTRONIC PHYSICIST

B.S. and graduate work in physics. 1 year electronic laboratory experience, 1 year full time and 1 year part time additional professional experience plus 4 years Navy radar maintenance. Prefer position in west or mid-west. Box 223 W.

JUNIOR ENGINEER

Graduating in radio engineering, March 1949. B.S. Age 24, married, no children. 3 years naval experience as radio technician, with accent on Airborne radar counter-measures. Desires position in electronic circuit design and development. Will consider foreign position. Box 224 W.

ENGINEER

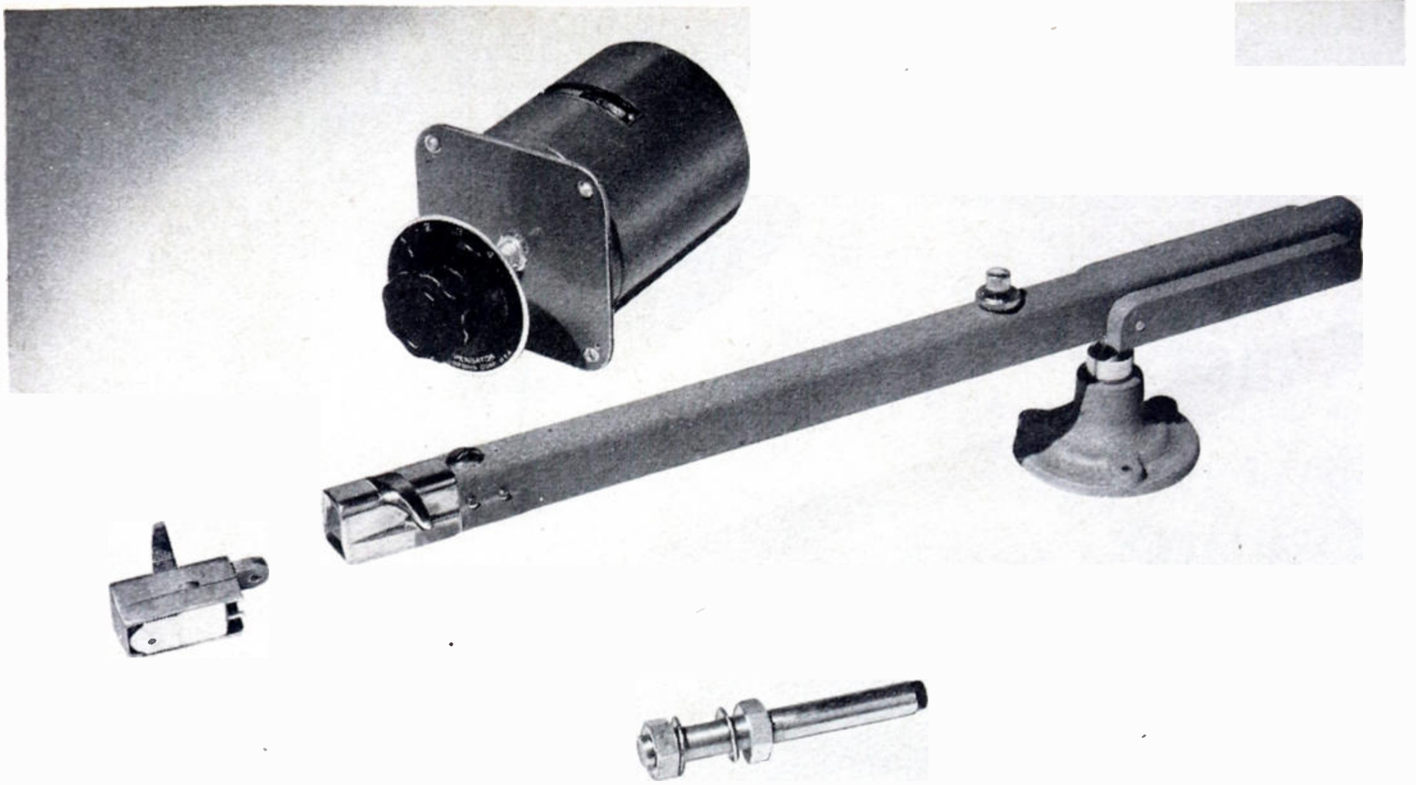
B.S.E.E. University of Wisconsin. 2 years Army communications experience and 1 year experience in electronic and control circuit design. Age 25. Single. Desires position in production or electronic and control circuit development. Box 225 W.

SALES OR FIELD ENGINEER

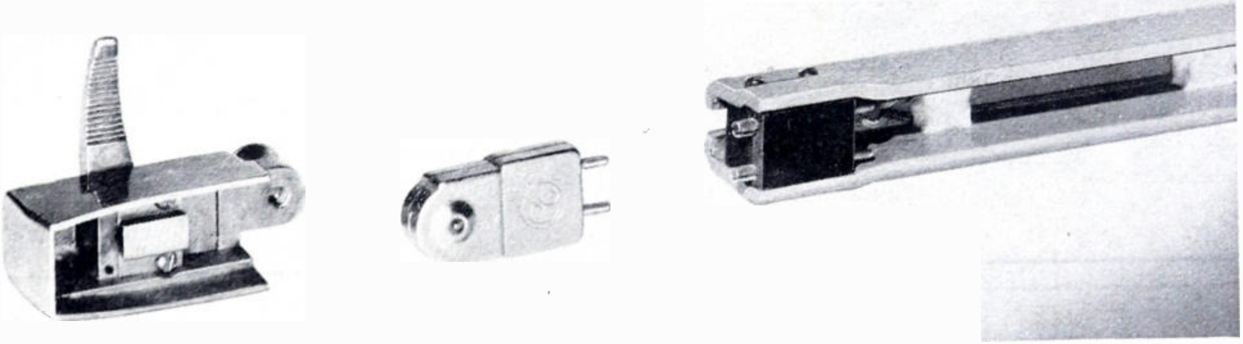
Young, hard hitting research and design engineer seeks inquiries from firms needing addition to sales or field engineering staff. Desires change to position which will allow combination of engineering and top sales talents. Have solid experience in all phases of electronics and communications. Also quality control and teaching experience along with B.E.E. degree. Married, 1 child. Will consider Baltimore, Md. or vicinity. Box 226 W.

SALES ENGINEER

Strong technical background, good sales personality. Desires position in New York City vicinity. Box 227 W.



You've got to be ready for Microgroove!



MICROGROOVE long-playing recordings are here to stay. This means that every broadcast station and recording studio must have quality equipment, especially for microgroove reproduction.

The new PRESTO type 153 reproducers include two separate Pickering diamond stylus heads for microgroove or regular recording, an exceptionally fine arm, and a 4-position compensating network.

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Roe, J. H., 315 Newton Ave., Collingswood, N. J.
Sloughter, G. S., R.D. 4, Ridgefield, Conn.
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Ware, W. H., Institute for Advanced Study, Princeton, N. J.
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Admission to Senior Member

Augustadt, H. W., 903 Boulevard, Westfield, N. J.
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Wersen, D. T., 3561 Military Ave., Los Angeles 34, Calif.
Wooley, R. L., 10 Bradford Ct., Syracuse 7, N. Y.

Admission to Member

Banerjee, S. K., High Frequency Laboratory, Ericsson Telephones, Ltd., Beeston, Nottingham, England
Berger, E. W., 1367 University Ave., Stockton, Calif.
Chamberlin, R. M., WGGG, 1230 Waldo Rd., Gainesville, Fla.
deVoogt, A. H., Radio-Section PTT, 6 Scheveningseweg, The Hague, Holland

(Continued on page 64A)

MEMBERSHIP

(Continued from page 62A)

Griss, C. H., 218 Maple, Perrysburg, Ohio
Haacke, E. M., Electrical News & Engineering, 347 Adelaide St., W., Toronto, Ont., Canada
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Proud, E., c/o Office of the Engineer in Chief, Posts & Telegraphs, Box 581, Nairobi, Kenya, East Africa
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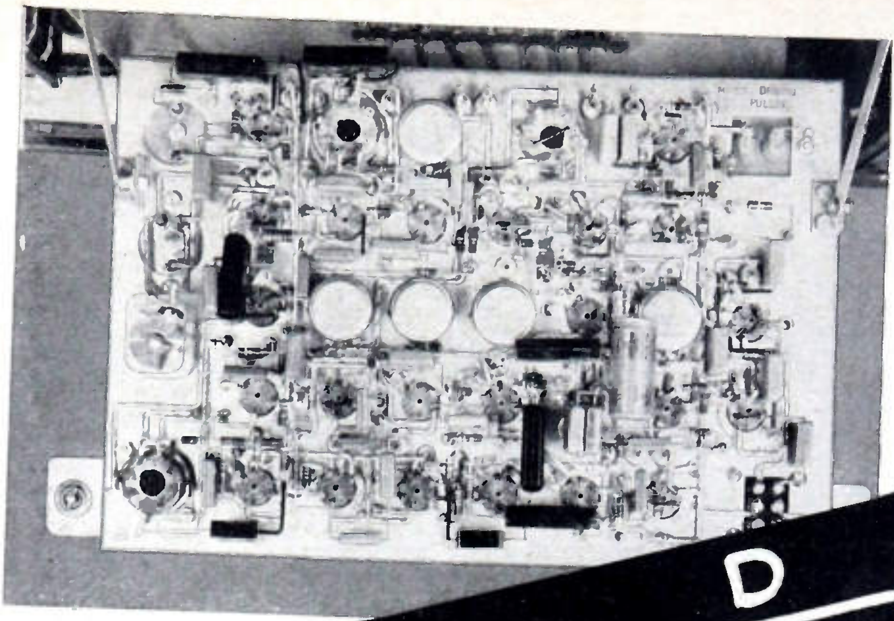
The following admissions to Associate were approved and were effective as of January 1, 1949:

Alfreds, L. J., Lane Head, Elborough Rd., South, Norwood, London, England
Ataman, A., 1306 S. Orchard, Urbana, Ill.
Barnes, O. C., 2735 Walker, Kansas City 2, Kans.
Bellow, F. T., c/o Radio Station WULA, Eufaula, Ala.
Bendett, R. M., 35-16—76 St., Jackson Heights, L. I., N. Y.
Bidell, F. W., East River Rd., Grand Island, N. Y.
Bright, T. F., 123 Lenox Ave., New Milford, N. J.
Bruce, L. G., 1942 S. Carpenter St., Chicago 8, Ill.
Carnagie, V. F., 411 State St., Brooklyn, N. Y.
Carvajal, H. H., 223 Avda. Brasil, Santiago, Chile
Cooper, R. J. G., 11831—96 St., Edmonston, Alta., Canada
Davis, J. C., Jr., 686 Pleasant St., Framingham Center, Mass.
Dunn, F. S., 2433 Ravenwood Ave., Dayton 6, Ohio
Ebling, W. F., Jr., 3466 Braddock St., Philadelphia, Pa.
Flynt, E. R., Box 3666, Georgia Institute of Technology, Atlanta, Ga.
Foss, W. L., 927 15 St., N.W., Washington 5, D. C.
Freeman, H., Mail Station K39, Sperry Gyroscope Company, Great Neck, L. I., N. Y.
Gabor, W. D., 142 Second St., Keyport, N. J.
Gibbon, W. F. H., 704 E. Stepney Pl., Inglewood, Calif.
Gnagey, L. B., 653 S. Miami St., West Milton, Ohio
Gober, R. H., 1430 Glenwood Ave., S.E., Atlanta, Ga.
Graham, N. L., 998 Center Ave., Lancaster, Pa.
Green, R., 2713 N. 34 St., Milwaukee 10, Wis.
Grometstein, A. A., 94 St. Botolph St., Boston 16, Mass.
Haimowitz, B., 1726 N. Paxon St., Philadelphia 31, Pa.
Hammann, W. M., 75 Converse Ave., Malden 48, Mass.
Harvey, D. M., USAF, Box 1346, Wright Field, Dayton, Ohio

(Continued on page 66A)



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PORTABLE SYNC GENERATOR
Type 5030-A

