Proceedings of the I·R·E

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February, 1952

Volume 40 Number 2

Convention Program and Summaries of Technical Papers appear on pages 212-234

PROCEEDINGS OF THE I.R.E.

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The Institute of Radio Engineers

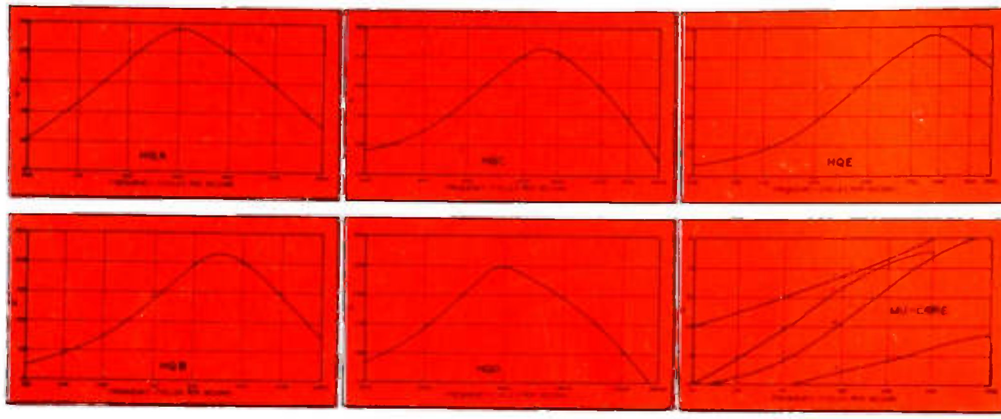
PERMALLOY DUST TOROIDS FOR MAXIMUM STABILITY...

The UTC type HQ permalloy dust toroids are ideal for all audio, carrier and supersonic applications. HQA coils have Q over 100 at 5,000 cycles... HQB coils, Q over 200 at 4,000 cycles... HQC coils, Q over 200 at 30 KC... HQD coils, Q over 200 at 60 KC... HQE (miniature) coils, Q over 120 at 10 KC. The toroid dust core provides very low hum pickup... excellent stability with voltage change... negligible inductance change with temperature, etc. Precision adjusted to 1% tolerance. Hermetically sealed.

HQA, HQC, HQD CASE
1 13/16" Dia. x 1 3/16" High

HQB CASE
1 5/8" x 2 5/8" x 2 1/2" High

HQE CASE
1/2" x 1 5/16" x 1 3/16" High



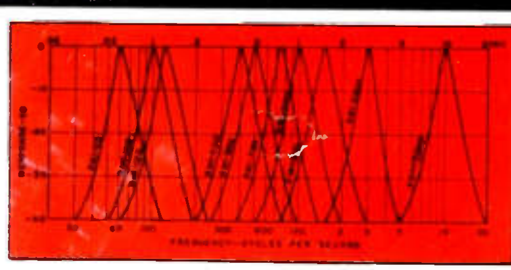
Type No.	Inductance Value	Net Price
HQA-1	5 mhy.	$7.00
HQA-2	12.5 mhy.	7.00
HQA-3	20 mhy.	7.50
HQA-4	30 mhy.	7.50
HQA-5	50 mhy.	8.00
HQA-6	80 mhy.	8.00
HQA-7	125 mhy.	9.00
HQA-8	200 mhy.	9.00
HQA-9	300 mhy.	10.00
HQA-10	.5 hy	10.00
HQA-11	.75 hy.	10.00
HQA-12	1.25 hy.	11.00
HQA-13	2. hy.	11.00
HQA-14	3. hy.	13.00
HQA-15	5. hy.	14.00

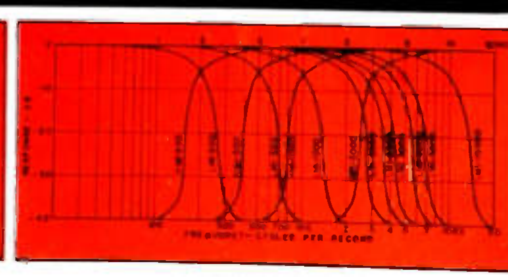
Type No.	Inductance Value	Net Price
HQA-16	7.5 hy.	$15.00
HQA-17	10. hy.	16.00
HQA-18	15. hy.	17.00
HQB-1	10 mhy.	16.00
HQB-2	30 mhy.	16.00
HQB-3	70 mhy.	16.00
HQB-4	120 mhy.	17.00
HQB-5	.5 hy.	17.00
HQB-6	1. hy.	18.00
HQB-7	2. hy.	19.00
HQB-8	3.5 hy.	20.00
HQB-9	7.5 hy.	21.00
HQB-10	12. hy.	22.00
HQB-11	18. hy.	23.00
HQB-12	25. hy.	24.00

Type No.	Inductance Value	Net Price
HQC-1	1 mhy.	$13.00
HQC-2	2.5 mhy.	13.00
HQC-3	5 mhy.	13.00
HQC-4	10 mhy.	13.00
HQC-5	20 mhy.	13.00
HQD-1	.4 mhy.	15.00
HQD-2	1 mhy.	15.00
HQD-3	2.5 mhy.	15.00
HQD-4	5 mhy.	15.00
HQD-5	15 mhy.	15.00
HQE-1	5 mhy.	6.00
HQE-2	10 mhy.	6.00
HQE-3	50 mhy.	7.00
HQE-4	100 mhy.	7.50
HQE-5	200 mhy.	8.00

UTC INTERSTAGE AND LINE FILTERS

FILTER CASE M
1 3/16 x 1 11/16,
1 5/8 x 2 1/2 High





These U.T.C. stock units take care of most common filter applications. The interstage filters, BMI (band pass), HMI (high pass), and LMI (low pass), have a nominal impedance at 10,000 ohms. The line filters, BML (band pass), HML (high pass), and LML (low pass), are intended for use in 500/600 ohm circuits. All units are shielded for low pickup (150 mv/gauss) and are hermetically sealed.

STOCK FREQUENCIES
(Number after letters is frequency)
Net Price $25.00

BMI-60	BMI-1500	LMI-200	BML-400
BMI-100	BMI-3000	LMI-500	BML-1000
BMI-120	BMI-10000	LMI-1000	HML-200
BMI-400	HMI-200	LMI-2000	HML-500
BMI-500	HMI-500	LMI-3000	LML-1000
BMI-750	HMI-1000	LMI-5000	LML-2500
BMI-1000	HMI-3000	LMI-10000	LML-4000
			LML-12000

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

What to SEE at

The Radio Engineering Show

March 3-6, 1952 at Grand Central Palace, New York

355 Exhibits of Radio-Electronic Equipment

Firm — Booth

Ace Engineering & Machine Co., Inc. Philadelphia 40, Pa. 350, 352
An Ace cell type shielded room, panel construction, showing construction details, power line entrances, filters, method of air conditioning, access doors for conveyor production.

Advance Electric & Relay Co., Burbank, Calif. 221
Electrical relays.

Aerovox Corp., New Bedford, Mass. 62
Capacitors, filters, resisitors.

Aircraft-Marine Products, Inc., Harrisburg, Pa. 475
Solderless terminals and connectors and automatic machines for applying them. Amplifilm high performance dielectric, high voltage capitrons, radar pulse networks (Capitron pulse forming networks), "Polyzol" A new dielectric Polymer.

Air-Marine Motors, Inc., Seaford, L.I., N.Y. 406
Subfractional horsepower motors 60, 400 cps, variable frequency. Centrifugal blowers and axial fan units—60,400 cps, and variable frequency. Control and low inertia motors.

Airpax Products Co., Baltimore 20, Md. 477
Choppers, vibrators, transformers, vibrator power supplies, vibrator inverters, and servo equipment using our choppers.

Airtron, Inc., Linden, N.J. 356
Flexible and rigid waveguides, dummy loads, directional couplers, mixer assemblies, duplexer assemblies, waveguide switches.

Alden Electronic & Impulse Recording Equipment Co., Westboro, Mass. 474
New recorders using electrosensitive recording paper including the Alden Magazine recorder, which records 3 channels on ¼" record; the Alden Multi Stylii Recorder which records 25 channels in 4½" width; and integral parts for those who want to build their own recorder such as helicies, paper holder feed mechanisms etc.

Alden Products Co., Brockton 64, Mass. N-3
Indicator lights, fuseholders, ac outlets, dial light sockets, tube cap connectors for all applications. Plugs and Sockets: non-interchangeable, detachable terminal, hi-voltage disconnect, miniature, cathode-ray tube, tuning eye. Components for test equipment: Sockets, adapters, test jacks, analyzer kits. Components for plug-in unit construction. Computer component, plugs and keys, line cords.

Allegheny Ludlum Steel Corp., Brackenridge, Pa. 25-26
See: Arnold Engineering.

Allied Control Co., Inc., New York 21, N.Y. 279
Electrical relays, and coils.

Alpha Metals, Inc., Brooklyn 1, N.Y. 326
Solders, Cen-Tri-Core energized rosin filled solder, wire solder, foil, powder, preforms.

Alpha Wire Corp., New York 13, N.Y. 471
Insulated and uninsulated copper wire and cable, government specification wire and cable, electronic hook-up wire, multiconductor intercommunication cable, test lead wire.

Altec Lansing Corp., New York 13, N.Y. 309, 311
Speech input equipment: consoles, remote mixers, plug-in amplifiers, microphones, transformers, speakers, amplifiers, intermodulation test equipment, AM-FM tuners, horns, driver units.

American Lava Corp., Chattanooga 5, Tenn. 64
Ceramic insulators for radio, television, radar, electronic components, wire communications, control equipment, and household appliances.

American Phenolic Corp., Chicago 50, Ill. 111, 112
Coaxial cables, polyethylene and teflon, 300 ohm twin-lead. rf connectors, AN Connectors and fittings, industrial connectors, AN conduit and fittings, television antennas, radio components and hardware, plastics for electronics.

American Structural Products Co., Industrial & Electronic Div., Toledo 1, Ohio. 89, 90
Technical glassware for electrical and electronics industry, principally cathode-ray bulbs.

American Television & Radio Co., St. Paul 1, Minn. 202
Light and heavy duty vibrators, vibrator power supplies, dc-ac inverters, and battery eliminators.

Amperex Electronic Corp., Brooklyn 1, N.Y. 10-12
Transmitting and power tubes, x-ray tubes and rectifiers, Geiger-Mueller tubes, vacuum capacitors, magnetrons, hydrogen thyratrons and various uhf types and subminiatures.

Ampex Electric Corp., Redwood City, Calif. 324
Tape recorders for instrumentation.

Amplifier Corp. of America, New York 13, N.Y. 317
Magnetic tape recorders, particularly a newly developed midget, battery-operated unit. Audio-amplifiers and regulated power supplies.

Anchor Metal Co., New York 13, N.Y. 217
Shurflo rosin core, and Shurflo special rosin core solders. Bar solders, solid wire preforms.

Andrew Corp., Chicago 19, Ill. N-9
VHF and UHF antennas, transmission lines, and assorted equipment.

Anton Electronic Labs., Inc., Brooklyn 6, N.Y. 390
Corona discharge voltage regulators. Geiger counter tubes and Geiger counter equipment. High intensity radiological monitoring instrument, the AN/PDR-32 which was developed for the Navy by the Anton.

Seven Million Dollars Worth of Components, Tools and Materials

Firm — Booth

Approved Electronic Instrument Corp., New York 6, N.Y. 314B
AM/FM tuner, TV pattern generator, FM/TV sweep signal generator, field strength meter, audio amplifiers, preamplifiers, power supply, and AM tuner.

Arco Electronics, Inc., New York 13, N.Y. 338
Molded Mica Capacitors, Steatite Encased Paper Tubular Capacitors, Variable Mica Compression Type Trimmers and Padders, and Silver High K Tubular and Disc Ceramics.

Arnold Engineering Co., Marengo, Ill. 25, 26
Permanent Magnets: Cast and sintered Alnico; cast cobalt; Remalloy and Vicalloy high permeability materials. Mo-permalloy powder Cores; Deltamax, Permalloy, Supermalloy, Mumetal, Silectron and 4750 tape wound cores; "C" cores of Silectron (grain oriented silicon steel), Permalloy and Deltamax; Permendur hot rolled bars, forgings and castings.

Arrow Electronics, Inc., New York 7, N.Y. 318
Jobber: Audio equipment and industrial equipment.

Atomic Instrument Co., Cambridge 39, Mass. 396A
Glow Transfer Counter, Rapid data Scaler and Printer, Automatic Scaler with built-in linear amplifier, calibrated reference sources, beta emitting. Linear amplifiers, scintillation head, complete model 1030 decade scaler, Libby-Kulp screen wall counter.

Audio Devices, Inc., New York 22, N.Y. 316
Audiodiscs—recording blanks. Audiotape—magnetic recording tape. Audiofilm—magnetic recording film. Audiopoints—recording and playback styli.

Audio & Video Products Corp., New York 19, N.Y. 485
Special magnetic tape recorders developed by the Ampex Electric Corp. for military and scientific research, data recording, telemetering, vibration studies, and shock analysis. Engineering representatives are in attendance for discussions of the applications of magnetic tape equipment for the storage of information. 100 KC tape recorder and special low wow and low flutter equipment.

Automatic Electric Sales Corp., Chicago 7, Ill. 290
Telephone type relays—sealed and non-sealed; aircraft and AN types; also regular. Rotary stepping switches of small size and high speed characteristics; extremely rugged and long lived. Manually operated lever key switches. Hermetically sealed rotary stepping switches.

Avery Adhesive Label Corp., Monrovia, Calif. 467, 468
Kum-Kleen pressure-sensitive labels and labeling methods for: product identification, printed circuits, trademarks, installation instructions, warning and inspection markers. Kum-Kleen electric label dispenser.

Avion Instrument Corp., Paramus, N.J. 333
Sweep Signal Generator, Magnetic data storage drum, Plug-in amplifier unit.

Ballantine Laboratories, Inc., Boonton, N.J. 100
Sensitive electronic volmeters; decade amplifiers, voltage multipliers, precision shunt resistors.

Barker & Williamson, Inc., Upper Darby, Pa. N-2
Coils capacitors, components, and test equipment.

(Continued on page 2A)

PROCEEDINGS OF THE I.R.E. February, 1952, Vol. 40, No. 2. Published monthly by the Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927. Printed in U.S.A.

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Beam Instruments Corp., New York 1, N.Y. Cossor Oscillographs, oscillograph cameras, and oscillograph motor drives. Best vacuo-junctions: uhf and standard types and special types to order. Sterling wire and cables: radio, multi conductor, coaxial and TV.	223
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Bendix Radio Division, Baltimore 4, Md. Type MRT-6 Command-Air Series mobile communications unit. Type MRT-3 Railmaster communications unit. Type AN/ARC-33 uhf Multi-Channel communications unit. Type MI-51-A Amspeaker. Type GDF-2 vhf direction finding system. MN-35D navigation receiver.	14-17
Scintilla-Magneto Div., Sidney, N.Y. Electrical connectors, ignition analyzers miscellaneous, ignition pieces.	14-17
Eclipse-Pioneer Division, Teterboro, N.J. Precision components for servo-mechanism and computing equipment including synchros, low inertia motors, gyros, and remote indicating-transmitting systems.	14-17
Berkeley Scientific Corp., Richmond, Calif. Electronic tachometer, direct reading frequency meter, double pulse generator, time interval meter, electronic decimal counting units, preset electronic counters, nuclear scalers, count rate meters.	399
Berlant Associates, Los Angeles 16, Calif. "Network" Magnetic tape recorders, "Professional" Magnetic tape recorders, microphone mixer preamplifiers.	314A
Beta Electric Corp., New York 29, N.Y. High voltage power supplies, kilovoltmeters, portable projection oscilloscopes, electronic microammeters.	S-21
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Boesch Mfg. Co., Inc., Danbury, Conn. Hand operated type toroidal winding machine. Automatic type toroidal winding machine with full equipment including shuttles. Automatic type toroidal winding machine with 12 inch shuttle. Automatic type toroidal winding machine with continuous winding assembly. Automatic type toroidal winding machine for winding miniature size coils. Automatic type toroidal winding machine with square window attachment.	306
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Bond Electronics Corp., Springfield, N.J. Precision wire-wound resistors, wire wound products, coils and coil assemblies.	503
Boonton Radio Corp., Boonton, N.J. Q-Meters, G-Meter, FM-AM signal generator, Univerter, and FM Signal generator.	276, 277
Borg Corp., Borg Equip. Div., Janesville, Wis. Precision ten-turn potentiometer, trade-named Micropot. Ten-turn counting dial-Microdial	272

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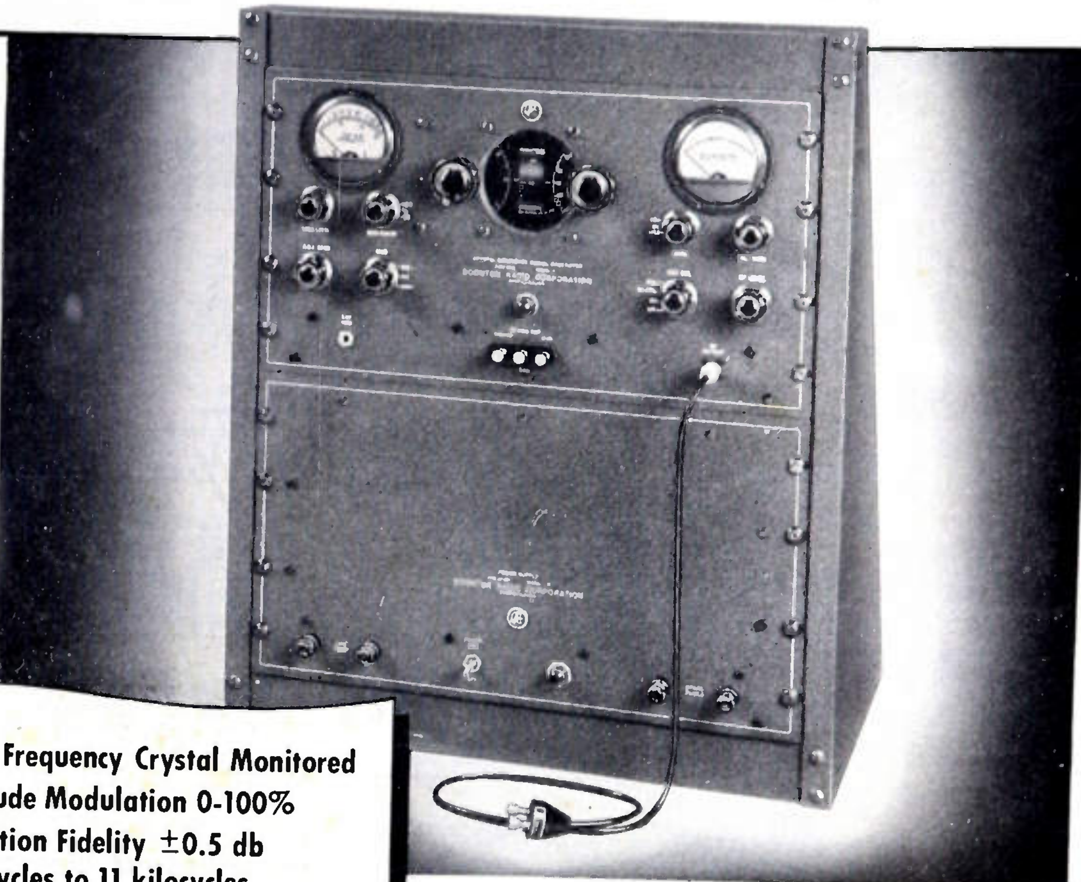
SIGNAL GENERATOR FOR OMNI-RANGE RECEIVERS

Type 211-A • Frequency Range 88-140 Megacycles



Output Frequency Crystal Monitored
Amplitude Modulation 0-100%
Modulation Fidelity ±0.5 db
30 cycles to 11 kilocycles
Negligible Spurious FM

SPECIFICATIONS

FREQUENCY RANGE: Master Oscillator: 88-140 megacycles in one range. Vernier frequency dial has 100 divisions and is coupled to the main tuning capacitor through a 120:1 gear drive. Each vernier division is equivalent to a 10 kc. change in frequency.

Crystal Controlled Frequencies: Either of two crystals 110.100 mc. and 114.900 mc., accurate to ±0.0035%, may be selected by a switch for use individually or in combination with the master oscillator to standardize its output frequency.

AMPLITUDE MODULATION CHARACTERISTICS: Two amplitude modulation ranges, 0-30% and 0-100%, are provided for use with the internal oscillator or a low distortion external oscillator. Distortion is 5% or less at 95% amplitude modulation.

Internal Audio Oscillator: Two modulating frequencies, 400 and 1000 cycles.

Modulation Amplifier: The internal modulating amplifier has the following characteristics:

Uniform response within ±0.5 db. 30 cycles to 11 kc.
Uniform response within ±0.1 db. 90 cycles to 150 cycles.
Uniform response within ±0.1 db. 9500 cycles to 10.5 kc.

Phase Distortion: (up to 60% amplitude modulation.)

Less than 0.25 degrees at 30 cycles.
Less than 10 degrees at 11 kc.

AUDIO TEST VOLTAGE: This instrument contains a demodulator or detector which supplies to front panel terminals a portion of the demodulated carrier.

SPURIOUS FM: Less than 1 kc. at 60% AM.

OUTPUT ATTENUATOR: Single ended piston type, adjustable from 0.2 volt to 0.1 microvolt. Output impedance as seen looking in at terminals of output cable is 26.5 ohms.