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(Continued from page 64A)

Hoff, I. M., Radio Station WRSW, Times Bldg., Warsaw, Ind.
Howe, R. B., 3539 Gillham Rd., Kansas City, Mo.
Huber, E. H., R.D. 2, Shawnee Rd., North Tonawanda, N. Y.
Humphrey, E.L., 1519 W. Estes Ave., Chicago 26, Ill.
Hunt, D. F., Moore School of Electrical Engineering, 200 S. 33 St., Philadelphia 4, Pa.
Jaffe, L. D., 315 Eastern Pkwy., Brooklyn, N. Y.
Jones, T. W., "Sunnerbroc," Sabot, Va.
Kamal, A. K., Department of Electrical Communication Engineering, Indian Institute of Science, Bangalore 3, India
Kean, D. W., 110A Forrestal St., China Lake, Calif.
Kilpinen, R. M., 507 Fullerton Pkwy., Chicago 14, Ill.
Knoerl, J. L., 806 Beach Rd., Buffalo 21, N. Y.
Kovalevski, N. N., 892 Prospect Ave., New York 59, N. Y.
Kramer, E. R., Harris-Magruder, Inc., 1835 K St., N.W., Washington 6, D. C.
Krocke, W. F., 513-C E. South St., Angola, Ind.
Lange, W. G., R.D. 1, Sanborn, N. Y.
Loewen, K. L., 915 N. Eleventh, Manhattan, Kan.
Maertz, W. A., 4723 Kenwood Ave., Chicago 15, Ill.
Mardiks, D., 3900 W. 53 St., Kansas City, Kan.
McAllister, J. A., USN, Sur-AS-Dev-Det., Key West, Fla.
McCool, W. A., 5609 Randolph St., Hyattsville, Md.
Mendenhall, W. S., Jr., 439 Garden Rd., Columbus 12, Ohio
Mitra, S. C., 158/2B Harish Mukherjee Rd., Calcutta 25, India
Mruk, W. F., Jr., 119 Ella St., McKees Rocks, Pa.
Murugasu, V., Ponnalai East, Chulipuram Post, Jaffna, Ceylon
Nagai, H., 936 E. 42 Pl., Chicago 15, Ill.
Neumark, J., T.R. Laboratory, American Television, Inc., 5050 N. Broadway, Chicago 40, Ill.
Newton, R. L., Radio Station WCON, The Atlanta Constitution, Atlanta 2, Ga.
Ogletree, W. A., 616 College Ter., Williamsburg, Va.
Owings, J. L., 503 Baird Rd., Merion Station, Pa.
Pewitt, B. D., 411 Ninth Ave., Haddon Heights, N. J.
Rajaratnam, T., 34 Venkatesa Naick St., George Town, Madras, India
Ransom, A. W., 4429 Duneden Ave., Deer Park, Cincinnati, Ohio
Ransome, R. L., 722 West 30, Houston 18, Tex.
Reeves, P. E., Box 231, Vapor Recovery Systems Company, Compton, Calif.
Roberts, J. M., 5927 S.E. 23 Ave., Portland 2, Ore.
Rozas, F. N., 1612 S. St. Sndrews Pl., Los Angeles 6, Calif.
Saunderson, J. J. G., c/o Central Radio Maintenance, Bryan St., Ballymena, North Ireland
Scholfield, W. D., 179 Bessborough Dr., Toronto, Ont., Canada
Schultz, D. G., 507 Fullerton Pkwy., Chicago 14, Ill.
Schumacher, H. W., 3280 N. Tenth St., Milwaukee 6, Wis.
Seetharam, H. R. B., c/o H. Ramanna, The Mysore Silk Filatures, Ltd., Mysore, Mysore State, India
Sesha-Iyer, C. S., 6 Saraswat Colony, Poona 2, Bombay Prov., India
Silverstein, M. M., 825 E. 49 St., Chicago 15, Ill.
Smith, R. W., Jr., Room 4000, Michelson Laboratory, China Lake, Calif.
Somayajulu, Y. V., Physics Department, Andhra University, Waltair, India
Sotton, V. O., c/o Philippine Consulate, 40 Exchange Pl., New York, N. Y.

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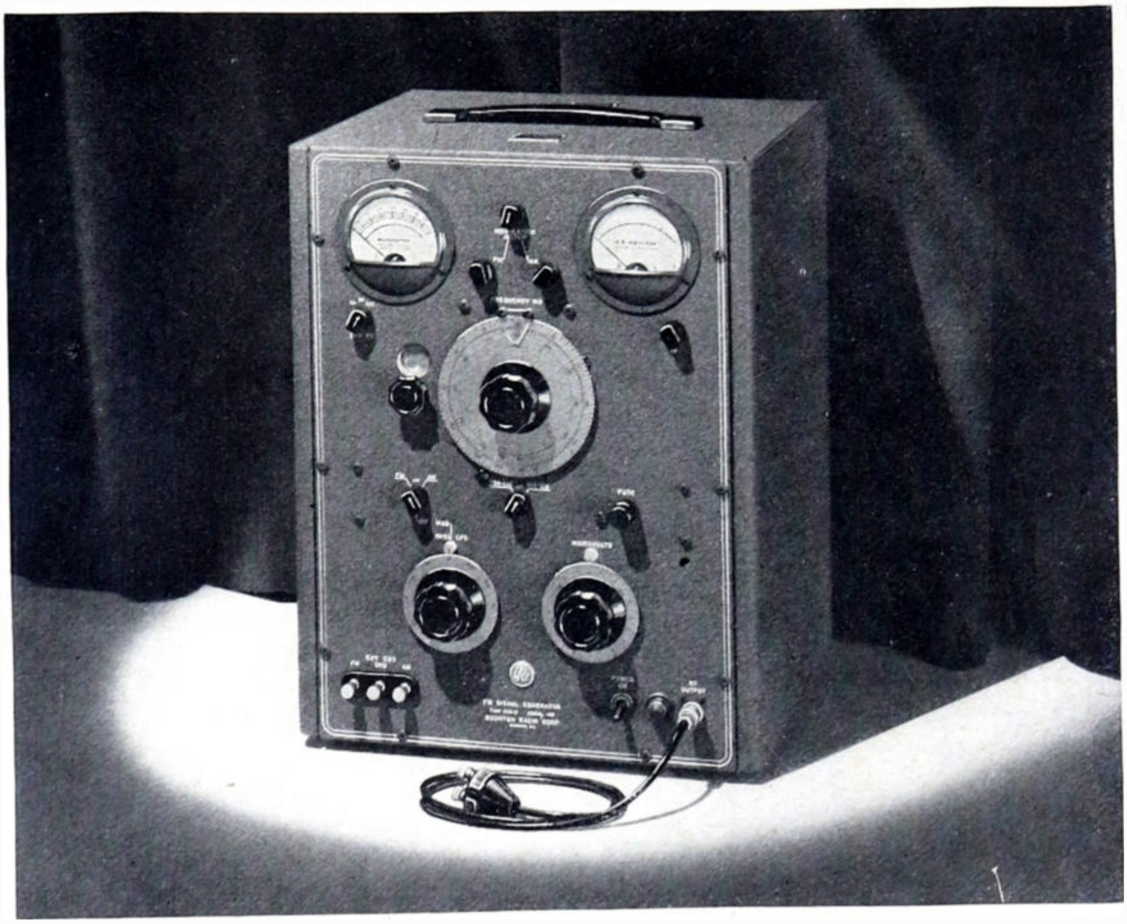

Membership

(Continued from page 66A)

Spiro, J., 605 W. 170 St., New York 32, N. Y.
Stello, P. E., 3619 W. 102 St., Inglewood, Calif.
Weedie, D. A., 608 S.E. 18 St., Portland 14, Ore.
Wahl, R. E., 2031 W. 61 Terrace, Kansas City, Mo.
Whalen, T. J., Jr., 162 Highland Ave., Buffalo 9, N. Y.
White, E. A., Baldwin Manor, Apt. C-11-4, Aberdeen, Md.
Wofford, J. H., 4317-A Montrose Blvd., Houston 6, Tex.
Zentner, J. E., 127 E. 73 St., New York, N. Y.
Zucker, M. J., 4910 N. Karlov Ave., Chicago 30, Ill.

The following transfers to Associate grade were approved to be effective as of January 1, 1949:

Anderson, T. N., 9 Cottage Lane, Huntington Station, N. Y.
Ariessohn, E. R., 96-11—65 Rd., Forest Hills, L. I., N. Y.
Aucremanne, M. J., 18 Whitney Ave., Floral Park, L. I. N. Y.
Baer, T. M., 875 West End Ave., New York 25, N. Y.
Brorein, W. J., Southern Bell Telephone & Telegraph Company, Division Engineer Office, 1630 Lynch Bldg., Jacksonville, Fla.
Brown, G. L., 416 West St., Stillwater, Okla.
Brown, I., 835 N. Seventh St., Camden, N. J.
Buerger, E. J., 325 Mohawk Ave., Scotia 2, N. Y.
Carpenter, J. W., 207 E. Second St., Austin, Tex.
Catterınole, G. B., ComAirLant Staff, Naval Air Station, Norfolk 11, Va.
Cerino, R., 278 First Ave., New York 9, N. Y.
Conroy, G. J., 2644 Glenmawr Ave., Pittsburgh 4, Pa.
Cummings, K. C., 2212 Aldrich Ave. S., Minneapolis 5, Minn.
Dawson, E. B., Savannah Armature Works, Inc., 126 W. Bay St., Savannah, Ga.
Dazey, M. H., 933 Richmond St., El Cerrito, Calif.
Dennison, P. R., 1539 Maple St., Pasadena 5, Calif.
Emmons, A. W., 1308 S. Ridgewood Ave., Daytona Beach, Fla.
Evans, W. E., Jr., 2281—12 Ave., Sacramento Calif.
Fee, D. R., 125 Main St., Orange, N. J.
Frisch, F. W., 418 Kelby St., Fort Lee, N. J.
Gartzke, F. J., 1812 Morningside Dr., Iowa City, Iowa
Gellert, J., 181 Lenox Rd., Brooklyn 26, N. Y.
Greenhill, C. W., 1144—22 St., San Diego 2, Calif.
Grim, W. M., Jr., Massachusetts Institute of Technology, Servo Laboratory Bldg. 32, Cambridge 39, Mass.
Hegde, K. S., Department of Tele-Communications, College of Engineering, Madras 15, India
Hofheimer, R. W., 5 W. 86 St., New York 24, N. Y.
Jesurun, M. M., 1263 Cole St., Columbus 5, Ohio
Keyes, W. R., Jr., 4021 Greenwood Ave., Oakland 2, Calif.
Kingsley, B., 968 Bronx Park South, New York 60, N. Y.
Knight, S. J., Jr., 3040 E. Exposition Ave., Denver 9, Col.
Kramer, S. I., 2340 Valentine Ave., New York 57, N. Y.
Krilanovich, N. J., 725 Fern Pl., N.W., Washington 12, D. C.
La Rosa, R., 267 Bedell Ter., West Hempstead, L. I., N. Y.
Lowell, R., Naval Research Laboratory, Code 3422, Washington 20, D. C.
Martin, F. C., Jr., U. S. Navy Electronics Laboratory, Code 413 C, Bldg. 128, San Diego 52, Calif.
McBrian, J. E., 33 Oldwood Rd., Port Washington, L. I., N. Y.
McFarlin, F. E., 600 S. 14 St., Ponca City, Okla.
Mendel, F. S., 1579 S. 60 St., Milwaukee 14, Wis.
Menes, M., 68 W. 107 St., New York 25, N. Y.
Palm, H. A., Jr., 5250 N. Spaulding Ave., Chicago 25, Ill.
Paschetto, D. D., 4301—56 St., Woodside, L. I., N. Y.

(Continued from page 70A)

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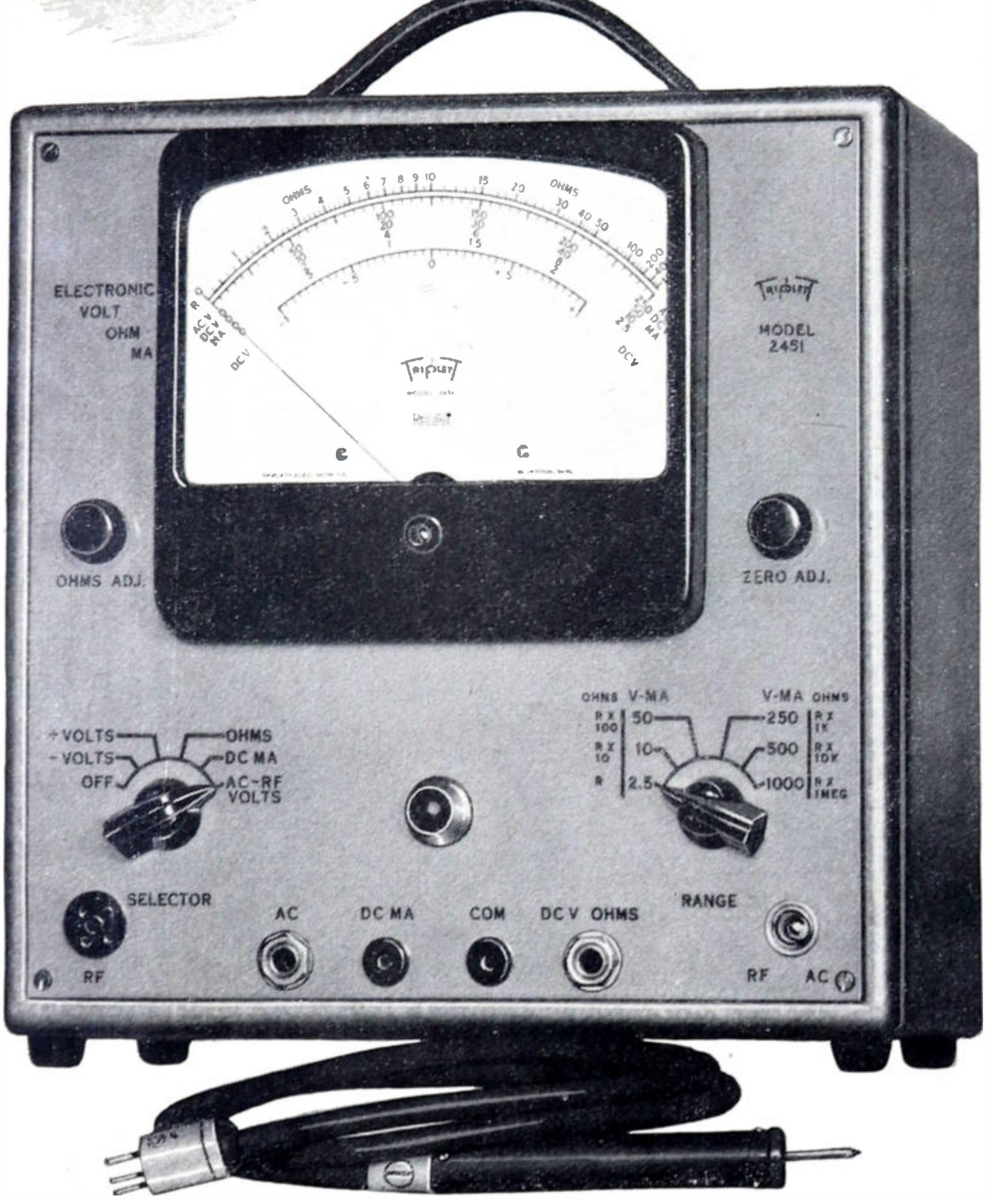
ELECTRONIC VOLT-OHM-MILLIAMMETER



MODEL 2451
DEALER NET ONLY

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This new vacuum tube Volt-Ohm-Milliammeter incorporates features previously found only in the most expensive equipment.

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D.C. MILLIAMPERES:	0-2.5-10-50-250-500-1000
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(Plugs in same socket without change of connection . . . frequencies up to 400 MC.)

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Perry, M. A., 565 Crown St., Brooklyn 13, N. Y.
Pughe, E. W., Jr., 3828 Brookside Rd., Toledo 6, Ohio
Reddeck, J. G., RCA Laboratories, Princeton, N. J.
Robinson, K. W., 33 Vandeventer Ave., Princeton, N. J.
Robison, W. C., Box 464 Mount Ayr, Iowa
Rosenberg, J. M., 1962 Hoover Ave., Oakland 2, Calif.
Sackler, A. A., c/o Ebasco Services Inc., Rm. 1356, 2 Rector St., New York 6, N. Y.
Scarbrough, A. D., Ave. F & 6 St., Yucaipa, Calif.
Schmitzer, R. W., 2014 E. Windsor Pl., Milwaukee 2, Wis.
Schwarzkopf, J., 51 Verdun Ave., New Rochelle, N. Y.
Seaward, P. E., U. S. Naval Electronics Laboratory, Bldg. 328, San Diego 52, Calif.
Sencenbaugh, D. W., Staff, Commander Submarine, Atlantic Fleet, New London, Conn.
Simmons, J. M., Jr., Box 966, Seagraves, Tex.
Smith, R. A., 989 Cliff Dr., Laguna Beach, Calif.
Sumpter, C. B., Jr., 4373 Fort Ave., Lynchburg, Va.
Sweet, R. G., 2035 Kenilworth Ave., Los Angeles 26, Calif.
Taber, J. E., 824 Doan Dr., Burbank, Calif.
Teeple, R. P., 1481 E. Rock Springs Rd., N.E., Apt. 1, Atlanta, Ga.
Tobin, M. W., 7 Stuyvesant Oval, Apt. 7-A, New York 9, N. Y.
Tuttle, F. M., 585 Page Ave., N.E., Atlanta, Ga.
Underwood, J. F., 1972 Welton Ave., New York 53, N. Y.
Van Nice, R. I., 5212 Forbes St., Pittsburgh 17, Pa.
Vreeland, R. W., 36 Magee Ave., Mill Valley, Calif.
Wade, O., 791 E. Washington St., Pasadena 6, Calif.
Waldron, T. J., 67 Richmond Hill, New Canaan, Conn.
Walker, B., 354 Ocean Ave., Brooklyn 26, N. Y.
Weber, J. W., 402 E. Walnut, Kokomo, Ind.
Weise, D. M., 2014 E. Genesse St., Syracuse, N. Y.
Wilson, R. M., Electronic Research & Development, Ansco, Binghamton, N. Y.
Zellmer, N. A., 701 W. Broadway, Brownfield, Tex.

The following transfers to Associate grade were approved to be effective as of December 1, 1948:

Berl, S., 409 Throop Ave., Brooklyn 21, N. Y.
Brearley, H. C., Jr., 703 Fountain Pl., Burlington, N. C.
Conley, P., Westinghouse Research Laboratory, East Pittsburgh, Pa.
Daggett, N. L., 44 Bay View St., Hyannis, Mass.
Garcia, A., 837 E. Mulberry St., Evansville, Ind.
McGill, D. R., 14612 Summit Dr., Edmonton, Alta, Canada
Y9kelson, B. J., 2301 Kings Highway, Brooklyn 29, N. Y.

The following admissions to Associate grade were approved to be effective as of December 1, 1948:

Appuhamy, F. A., "San Jose," Dankotvwa, Ceylon
Ardelian, T. E., Fur Trade Department (Radio), Hudson's Bay Co., Winnipeg, Mani., Canada
Barnett, H. A., Jr., Radio Station WIST, Charlotte 2, N. C.
Beyer, A. R., 2019 Dunstan, Holiston 5, Tex.
Christopher, O., Box 68, Air Base Branch, Belleville, Ill.
Cooper, L. V., 1930 AACS, APO 731, c/o Postmaster, Seattle, Wash.
Cromlich, R. M., Box 171, New Baden, Ill.
Da Prato, B. F., 505 E. Broad St., Angola, Ind.
Dickens, J. A., E. 56 & Easton Blvd., Des Moines 17, Iowa
Dickinson, I. E., McClatchy Broadcasting Co., 7 & Eye Sts., Sacramento, Calif.
Elam, F. E., 1345 Madison St., N.W., Washington 11, D. C.

(Continued on page 72A)

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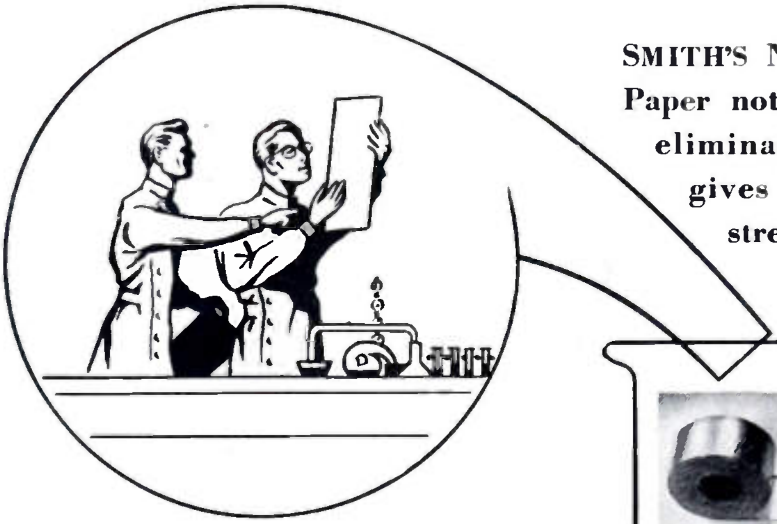
SMITH'S New-type Metallized Condenser Paper not only reduces space factor by eliminating the use of electrodes, but gives longer life, higher dielectric strength, and ends forever the disastrous effects of breakdowns.



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It's here! Today . . . now . . . Smith Paper, Inc. makes available its new Metallized Condenser Paper that bids fair to revolutionize the entire electric and electronic field. Smith Metallized Condenser Paper makes possible the first one-layer condenser with an .0003" dielectric material. It also makes possible a 75% saving in space factor over most conventional capacitors. And because of its self-healing characteristics it almost completely eliminates the factor of conducting particles and the usual serious effect of a breakdown.

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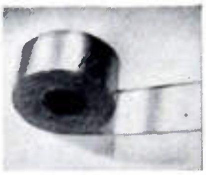
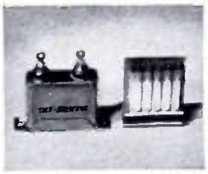
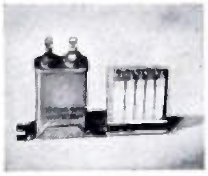


Metallized Paper



1 MFD—400 WVDC



4 MFD—150 WVDC

Capacitor samples courtesy Solar Manufacturing Corp., North Bergen, N. J.

Self-healing on breakdowns. Another of the outstanding properties of metallized paper is its capacity to self-heal on a breakdown. In other words, a capacitor wound with metallized paper may be brought to a breakdown voltage, but the effects of the breakdown are sufficient to cause a re-insulation around the breakdown area so that the capacitor is satisfactory for continued use. Numerous breakdowns do not appear to impair this self-healing characteristic. Smith Paper, by taking advantage of this property, is able to furnish a metallized paper devoid of particles which are conducting at the usual test voltages.

Automatically Controlled. Today equipment has been perfected for automatically controlling and continuously recording (where necessary) such properties as lacquer thickness and consistency, thickness of the metal layer, color and resistance of the metal layer, width of margin, etc. — all important characteristics of the product, the close control of which is essential for the best design and manufacture of metallized paper capacitors.

Complete facts available. All facts on Smith Metallized Condenser Paper as it applies to your industry may be had on request. Simply address Smith Paper, Inc., Lee, Massachusetts. There is no obligation.