The Type 211-A Signal Generator is specifically designed for the testing and calibrating of omni-range radio receiving equipment. It is also well suited for laboratory and development work where a precision type amplitude modulated R. F. signal source is required.

Careful consideration has been given to the location of panel controls with respect to function and degree of use. The main frequency dial is located in the center of the panel, with the vernier dial to the left in close proximity, utilizing the same fiducial for simplicity and ease of operation. Symmetrically located to the right of the frequency dial is the output attenuator dial, directly calibrated in microvolts. The center panel enclosure embodies those controls which the operator will have the greatest occasion to use, permitting rapid, accurate settings to be made with maximum convenience.

The calibration accuracy of the frequency dial settings is ±0.25% at any point; however since crystal controlled frequencies are also available within the instrument, zero beats may be obtained from which the output frequency may be standardized to an accuracy of about ±0.025% by slipping the vernier frequency dial with respect to the main frequency dial. This feature permits the identification and checking of channel frequencies differing by as little as 100 kc.

Write today for complete details!

DESIGNERS AND MANUFACTURERS OF THE Q METER · QX CHECKER
FREQUENCY MODULATED SIGNAL GENERATOR · BEAT FREQUENCY
GENERATOR AND OTHER DIRECT READING INSTRUMENTS

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

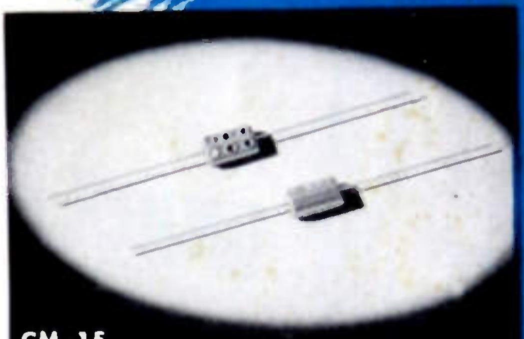
Capacity

COMES IN MANY SIZES

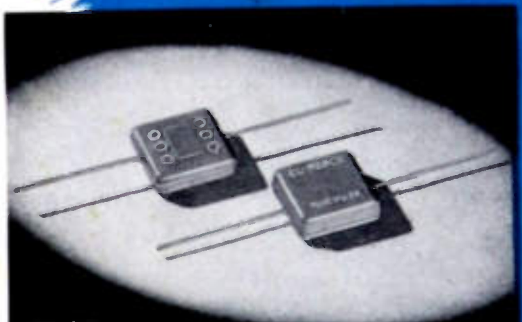
Whether you're talking in the simple terms of drinking water . . . a drink for yourself . . . the needs of a construction crew . . . or designing the latest in electronics equipment . . . capacity is important on every job. El-Menco Silvered-Mica Capacitors meet exacting requirements over a wide range . . . from the tiny CM-15 (2-525 mmf. cap.) to the mighty CM-35 (3300-10000 mmf. cap.).

The safety factor of a half-filled jug is built into every El-Menco Capacitor. Each unit is factory-tested at *double* its working voltage. You are assured of dependability in every application. El-Menco Capacitors offer peak performance for all specified military capacities and voltages.

For higher capacity values — which require extreme temperature and time stabilization — there are no substitutes for El-Menco Silvered-Mica Capacitors.



CM-15



CM-35

Write on your business letterhead for catalog and samples.

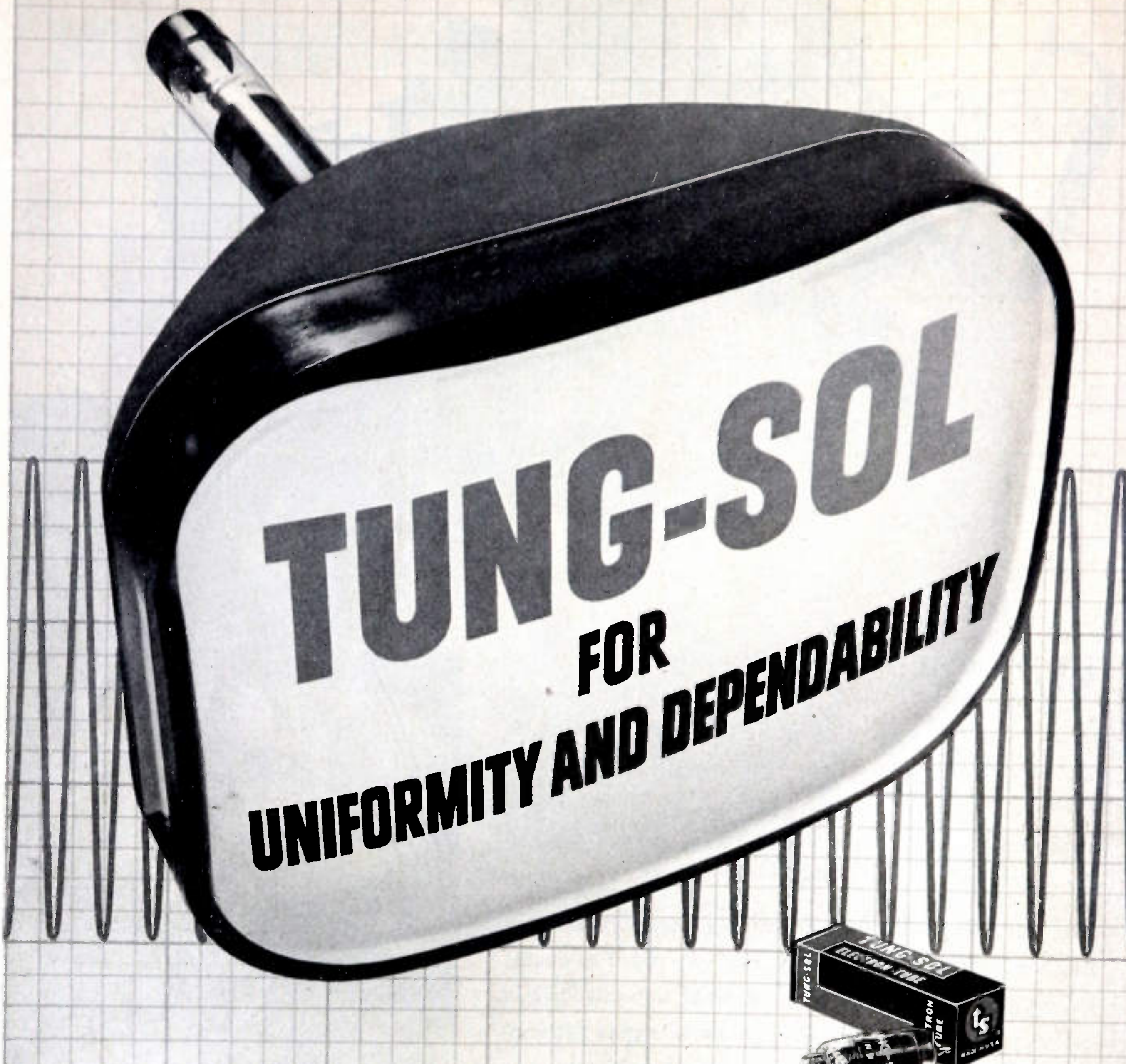
Jobbers, Retailers, Distributors—For information communicate direct with Arco Electronics, Inc., 103 Lafayette St., New York, N. Y.

MOLDED MICA **El-Menco** MICA TRIMMER

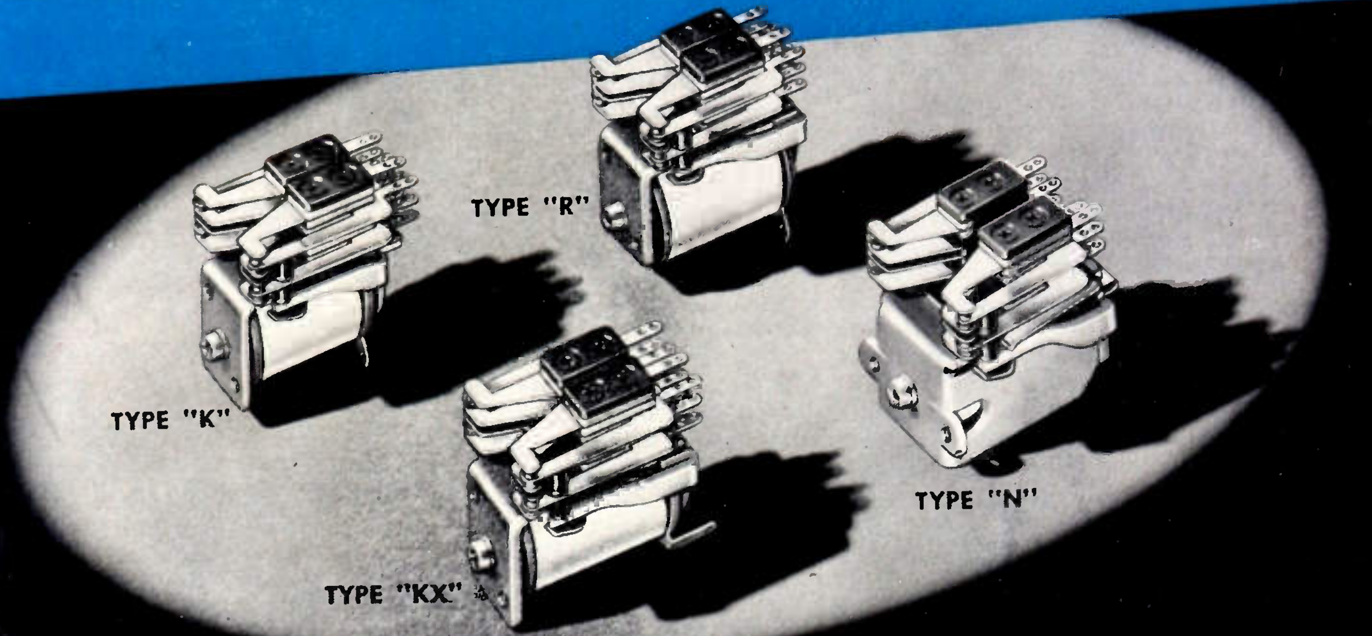
CAPACITORS

Radio and Television Manufacturers, Domestic and Foreign, Communicate Direct With Factory—

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


TUNG-SOL
FOR
UNIFORMITY AND DEPENDABILITY

THESE FOUR CLARE RELAYS offer designers *everything* they demand from small size, lightweight relays!



Special Features

which make these relays first choice of exacting design engineers

THE CLARE Type "K" RELAY

Fast operation—adequate contact pressure—high resistance to shock and vibration—long life—small size—$1\frac{1}{2}$" x $1\frac{1}{4}$" x $\frac{13}{16}$".

THE CLARE Type "KX" RELAY

Same features as Type "K" but adds greater operating range and increased sensitivity by use of a slightly longer coil which can be safely wound to a maximum resistance of 8000 ohms. Size: $1\frac{39}{64}$" x $1\frac{15}{32}$" x 1".

THE CLARE Type "R" RELAY

Same basic features as Type "K" but adds still further sensitivity and operating range to the Type "KX" by the use of a coil which is not only longer but of increased diameter to provide even greater winding space. Size: $1\frac{7}{16}$" x $1\frac{3}{64}$" x 1".

THE CLARE Type "N" RELAY

Same basic features as Type "K" but designed for operation on extremely low power. This results from a close-coupled magnetic circuit, generous use of magnetic iron and unusually efficient coil design. Size: $1\frac{5}{8}$" x $1\frac{5}{8}$" x $1\frac{5}{32}$".

The Clare Type "K" Relay was the first of the famous CLARE line of small, lightweight, telephone-type relays. It is still the mainstay of design engineers who must have superior relays to operate in extremely small space.

Its fast operation, adequate contact pressure, high resistance to shock and vibration, long life and complete all-around dependability have met many complex requirements.

The other three relays shown here supply all these advantages of the Type "K", but each adds certain other features to meet such specifications as a greater operating range, greater sensitivity or operation on extremely low power. All except the type "N" retain the reed armature suspension of special alloy which engineers recognize as one of the subtler reasons for the superior performance of these small CLARE Relays.

Clare sales engineers are located in principal cities to give you first hand information on the entire line of CLARE Relays . . . all of them designed to meet the most exacting relay requirements. Call them or write to C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials, Ltd., Toronto 13. Cable Address: CLARELAY.

All relays available in hermetically sealed form

All of these four CLARE small relays are available enclosed in hermetically sealed steel containers. Connection diagrams are permanently imprinted on the base of the enclosure by silk screen process.

See us at Booth No. 204 1952 Radio Engineering Show

CLARE RELAYS

. . . First in the Industrial Field

JOHNSON electronic components
...accepted by the Industry!

Types C & D

Rugged, dependable and low cost capacitors. Available in 52 standard single and dual models from 12 to 496 mmfd. and voltage ratings 3,500 to 11,000 volts. Extreme rigidity assured by heavy aluminum end frames, .051" plates, 5/16" tie rods. All are equipped with ¼" cadmium plated shafts, laminated phosphor bronze rotor contacts and Steatite high frequency insulators. Mounting brackets furnished for normal or inverted mounting. Panel space required: Type C, 5½" wide x 5⅜" high; Type D, 4¼" wide x 4" high.

Type L Variables

Built for the gruelling applications where dependable performance is a must. Tie rods soldered directly to ceramic end frames. Rotor and stator assemblies full soldered construction. No parts can work loose — capacity can't fluctuate. Metal parts are brass, plated with corrosion resistant Bright Alloy. Numerous sizes single, dual, butterfly and differential types from 11 to 200 mmfds. available with .030" spacing. Manufactured with .020", .060" and .080" spacing to special order in production quantities. Panel space required is 1⅜" square.

Type M Miniatures "Space Savers"

Smallest air variables ever built. Manufactured to extremely close tolerances these Lilliputs are quality condensers. Features of their construction are: Steatite insulation, split sleeve bearings, beryllium copper rotor contacts. Finish of metal parts is nickel plate. Capacitances of stock models range from 1.5 to 19 mmfd. in single, butterfly and differential types. Mounting area required is ¾" x ⅝". Flats on single hole mounting bushing keep condenser accurately positioned. 3/16" diameter shaft is slotted, may be tuned with either knob or screw driver.

JOHNSON manufactures a tube socket for virtually every transmitting and industrial application. Characteristics of all types are: functional design, low loss insulation and permanently positive contacts.

BAYONET SOCKETS

The 123-211SB "fifty watt" socket pictured is an example of JOHNSON bayonet type transmitting sockets. Steatite base insulation extends under contacts to prevent arcs to ground. Double beryllium copper filament contacts handle high current with minimum heating. Heavy aluminum shell provides rigid support for the tube in any position.

SEPTAR SOCKETS

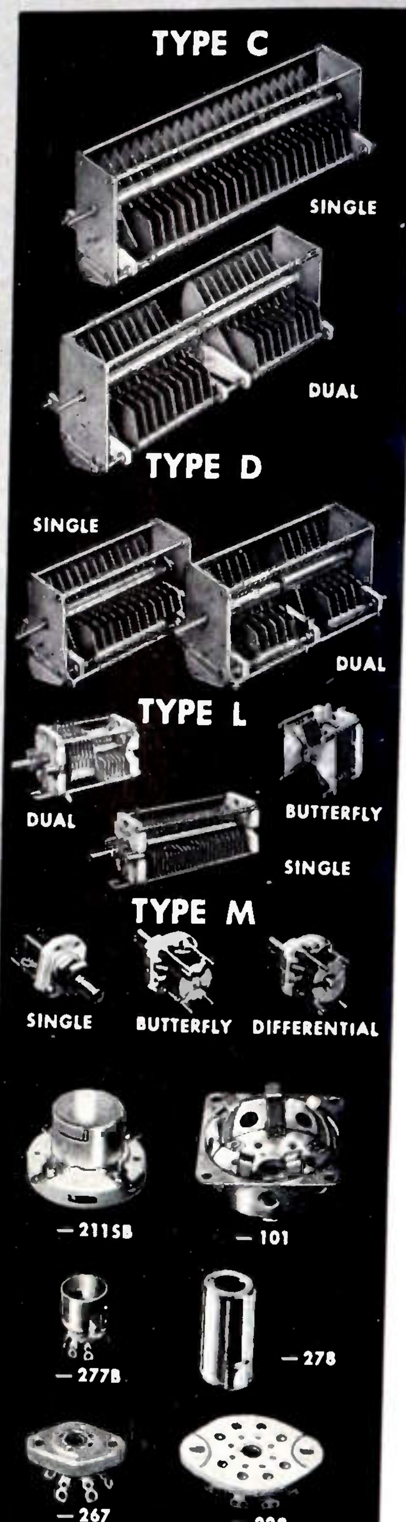
The 122-101 socket accommodates all Septar based VHF tubes such as B29's and B32's. The ventilated base shield is designed for mounting of button mica bypass condensers at the tube terminals. Contacts and contact springs are silver plated and recessed in the ceramic insulation to prevent axial movement. Contacts and retaining springs hold tube securely.

MINIATURE SOCKETS

For 7 pin miniature tubes JOHNSON offers the 120-277B miniature socket equipped with shield base and 120-267 for applications not requiring shields. Both have Steatite insulation with floating phosphor bronze silver plated contacts. Lengths of shields available for the 120-277B are 1⅜", 1¾" and 2¼".

WAFER SOCKETS

JOHNSON wafer sockets are simple and sturdy. Insulation is grade L-4 or better glazed and impregnated Steatite. Contacts are brass with steel retaining springs and cadmium plated. Rivet heads are recessed and mounting holes bossed to permit sub-chassis mounting. Locating grooves facilitate tube insertion. Available for 4 pin 5, 6, 7 pin med. and octal based tubes.



JOHNSON... *a famous name in Radio!*

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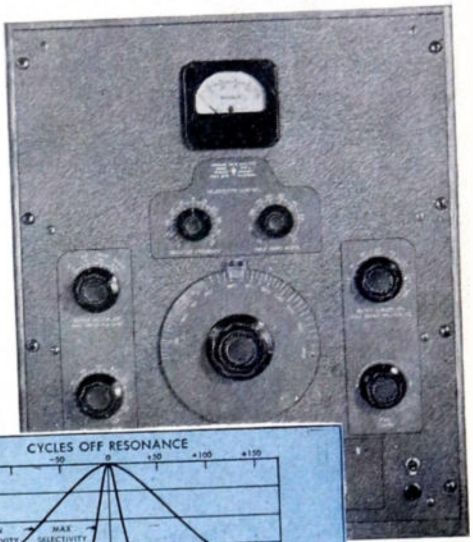
. . . . 20 cps to 20 kc

For MEASURING INDIVIDUAL WAVE COMPONENTS



-*hp*- 300A HARMONIC WAVE ANALYZER (right) is a selective voltmeter measuring the value of individual components of complex waves. Its variable selectivity is the key to its speed and versatility of operation. When wave components are close together, a unique selective amplifier can be narrowed to accept only desired components. When components are far apart, selectivity may be broadened to speed measuring without sacrificing accuracy. This feature is also important when measuring waves (such as in sound tracks) where some FM is present. The equipment is also ideal for analysis of noise, broadcast amplifier and network characteristics, recording devices, rotating machinery, hum and for all types of audio distortion measurements.

-*hp*- 300A is direct reading, covers the audio spectrum from 30 cps to 16,000 cps, and makes possible full-scale readings with inputs of 0.001 to 500 volts. Selectivity may be varied between limits shown in Fig. 2.

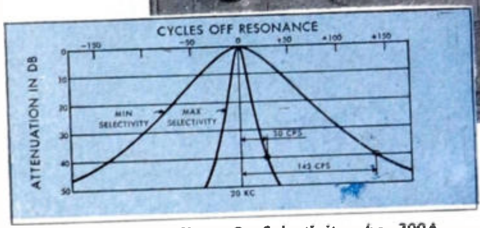


Figure 2. Selectivity, -hp- 300A

For TRANSIENT and FREQUENCY RESPONSE

-*hp*- 210A SQUARE WAVE GENERATOR provides a convenient, rapid method of determining transient and frequency response in a single measurement. It is widely used for testing receivers, video amplifiers, networks and transmitters; to measure time constants or provide a time base; to check cathode sweep circuits, indicate phase shift, transient effects or frequency response, to generate harmonics or control electronic switchers.

The 210A is an excellent, easy-to-use source of square waves for production line tests and laboratory use. High-quality square waves are generated over frequencies from 20 cps to 10 kc, and the equipment provides usable square waves up to 100 kc.

LOW COST DISTORTION ANALYZERS

-*hp*- 320A/B DISTORTION ANALYZERS are simple, low-cost devices for determining total harmonic distortion in audio frequency apparatus. They are particularly useful for high-speed production tests. -*hp*- 320A operates at two fixed frequencies: 400 and 5,000 cps. -*hp*- 320B operates at five fixed frequencies: 50, 100, 400, 1,000 and 5,000 cps. Both models require an external detector.