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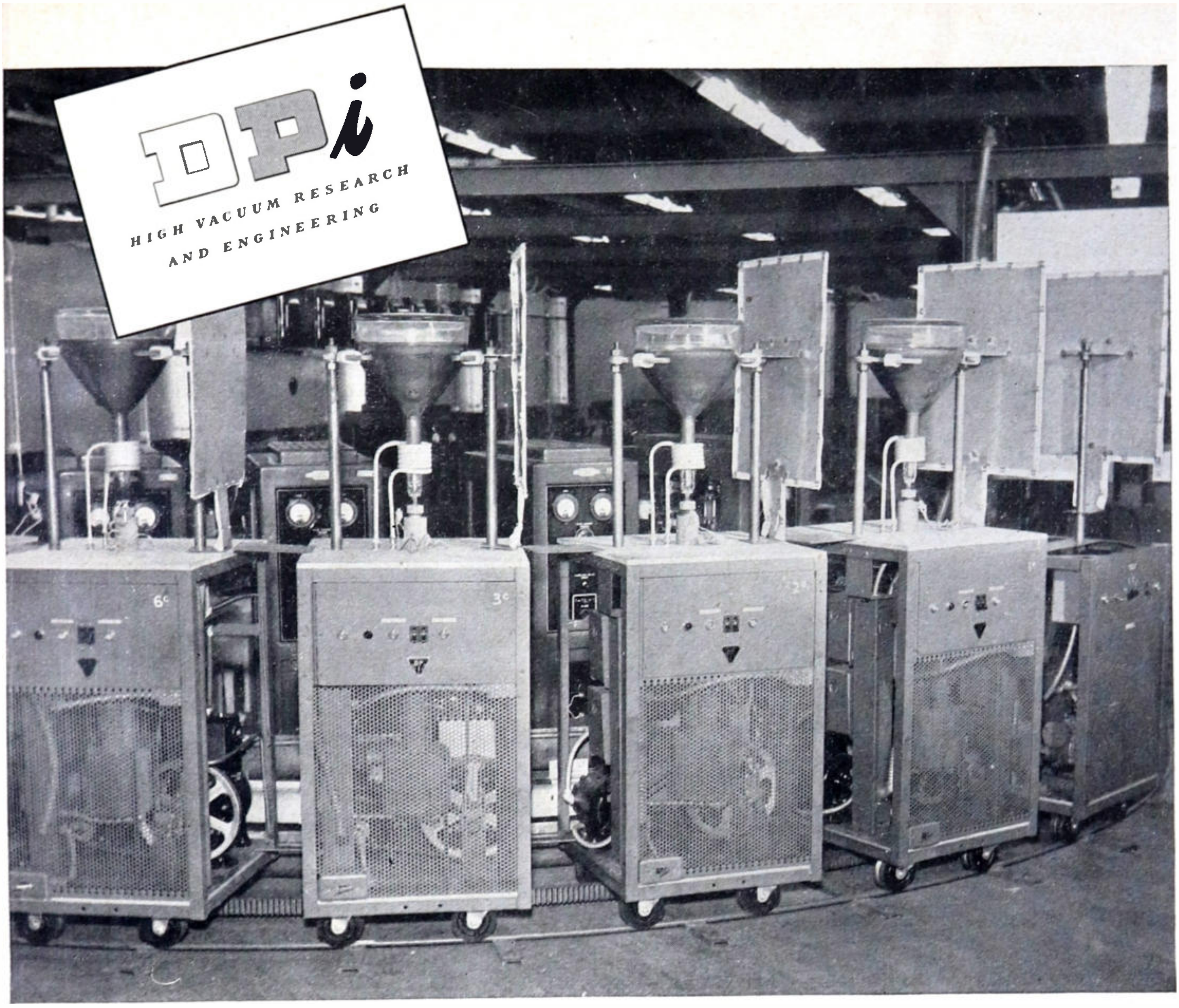
Freeman, G. A., Jr., 2117 Ave. L, Galveston, Tex.
Galbraith, A., 43 Crestmont Rd., Bangor, Me.
Graham, J. M., Rm. 390, General Electric Company, Schenectady, N. Y.
Isaacs, S. J., 2043 Holland Ave., New York 60, N. Y.
Kelley, N. J., 123 Ontario St., Toledo 2, Ohio
Kothe, E. W., 68 Prince St., Hastings-on-Hudson, N. Y.
Kovner, H., 400 E. 57 St., New York 22, N. Y.
Lorenz, L. H., 1250 Wakefield Dr., Houston 18, Tex
Margulius, M. P., Acoyte 143-VII, Buenos Aires, Argentina
Massie, H. B., 819 S. Lawndale Ave., Chicago 24, Ill.
McCarrell, S. G., 8925 Bishop St., Chicago, Ill.
McNamara, T. R., 254-31—84 Rd., Floral Park, L. I., N. Y.
Padilla, D. J., R-217—D-13 St., Sandia Base, Albuquerque, N. Mex.
Perlman, I. B., 28 Windermere Dr., Bronxville Hgts., Yonkers, N. Y.
Perlman, A., 160 Beach 82 St., Rockaway Beach, L. I., N. Y.
Reedy, J. F., 4652 Kenmore Ave., Chicago 40, Ill.
Roberts, J. T., Barnard Hall, North Brother Island, New York 54, N. Y.
Rynder, L. S., 2372 Franklin Ave., Toledo 10, Ohio
Shackelford, W. T., III, 8727 Loch Bend Dr., Towson 4, Md.
Sterne, H. J., 7071 N. Wilbur St., Portland 3, Ore.
Tompson, R. N., Department of Mathematics, University of Nevada, Reno, Nev.
Warner, G. A., 249-37—51 St., Little Neck, L. I., N. Y.
Whitehurst, C. H., Jr., 201 N. Highland Ave., Winter Garden, Fla.
Wofford, W. D., Jr., 1502 Elmen, Houston 19, Tex.
Wolfe, R. E., 9306 Harford Rd., Baltimore 14, Md.



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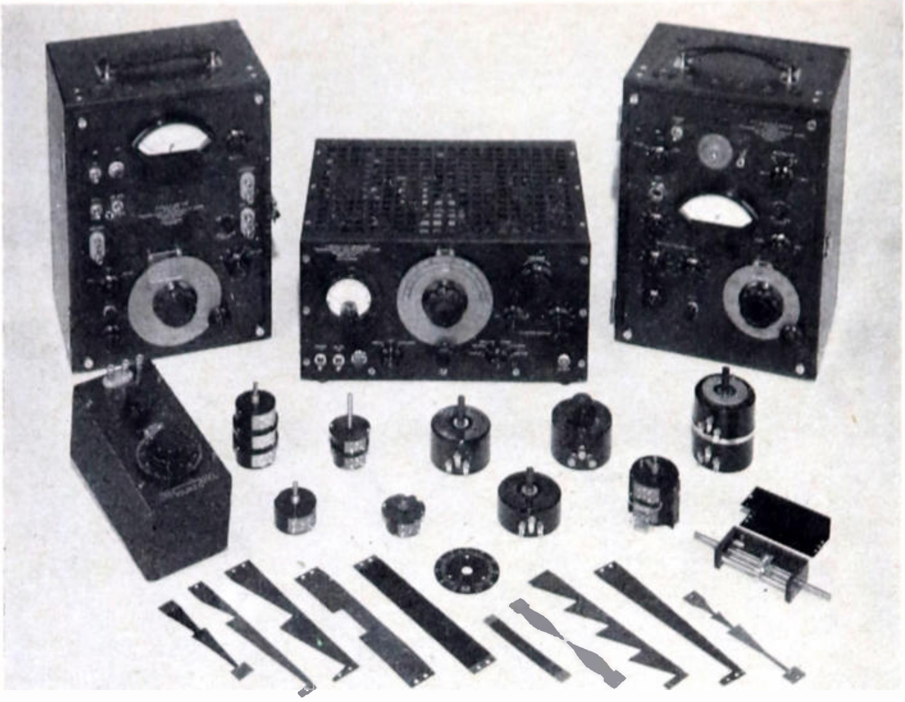


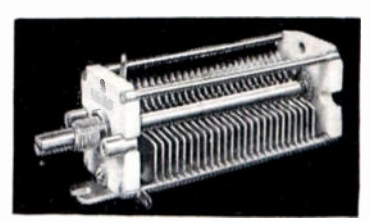
What to SEE at the 1949 Radio Engineering Show

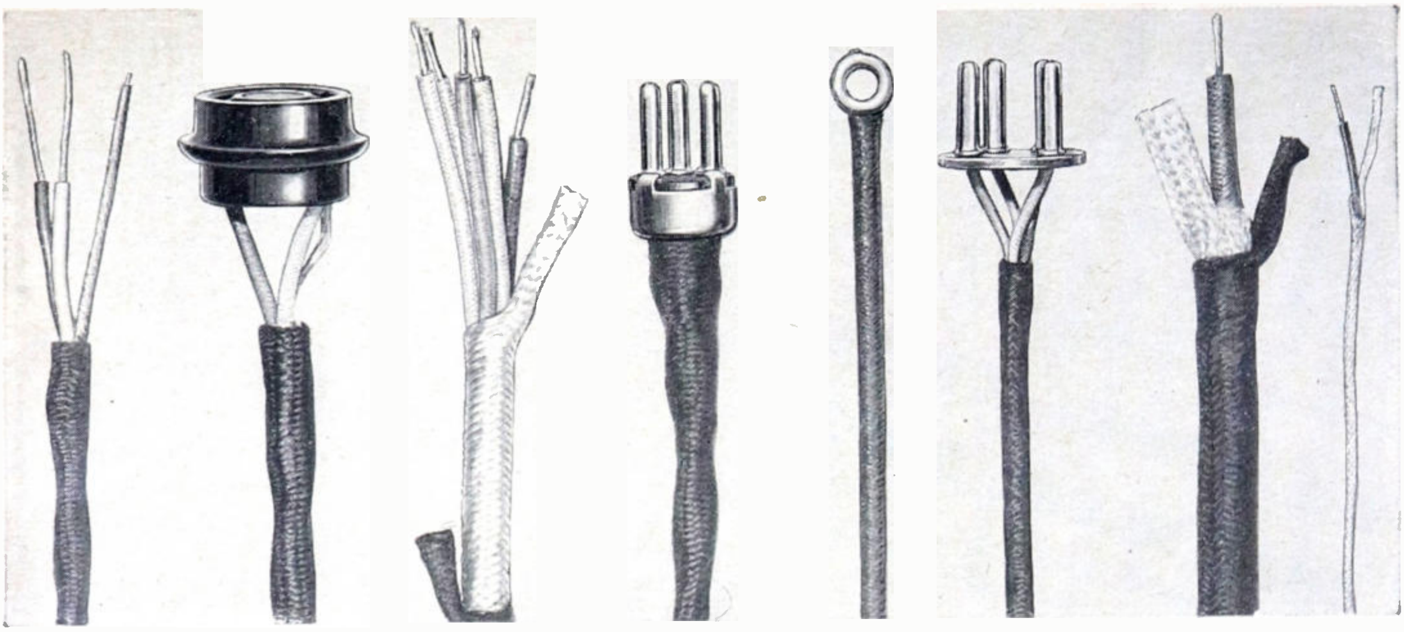
Directory of 200 Exhibits

(Continued from page 22A)

Firm	Booth
Glaser Lead Co., Inc., Brooklyn, N. Y. Solders and fluxes, lectron rosin core solders.	260
Gray Research & Development Co., Hartford, Conn. Variable speed turntables, sound effects turntable, transcription, turntables, New LP tone arm, transcription tone arms.	X
Graybar Elec. Co., Inc., New York, N. Y. Distributing Machlett and Western Electric vacuum tubes.	96 & 97
Haydu Brothers, Plainfield, N. J. Small metal stampings, precision machined parts, drawn and pressed parts, screw machine parts, formed wire parts, gas, oxygen and air burners.	211
Helipot Corp., So. Pasadena, Calif. Helipot potentiometers, Beckman Duodials, Continuous rotation potentiometers, Multi-tap Potentiometers	243
Hewlett-Packard Co., Palo Alto, Calif. Electronic test equipment and will feature 805A Slotted Line, 430A Microwave power meter, 614A Signal generator, new instruments.	40 & 41
Hickok Electrical Inst. Co., Cleveland, Ohio Electrical indicating meters and electronic test equipment.	232
Hopkins Engineering Co., Detroit, Mich. Stoddart Aircraft Radio Co. products.	56
Indiana Steel Products Co., Chicago, Ill. Permanent magnets, cast and sintered magnetic tape recording heads, magnetic recording tape.	279

(Continued on page 76A)





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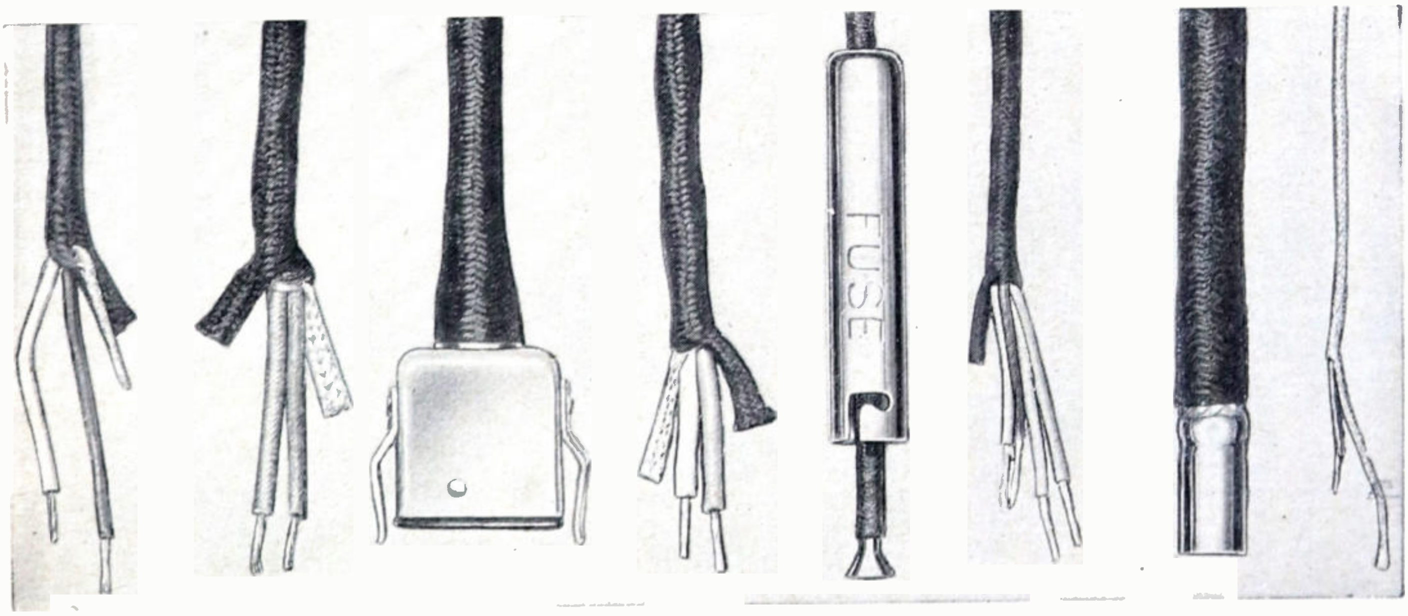
''IN BUSINESS SINCE 1904''

LENZ ELECTRIC MANUFACTURING CO. • 1751 No. Western Avenue, Chicago 47, Illinois



What to SEE at the 1949 Radio Engineering Show

Directory of 200 Exhibits

(Continued from page 74A)

Firm	Booth
Industrial Instruments, Inc., Jersey City, N. J. Auto bridge, automatic component sorter, wide range step attenuator, wheatstone bridges, resistance decades, capacity decades.	295
Industrial Products Co., Danbury, Conn. Coaxial connectors, RF components, panel connectors, antennas	223–226
Industrial Television Inc., Clifton, N. J. Remote control public view television receivers, hotel television systems, TV station monitors, antenna distribution systems.	285
International Nickel Co., Inc., New York, N. Y. Applications of nickel and nickel alloys in electronic industries.	98 & 99
International Resistance Co., Philadelphia, Pa. Composition Carbon Resistors, high stability resistors, carbon resistors, high frequency resistors, high voltage resistors, precision resistors.	102
J-B-T Instruments, Inc., New Haven, Conn. Vibrating Reed frequency meters, and testers, rotary selector switches, appliance temperature testers, Shurite AC and DC Panel and Pocket Meters.	254
J. F. D. Manufacturing Co., Inc., Brooklyn, N. Y. TV-FM Antennas, Brackets, Accessories, radio parts and accessories.	266
Howard B. Jones Div., Cinch Mfg. Corp., Chicago, Ill. Tube Sockets, Multi-Contact Plugs and Sockets, Terminal strips, terminals.	257

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Firm — **Booth**

Karp Metal Products Co., Inc., Brooklyn, N. Y. — 49 & 50
Sheet Metal Cabinets, Housings and consoles for electronic equipment.

Kay Electric Co., Pine Brook, N. J. — 22
Wide range sweeping oscillators, ultra short pulse generators, VHF spectrum analyzer, television test equipment, sound spectrograph.

Kenyon Transformer Co., Inc., New York, N. Y. — Y
Transformers and reactors for broadcast, commercial, and government service.

Kester Solder Co., Chicago, Ill. — 253
Kester "Resin-Five" Core solder—other Kester flux core solders, solid wire and bar solders, also solder preforms and fluxes.

Kimble Glass Div., Owens-Illinois Glass Co., Toledo, Ohio — 90
Kimble Glass Cathode Ray bulbs for the television industry.

Kings Electronics Co., Inc., Brooklyn, N. Y. — L
Connectors, antennas, variable condensers.

James Knights Co., Sandwich, Ill. — I
Quartz Crystals, crystal holders, frequency standards.

LaPointe Plascamold, Inc., Unionville, Conn. — 306
Television Antennas, Masts, Towers, and Accessories.

Lavoie Labs, Morganville, N. J. — 87 & 88
Oscilloscopes, Signal Generators, Wave meters Sound Embossers.

Linde Air Products Co., New York, N. Y. — 51
Linde Xenon, Krypton, Argon, Neon, Helium, rare gas mixtures, synthetic sapphire boules, rods, balls, centerless-ground rounds. Synthetic rutile.

Littelfuse, Inc., Chicago, Ill. — 273
Instrument Fuses, indicating lights, high voltage fuses, circuit breakers.

Lord Mfg. Co., Erie, Pa. — 317
Rubber Insulation and Vibration Mountings

M B Mfg. Co., Inc., New Haven, Conn. — 250
Vibration testing equipment, vibration pick-up, vibration isolators.

McGraw-Hill Publishing Co., New York, N. Y. — 200
Electronics, Nucleonics, McGraw-Hill Book Co.

Machlett Laboratories, Inc., Springdale, Conn. — 96 & 97
Electron tubes for broadcast industry and special purposes.

Magnecord, Inc., Chicago, Ill. — W
Magnecord tape recording and reproducing sytems and equipment.

P. R. Mallory & Co., Inc., Indianapolis, Ind. — 59 & 60
Television and radio components, inductuner, midgetrol, capacitors, switches, resistors.

Marion Electrical Instrument Co., Manchester, N. H. — 201
Hermetically sealed electrical indicating instruments, hermetically sealed elapsed time indicators.

Measurements Corp., Boonton, N. J. — 237
Signal generators, megacycle meters, square wave generators, pulse generators, special television test equipment.

Microwave Equipment Co., Verona, N. J. — 305
Microwave Equipment, Waveguide components and test equipment.

August E. Miller (Laboratories), North Bergen, N. J. — 308
Frequency Control Units .001% to .0001% accuracy, Miniature Crystal Ovens, Quartz Crystals and electronic telegraph equipment.

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What to SEE at the 1949 Radio Engineering Show

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Firm	Booth
Millivac Instruments, New Haven, Conn. High Impedance Millivoltmeter for Television, FM and Radar Frequencies. High Impedance DC Millivoltmeter. Multimeters, Ultra-sensitive DC amplifiers.	K
Mycalex Corp. of America, Clifton, N. J. Glass-bonded mica, the "Mycalex" products.	82
National Carbon Co., Inc., New York, N. Y. Radio batteries, hearing aid batteries, batteries for radiation detectors.	52 & 53
National Company, Inc., Malden, Mass. Commercial, communications receivers, electronic components and assemblies, including television receivers.	83 & 84A
National Research Corp., Cambridge, Mass. High vacuum equipment, diffusion pumps, vacuum gauges, valves, special equipment.	217
National Research Council, Ottawa, Canada Nuclear Instrumentation.	336
National Technical Labs, So. Pasadena, Calif. Mx Series Portable Radiation Instruments both Ionization and GM Tubes.	337
Northern Radio Co., Inc., New York, N. Y. Complete line of frequency shift equipment such as: Diversity receivers, F. S. Keyers, Converters, Tone Keyers and Filters.	275 & 276
Nuclear Inst. & Chem. Corp., Chicago, Ill. Scalers, geiger counters, accessories.	330
Nucleonics, New York, N. Y. Magazines.	200
Oak Ridge Antennas, New York, N. Y. Television Antennas.	274A

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What to SEE at the 1949 Radio Engineering Show

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Firm	Booth
Panoramic Radio Corp., Mount Vernon, N. Y. Scanning instruments covering sonic and radio frequency ranges.	241 & 242
Par-Metal Products Corp., Long Island City, N. Y. Relay racks, cabinets, panels, chassis, metal parts.	212
Patton-MacGuyer Co., Providence, R. I. Terminal attaching machine in operation. Terminals for electric wires.	294
Philco Corp., Philadelphia, Pa. Microwave televisicn relay equipment, loran, mobile radio communications systems, laboratory products.	44
Philmore Mfg. Co., New York, N. Y. Television Front Ends.	213
Polarad Electronics Co., New York, N. Y. Television and test equipment.	J
Polytechnic Research and Dev. Co., Inc., Brooklyn, N. Y. Microwave test equipment, variable attenuators, slotted sections and probes and frequency meters of novel design.	261
Potter & Brumfield Sales Co., Chicago, Ill. Relays.	252
Precision Apparatus Co., Inc., Elmhurst, L. I. Wide band sweep signal generator, 5" cathode ray oscillograph multi-range high sensitivity vacuum tube voltmeter.	222
Premier Crystal Labs., Inc., New York, N. Y. Quartz crystal units quartz crystal blanks, crystal controlled oscillators, Temperature control Crystal Units.	318
Presto Recording Corp., Paramus, N. J. Gear driven recorders and transcription turntables, microgroove recorders, amplifiers, blank discs, microgroove turntables.	25 & 26
Pyramid Electric Co., Paterson, N. J. Electrolytic capacitors, paper capacitors, radio-noise suppressors.	208
Radiart Corp., Cleveland, Ohio Vibrators, television antennas, car aerials, vibrator power supplies, inverters, heavy duty vibrators.	296
Radio Corp. of America, Camden, N. J. Broadcast Equipment (AM-FM-TV), Tubes, parts and test equipment.	4–9
Radio-Electronics, New York, N. Y. Radio-Electronics Magazines, Gersback library of radio technical books.	322
Radio Engineering Labs., Long Island City, N. Y. Complete Broadcast Systems FM Broadcast Transmitters Serrasoid Modulator, Studio to Transmitter Link Equip., Educational Transmitter.	323–326
Radio Magazines, Inc., New York, N. Y. Magazines.	94A
Rangertone, Inc., Newark, N. J. Hi Fidelity Broadcast tape recorders.	V
Raytheon Mfg. Co., Waltham, Mass. Receiving, Special Purpose, Subminiature and Microwave Tubes, Microwave communications equipment, and stabilizers.	13
Reeves Instrument Corp., New York, N. Y. Reeves Electronic Analog Computer, Servo mechanisms, components (motors, amplifiers, mechanical parts) verticle gyro.	35
Rek-O-Kut Company, Inc., Long Island City, N. Y. "Challanger" disc recorder, LP-12–dual speed transcription turntable, T-12-12" dual speed transcription turntable.	H

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What to SEE at the 1949 Radio Engineering Show