ANALYZER	PRIMARY USES	FREQUENCY RANGE	CHARACTERISTICS	PRICE
-hp- 300A	Wave form analysis.	30 cps to 16 kc	Variable selectivity, measuring range 1 mv to 500 v. 5% accuracy.	$625.00†
-hp- 320A	Measures total harmonic distortion at 2 fixed frequencies.	400 cps and 5 kc	Requires separate detector.	$75.00*
-hp- 320B	Measures total harmonic distortion at 6 fixed frequencies.	50, 100, 400 cps; 1, 5 and 7.5 kc	Same as above.	$150 00*
-hp- 330B	Measures total distortion, frequency tunable.	20 cps to 20 kc	Includes input amplifier and VTVM.	$395.00†
-hp- 330C	Similar to 330B. For FM broadcast measurements.	20 cps to 20 kc	VTVM has special characteristics to meet F C.C. requirements.	$425 00†
-hp- 330D	Similar to 330B. For AM and FM broadcast measurements.	20 cps to 20 kc	Includes AM detector and special meter to meet F.C C. requirements.	$440 00†

SQUARE WAVE GENERATOR

-hp- 210A	For rapid determination of transient and frequency response.	20 cps to 10 kc	Output 50 v peak-to-peak. 1,000 ohm impedance.	$150 00*

*Rack mounting available at $5.00 extra cost.
†Rack mounting available at no extra cost.

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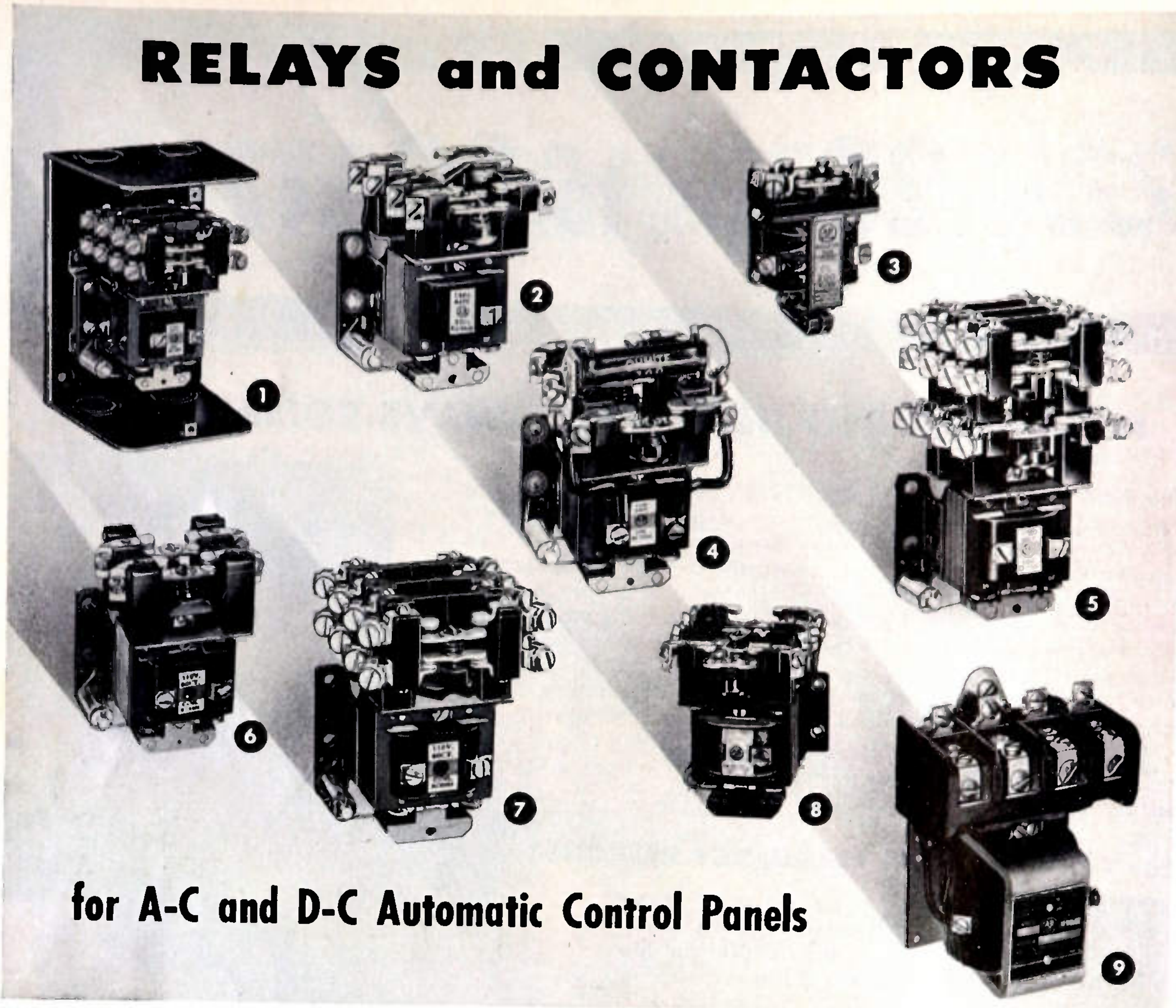
hp INSTRUMENTS ***Complete Coverage***

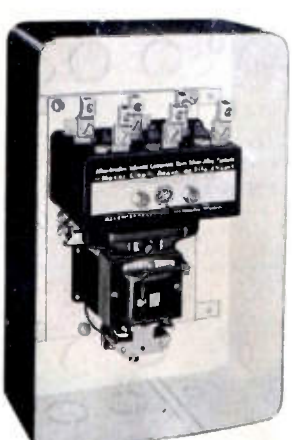


Do you need small relays from one to eight poles . . . or contactors up to 900 amperes? You will find these units, and many more, in the Allen-Bradley line, factory-tested for millions of maintenance-free operations.

Type BX universal relays have interchangeable normally open and normally closed contacts. No assembling to change from normally open to normally closed contacts. A few of the relays in the Allen-Bradley line are illustrated above.

Write for Bulletins 700 and 200.

1. Type BX Universal A-C Relay—in Enclosure
2. Type BM Mechanically Held Relay—No Hum
3. Type CL Low Coil Current Relay
4. Type BA 3-Wire Thermostat Relay—Open
5. Type BX 8-Pole Universal Relay—Open
6. Type B 2-Pole A-C Relay—Open
7. Type BX Universal A-C Relay—Open
8. Type BM Mechanically Held Relay—Open
9. Bulletin 202 2-Pole D-C 25-Ampere Relay



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• TYPE F-211

For .096" hole, order F261

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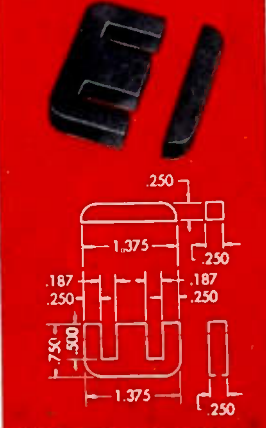
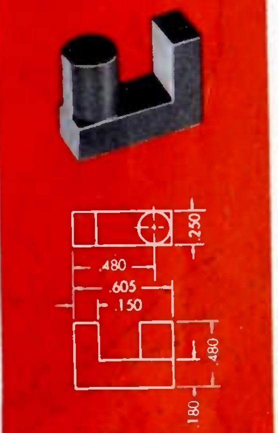


• TYPE F-167



Ferramics offer many important advantages as an electro-magnetic core material. The result has been wide adoption of this material in commercial and military electronic applications. We would welcome an opportunity to tell you how Ferramics can improve your components. For complete information call or write today.

PROPERTY	UNIT	TYPE OF FERRAMIC MATERIAL					
		B-90	C-159	E-212	H-419	I-141	J-472
Initial permeability at 1mc/sec	—	95	220	750	850	600	330
Maximum permeability	—	183	710	1710	4300	1010	750
Saturation flux density	Gauss	1900	3800	3800	3400	1540	2900
Residual magnetism	Gauss	830	2700	1950	1470	660	1600
Coercive force	Oersted	3.0	2.1	0.65	0.18	0.40	.80
Temperature coefficient of initial permeability	%/°C.	0.04	0.4	0.25	0.66	0.3	0.22
Curie point	°C.+	260	330	160	150	70	180
Volume resistivity	Ohm-cm	2×10^5	2×10^3	4×10^5	1×10^4	2×10^5	—
Loss Factor: at 1 mc/sec	—	.00016	.00007	.00008	.00030	.0003	.000055
at 5 mc/sec	—	.0011	.0008	.002	.00155	.005	—
at 10 mc/sec	—	—	—	—	.00275	—	—

High frequency materials are available up to approximately 150 megacycles; write for details.

• TYPE F-108

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- **TYPE F-219** 8 — 32 x .500
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- **TYPE F-244** 1/4 — 28 x .750
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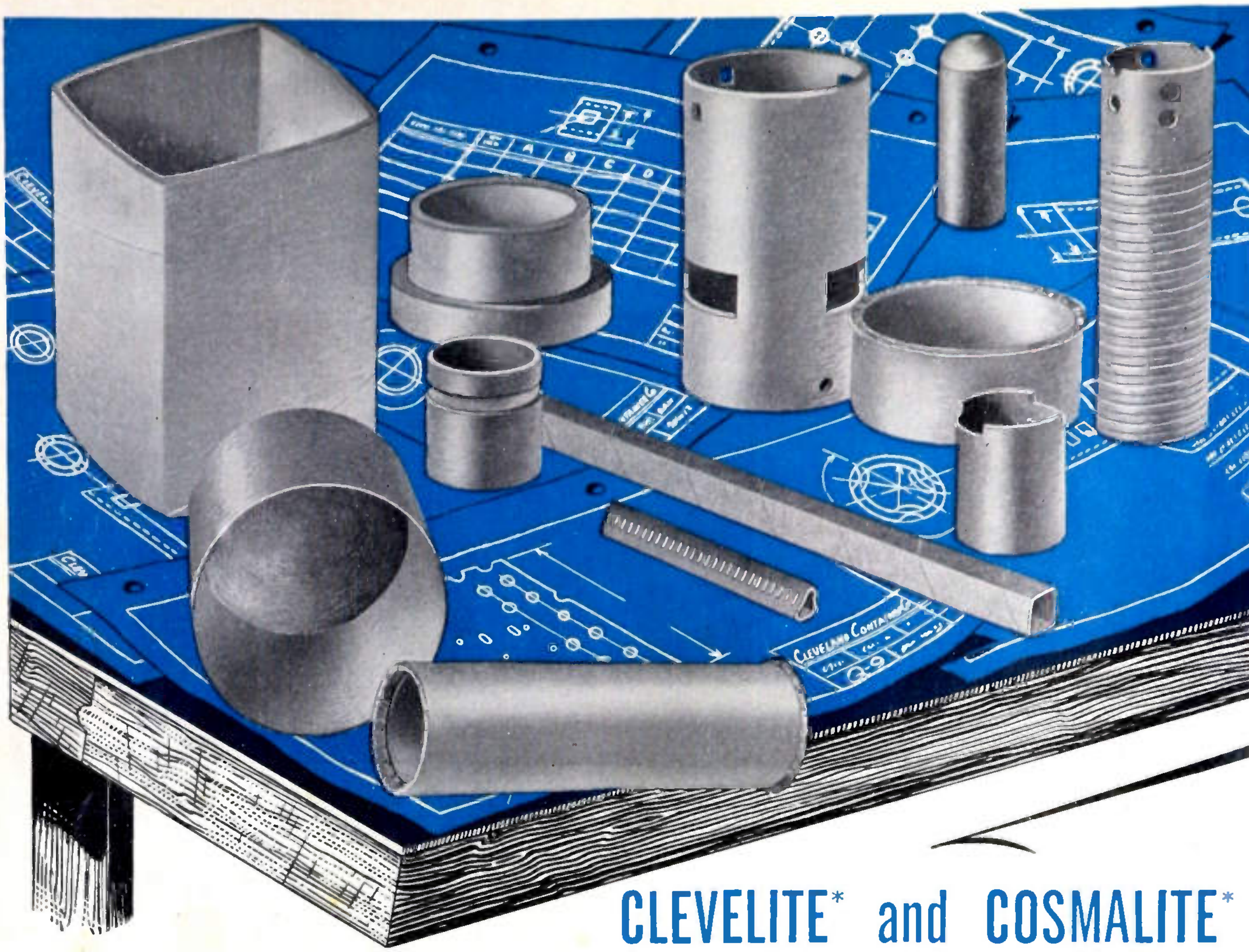

NEW! FOR VHF-UHF... **PRD** TYPE 907

sweep frequency generator

FREQUENCY RANGE:
35 TO 900 MEGACYCLES

MINIMUM OUTPUT VOLTAGE:
1 VOLT

DIRECT READING FREQUENCY DIAL:
CONTINUOUSLY VARIABLE

OUTPUT IMPEDANCE:
75 OHMS-BNC CONNECTOR

MINIMUM SWEEP WIDTH ABOVE 60 MC/S:
20 MC/S

The Type 907 is a fundamental oscillator which can be swept in frequency over a band of not less than 10 mc/s for a center frequency of 35 mc/s. The sweep width is greater than 20 mc. for carrier frequencies above 60 mc/s. Output is continuously variable over a voltage range of 10 microvolts to 1 volt. Internal blanking circuits provide a "true zero" base line for an oscilloscope display.

For further information concerning this instrument and additional UHF-VHF equipment, address inquiries to Dept. R1, or visit us at the IRE Show, Booths 268-269.

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Type 396-A Balun
Balance-Unbalance Transition

provides a low VSWR transition between 50 ohm unbalanced to 300 balance transmission line over a frequency range of 470 to 890 mc/s.

Type 904 VHF-UHF Noise Generator

permits direct measurements of noise factors as high as 20 db for r-f amplifiers and receivers operating from 10 to 1000 mc/s.

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is a high Q Frequency Meter covering the band of 470 to 890 mc/s.

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The newest addition to Sperry's Microline* is Model 296B Microwave Receiver for laboratory use. This instrument is an important addition to the microwave laboratory where a good secondary standard of attenuation is required.

The versatility of Model 296B permits measurements to be made at all microwave and UHF frequencies. In addition to its use as a secondary standard of attenuation, this receiver has many other uses . . . one of the more important being antenna pattern measurements.

NEW SPERRY MICROLINE RECEIVER FOR ACCURATE MEASUREMENTS AT MICROWAVE FREQUENCIES

Model 296B consists of a 30 mc pre-amplifier, IF amplifier and precision 30 mc waveguide below cut-off attenuator. Included in the receiver is a well-regulated klystron power supply. Klystron stability is assured by self-contained, automatic frequency control circuitry.

Our Special Electronics Department will be happy to give you further information on this instrument as well as other Microline equipment.

*T. M. REG. U. S. PAT. OFF.

Model 296B Microwave Receiver

30 Mc Amplifier Gain: 70 db + 30 db preamp gain — 15 db insertion loss.

IF Bandwidth: 1.8 Mc.

Attenuator: Insertion loss 15 db; 80 db attenuation range with detent positions at 10 db steps.

Local Oscillator Power Supply: Beam supply 600 to 800 volts 50 ma, continuously variable, positive grounded. Reflector supply continuously variable from —10 to —500 volts with respect to cathode.

Accessories Supplied: One pre-amplifier, one pre-amplifier power cable, one klystron power cable, two 30 Mc IF cables.

Accessories Needed: Local Oscillator Klystron and a mixer.

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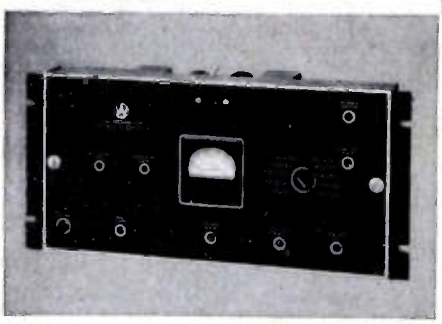

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Pan American World Airways System Selects WILCOX Type 408A Transmitter and Type 305A Receiver for Idlewild International Airport VHF Installation

New features offered in fixed frequency equipment for the 118-132 mc/s band:

50 WATT TYPE 408A TRANSMITTER

1. Automatic coaxial transfer relay permits operation of transmitter and receiver from same antenna.
2. Compact design requires only 8¾ inches of rack space.
3. Only thoroughly proven octal and transmitting type tubes are used in R.F. stages.
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5. .005% frequency stability without temperature control.

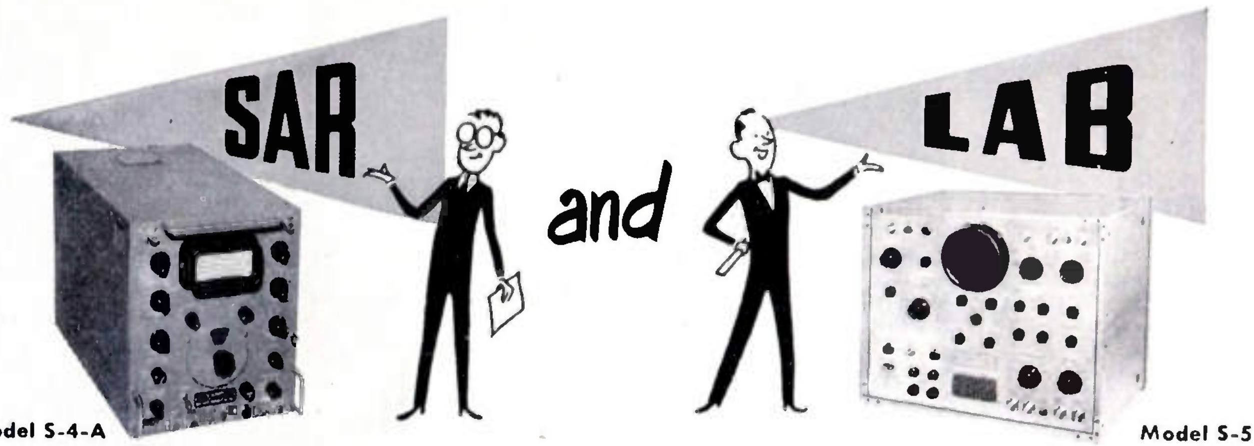
TYPE 305A RECEIVER

1. New noise limiter means better reception. With a pulse type noise 33 times as strong as the desired signal, the receiver output is clearly intelligible.
2. Spurious frequency responses are all at least 100 db *below* the desired signal.
3. Selectivity permits 100 Kc. adjacent channel operation.
4. The front panel is easily removable, exposing all mounted parts for inspection and maintenance.
5. Simple, conventional circuits minimize the number and types of tubes and require no special training, techniques, or test equipment.

Write Today for complete information and specifications on the Wilcox Type 408A Transmitter and the Type 305A Receiver.

WATERMAN PIONEERING

PULSESCOPE®



Model S-4-A

Model S-5-A

PULSESCOPES are Oscilloscopes to portray the attributes of the pulse: such as shape, amplitude, duration and time displacement. Both of the PULSESCOPES have Video amplifiers with frequency response up to 11 megacycles with Video delay of 0.55 microseconds and pulse rise and fall time better than 0.07 microseconds.

S-4-A SAR PULSESCOPE—Video Sensitivity 0.5vp to p/in. S Sweep 80 cycles to 800KC, either trigger or repetitive. A Sweep 1.2 microseconds to 12,000 microseconds. R Delay 3 microseconds to 10,000 microseconds directly calibrated on precision dial. R Pedestal (or Sweep) 2.4 microseconds to 24 microseconds. Internal Crystal Markers 10 microseconds and 50 microseconds. Size $9\frac{1}{8}$ x $11\frac{1}{4}$ x $17\frac{1}{4}$″. Weight: Less than 32 pounds.

S-5-A LAB PULSESCOPE—Video Sensitivity 0.1vp to p/in. Sweep 1.2 microseconds to 120,000 microseconds with 10 to 1 expansion. Sweep either trigger or repetitive. Internal Markers synchronized with Sweep from 0.2 microseconds to 500 microseconds. Trigger Generator and built-in precision amplitude calibrator. Completely cased. Size: $16\frac{1}{2}$ x $14\frac{1}{8}$ x $14\frac{1}{2}$″. Weight: Less than 60 pounds.

WATERMAN RAYONIC TUBE DEVELOPMENTS

3 SP

Since the introduction of Waterman RAYONIC 3MP1 tube for miniaturized oscilloscopes, Waterman has developed a rectangular tube for multi-trace oscilloscopy. Identified as the Waterman RAYONIC 3SP, it is available in P1, P2, P7 and P11 screen phosphors. The face of the tube is $1\frac{1}{2}$″ x 3″ and the over-all length is $9\frac{1}{4}$″. Its unique design permits two 3SP tubes to occupy the same space as a single 3″ round tube, a feature which is utilized in the S-15-A TWIN-TUBE POCKETSCOPE. On a standard 19″ relay rack, it is possible to mount up to ten 3SP tubes with sufficient clearances for rack requirements. All RAYONIC cathode ray tubes are available in P1, P2, P7 and P11 phosphors. We are authorized to supply 3SP1, 3JP1 and 3JP7 with JAN stamp. All RAYONIC tubes listed below operate on 6.3 volts heater with .6 amp. current.