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Firm	Booth
John F. Rider Publisher, Inc., New York, N. Y. Technical Literature and electronic equipment.	312
Sanborn Co., Cambridge, Mass. Galvanometer assemblies, photographic type oscillographs, single and multichannel direct-writing recorders and associated amplifiers.	S
Carl W. Schutter, Rockville Centre, N. Y. Radar Components, consisting of choke flanges, cover flanges, special flanges, rigid waveguide, brass and aluminum, adaptors, E & H Elbows.	269
Hermon Hosmer Scott, Inc., Cambridge, Mass. Dynamic Noise suppressors, Amplifiers, Sound Level Meters, Laboratory equipment.	V
Shallcross Mfg. Co., Collingdale, Pa. Instruments, Resistors, Switches.	281
Sherron Electronics Co., Brooklyn, N. Y. Multiwave shape generator, comparison oscilloscope, oscillo-synchroscope, electronic voltage regulated supply, RF cathode ray null detector	C
Shure Brothers, Inc., Chicago, Ill. Microphones, phonograph pickup cartridges, phonograph pickups, wire recording heads, tape recording heads, microphone accessories.	264
Sigma Instruments, Inc., Boston, Mass. Sensitive relays for AC and DC., polarized relays for general applications and for high speed keying such as in teletype applications.	209
Simpson Electric Co., Chicago, Ill. Electrical Indicating Instrument, and radio test equipment.	86
Smith Paper, Inc., Lee, Mass. Metallized Condenser Paper, Kraft and Hemp Condenser, Kraft and Benares Electrolytic papers, coil and tan papers.	M
Sola Electric Co., Chicago, Ill. Sola Constant Voltage transformers.	21
Solar Mfg. Corp., North Bergen, N. J. Electrolytic Capacitors, various types of paper capacitors, including metallized paper types, radio noise filters and capacitor analyzer.	101
Somerset Labs., Inc., Union City, N. J. Noise suppressing preamplifier for use in record reproducing systems including models for all uses. Amplifier for music reproduction incorporating dynamic noise suppression.	E
Sorensen & Co., Inc., Stamford, Conn. AC line voltage regulators, nobatrons, 400 cycle regulator.	14
Sperry Gyroscope Co., Div., The Sperry Corp., Great Neck, N. Y. Microwave Measuring Instruments, Klystron tubes.	18–20
Sprague Electric Co., North Adams, Mass. Ceramic, Electrolytic, Mica and Paper Dielectric Capacitors, Fixed Wire Wound Resistors, Filters, high voltage networks.	57 & 58
Standard Piezo Co., Carlisle, Pa. Quartz crystal and frequency control units.	310
Standard Transformer Corp., Chicago, Ill. Transformers, Reactors, Power Packs, Transmitters.	263
Star Expansion Products Co., New York, N. Y. Electronic accessories, (Star miniature tube pin straighteners and Star miniature socket wiring plugs), precision die castings, toggle bolts.	P
Stoddart Aircraft Radio Co., Detroit, Mich. Noise and Field intensity Meter and Radiation Monitor.	56

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What to SEE at the 1949 Radio Engineering Show

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Firm	Booth
Super Electric Products Corp., Jersey City, N. J. Complete line of AM, FM and TV Coils, IF, RF Chokes, Transformers, etc., in both Audio and Radio Frequency.	Z
Superior Electric Co., Bristol, Conn. New low Capacity Stabiline voltage regulators, new stabiline voltage regulators. Powerstat Variable transformers.	108–110
Surprenant Mfg. Co., Boston, Mass. Television Line Cable, Special Coaxial Cable, Antifungus Wire, insulation, plastics.	38
Sylvania Electric Prod., Inc., New York, N. Y. Germanium Diodes, Germanium Triodes, Cathode Ray Tubes, Microwave Equipment, Radio Receiving Tubes, Radio Tube Parts.	104–106
Synthane Corp., Oaks, Pa. Laminated Plastic Products, Sheets, Rods, Tubes, Molded Laminated, Molded macerated, fabricated parts.	Q
Tech Laboratories, Inc., Palisades Park, N. J. Attenuators, variable and fixes, precision attenuators, potentiometers, Mixers, Switches, audio, rf and midget, silver soldering pot.	206
Technology Inst. Corp., Waltham, Mass. Precision, linear and functional potentiometers, audio Z-Angle Meter, R-F Z Angle Meter, R-F Power Oscillator, Phase-Angle Meter.	N
Tektronix, Inc., Portland, Ore. Three types of wide band oscilloscopes, two types of square wave generators.	238
Tel-Instrument Co., Inc., E. Rutherford, N. J. Type 1000 Teladapter, Type 1200 R-F-Sweep Generator, Type 1900 Multi-Frequency Generator, Type 2000 Bar and Dot Generator.	T
Tele-Tech, New York, N. Y. Magazines.	286
Terminal Radio Corp., New York, N. Y. High Quality Sound Equipment, Test Equipment, Miscellaneous radio parts.	287
Test Equipment Sales, Philadelphia, Pa. Test Equipment.	216A
Times Facsimile Corp., New York, N. Y. Facsimile Communications equipment, tuning fork controlled frequency standards, tube clamps	23 & 24
Tracerlab, Inc., Boston, Mass. Equipment for radioactivity research and for radiation health protection, including the Autoscaler, The "64" Scaler, Automatic Sample Changer.	332
Transvision, Inc., New Rochelle, N. Y. Receiver kits combining television, FM and AM, Sweep Signal Generators, Field Strength Meter, Booster, Antennas.	203 & 204
Truscon Steel Co., Youngstown, Ohio Technical information on selfsupporting and Uniform Cross-Section Guyed Radio Towers, Copper Mesh Ground Screen.	202
United Electronics Co., Newark, N. J. High Vacuum Transmitting tubes, gas rectifiers and thyratrons, vacuum capacitors.	249
U. S. Air Force, Washington, D. C. A 60′ exhibit depicting the work of Air Force Air Communications.	3rd floor
U. S. Army Signal Corp. exhibit.	3rd Floor
U. S. Navy Electronic exhibit.	3rd Floor
United Transformer Co., New York, N. Y. Transformers, reactors, chokes, voltage regulators, filters (audio, carrier, supersonic) high Q coils, filter coils, stepdown transformers, equalizers.	81

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What to SEE at the 1949 Radio Engineering Show

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Firm	Booth
VEE-DX, Unionville, Conn. Television Antennas, Masts, Towers and accessories.	306
Veterans Administration Demonstration of disabeled veterans working ability.	248A
Victoreen Instrument Co., Cleveland, Ohio Instrumentation in Nuclear Physics.	328 & 329
Video Television, Inc., New York, N. Y. Television Antennas.	274A
Vision Research Lab., Inc., Richmond Hill, N. Y. Vision Front End, Model 701, Vision Tele-Boosters, Model TVA, TVX TVL—The FM TeleTuner, Test equipment, tele-marker.	218
Waldes Kohinoor, Inc., Long Island City, N. Y. Truarc Retaining Rings.	230
Waterman Products Co., Inc., Philadelphia, Pa. Portable oscilloscopes, as well as regular type, together with associated equipment. Rayonic cathode ray tubes will also be exhibited.	216
Webster-Chicago Corp., Chicago, Ill. Automatic Record Changers, Wire Recorders.	256
Webster Electric Co., Racine, Wis. Pickup cartridges and tone arms for LP playing equipment, magnetic cartridges, retractable cartridges, standard cartridges.	290
Welwyn Electronic Components, Inc., New York, N. Y. High stability carbon resistors, miniature wirewound resistors, pyromatic resistors, vitreous enameled resistors.	89
Western Lithograph Co., Los Angeles, Calif. E-Z Code Wire Markers, Business Systems, Cable, Conduit and Pipe Markers, High Production Identification Methods.	107
Westinghouse Electric Corp., Pittsburgh, Pa. Point-to-point radio, oil insulated selenium rectifiers, tubes, transformers, instruments and oomphometer, cores, magnetic comparator, circuit breakers, contactors.	62–69
Weston Elec'l. Inst. Corp., Newark, N. J. Electronic and radio service instruments.	282 & 283
S. S. White Dental Mfg. Co., New York, N. Y. Flexible shafts and resistors.	251
Wind Turbine Co., West Chester, Pa. A tower section and a VHF Antenna.	42
Workshop Associates, Newton Highlands, Mass. High frequency antennas.	39

News—New Products

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

New Shock and Vibration Control Material

A new, all-metal resilient material providing dual protection against shock and vibration has been developed by **Vibrashock**, of **Robinson Aviation, Inc.**, Teterboro Air Terminal, Teterboro, N. J.

Described by the manufacturer as "Met-L-Flex" mounting systems, this type of isolation offers uniform performance under temperature extremes; is less subject to permanent set; permits wide loading range; inter-wire friction providing

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News—New Products

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

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inherent high damping action; resistance to aging or deterioration; nonlinear load-deflection characteristics providing shock protection as well as vibration isolation.

This company now has available complete mounting bases, ready for installation of equipment. These may be procured in standard form factors, or by custom fabrication conforming to Government performance specifications. Detailed information supplied on request.

New Enterprise

An established English firm, Welwyn Electrical Laboratories, Ltd., has recently opened a New York office.

Operating in this country as **Welwyn Electronic Components, Inc.**, 234 E. 46 St., New York 17, N. Y., this firm specializes in solving difficult resistor and capacitor problems.

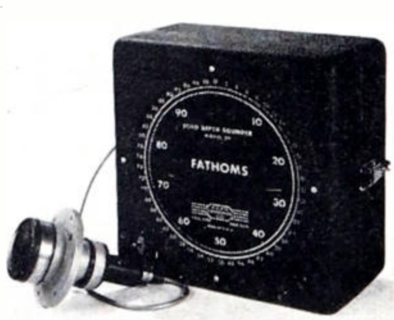
Welwyn high-stability carbon resistors are made in 1% tolerance from 100 ohms to 10^4 ohms. Claimed by the manufacturer to be individually noise tested before shipment, these resistors are outstanding for their quietness of operation in high-gain audio and nucleonic applications as they approach wirewound resistors in noise-free operation. The reactive component is nil to frequencies above the supersonic. In low-resistance values, they can be supplied in nonreactive form for operation at frequencies of 100 mc.

Above 10^4 ohms and up to 10^9 ohms, they supply a pyromatic resistor with proportionate high resistance values.

In wire-wound types, this firm produces vitreous-enamelled and cement-coated power resistors. Outstanding is a 3-watt vitreous resistor capable of full dissipation, but of miniscule physical size.

The aforementioned is completely described in Welwyn's latest catalog, available to firms requesting it on their corporate stationery.

Recent Catalogs



Kaar Engineering Co., Middlefield Road, Palo Alto, Calif. have announced literature descriptive of their Electronic Echo Depth Sounder. Interesting to the owner of small fishing fleets and yachts, it will operate over depths of 100 or more fathoms. Of small physical size, with only two units to install and low power drain, it works on ultrasonic principles.

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

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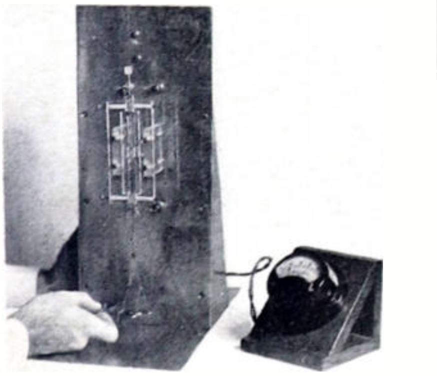
A New Aspect of Precision Instrumentation

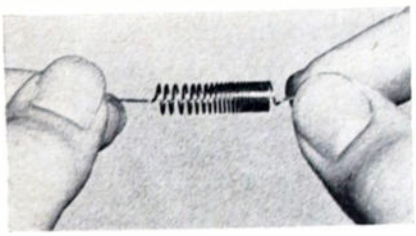
The National Bureau of Standards, U. S. Department of Commerce, Washington 25, D. C. has announced a new technique in transducers, for translating minute displacements of motion into variations of resistance, voltage, or current. Still under development, it has already been perfected to a point where laboratory engineers will find the process worth while to investigate.