3 MP

TUBE	PHYSICAL DATA			TYPICAL VOLTAGES				DEFLECTION FACTOR V/IN.		MAX. VOLTS	
	Face	Length	Base	Anode #3	Anode #2	Anode #1	Grid #1	D1 to D2	D3 to D4	Anode #3	Anode #2
3JP	3 inch Round	10 inches	Medium Diheptal 12 Pin	3000	1500	300 to 515	−22.5 to −67.5	127 to 173	94 to 128	4000	2000
				4000	2000	400 to 690	−30 to −90	170 to 230	125 to 170		
3MP	3 inch Round	8 inches	Small Duodecal 12 Pin		1000	200 to 350	0 to −68	140 to 190	130 to 180		2500
					2000	400 to 700	0 to −126	280 to 380	260 to 360		
3SP	1½ x 3 inches	9.12 inches	Small Duodecal 12 Pin		1000	165 to 310	−28.5 to −67.5	73 to 99	52 to 70		2750
					2000	330 to 620	−58 to −135	146 to 198	104 to 140		

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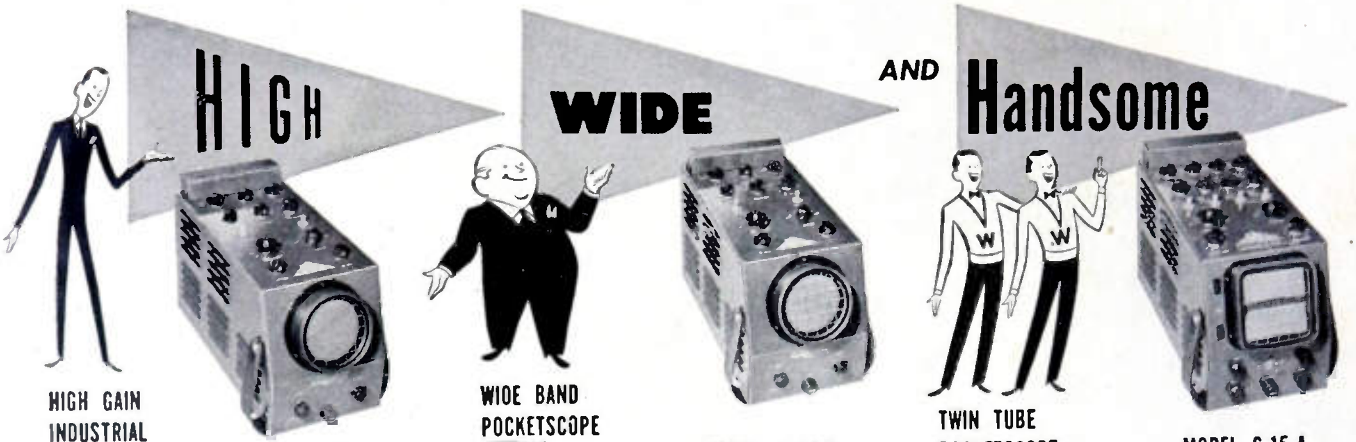
THE WATERMAN LINE-UP

POCKETSCOPE®



HIGH GAIN INDUSTRIAL POCKETSCOPE — MODEL S-14-A

WIDE BAND POCKETSCOPE — MODEL S-14-B

TWIN TUBE POCKETSCOPE — MODEL S-15-A

HI, WIDE and HANDSOME POCKETSCOPES are characterized by small size, light weight, and outstanding electrical performance. All units have frequency compensated attenuators as well as non-frequency discriminating gain controls. All units have both periodic and trigger sweeps from ½ cycle to 50KC. The amplifiers are direct coupled thus frequency response starts from 0 cycles. No peaking coils are used, thus, the transient response is good. Full expansion of trace, both vertical and horizontal, is built in. Means for amplitude calibration are provided. DC coupling in POCKETSCOPES provides unusual stability of the trace, regardless of the line voltage changes or variations of impedances in the input circuit. The HI, WIDE and HANDSOME POCKETSCOPES are the outgrowth of Waterman pioneering of the first commercial miniature oscilloscope, which has proved to be useful and reliable over a period of years. Combination filter and graph screens are used for better visibility, thus traces can be observed even under high ambient light conditions. Binding posts for convenience of connections, with an effective shield, are used. S-14-A has sensitivity of 10 mv/inch with pass band above 200KC. S-14-B has sensitivity of 50 mv/inch with pass band above 1 megacycle. S-15-A is similar to S-14-A except that it has two independent CR Tubes for multi-trace oscilloscope work. Accessories such as carrying cases and probes are available.

S-11-A

The Model S-11-A Industrial & Television POCKETSCOPE is a small, compact, lightweight instrument for observation of repetitive electrical circuit phenomena. The Industrial & Television POCKETSCOPE is a complete cathode ray oscilloscope incorporating the cathode ray tube, vertical, horizontal, and intensity amplifiers, linear time base oscillator, blanking, synchronization means and self-contained power supply. The Industrial & Television POCKETSCOPE can be used, not only for AC measurements, but for DC as well, inasmuch as it has vertical and horizontal amplifiers which are capable of reproducing faithfully within —2 db, from 0 to 200KC. The sensitivity of the vertical and horizontal amplifiers is high and is in the order of 100 mv rms/in.

S-12-B

Model S-12-B RAKSCOPE has the features of S-11-A POCKETSCOPE, plus. The RAKSCOPE is JANized and the government model number is OS-11. The Sweep, from 5 cycles to 50KC is either repetitive or triggered. Vertical and horizontal amplifiers are 50 millivolts rms per inch with band pass from 0 to 200KC. Special calibrating circuitry is provided for frequency comparison. Both the vertical and horizontal amplifiers are identical and use no peaking. The panel is only 7″ high and the scope fits standard rack. The functional layout of the control permits ease of operation.

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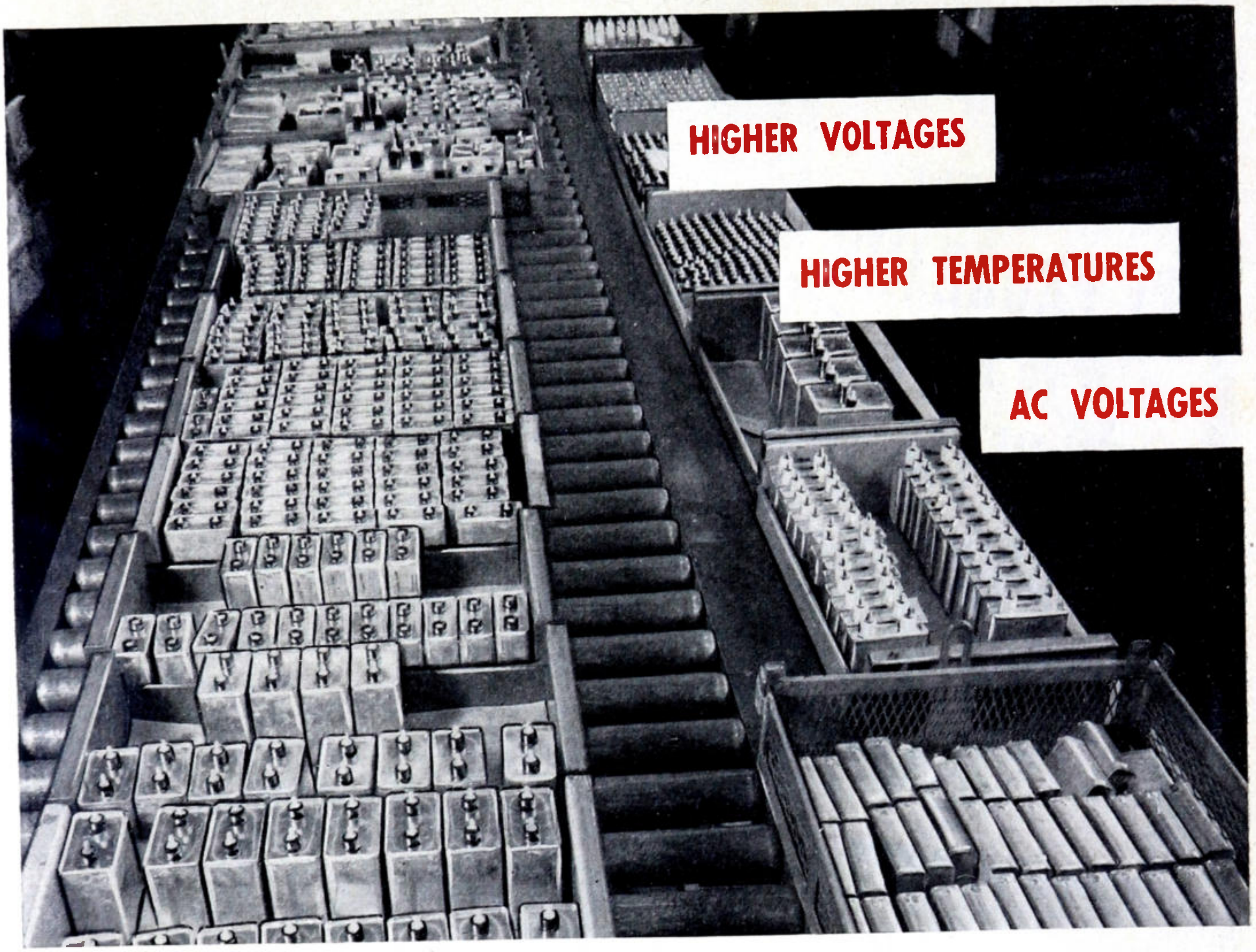
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AC VOLTAGES

WHBF's TV tower, with an overall height of 482 ft., was mounted on a specially constructed substructure 61 ft. high. Tower is designed to mount station call letters on all 4 sides, and carries an RCA custom-built, 5-section, Super Turnstile antenna.

Here is a situation that called for initiative and foresight—as well as unique design-engineering.

WHBF owns a downtown site on which they will erect a five-story building when material allocations permit. In the meantime, their TV license would be in disuse without proper antenna support. The problem was put up to Blaw-Knox . . . the solution is shown above—a permanent "tax-paying" base around which WHBF will eventually erect its new quarters.

BLAW-KNOX DIVISION OF BLAW-KNOX COMPANY
2037 Farmers Bank Building, Pittsburgh, Pa.

Distributed by Graybar

BLAW-KNOX *Antenna* TOWERS

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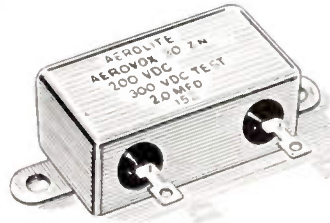
Series P92ZN Aeralene-impregnated metallized-paper capacitors are modified plastic-tubular, duranite-end-sealed units in paper cases. Operating temperatures of −30°C. to +100°C. 200, 400 and 600 V. D.C. 0.01 to 2.0 mfd.

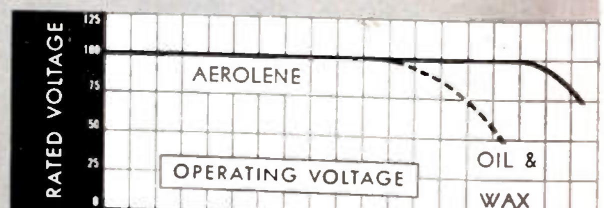
Series P123ZNG Aeralene-impregnated metallized-paper capacitors housed in tubular metal cases with vitrified ceramic terminal seal. Operating temperature range of −55°C. to +100°C. at full rating: to +125°C. at 75% of voltage rating. 200, 400 and 600 V.D.C. .0005 to 2.0 mfd.



Series P30ZN Aerolene-impregnated metallized-paper capacitors housed in "bathtub" metal cases with vitrified or glass terminal seals. Operating temperature range of −55°C. to +100°C. at full rating; to +125°C. at 75% of voltage rating. 3.0 mfd. to 15.0 mfd. for 150 V.D.C. units. 2.5 mfd. or less for 300 V.D.C. units.

Once again, Aerovox is privileged to blaze the capacitor-development trail. For these high-temperature metallized-paper capacitors are definitely Aerovox "firsts" in *conception, production and application.*

Their truly phenomenal acceptance is due to (1) The *Space Factor*, especially when *miniaturization* is a prime consideration; (2) *Reliability*, particularly in meeting voltage peaks or surges, by taking advantage of their self-healing characteristics; and (3) *Wide Operating Range*, from sub-zero to elevated temperatures.



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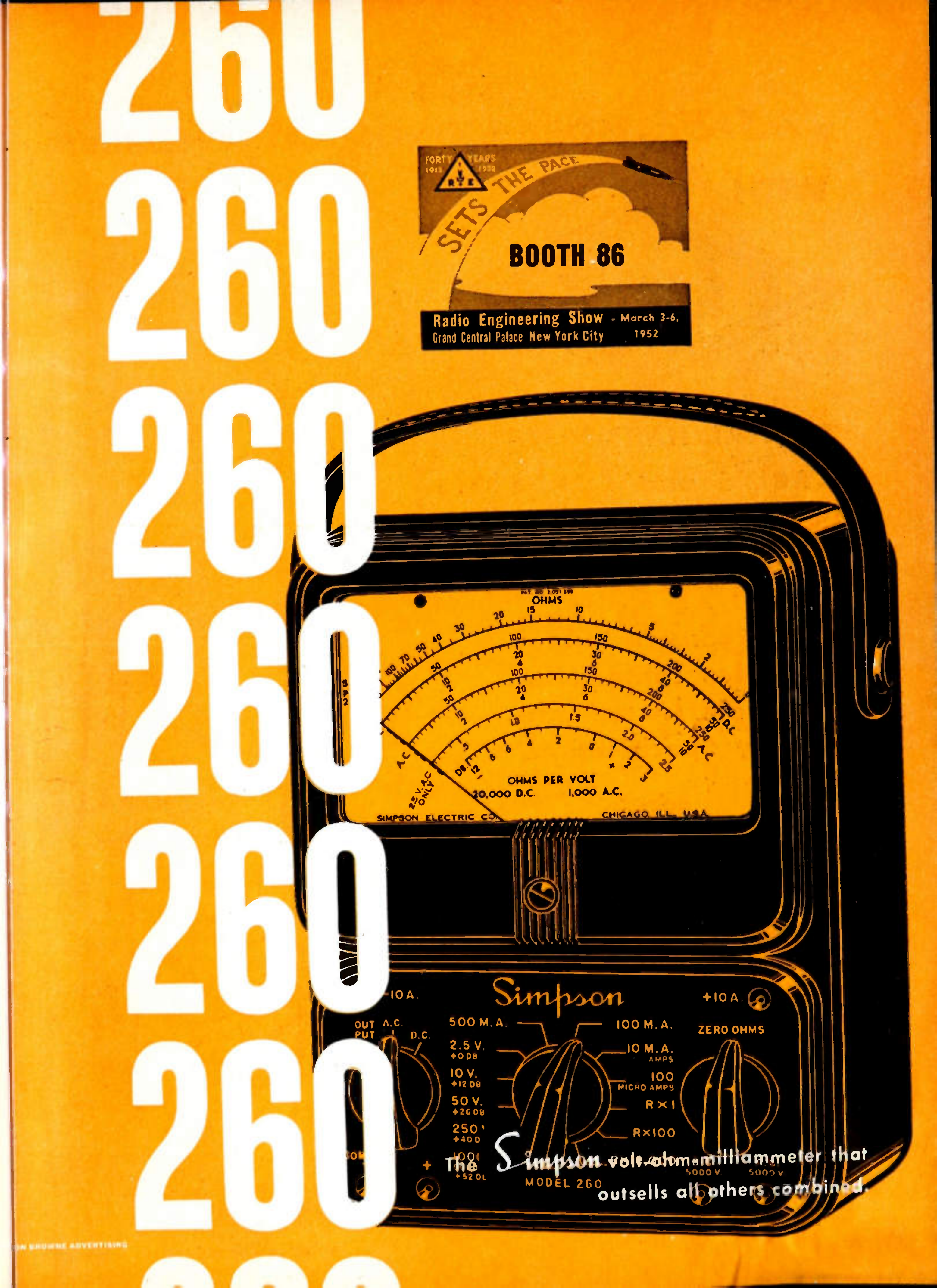
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designed for MINIATURIZATION, RUGGEDIZATION

Erie Ceramicons fulfill all the requisites for efficient by-passing—compact design, low inductance, and conservative 500 volt D. C. rating. Erie Resistor offers the most complete line of ceramic by-pass units available. Each design has been thoroughly proven in domestic and military equipment.

Eighteen popular styles in ceramic capacitors are shown above. Feed-Thru's are supplied in values up to 2000 mmf, Stand-Off units up to 5000 mmf, Tubular and Disc units up to .01 mfd. Also shown above are two Silver Button Micas representing the 370 series for values up to 1000 mmf and the 4700 series for values up to 6000 mmf. Write for samples to meet your specific requirements.

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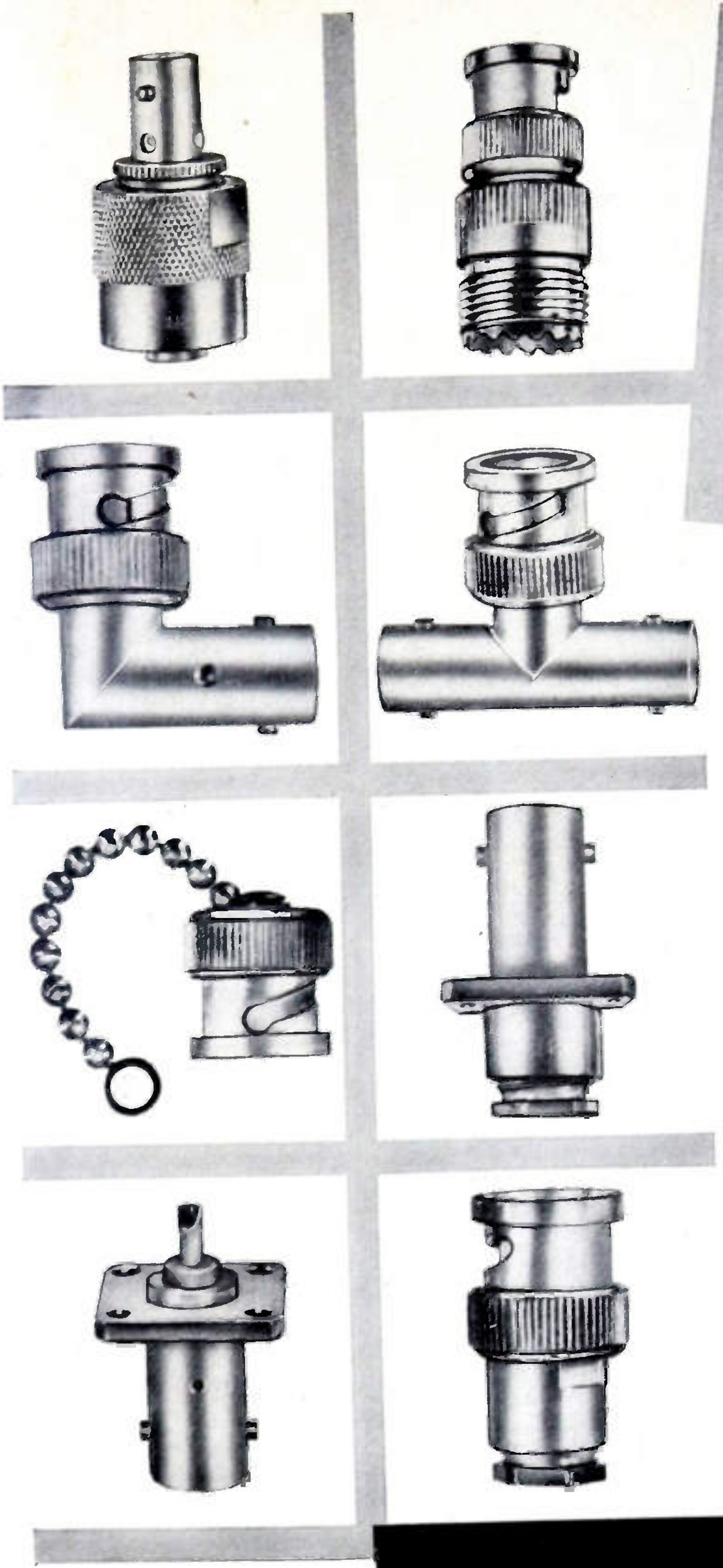

BNC CONNECTORS BY KINGS

The B N C Connectors shown are small, lightweight Connectors designed for use with small cables such as RG-58/u, RG-59/u and RG-71/u. Widely used for video and aircraft test equipment, they are recommended for frequencies as high as 3000 M.C., where impedance matching is important. The B N C series is used successfully in the region of microwave frequencies.

Whether your connector requirements call for the B N C series, N series, the new C series or special adaptations of standard connectors, you can rely on Kings. Our staff of highly specialized engineers invite your inquiries.

SETS THE PACE

BOOTH 291

Radio Engineering Show - March 3-6, 1952
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KINGS *Electronics* CO., INC.

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IN CANADA: ATLAS RADIO CORP., LTD., TORONTO


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3 AMP 125VAC
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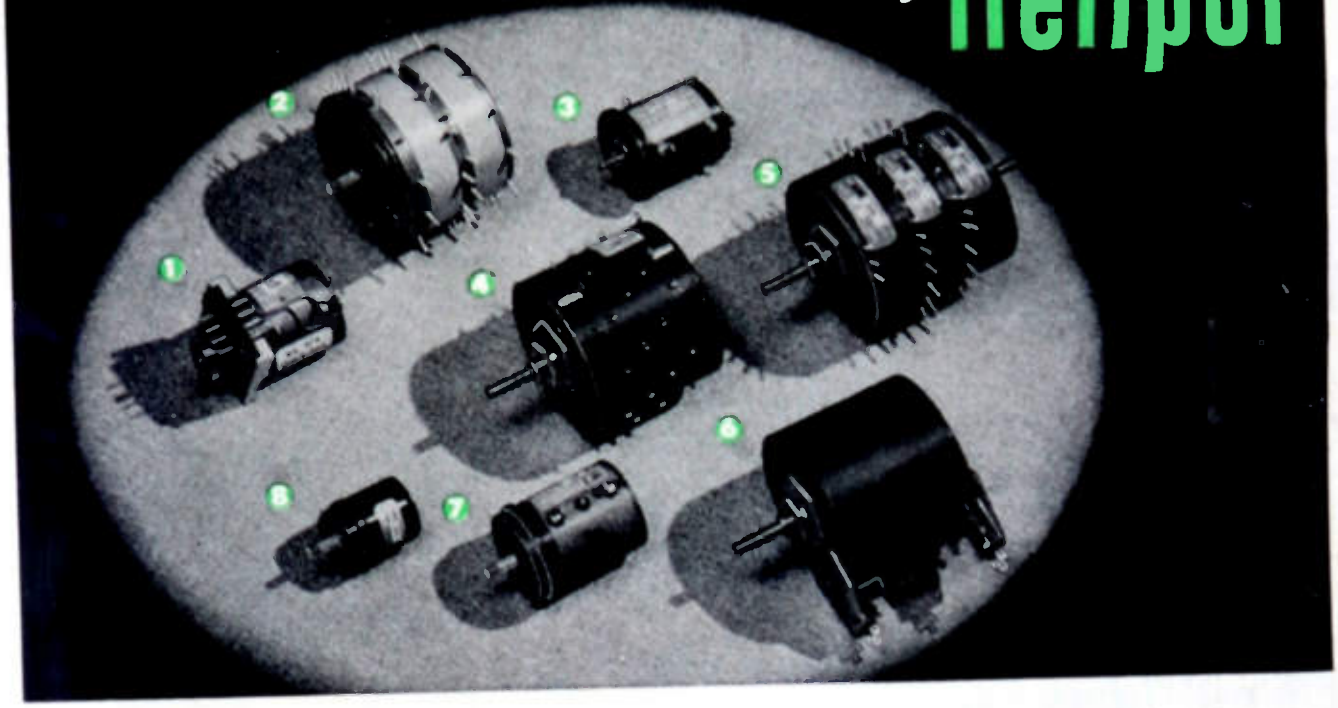


Precise Accuracy + Maximum Versatility + Space-saving Compactness

The potentiometers illustrated above are typical examples of the tough problems HELIPOT engineers are solving every day for modern electronic applications. If you have a problem calling for utmost precision in the design, construction and operation of potentiometer units—coupled with minimum space requirements and maximum adaptability to installation and operating limitations—bring your problems to HELIPOT. Here you will find advanced "know-how," coupled with manufacturing facilities unequaled in the industry!

The HELIPOTS above—now in production for various military and industrial applications—include the following unique features . . .

1. This 10-turn HELIPOT combines highest electrical accuracies with extremes in mechanical precision. It features zero electrical and mechanical blacklash...a precision-supported shaft running on ball bearings at each end of the housing for low torque and long life . . . materials selected for greatest possible stability under aging and temperature extremes . . . special mounting and coupling for "plug-in" convenience . . . mechanical and electrical rotation held to a tolerance of ½° . . . resistance and linearity accuracies, ±1% and ±0.025%, or better, respectively.

2. *This four-gang assembly of Model F single-turn potentiometers has a special machined aluminum front end for servo-type panel mounting, with shaft supported by precision ball bearings and having a splined and threaded front extension. Each of the four resistance elements contains 10 equi-spaced tap connections with terminals, and all parts are machined for greatest possible stability and accuracy.*

3. This standard Model A, 10-turn HELIPOT has been modified to incorporate ball bearings on the shaft and a special flange (or ring-type) mounting surface in place of the customary threaded bushing. This HELIPOT also contains additional taps and terminals at the ¼- and 9¾-turn positions.

4. *This standard Model B, 15-turn HELIPOT has a total of 40 special tap connections which are located in accordance with a schedule of positions required by the user to permit external resistance padding which changes the normally-linear resistance vs. rotation curve to one having predetermined non-linear characteristics. All taps are permanently spot-welded and short out only one or two turns on the resistance element—a unique HELIPOT feature!*

5. This six-gang assembly of standard Model F single-turn potentiometers has the customary threaded bushing mountings, and has shaft extensions at each end. The two center potentiometers each have 19 equi-spaced, spot-welded tap connections brought out to terminals. *Each tap shorts only two turns of .009" diameter wire on the resistance element.*

6. *This Model B, 15-turn HELIPOT has been modified to incorporate, at the extreme ends of mechanical and electrical rotation, switches which control circuits entirely separate from the HELIPOT coil or its slider contact.*

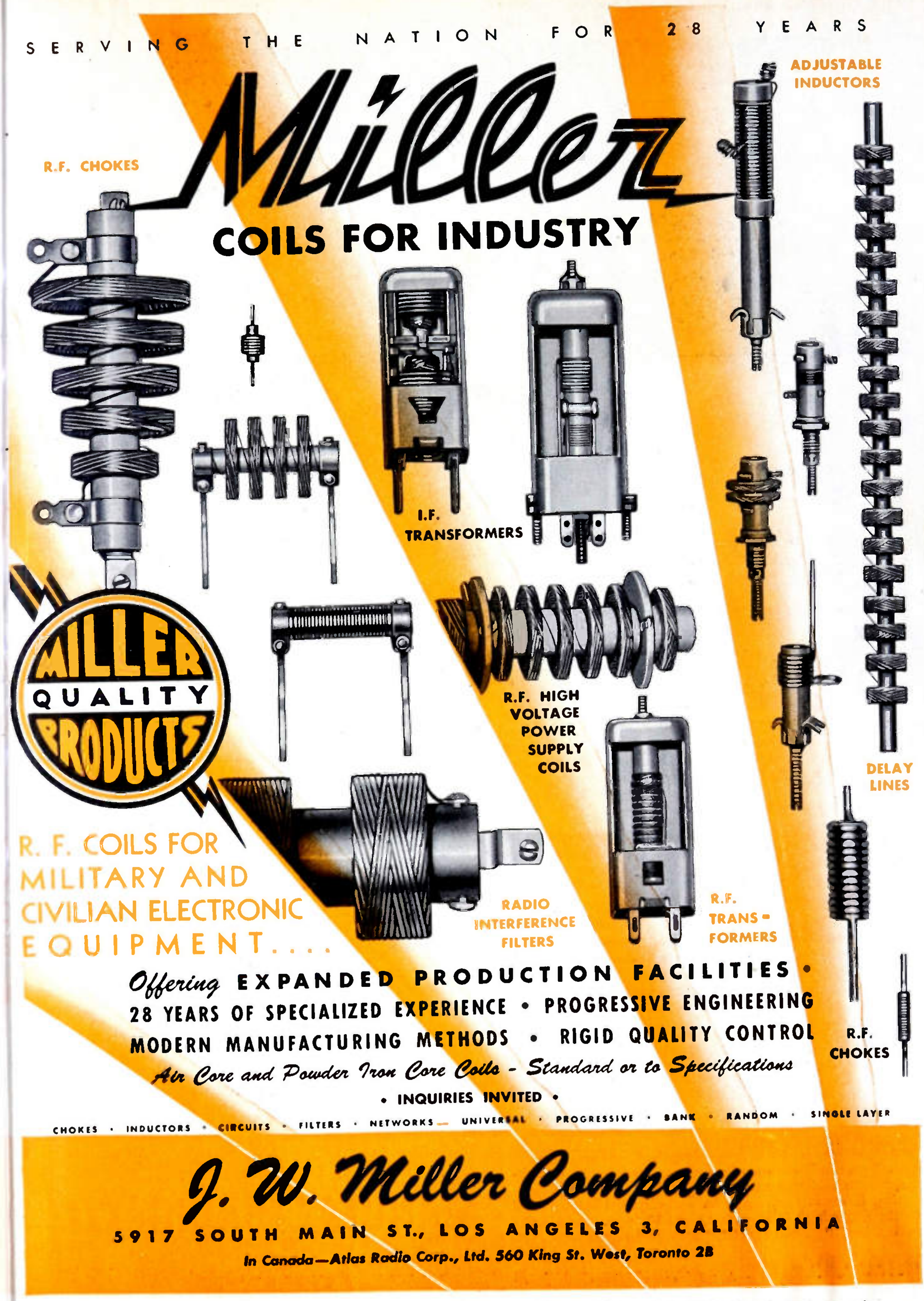
7. This 10-turn HELIPOT has many design features similar to those described for unit No. 1, plus the following additional features . . . a servo-type front end mounting . . . splined and threaded shaft extension . . . and a center tap on the coil. All components are machined to the highest accuracy, with concentricities and alignments held in some places to a few *ten-thousandths* of an inch to conform to the precision of the mechanical systems in which this HELIPOT is used. *Linearity accuracies frequently run as high as ±0.010%!*

8. *This single-turn Model G Potentiometer has been modified to incorporate a ball bearing shaft and a servo-type front end mounting. Special attention is given to contact designs and pressures to insure that starting torque does not exceed 0.2 inch-ounces under all conditions of temperature.*

The above precision potentiometers are only typical of the hundreds of specialized designs which have been developed and produced by HELIPOT to meet rigid customer specifications. For the utmost in accuracy, dependability and adaptability, bring your potentiometer problems to HELIPOT!

Representatives in all major areas of the United States. Export agents: Frathom Co., 55 W. 42nd St., New York 18.

THE Helipot CORPORATION, SOUTH PASADENA 6, CALIFORNIA



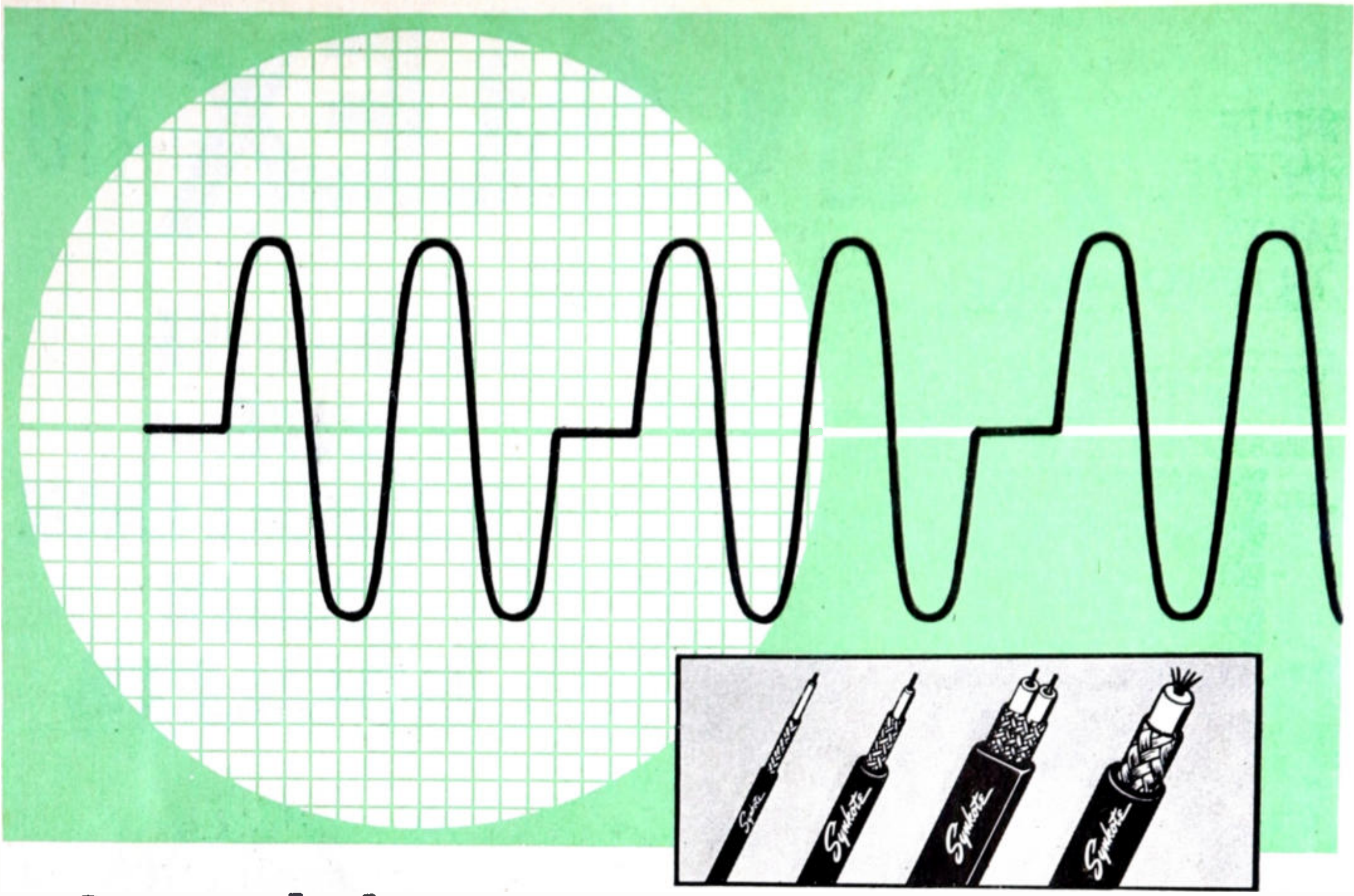
See our display in Booth 454—I.R.E. Show Grand Central Palace, New York City, March 3, 4, 5 and 6

What to see at the Radio Engineering Show

(Continued from page 2A)

Firm	Booth
W. H. Brady Co., Chippewa Falls, Wis. Self-sticking wire markers, special labels, safety signs, masks, and gaskets.	258
Brentano's Technical Dept., New York 19, N.Y. Latest technical books of all publishers in the related fields of radio, television, electronics, nuclear physics, and related mathematical subjects.	259
British Industries Corp., New York 13, N.Y. Garrard record changers and phono equipment, Ersin Multicore solder, KT66 power amplifying tube, leak "Point-One" amplifiers, Douglas and MacAdie automatic coil winding machines, Wharfedale speakers, Avometers (test equipment).	270
Browning Laboratories, Inc., Winchester, Mass. FM Tuner; FM-AM Tuners; FM Modulation monitor; oscillosynchroscope; synchroscope.	S-12, S-13
Brujac Electronic Corp., New York, N.Y. Electronic measuring instruments and components.	S-18, S-19
Brush Development Co., Cleveland 14, Ohio Cabinet rack with one type BL-246 Six Channel Combination Oscillograph, two type BL-962 dc Amplifiers and one type BL-360 Universal Strain Amplifier. One BL-928 Dual Channel Amplifier with type BL-201AR Single Channel Oscillograph with slow speed paper drive and Event Marker and BL-933 Takeup Reel. One BL-312 Universal Strain Bridge Switch and BL-222 Dual Channel Electric and Ink Writing Oscillograph. Piezoelectric crystal and ceramic elements. Laboratory generator and transducer.	70, 71, 72
Burlington Instrument Co., Burlington, Iowa Electrical indicating instruments, ac and dc, both standard and hermetically sealed.	228
Burroughs Adding Mach. Co., Philadelphia 23, Pa. Pulse control units—a line of standardized electronic building blocks: Pulse generators, flip flops, coincidence detectors, mixers, pulse gaters, etc., that can be combined in standard relay racks, operated from standard voltages, to form complex circuits.	413
Bussmann Mfg. Co., St. Louis 7, Mo. Fuses, fuse clips, fuse blocks and fuse holders.	371
CBS-Remington-Rand, Vericolor Hall See: Columbia Broadcasting	380
C. G. S. Laboratories, Inc., Stamford 1, Conn. S-Band Oscillator Cavity. Peak pulse voltmeter "Increductor," line of variable controlled inductors.	N-4
C&H Supply Company, Seattle 8, Wash. Metal-Cal identification name plates, and circuit diagrams.	346
Caldwell-Clements, Inc., New York 17, N.Y. *TELE-TECH*—Technical magazine of the radio-television-electronic industries. ***RADIO & TELEVISION RETAILING***—Magazine of radio-TV distribution and maintenance.	253
The Calidyne Co., Winchester, Mass Vibration Test Equipment—electro-dynamic Shakers—25, 450, and 2,500 pound force output. Electro-dynamic vibration pickup and Accelerometer calibration systems, accelerometers and accelerometer couplers, Vibrascope, Calivolters, vibration meter, and signal monitor for velocity signal generators.	273
Cambridge Thermionic Corp., Cambridge 38, Mass. Soldering terminals, terminal boards, IF and rf coils, electronic hardware (handles, shaft locks, tube clamps, standoffs, etc.) insulated standoffs and feed-thru's (ceramic and phenolic) slug-tuned coil forms (ceramic and phenolic).	287

(Continued on page 68A)



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MYCALEX is a highly developed glass-bonded mica insulation backed by a quarter-century of continued research and successful performance. Both pioneer and leader in low-loss, high frequency insulation, MYCALEX offers designers and manufacturers an economical means of attaining new efficiencies, improved performance. The unique combination of characteristics that have made MYCALEX the choice of leading electronic manufacturers are typified in the table for MYCALEX grade 410 shown below. Complete data on all grades will be sent promptly on request.

MYCALEX is efficient, adaptable, mechanically and electrically superior to more costly insulating materials

- **PRECISION MOLDS TO EXTREMELY CLOSE TOLERANCE**
- **READILY MACHINEABLE TO CLOSE TOLERANCE**
- **CAN BE TAPPED THREADED, GROUND, SLOTTED**
- **ELECTRODES, METAL INSERTS CAN BE MOLDED-IN**
- **ADAPTABLE TO PRACTICALLY ANY SIZE OR SHAPE**

MYCALEX is available in many grades to exactly meet specific requirements

CHARACTERISTICS OF MYCALEX GRADE 410

Meets all the requirements for Grade L-4A, and is fully approved as Grade L-4B under Joint Army-Navy Specification JAN-1-10

Power factor, 1 megacycle	0.0015
Dielectric constant, 1 megacycle	9.2
Loss factor, 1 megacycle	0.014
Dielectric strength, volts/mil	400
Volume resistivity, ohm-cm	1×10^{15}
Arc resistance, seconds	250
Impact strength, Izod, ft.-lb/in. of notch	0.7
Maximum safe operating temperature, °C	350
Maximum safe operating temperature, °F	650
Water absorption % in 24 hours	nil
Coefficient of linear expansion, °C	11×10^{-6}
Tensile strength, psi	6000

MYCALEX is specified by the leading manufacturers in almost every electronic category

Mycalex 410
Tuning Coil Form

Mycalex 410
Tuning Switch Plate

Mycalex 410 Terminal Base and Cap Assembly for Fire Detection Equipment

Mycalex 410
Rotary Switch Stator

Mycalex 410
Solenoid Type Coil Form

Mycalex 410
Tuning Stator Plate

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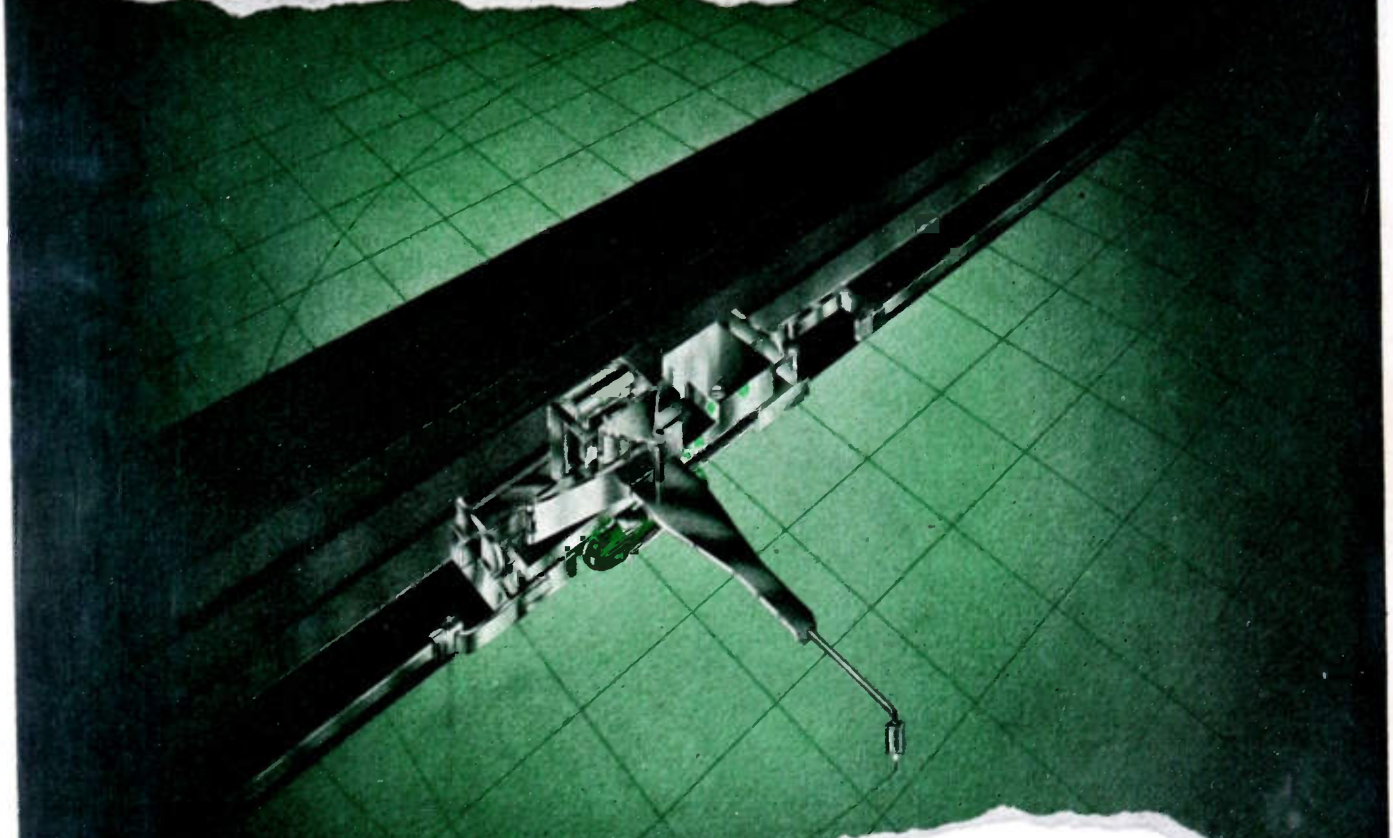
A Pen That Automatically Records Two Variables With Precision Accuracy.