Operating on the change of resistance principle, from a coiled or distended spring, the technique lends itself to bridge constructions so that the change in output, electrically, from mechanical displacement, is double the applied voltage, as one set of arms of the bridge is increased while the second set is decreased, in their respective conductivity.



Springs are coiled of nickel-alloy wire, gold-plated to a thickness of 1/10 mil. When in their rest state, they offer the resistance of a cylindrical tube, but as tension increases the resistance changes to that of the linear wire itself. Thus displacements as small as 1 Y10⁻⁶ are possible to translate into electrical currents and are indicated upon a galvanometer scale.

When actuated by another transducer, such as a bi-metallic strip electrical outputs can be easily measured from small changes of temperature, etc. Then the technique lends itself to accelerometers, pressure indicating elements, strain gauges, electric weighing devices, voltage regulators, dc to ac inverters, and many other applications.



Shown is a spring, partially stretched, and also a bridge element for electrically

(Continued on page 86A)

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

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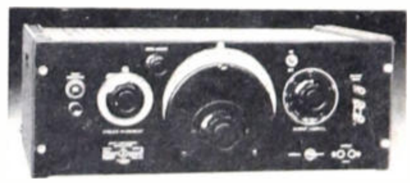
indicating small mechanical displacements. Still under development, this description is being published to appraise our readers of the strides taking place within the industry itself, for more precise mensuration techniques.

Beat Frequency Oscillator

General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass., has announced the post war redesign of their familiar 913 A beat frequency oscillator, now known as Type 1304 A. The same range is covered with many improvements added. Calibration is now done at line power frequency, through the means of a neon-lamp indicator.



The new gear-driven illuminated logarithmic dial, with improved bearings for alignment accuracy, improves the application use of the instrument. A decremeter adjustment of ±50 cps allows for accurate readings to be made at any point along the entire frequency range of 20 to 20,000 cps. High output at low distortion, 0.3 watt at less than 0.25% across the 600-ohm terminals is the maker's rating, although as much as 1 watt can be obtained from the instrument at 1% distortion. Drift is less than 7 cps, most of which occurs in the first hour of operation, with stability achieved after about two hours of use. Relay-rack and table models are both available.

Three RCA Improvements

The RCA-812A power triode (superseding the 812) features a zirconium-coated plate having radiating fins to give greater dissipation capability, grid and plate leads designed to have low rf loss, and a strengthened top-cap assembly with ceramic collar.

Because of its high perveance, the 812A can be operated at high efficiency and low driving power. For example, a pair of 812A's under ICAS conditions in unmodulated class C service can be operated with a plate input of 520 watts and the low driving power of 13 watts. Operation with maximum ratings is permissible up to 30 Mc, and with reduced ratings to 100 Mc.

The RCA-672-A mercury-vapor thyratron, which supersedes the 672, has the following modifications: (1) heater current is reduced from 6 to 5 amperes, (2) maximum peak forward and inverse anode voltage ratings are increased from 1500 to 2500 volts, (3) maximum peak anode current is increased from 30 to 40 amperes, (4) maximum average anode current is increased from 2.5 to 3.2 amperes.

The RCA-4×500A power tetrode is radiator-cooled, and is intended for vhf operation as a power amplifier and oscillator. The maximum plate dissipation is 500 watts.

Small in size, the 4×500A features low-inductance leads, low grid-plate capacitance, effective shielding between grid

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

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and plate circuits, and a grid terminal located at the center of the filament end of the tube to facilitate its use in coaxial circuits.

Technical bulletins with detailed information will be supplied on request to the **Tube Dept., Radio Corporation of American,** Harrison, N. J.

Plant Expansions

ALLENTOWN, Pa., witnessed the opening in October of a new manufacturing plant by the **Western Electric Co.** where precision electronic products are made by a staff expected to reach 2500 employees. WE has built a factory as modern as the electronic art itself, where quality will be the watchword of the production staff. A controlled-conditions plant, every precaution has been taken to insure cleanliness and dust-free atmosphere, as these are prerequisites in the manufacturing of small vacuum tubes, a large part of the output from this new factory in the Lehigh Valley.

CHICAGO, Ill., is the locus of the new plant of *Lectrohm, Inc.*, long-time maker of vitreous-enameled resistors and electrically heated soldering pots. In the rapidly expanding Clearing Industrial District of Chicago, the new address for Lectrohm is 5939 Archer Ave., Chicago 38, Ill. These new and larger quarters will speed production and delivery of the products of this company long active in the radio and electric fields.

NEW YORK, N. Y., has just been selected for the new *Fairchild Recording Equipment Corporation*, with offices at 30 Rockefeller Plaza, and the locus of a new manufacturing plant yet to be decided upon, but probably in the Long Island City vicinity. This new enterprise has been formed to take over the sale of the disk recording equipment of the Fairchild Camera & Instrument Corporation, and also to market a new professional-quality studio-type magnetic-tape recording and playback device, which was developed by Dr. D. G. C. Hare, recently president of the Deering-Milliken Research Trust, who is serving as a consultant to the new firm. Other audio items will shortly find their way into the new Fairchild line, too.

TOLEDO, Ohio, recently saw the expansion of the *Kimble Glass• Co.* plant, where direct-viewing television glass bulbs are made in quantity by automatic machinery. The rapid expansion of the interest in television has caused a continuing shortage of glass envelopes for the cathode-ray tubes used for picture reproduction, and the Kimble organization have increased their production capacity tremendously by this rapid conversion from handmade envelopes to machine-fabricated units, with progress once estimated to take years condensed into a few months. Kimble is a division of Owens-Illinois Glass Company.

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News—New Products

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

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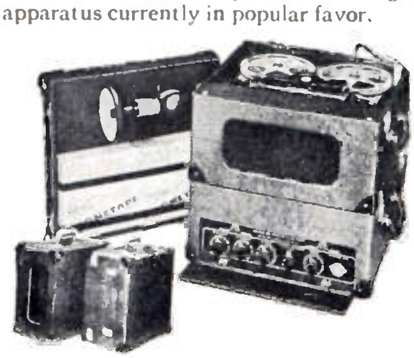
DOVER, N. H., welcomes the move from Brooklyn, N. Y. of *Clarostat Mfg. Co.*, one of the oldest makers of volume controls and resistance devices in the radio industry. The new plant was opened in November, affording over ¼ million square feet of space for offices, manufacturing and warehouse facilities, under one roof in a five-story building. For over a quarter century in Brooklyn, the move to New England is a departure for this firm, designed to increase operational efficiency by combining the several plants into which Clarostat was forced when expanding to meet with the increase in business that the past war brought to them.

Recent Catalogs

Herman Hosmer Scott, Inc., 385 Putnam Ave., Cambridge, Mass., announce literature describing their "Little-Wonder" type 110-A dynamic noise suppressor. Small physical size and remote mounting of the control from the chassis contribute to ease of installation of the unit in existing phonographs. Possessing noise-suppressing features superior to the frequency range of most current recordings, this unit includes a matched crystal pickup, for installation when it is used. Packaged units of this type lend themselves to easy merchandising of apparatus currently in popular favor.



Amplifier Corp. of America, 398 Broadway, New York 13, N. Y. have announced descriptive literature detailing their newest magnetic-tape recording equipment. This unit incorporated a twin recording track, on one tape, with instantaneous magnetically controlled reverse for continuous operation. This is similar to the shuttling of a typewriter ribbon. Mounted in separate cases for ease in carrying, the unit dovetails together when in operation, and includes storage space for microphone, connecting cables, and reels of tape.

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News—New Products

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

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High-Fidelity FM Tape Recorder

Described as having been "designed expressly to the highest requirements of FM broadcast application," the Ampex Tape Recorder with a 35-minute transcribing capacity at 30 inches of tape per second is being marketed by **Ampex Electric Corp.**, 131 Howard Ave., San Carlos, Calif.

Frequency response is within ±1 db between 30 and 15,000 cps. Over-all distortion is 4% intermodulation at peak meter reading. At 10 db above peak meter reading the total rms harmonic distortion does not exceed 5%. Unweighted noise level is 60 db below 5% harmonic distortion for the system as a whole—including both amplifiers, bias, erase, and stray pick-up. Accuracy of playback timing is ±0.2%.

This recorder has separate heads for erase, record, and playback; separate record and playback amplifiers; a marking device for spotting tape when editing; relay and pushbutton controls for remote operation; forward and rewind speeds of over 300 in sec; and head gaps aligned to permit interchangeability of tapes without high-frequency loss.

Recent Catalogs

• • • Application notes concerning "Adjustment of Filament Voltage of RCA-1B3-GT by Observation of Filament Temperature," AN-134; "Single-Section Filament Operation of Types 3S4 and 3V4," AN-135; "Overload Protection for the Horizontal Deflection Circuit in Television Receivers," AN-136; "Reduction in Peak Inverse Voltage Rating of Type 1B3-GT," AN-137, all by **Tube Department, RCA**, Harrison, N. J. In requesting these bulletins, specify the identifying code letters and numbers immediately following captions.

(Continued on page 90A)

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 89A)

High-Frequency Heating Unit

The new 2R Thermax Red Head, a high-frequency heating unit which has an output of 2.5 kw and will raise the temperature of 1¾ lbs of average general-purpose material from 80° to 250°F in one minute, reducing mold pressure as much as 80%, has been designed and produced by the **Thermex Div., Girdler Corp.**, Louisville, Ky.

The standard lower electrode is a 10″×12″ aluminum plate. The upper electrode is 10″×12″ stainless steel wire mesh. Spacing between the electrodes is adjustable from ¼″ to 5″, and the drawer accepts preforms up to 4¾″ in thickness. It operates on 230 or 460 volts, 60 cps, three-phase power supply. Rated output requires 5.6 kva input at 90% power factor. One WL437 hf power oscillator and six 872A vapor rectifiers comprise the tube complement. Net weight, 800 lbs; height, 55″; depth, 27¼″; width, 17¼″.

Ultra-Short Pulse Generator

The Mega-Pulser, a new ultra-short pulse generator that provides a pulse with a spectrum which, the manufacturer claims, more than covers the present video-frequency range, is being marketed by the **Kay Electric Co.**, Pine Brook, N. J.

(Continued on page 91A)

News—New Products

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

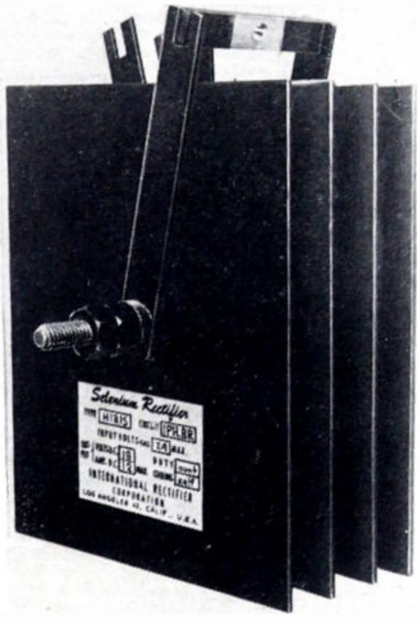
(Continued from page 90A)

Its features are: pulse width, 0.025, 0.05, 0.1, and 0.25 microsecond; pulse amplitude, 100 volts at 50 ohms impedance; pulse shape, flat-top pulses 0.05 microsecond and greater; pulse rise and fall time less than 0.01 microsecond.

It also triggers from an internally or externally provided pulse, and provides a positive or negative pulse. The output pulse is delayed approximately 0.25 microsecond to allow observation on an oscilloscope.

Heavy Duty Selenium Rectifier

The Type "H" rectifier with individual selenium elements, measuring 6¼"×7¼" and rated at 12 amperes (single-phase bridge), is being manufactured by **International Rectifier Corp.**, 6809 S. Victoria Ave., Los Angeles 43, Calif.



Features incorporated in these assemblies are: (1) interlocking assemblies of rectifier components, (2) conservatively rated terminals, (3) special moisture proofing.