You are looking at a plotting pen of one of the Model 205 Series Variplotter Plotting Boards.

With this self-balancing potentiometer type recorder, you can plot on a 30-inch square plotting surface a precise, graphic representation of one variable DC voltage as a function of a second variable DC voltage. It is also possible to plot two sets of two independent variables simultaneously, both using the full surface of the plotting board.

The Variplotter permits you to do this with speed and great accuracy — less than the width of the pen line itself. It produces a large, clear, permanent presentation that is easily interpreted.

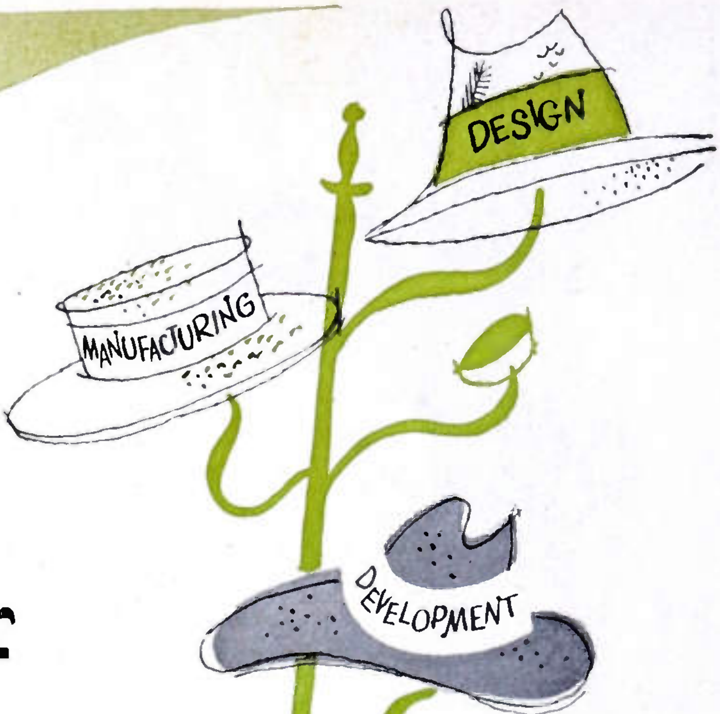
Permit us to forward to you complete data on the new Variplotters and their accessories. We will also be very happy to study your plotting problems and make recommendations. Simply contact Electronic Associates, Inc., Long Branch, New Jersey.

ELECTRONIC ASSOCIATES *Incorporated*

Variplotter

Model 205 Series Plotting Boards

Booth 98 and 99—Radio Engineering Show, New York City



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Standard Coil's experience-proved design, development and manufacturing facilities are at your command. In addition, coast-to-coast manufacturing locations assure you of diversified, prompt, efficient, economical production for any civilian or military electronic work. Give us a call . . . write or wire!

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Picture I. F. Transformers
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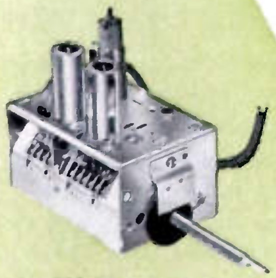
Sound I. F. Transformers
Sound Discriminator Transformers
Horizontal Oscillator Coils
Horizontal Linearity Control Coils
Width Control Coils
I. F. Strips   Flyback Coils

Radio and Miscellaneous Components

I. F. Transformers
R. F., Oscillator & Solenoid Coils
Antenna Loops
Ferrite Core Antennas
Permeability Tuning Pre-selector Assemblies
Miscellaneous Electro-Mechanical Assemblies



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"The Standard Tuner" designed and developed by Standard Coil and now used as original equipment in more than 5 million television sets.

Standard COIL PRODUCTS CO. INC.

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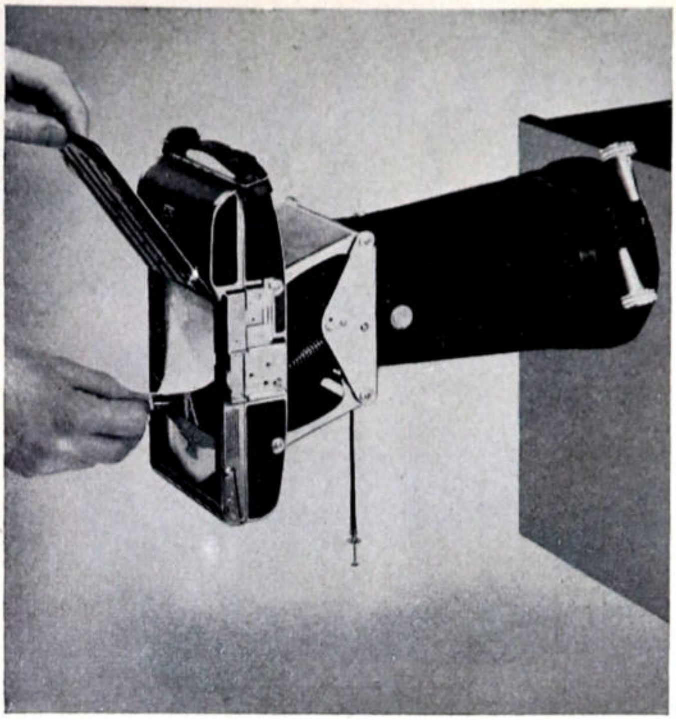
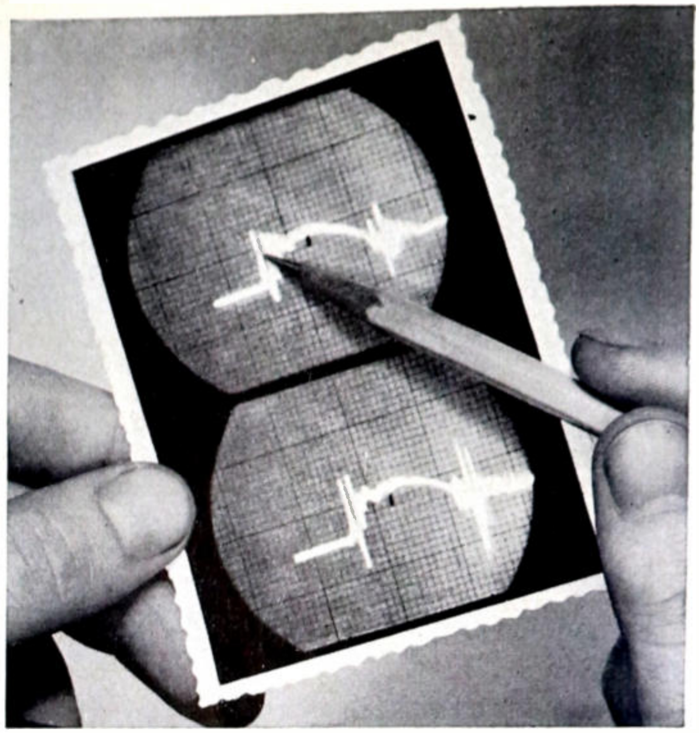




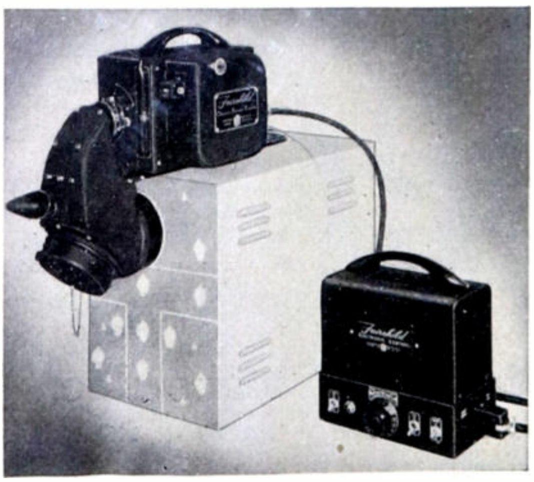
EASY TO TAKE...EASY TO READ

with the Fairchild-Polaroid® Oscilloscope Camera

ELECTRONIC CONTROL GIVES INFINITE SPEED VARIATION

with the Fairchild Oscillo-Record Camera

For both still and continuous-motion recording on 35-mm film or paper, the Fairchild Oscillo-Record Camera is ideal. Continuously variable speed control is provided through electronic regulation. There are no belts or pulleys. Film is sprocket-driven so there is no slippage. Top-of-scope mounting eliminates need for tripod and keeps scope controls easily accessible. There is provision for three film lengths—100, 400, or 1000 feet.



Fairchild's adaptation of the Polaroid-Land camera gives you more than just a fast photographic print of an oscilloscope image. The print is *exactly* half-size for easy measurement of values, especially when a grid is used. And you see the image exactly as it appears on the scope—not reversed. Each 3¼ x 4¼ print records two images.

Moreover, it takes only two minutes (less, if you're fast) to set up the camera, snap the shutter, and pull the tab. Then you wait one minute more and remove the finished print. It's as simple as 1-2-3. No focusing, no special training required.

Full information about the Fairchild-Polaroid and Fairchild Oscillo-Record Cameras is available on request. Write today to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, New York, Department 120-17C.*

Visit our exhibit at Booth 238-239 at the I. R. E. show

FAIRCHILD

OSCILLOSCOPE RECORDING CAMERAS

PRECISION

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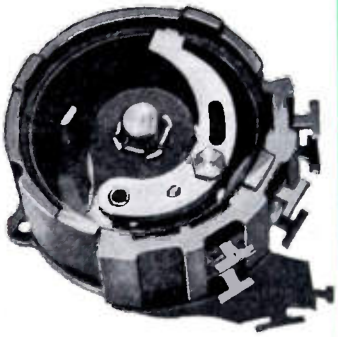
At the *very heart* of highly critical equipment such as electronic computers, electronic gunsights and radar assemblies, the control requirements call for outstanding electrical and mechanical precision. Indeed, from single section to as many as twenty sections, the precision controls must track with mathematical accuracy.

Clarostat Series 42 Controls fully meet these requirements. Thus *the climax in precision controls.*

Clarostat has made the major portion of such precision controls in use today. Many were supplied to the armed forces in World War II.
Many more have been supplied for civilian purposes since then. And now, *based on an unparalleled experience background,* Clarostat engineers offer you further refinements in their latest Series 42 design.

You can stand pat with CLAROSTAT



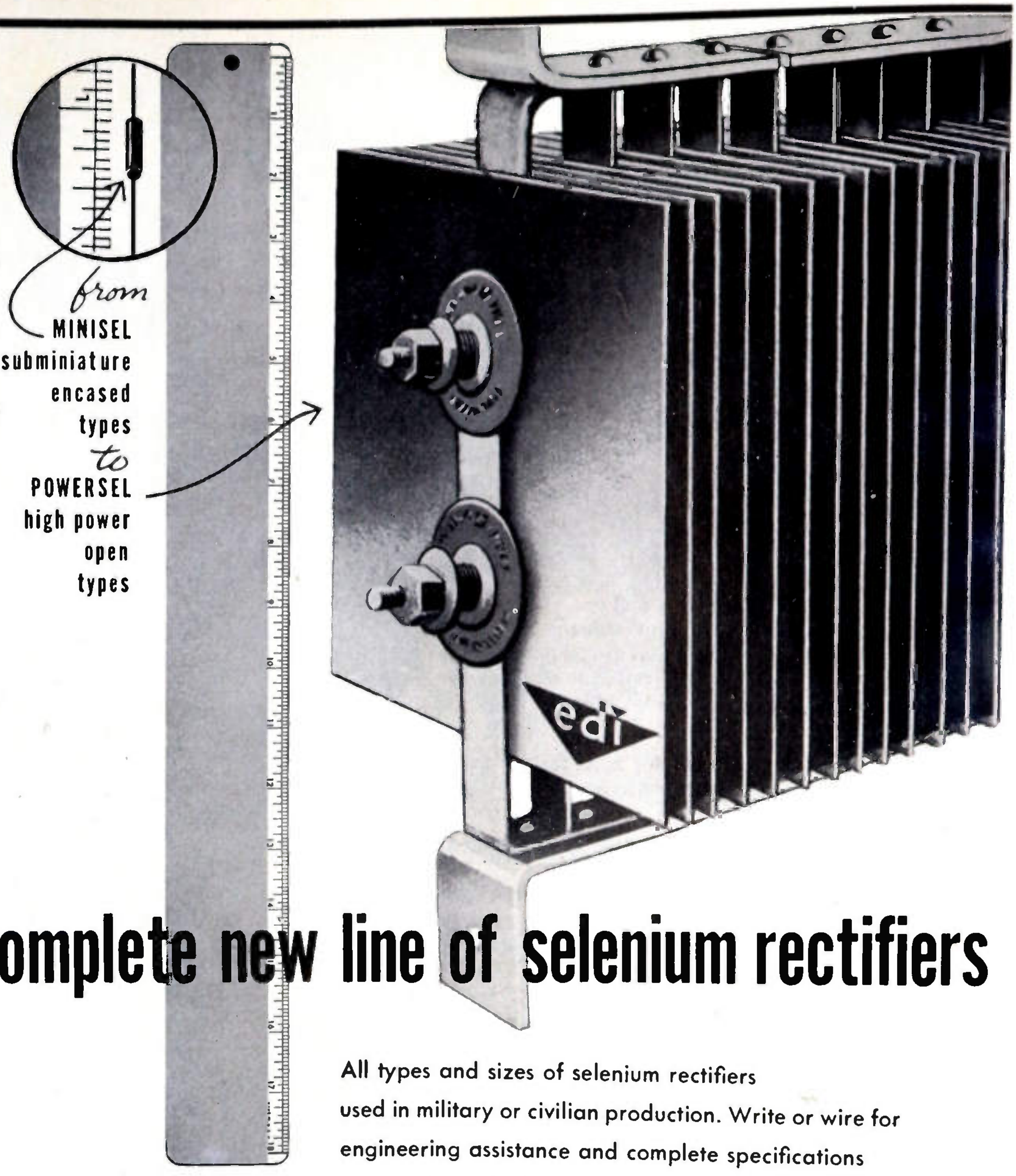
New Clarostat Series 42 potentiometer. Available in single and multiple assemblies up to 20 sections. Precision windings to plus/minus 0.5% and better. Positive contact rotor, smooth rotation, minimum wear. Perfect tracking of all units in assembly. No backlash or play. Rotor of each potentiometer mounted on centerless-ground shaft passing through all sections.

Engineering Bulletin No. 142 sent on request. And remember, when your control or resistor requirements call for *quality, quantity* and *economy,* you can meet them with Clarostat's engineering and production facilities. *Submit that problem!*

Controls and Resistors

CLAROSTAT

CLAROSTAT MFG. CO., INC., DOVER, NEW HAMPSHIRE
In Canada: Canadian Marconi Co., Ltd., Toronto, Ontario



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All types and sizes of selenium rectifiers used in military or civilian production. Write or wire for engineering assistance and complete specifications on your individual rectifier problems. No cost or obligation.

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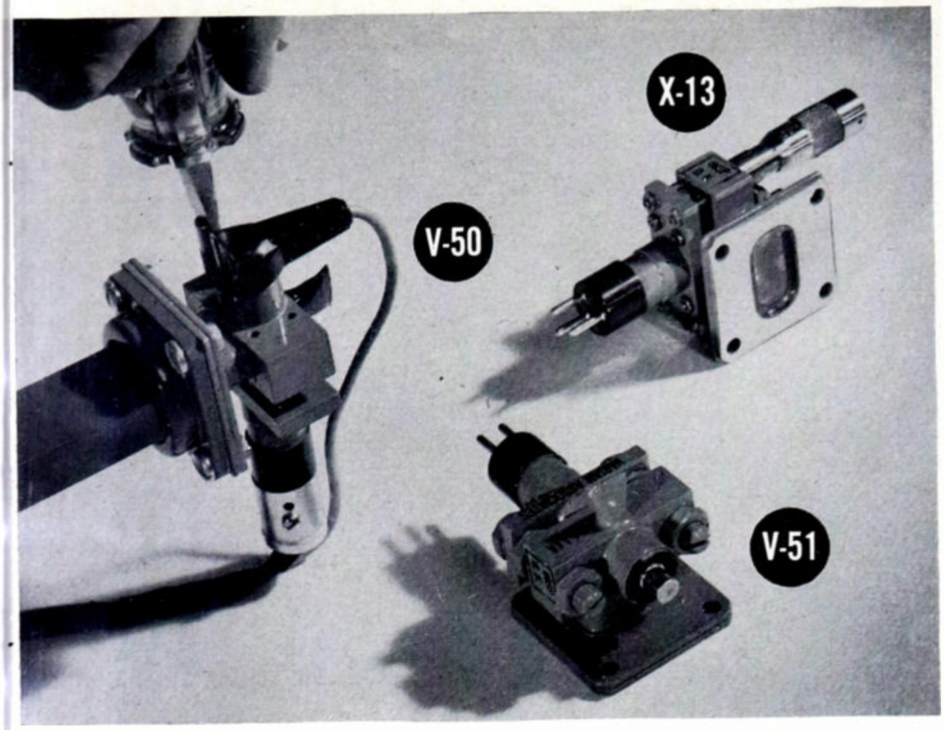

VARIAN associates

REFLEX KLYSTRONS for RUGGED SERVICE

8.5 to 10.0 kmc

Development based on the widely-used Varian X-13 klystron has produced two new Varian tubes with unusual possibilities for X-band applications involving extreme shock and vibration.

V-50 RUGGED, TUNABLE RADAR LOCAL OSCILLATOR. Here is a tube capable of withstanding severe vibration and shocks well beyond 30 times gravity It is tunable with extreme smoothness over the band from 8.5 to 10.0 kmc, and can be used with conventional afc circuits. Power output is 25 milliwatts, minimum, with a resonator voltage of 300 volts. The output connector mates with UG39/U flange (1 x ½″ waveguide).

V-51 RUGGED RADAR L. O. OR LOW-POWER TRANSMITTER. Lock-nut tuning enables the Varian V-51 klystron to withstand even rougher treatment than the V-50. Frequency range, application, and construction are otherwise similar. Tuning is easily done in the field with a standard open-end wrench. This tube is capable of 75 milliwatts, minimum, at 350 volts on the resonator. The output connection also mates with a UG39/U flange.

X-13 GENERAL-PURPOSE X-BAND SIGNAL SOURCE. A versatile, stable, reliable, laboratory-type signal source, the familiar Varian X-13 klystron tunes readily with a built-in micrometer device over a wide frequency range of 8.2 to 12.4 kmc. The X-13 is not intended for rugged service. It delivers well over 100 milliwatts at a resonator voltage of 500 volts. Output connection is a UG39/U flange.

Send for your copies of data sheets giving full information about this group of X-band Varian klystrons. There is a Varian Associates field representative nearby to assist on any application problems you may have.

Meet MIL-T-5422
and AN-E-19
specifications

SEE THESE TUBES AT
IRE NATIONAL CONVENTION
NEW YORK, MARCH 3 TO 6
BOOTH 55

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see next two pages

SIZE-SPACE-WEIGHT- AND COST

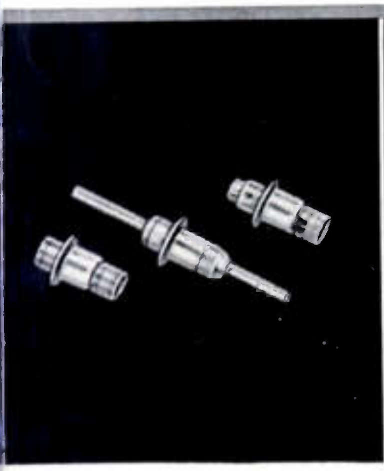
MILITARY ELECTRONIC GEAR

MINIATURE SWITCHES

Centralab's new miniature Series 20 and Series 30 switches have been specifically designed to meet the modern trend toward greatly reduced size for high-frequency, low-current applications. Extremely compact design and small size, plus availability of separate sections and index assemblies, provide an adaptability that is invaluable to design engineers and manufacturers. For complete information on the new Centralab Miniature Series 20 and Series 30 Switch line . . . multi-pole, multi-position, multi-section models or combinations with attached line switches and variable resistors, mail the coupon today. *Manufacturer's samples promptly.* Bulletins 42-156 and 42-157.

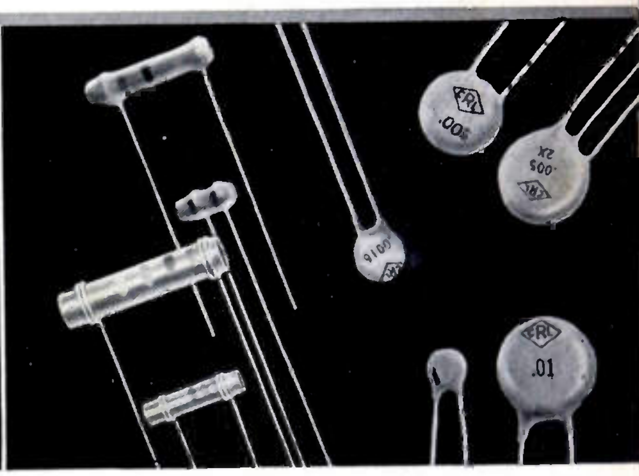
New Centralab Series 20 miniature switch, single steatite section. Available in 2 to 11 positions with stops, or 12 position continuous rotation—and with multiple sections.