The manufacturer claims that the elements have stable characteristics. Leakage current is less than 1 ma/cm² at 25 volts rms in the reverse direction. The type "H" is especially suitable for all applications requiring output in access of 10 amperes.

Recent Catalogs

• • • A bulletin and an accompanying news release describing a chassis punch driven by screw action that cuts square or round holes to facilitate mounting angular parts on sheet metal, by **Pioneer Broach Co.**, 1424 S. Main St., Los Angeles 14, Calif.

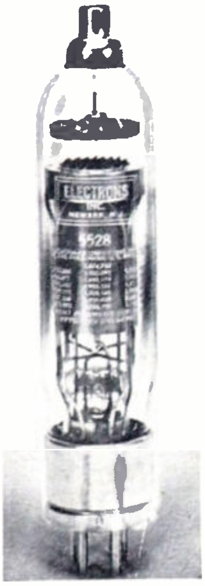
(Continued on page 93A)

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

(Continued from page 91A)

Grid Controlled Rectifier Tube

The new Type 5528 (C6L) a fast de-ionizing, temperature-free grid-controlled rectifier tube has been developed by **Electrons, Inc.**, 127 Sussex Ave., Newark, N. J.



Pertinent electrical characteristics of the type 5528 are: 60-second heating time; 500 volts peak inverse rating; 6.4 amperes steady dc; 77 amperes peak.

This tube is designed especially for high-power servomotor control or inverter applications which must operate at up to 800 cps. and where reliability and stable characteristics are required over wide ambient temperature ranges.

Inquiries should be directed to James H. Burnett at Electrons Inc.

Recent Catalogs

• • •The FTL-10-A, a 23-voice-channel radio link system, especially advantageous where terrain conditions or water paths make wire installation costly or impossible; The FTL-11-A, an Fm uhf radio link, used to relay program material from the broadcast station to the transmitter, eliminating wire or cable circuits; the FTL-12A, a complete television broadcast transmitter monitor for AM and FM signals; the FTL-13A, an FM uhf broad-band radio link for multichannel telephone service; the FTL-15-A, an all-metal dummy antenna for FM broadcast transmitters which accurately measures and safely dissipates large amounts of vhf rf power by means of an all-metal dissipating element through which water flows, all by **Federal Telecommunication Laboratories, Inc.**, 67 Broad St., New York 4, N. Y.

Is Your File of Standards Complete?

For your convenience a list of up to date IRE Standards and the ASA Standards sponsored by IRE is given below. The coupon at the bottom may be used for your convenience in ordering standards which you may not have. The standards are punched for 3 ring book and can be permanently filed conveniently.

Current IRE Standards

Price

1) **Standards on Electroacoustics, 1938 Definitions of Terms, Letter and Graphical Symbols, Methods of Testing Loud Speakers.** (vi + 37 pages, 6 x 9 inches).... $0.50

2a) **Standards on Electronics: Definitions of Terms, Symbols, 1938.** A Reprint (1943) of the like-named section of "Standards on Electronics, 1938." (viii + 8 pages, 8½ x 11 inches) ... $0.20

2b) **Standards on Electronics: Methods of Testing Vacuum Tubes, 1938.** A Reprint (1943) of the like-named section of "Standards on Electrons, 1938." (viii + 8 pages, 8½ x 11 inches) ... $0.50

3a) **Standards on Transmitters and Antennas: Definitions of Terms, 1938.** A Reprint (1942) of the like-named section of "Standards on Transmitters and Antennas, 1938. (vi + 10 pages, 8½ x 11 inches) $0.20

3b) **Standards on Transmitters and Antennas: Methods of Testing, 1938.** A Reprint (1942) of the like-named section of "Standards on Transmitters and Antennas, 1938." (vi + 10 pages, 8½ x 11 inches) $0.50

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4b) **Standards on Radio Receivers: Methods of Testing Broadcast Radio Receivers, 1938.** A Reprint (1942) of the like-named section of "Standards on Radio Receivers, 1938." (vi + 20 pages, 8½ x 11 inches) ... $0.50

4c) **Standards on Radio Receivers: Methods of Testing Frequency-Modulation Broadcast Receivers, 1947.** (vi + 15 pages, 8½ x 11 inches) .. $0.50

5a) **Standards on Radio Wave Propagation: Definitions of Terms, 1942.** (vi + 8 pages, 8½ x 11 inches) .. $0.20

Price

5b) **Standards on Radio Wave Propagation: Measuring Methods, 1942.** Methods of Measuring Radio Field Intensity, Methods of Measuring Power Radiated from an Antenna, Methods of Measuring Noise Field Intensity. (vi + 16 pages, 8½ x 11 inches) $0.50

5c) **Standards on Radio Wave Propagation: Definiitons of Terms Relating to Guided Waves, 1945.** (iv + 4 pages, 8½ x 11 inches).. $0.20

6a) **Standards on Facsimile: Definitions of Terms: 1942.** (vi + 6 pages, 8½ x 11 inches).... $0.20

6b) **Standards on Facsimile: Temporary Test Standards, 1943.** (iv + 8 pages, 8½ x 11 inches) .. $0.20

7) **Standards on Piezoelectric Crystals: Recommended Terminology, 1945.** (iv + 4 pages, 8½ x 11 inches) ... $0.20

8a) **Standards on Television: Methods of Testing Television Transmitters, 1947.** (vi + 18 pages, 8½ x 11 inches) ... $0.75

8b) **Standards on Television: Methods of Testing Television Receivers, 1948.** (vi + 32 pages, 8½ x 11 inches) .. $1.00

8c) **Standards on Television: Definitions of Terms, 1948.** (iv + 4 pages, 8½ x 11 inches)... $0.50

9) **Standards on Antennas, Modulation Systems, and Transmitters: Definitions of Terms, 1948.** (vi + 25 pages, 8½ x 11 inches) ... $0.75

10) **Standards on Abbreviations, Graphical Symbols, Letter Symbols, and Mathematical Signs, 1948.** (vi + 21 pages, 8½ x 11 inches) .. $0.75

11) **Standards on Antennas: Methods of Testing, 1948.** (vi + 18 pages, 8½ x 11 inches) .. $0.75

Normas Sobre Receptors de Radio, 1938.* A Spanish-language translation of "Standards on Radio Receivers, 1938," by the Buenos Aires Section of The Institute of Radio Engineers. (vii + 64 pages, 6 x 9 inches) Two Argentine Pesos (Postpaid)

* *Not carried in stock at IRE Headquarters in New York. Obtainable only from Señor Domingo Arbó, Editor of Revista Telegrafica, Peru, 165, Buenos Aires, Argentina.*

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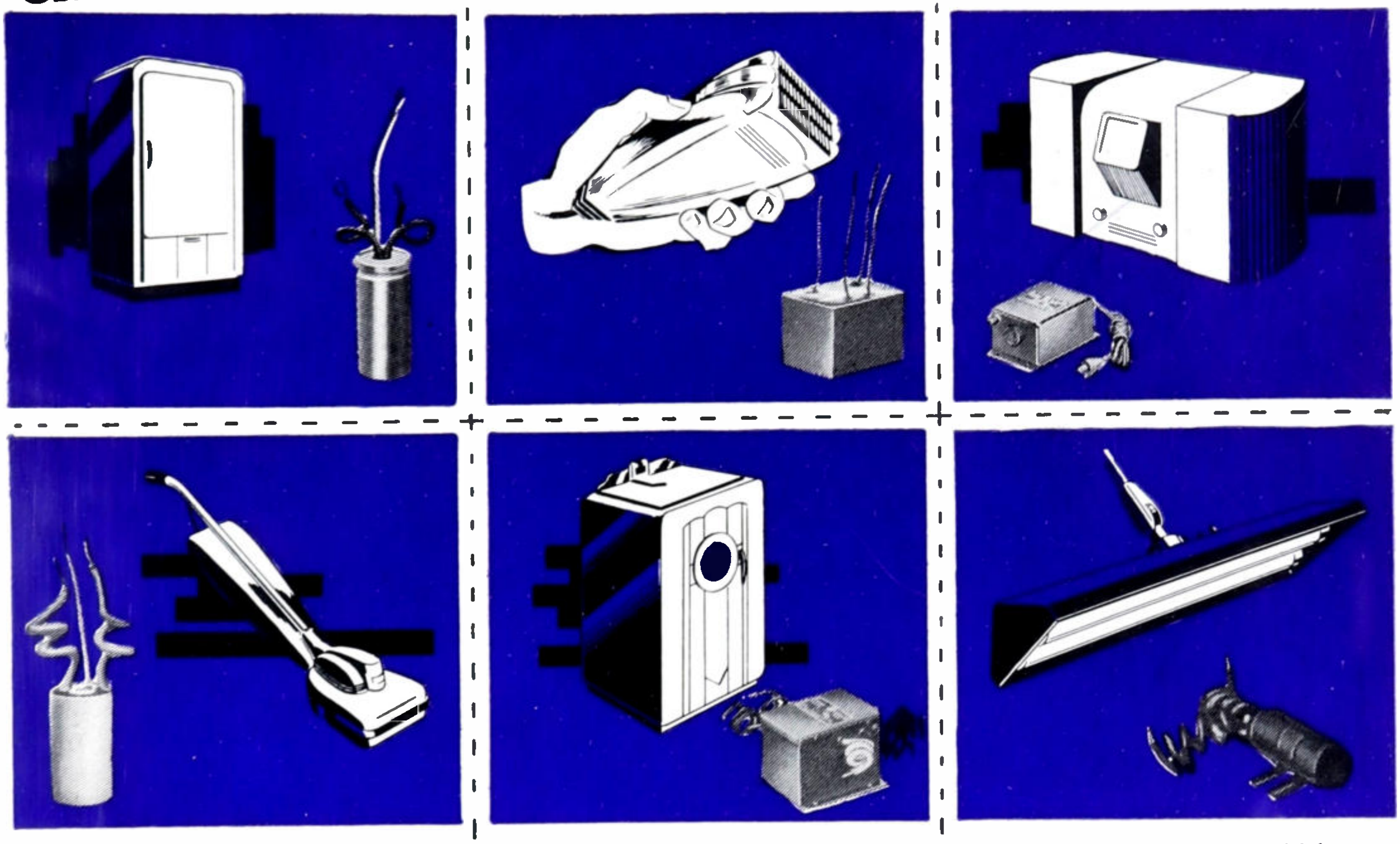

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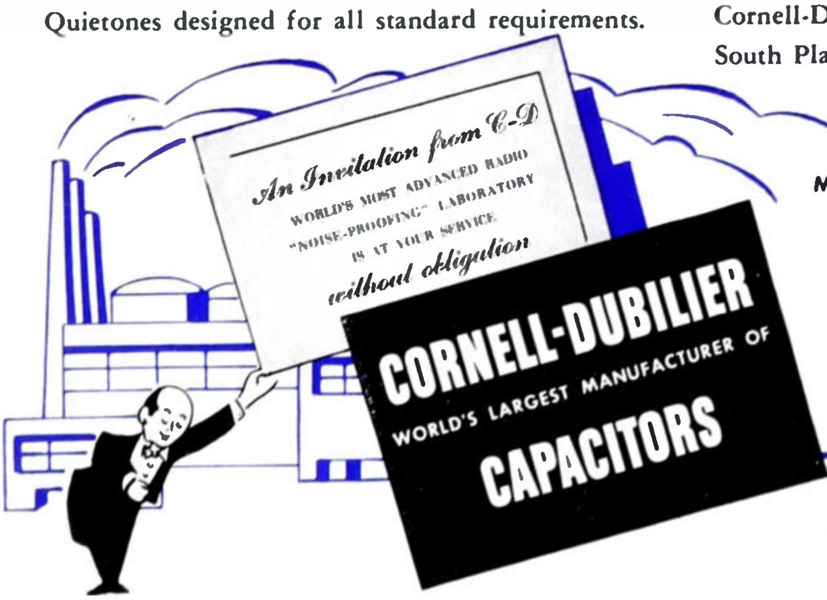

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