Here's standard Series 20 miniature switch with standard shaft and phenolic section with off-on switch added. Also available with multiple sections.

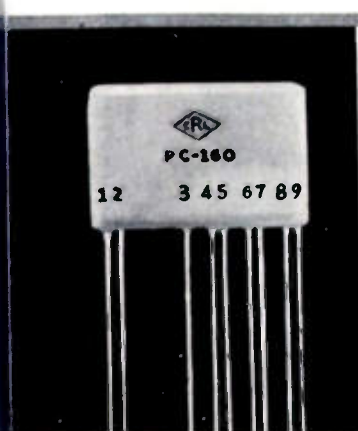


NEW Eyelet-Mounted Feed-through Ceramic Capacitors are exceptionally small. Capacities range from 25 to 3000 mmf., Voltage rating. 500 V. D. C. W. Check No. EP-15 in coupon.

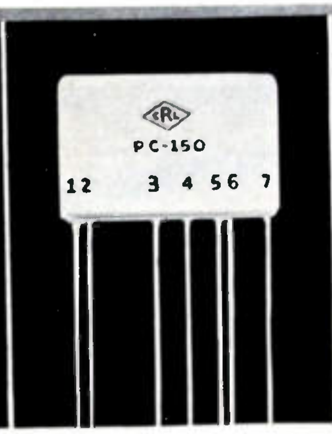
Centralab's Type 850 high voltage ceramic capacitors are especially designed for high voltage, high frequency circuits. Centralab's Type 950 high accuracy ceramic capacitors are especially developed for exacting electronic applications. Bulletins: 42-102 and 42-123.



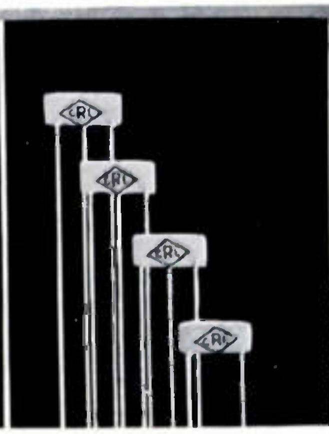
Ceramic Disc Hi-Kap Capacitors have very high capacity in extremely small size. Bulletin No. 42-4R. TC Tubulars (Temperature Compensating)—TCZ units show no capacity change over wide range of temperature; TCN's vary capacitance according to temperature. See Bulletin No. 42-18. BC (By-pass Coupling) Tubulars . . . for general circuit use. See Bulletin No. 42-3.



50% less soldered connections with Centralab's new Pendet . . . 5 capacitors and 4 resistors in a single plate . . . couples diode-triode and pentode tubes in output stage of AC-DC sets. Technical Bulletin 42-149.



50% less soldered connections with Centralab's Audet . . . furnishes all values of all components generally found in the output stage of AC-DC radio receivers. Technical Bulletin 42-129.



Tiny plate capacitor, resistor, and resistor-capacitor units. Readily fit all types of miniature and portable electronic equipment. Technical Bulletin 42-24.

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☐ 42-3	☐ 42-102	☐ 42-130	☐ 42-158
☐ 42-4R	☐ 42-123	☐ 42-149	☐ EP-15
☐ 42-18	☐ 42-126	☐ 42-156	
☐ 42-24	☐ 42-129	☐ 42-157	

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A page from the note-book of Sylvania Research

Pre-Determining CRT Electrode Configurations Saves Experimental Costs

An electrolytic tank and an ingenious plotting system give engineers at Sylvania's Research Laboratories great flexibility in the design of electron guns for cathode ray tubes.

An enlarged scale model of a vacuum tube electrode system is immersed in an electrolyte and voltages in proper ratio are applied to the electrodes. The potential distribution which results is that of the original electrode system in the vacuum tube.

With this tank various electrode configurations can be investigated and results plotted without expensive and time consuming assembly of endless sample tubes. With data gathered desired tubes having predetermined characteristics can be constructed for further tests. Development of such methods and equipment by Sylvania contributes much to the continuous improvement of Sylvania cathode ray tubes for television and other applications.

In the electrolytic wedge tank shown, the potential distributions in electrode systems of rotational symmetry are measured. Here, the sides of a wedge section of the electrode system are represented by the top and bottom surfaces of the water; the axis, by the water line; the cylindrical electrodes, by flat electrodes (as the wedge angle is small).

SYLVANIA

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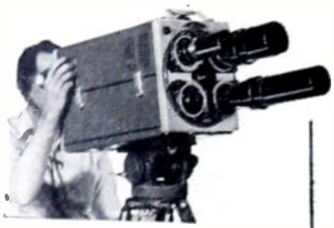

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Precision-built, lightweight, fast-handling. Push-button turret, remote iris control, remote focus and range selection. Easiest to service.

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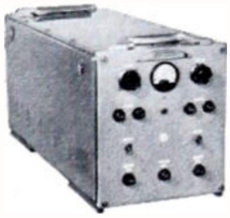
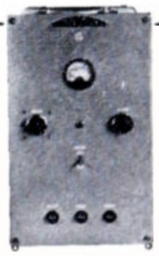
Touch-identified controls. 8½" monitor tube. Split or single headphone intercom system. CRO views horizontal, vertical, and vertical sync block. Iris control.



Camera Power Unit

Rugged, dependable, compact. Matched to other units in GPL chain. Standard relay panels swing out for maintenance.



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Affords maximum circuit reliability without operator adjustment. Binary counters and delay lines, stable master oscillator. Built-in power supply.

Complete TV Station Installations from Camera to Antenna

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Full studio flexibility anywhere. Control can view, preview, fade, dissolve, etc. Views any of 5 inputs, 2 remotes, outgoing line. Twin fade levers.

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Portable sync unit. No need for special phasing facilities. Projects rear-screen or "direct in." Ideal for remote origination of film. Relieves load on Telecine.

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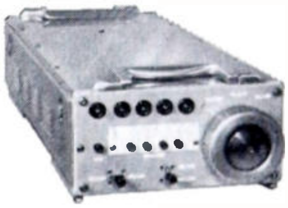

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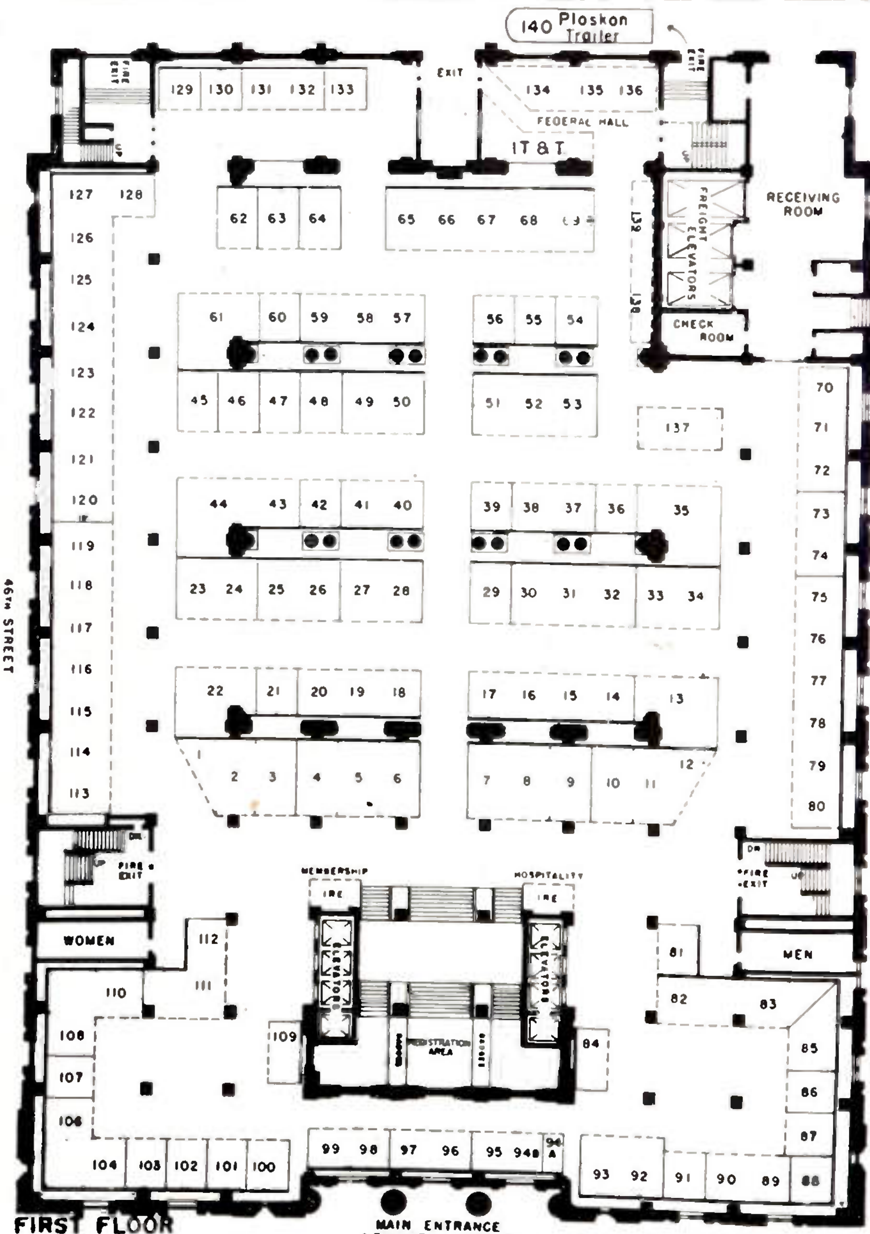
TV Camera Chains • TV Film Chains • TV Field and Studio Equipment • Theatre TV Equipment

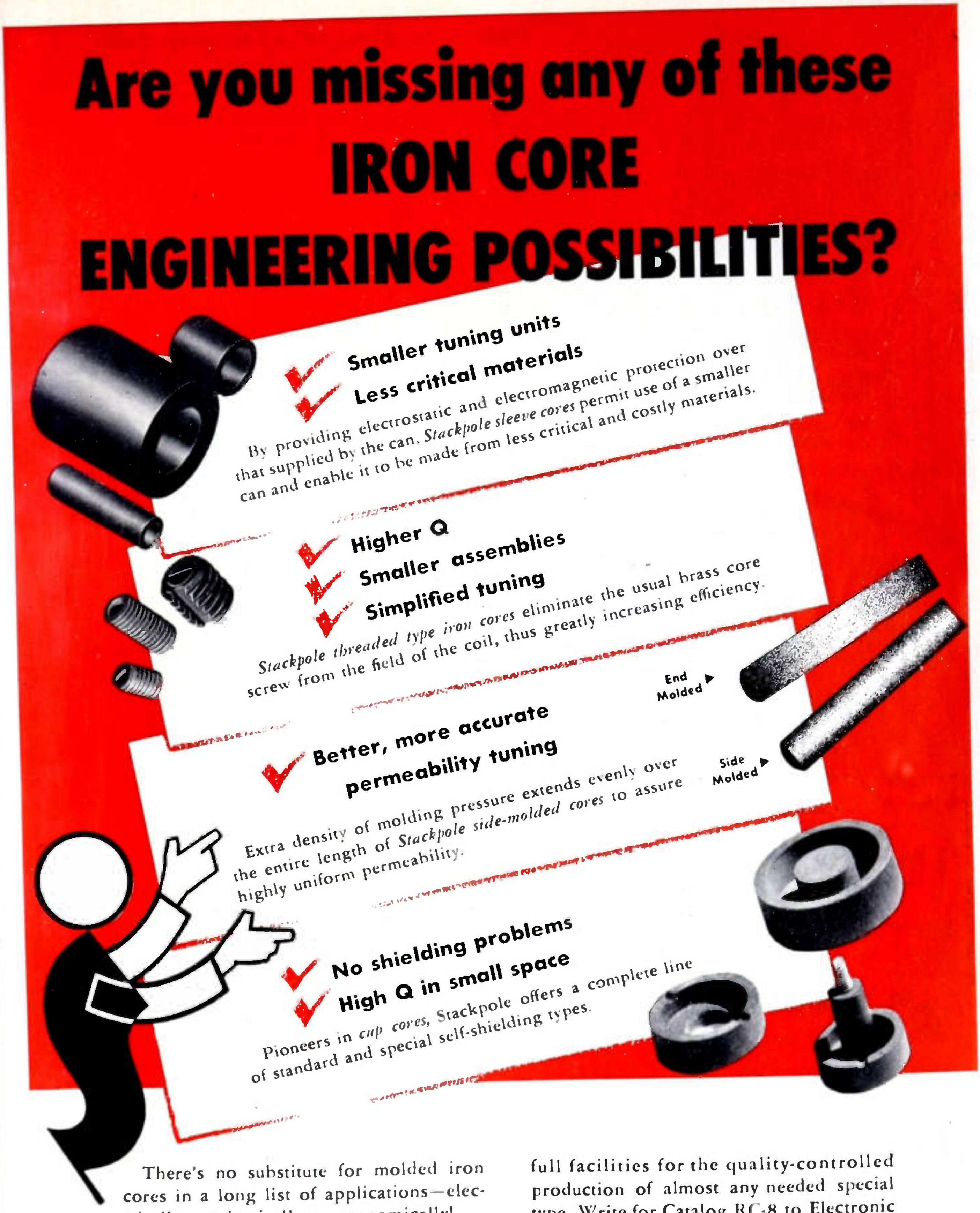
What to see at the Radio Engineering Show

(Continued from page 46A)

Firm	Booth
Cannon Electric Co., Los Angeles 31, Calif. Multicontact electrical and electronic connectors. Glass sealed and rubber sealed connectors. Connectors for waterproof applications. Miniature types.	249, 250
Capitol Radio Engineering Institute, Washington 10, D.C. Complete library of CREI home study courses in practical radio engineering, practical television engineering, and television and FM servicing.	357
Carboloy Dept., General Electric Co., Detroit 32, Mich. Permanent magnets.	61
Cargo Packers, Inc., New York 14, N.Y. Various types of packaging designs for instruments and equipment in the electronic industry which are in conformance with military requirements.	482
Centralab Div. of Globe-Union, Inc., Milwaukee 1, Wis. Ceramic capacitors, printed electronic circuits, controls, switches, special ceramics, components	232, 233
Century Geophysical Corp., Tulsa 6, Okla. Recording oscillographs, strain amplifiers, galvanometers, etc.	354
Chicago Rivet & Machine Co., Bellwood, Ill. Automatic rivet setting machines—setting singly and multiple for tubular type and split type rivets. Tubular and split rivets.	N-7
Chicago Telephone Supply Corp., Elkhart, Ind. Variable resistors, both carbonaceous and wirebound, and associated switches.	450
Cinch Mfg. Corp., Chicago 24, Ill. Electrical connecting devices.	255, 256
C. P. Clare & Co., New York 17, N.Y. Relays for electric, electronic, and industrial uses, hermetically sealed types. Stepping switches, lever and push keys.	204
Clarostat Mfg. Co., Inc., Dover, N.H. Wire-wound resistors, variable and fixed. Carbon resistors, variable. Attenuators. Resistance Devices.	254
Cleveland Container Co., Cleveland 11, Ohio CLEVELITE and COSMALITE paper base laminated phenolic tubing. Coil forms and deflection yoke sleeves. Kraft, acetate and combination tubing for electronic applications.	207

(Continued on page 70A)

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By providing electrostatic and electromagnetic protection over that supplied by the can, Stackpole sleeve cores permit use of a smaller can and enable it to be made from less critical and costly materials.
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Side Molded
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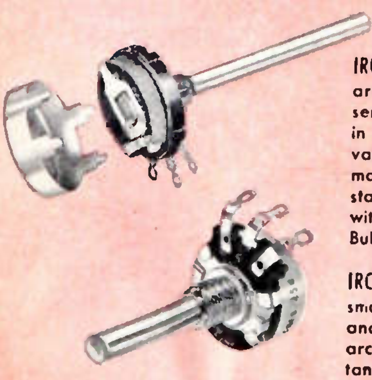
Pioneers in cup cores, Stackpole offers a complete line of standard and special self-shielding types.

A complete line *for resistors too!*

Unusual *combinations of characteristics* required in today's critical electronic circuits demand a complete range of resistor types. Specializing in resistors, IRC makes the widest line in the industry. This means ease of procurement—a single dependable source of supply for *all* your resistance needs. It also means unbiased recommendations—no substitution of units "just as good". IRC's complete line of products; complete research and testing facilities; complete network of licensees for emergency production—all add up to complete satisfaction for you.

CONTROLS



IRC Type W Wire Wound Controls are designed for long, dependable service and balanced performance in every characteristic. These 2-watt variable wire wound units provide maximum adaptability to most rheostat and potentiometer applications within their power rating. Catalog Bulletin A-2.

IRC New Type Q Controls feature small 15/16" size, rugged construction and superior performance. Increased arc of rotation permits same resistance ratios successful in larger IRC Controls. Catalog Bulletin A-4.

PRECISION RESISTORS

IRC Precision Wire Wounds offer a fine balance of accuracy and dependability for close-tolerance applications. Extensively used by leading instrument makers, they excel in every significant characteristic. Catalog Bulletin D-1.

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IRC Sealed Precision Voltmeter Multipliers are suitable and dependable for use under the most severe humidity conditions. Each consists of several IRC Precisions mounted and interconnected, encased in a glazed ceramic tube. Catalog Bulletin D-2.

is essential

INSULATED COMPOSITION and WIRE WOUND RESISTORS

IRC Advanced Type BT Resistors meet and beat JAN-R-11 Specifications at ⅓, ½, 1 and 2 watts—combine extremely low operating temperature with excellent power dissipation. Catalog Bulletin B-1.

IRC Type BW Wire Wound Resistors are exceptionally stable, inexpensive units for low range requirements. Have excellent performance records in TV circuits, meters, analyzers, etc. Catalog Bulletin B-5.

IRC Type BTAV High Voltage Resistors, developed for use as discharge resistors in fluorescent "Quick Start" ballasts, withstand momentary peak surge of 6000 volts. Also suited to TV bleeder circuits. Catalog Bulletin B-1.

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IRC Type MP High Frequency Resistors afford stability with low inherent inductance and capacity in circuits involving steep wave fronts, high frequency measuring circuits and radar pulse equipment. Available in sizes from ¼ to 90 watts. Catalog Bulletin F-1.

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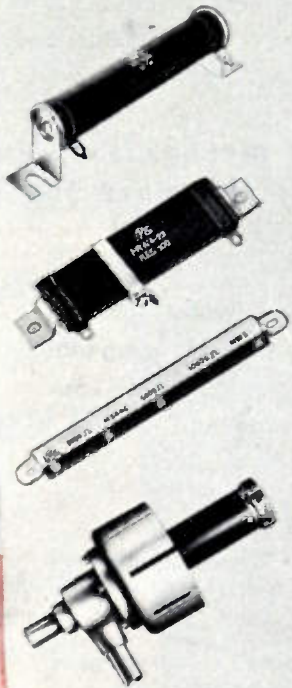
IRC Type MVX High Ohmic, High Voltage Resistors meet requirements for a small high range unit with axial leads. Engineered for high voltage applications, MVX has exceptional stability. Catalog Bulletin G-2.

IRC Type MPM High Frequency Resistors are miniature units suitable for high frequency receiver and similar applications. Stable resistors with low inherent inductance and capacity. Body only ⅜" long. Catalog Bulletin F-1.

POWER RESISTORS



IRC Fixed and Adjustable Power Wire Wounds give balanced performance in every characteristic—are available in a full range of sizes, types and terminals for exacting, heavy-duty applications. Catalog Bulletin C-2.

IRC Type FRW Flat Wire Wound Resistors fulfill requirements of high wattage dissipation in limited space—may be mounted vertically or horizontally, singly or in stacks. Catalog Bulletin C-1.

IRC Type MW Wire Wound Resistors offer low initial cost, lower mounting cost, flexibility in providing taps, and saving in space. Completely insulated against moisture. Catalog Bulletin B-2.

IRC Type LP Water-Cooled Resistors for TV, FM and Dielectric Heating Applications. Cooled internally by high velocity stream of water; adjustable to local water pressure and power dissipation up to 5 K.W.A.C. Catalog Bulletin F-2.

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☐ Bulletin B-1 (BT)	☐ Bulletin C-1 (FRW)	☐ Bulletin F-2 (LP)
☐ Bulletin B-1 (BTAV)	☐ Bulletin C-2 (PWW)	☐ Bulletin G-1 (MV)
☐ Bulletin B-2 (MW)	☐ Bulletin D-1 (WW)	☐ Bulletin G-2 (MVX)
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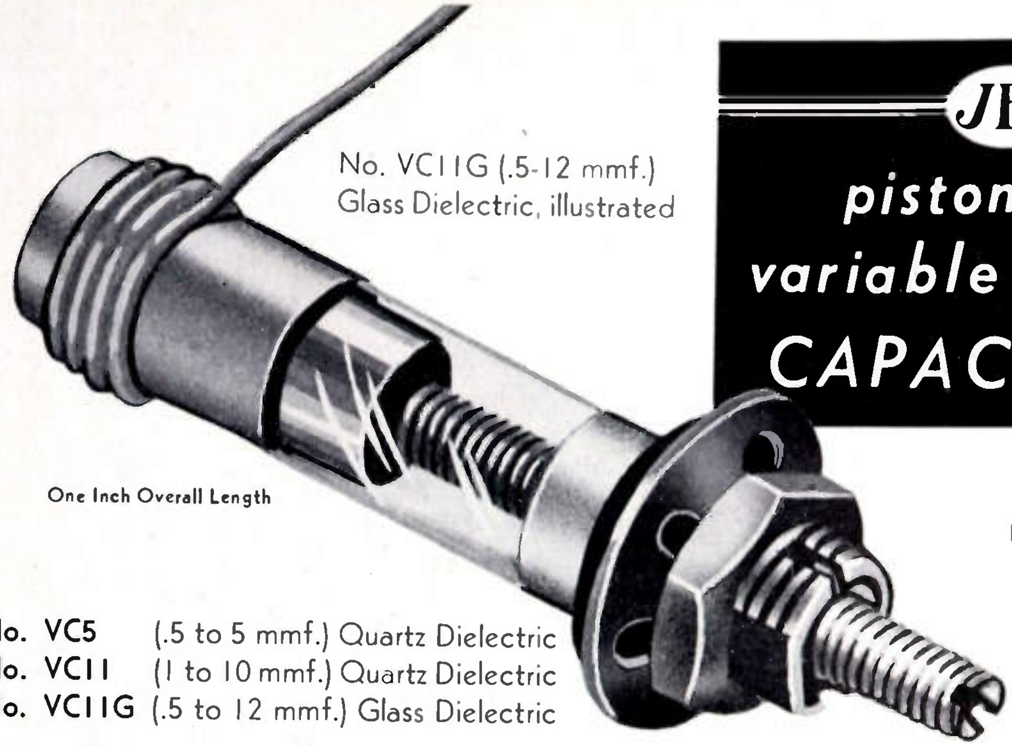

No. VC11G (.5-12 mmf.)
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JFD

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One Inch Overall Length

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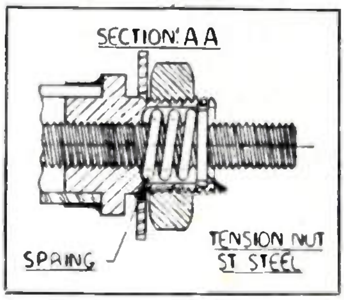
Built with precise mechanical concentricity and high electrical accuracy to meet the most exacting government specifications



Cutaway showing spring loading and antiback-lash construction.

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Tubular in design, they deliver continuously uniform change of capacitance in relation to rotation of the invar piston.

JFD Capacitors permit unusually precise adjustments without mechanical or electrical backlash or disturbance from severe vibration. Perfect resetability is achieved through use of precision mechanical parts. Approximately one inch in length, unique JFD Piston Capacitors also save space in tight quarters, mount with ease on any panel.

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- *One-piece spring loaded piston and screw made of special invar alloy having extremely low temperature coefficient of expansion.*
- *Silver band fused to exterior of precision drawn quartz or glass tube serves as stationary electrode.*
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JFD Electronics

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PRECISION POINT-TO-POINT COMMUNICATION EQUIPMENT

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Urgent yesterday, today even higher speed, fidelity and dependability—under even tougher conditions—are *vital*. Only continuing advance in modern precision point-to-point communication equipment can accomplish these feats.

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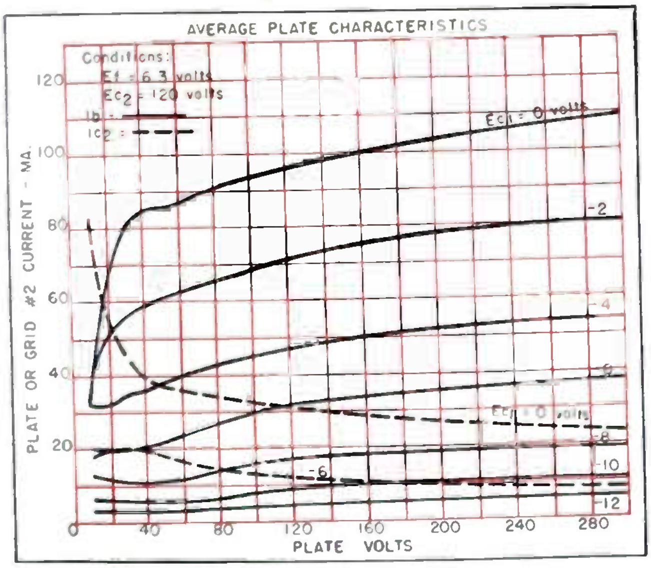
MINIATURE PENTODE For Computer Applications

The Raytheon 6AN5 was the first of its kind — the first with low interface resistance to avoid *"sleeping sickness"*. It remains the first choice of designers of dependable, long lived computing devices.



The Raytheon 6AN5 has been in continuous production for over two years. This means maximum reliability, minimum failures.

Important characteristics of the 6AN5 drop less than 10% in 5000 hours *under on, off, or flip-flop conditions.*

The Raytheon 6AN5, providing high efficiency with low plate voltage is also recommended for such services as

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For Industrial and Research Use

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FILAMENT SUPPLY: 6.3 volts AC, 10 Amp., CT.

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½% for load at 150 volts.
RIPPLE: 5 millivolts RMS.

VOLTS	CURRENT	MODEL
100-325 0-150 Bias 6.3 AC.CT.*	0-150 Ma. 0-5 Ma. 10 Amp.	131
200-500 6.3 AC.CT.	0-200 Ma. 6 Amp.	245
0-300 0-150 Bias 6.3 AC.CT.	0-150 Ma. 0-5 Ma. 5 Amp.	315
0-500 6.3 AC.CT.	0-300 Ma. 10 Amp.	500R
#1 200-500 #2 200-500 #3 6.3 AC.CT. #4 6.3 AC.CT.	0-200 Ma. 0-200 Ma. 6 Amp. 6 Amp.	510
0-500 0-150 Bias 6.3 AC.CT.	0-200 Ma. 0-5 Ma. 10 Amp.	515
#1 0-500 #2 0-500 #3 6.3 AC.CT. #4 6.3 AC.CT.	0-200 Ma. 0-200 Ma. 10 Amp. 10 Amp.	600

VOLTS	CURRENT	MODEL
0-500 0-150 Bias 6.3 AC.CT.	0-300 Ma. 0-5 Ma. 10 Amp.	615
#1 0-600 #2 0-600 #3 6.3 AC.CT. #4 6.3 AC.CT.	0-200 Ma. 0-200 Ma. 10 Amp. 10 Amp.	800
0-600 0-150 Bias 6.3 AC.CT.	0-200 Ma. 0-5 Ma. 10 Amp.	815
0-1000-Ripple 10 mv. 6.3 AC.CT.	0-50 Ma. 10 Amp.	1020
0-1200-Ripple 10 mv. 6.3 AC.CT.	0-20 Ma. 10 Amp.	1220
200-1000-Ripple 20 mv.	0-500 Ma.	1250
0-1000-Ripple 20 mv.	0-500 Ma.	1350
Specify your voltage and current requirements. Regulation available .5%, .1%, .01%.		SPECIAL SERIES

*All AC Voltages are unregulated. All units are metered except Models 131 and 315.

All units designed for relay rack mounting or bench use.

The Kepco Voltage Regulated Power Supplies are conservatively rated. The regulation specified for each unit is available under all line and load conditions, within the range of the instrument. Write for specifications.

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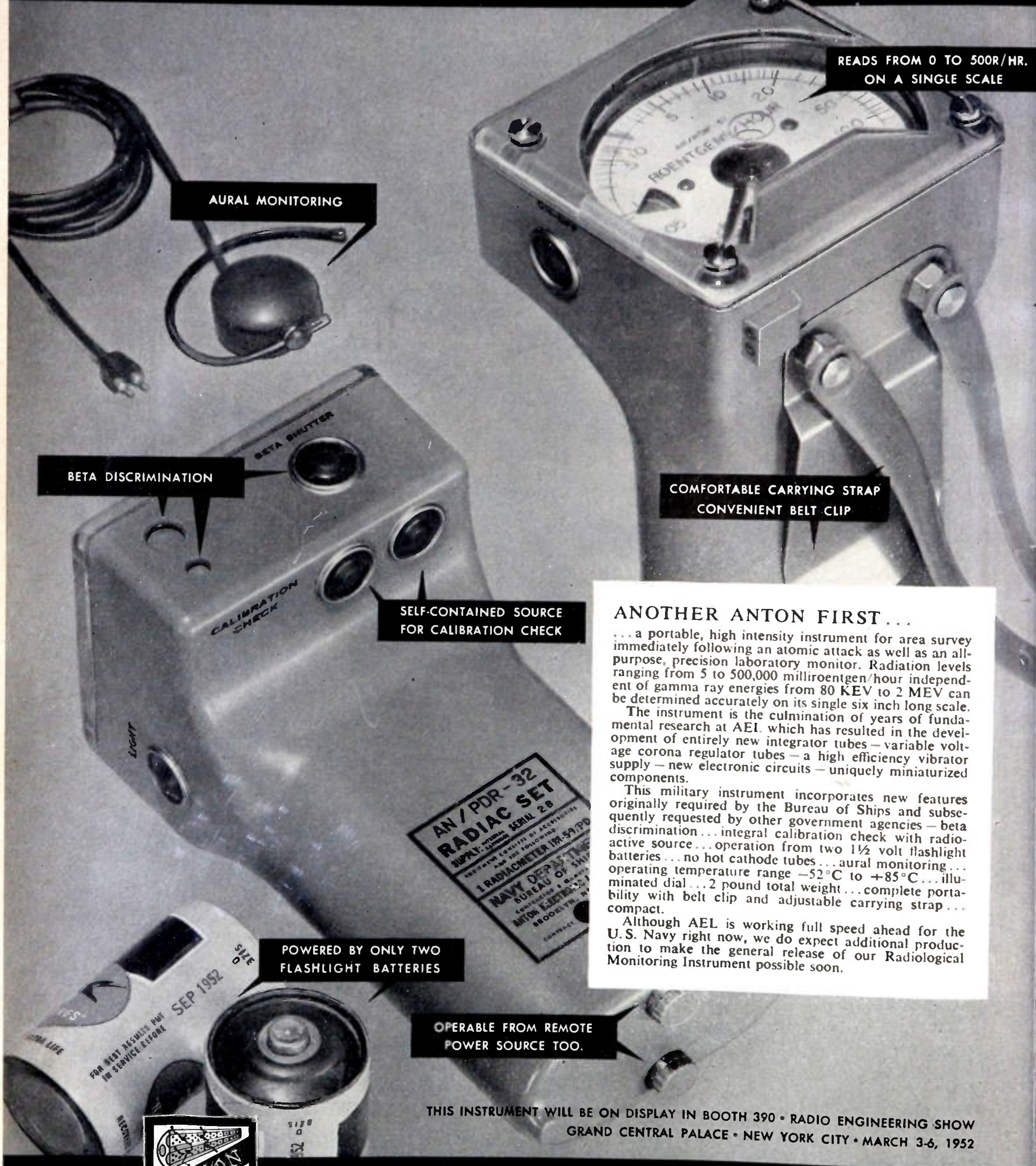
The HIGH INTENSITY RADIOLOGICAL MONITORING INSTRUMENT

conceived and developed by

ANTON ELECTRONIC LABORATORIES for the U.S. NAV



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. . . a portable, high intensity instrument for area survey immediately following an atomic attack as well as an all-purpose, precision laboratory monitor. Radiation levels ranging from 5 to 500,000 milliroentgen/hour independent of gamma ray energies from 80 KEV to 2 MEV can be determined accurately on its single six inch long scale.

The instrument is the culmination of years of fundamental research at AEL which has resulted in the development of entirely new integrator tubes – variable voltage corona regulator tubes – a high efficiency vibrator supply – new electronic circuits – uniquely miniaturized components.

This military instrument incorporates new features originally required by the Bureau of Ships and subsequently requested by other government agencies – beta discrimination . . . integral calibration check with radioactive source . . . operation from two 1½ volt flashlight batteries . . . no hot cathode tubes . . . aural monitoring . . . operating temperature range −52°C to +85°C . . . illuminated dial . . . 2 pound total weight . . . complete portability with belt clip and adjustable carrying strap . . . compact.

Although AEL is working full speed ahead for the U.S. Navy right now, we do expect additional production to make the general release of our Radiological Monitoring Instrument possible soon.

THIS INSTRUMENT WILL BE ON DISPLAY IN BOOTH 390 • RADIO ENGINEERING SHOW GRAND CENTRAL PALACE • NEW YORK CITY • MARCH 3-6, 1952

ANTON ELECTRONIC LABORATORIES, INC.

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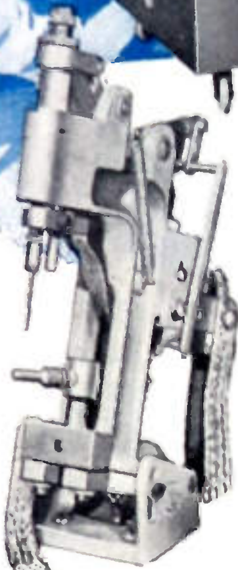
What to see at the Radio Engineering Show

(Continued from page 70A)

Firm	Booth
Cornell-Dubilier Electric Corp., S. Plainfield, N.J. Capacitors, Vibrators, Antennas, Converters (Vibrator powered).	73, 74
Corning Glass Works, Corning, N.Y. All-glass television bulbs (including new non-glare cylindrical face plate). Special glasses widely used for glass-to-metal seals. Metallized glassware for glass-to-metal seals. Bulbs and tubing for electronic and cathode-ray applications.	30-32
Cossor (Canada) Ltd., Halifax, Nova Scotia, Canada. See: Beam Instruments.	223
R. W. Cramer Co., Inc., Centerbrook, Conn. Synchronous motors (small 1 rps-¼ rpd output speeds). Dual motors, chart drive motors, interval timers, reset timers, time delay relays, cycle timers, reversing timers, pulse timers, duplex timers, percentage timers, multi-contact timers, running time meters, special timers. Hermetically sealed instruments: circuit reclosing relay, miniature time delay relays, running time meters.	416, 417
Crucible Steel Co. of America, New York 17, N.Y. Permanent magnets in a variety of sizes and shapes, tool, stainless and special purpose steels, including stainless tubing.	504
Curtis Dev. & Mfg. Co. See: Wally Swank	221
H. L. Dalis, Inc., New York, N.Y. Audio and industrial equipment.	320B
The Daven Co., Newark 4, N.J. Attenuators, audio and rf; output power supply, switches, audio and rf; attenuation network, laboratory and test sets, rf attenuation boxes, communication equipment, Jan-R-93-precision wire wound resistors, signal generator, transmission measuring sets.	94B, 95
Bryan Davis Publishing Co., Inc., New York 17, N.Y. *Television Engineering* magazine, *Service* magazine.	288
Tobe Deutschmann Corp., Norwood, Mass. Capacitors, metallized paper, oil and wax impregnated paper, molded paper, high temperature. Filters: Radio TV-Noise suppression, low pass, high pass, band pass, audio. Special products, pulse forming networks, delay lines.	343
Dial Light Co. of America, Inc., New York 3, N.Y. Warning signal indicator, and pilot light assemblies.	46
Diamond Mfg. Co., Wakefield, Mass. Rf coaxial cable connectors and microwave components.	489
Distillation Products Industries, Rochester 3, N.Y. Philips Vacuum gauge, magnetron tube pumping dolly, LD-01 Halogen Sensitive Leak detector, various types of thermocouple vacuum gauges, new designs in crystal coating equipment, the MB-10 booster pump and valve assembly.	241, 242
Wilbur B. Driver Company, Newark 4, N.J. Special alloys for electronic applications.	340, 342
Dumont Electric Corp., New York 34, N.Y. Tubular paper, dry electrolytic, metal clad, bathtub and upright capacitors. "Zero-Loss durene capacitors." "Thermofilm" dielectric capacitors.	403
Allen B. DuMont Laboratories, Inc., TV Transmitter Division, Clifton, N.J. Sync generator; monochrome scanner; universal color scanner; model studio control room; studio camera with mobile-mount dolly; master control equipment; video switching and mixing equipment.	120-123
Allen B. DuMont Laboratories, Inc., Cathode-Ray Tube Div., Clifton, N.J. Television picture tubes including Selfocus types, cylindrical faceplate tubes, low voltage electrostatic focus types, and the 30BP4 Teletron.	124, 125

(Continued on page 86A)



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ALL RANGES WITH THIS ONE CONTROL

Model 630

Just *one* knob—extra large—easy to turn—flush with the panel, controls all ranges. This one knob saves your time—minimizes the chances of "burn-outs" because you don't have to remember to set another control. You can work fast with Model 630 with your eyes as well as your hands. Look at that scale—wide open—easy to read, accurately. Yes, this is a *smooth TV* tester. Fast, safe, no projecting knobs, or jacks, or meter case. Get your hand on that single control and you'll see why thousands of "Model 630's" are already in use in almost every kind of electrical testing

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FOR THE MAN WHO TAKES PRIDE IN HIS WORK

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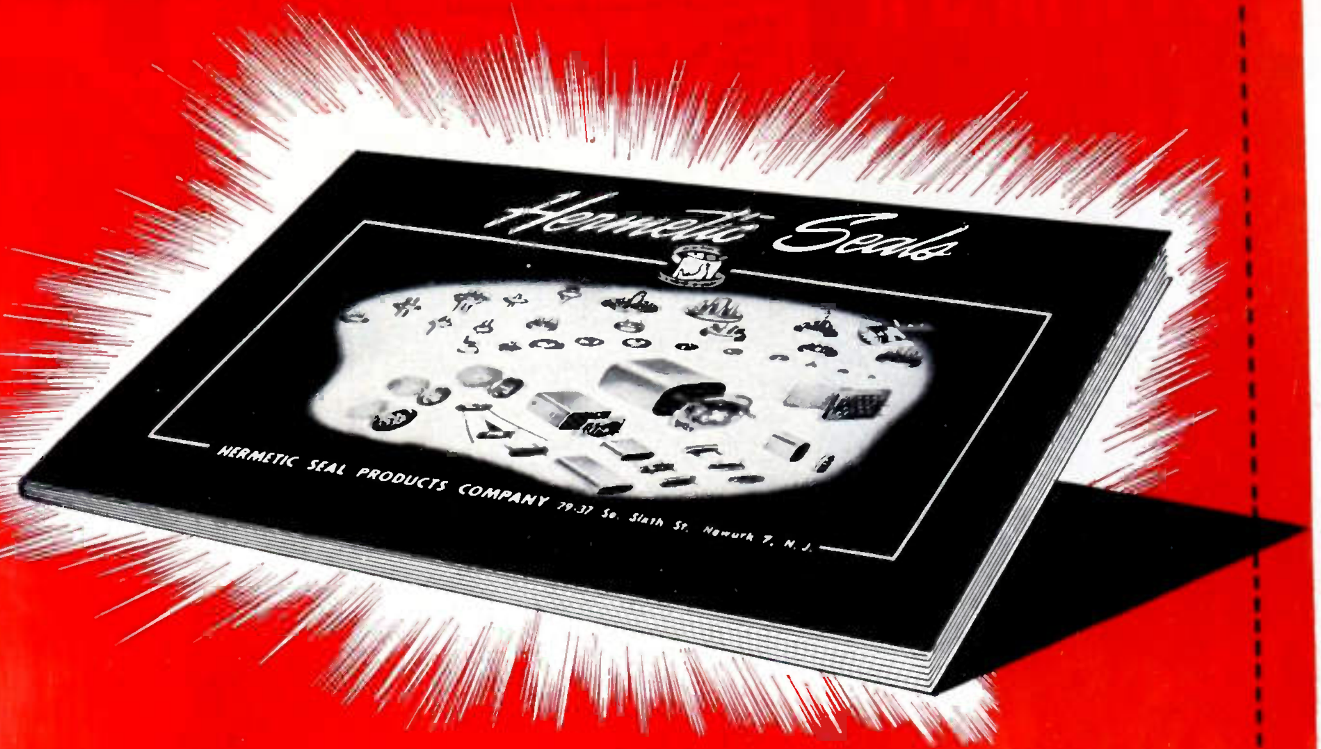

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What to see at the Radio Engineering Show

(Continued from page 82A)

Firm	Booth
Allen B. DuMont Laboratories, Inc., Instr. Div., Clifton, N.J. Cathode-ray oscillographs: Type 322 dual-beam; Type 303; Type 294-A; Type 304-HR; Type 334-A; CRO accessories. Oscillograph record cameras: Type 321; Type 295; Type 296; Type 297. Industrial Cathode-ray Tubes: Types K1065; K1056, K1101, K1052, K1098, K1080, K1105, K1084, 3RP-A, 5YP-, 5SP-.	126-128
DX Radio Products Co., Inc., Chicago 47, Ill. IF, rf, and discriminator transformers, rf choke coils, toroid off-center coils, toroid reactors, toroid deflection yokes, toroid filters, hermetic sealed focus coils, focus coils, special deflection yokes, special coil assembly, crystals, speakers, horizontal output transformer, ion traps, deflection yokes, TV tuner (Rotorette), quartz crystals.	345A
Dyna-Labs., Inc., New York 13, N.Y. D-79 Gaussmeter, to measure flux in small gaps, miniature magnetic earphones.	409
Hugh H. Eby, Inc., Philadelphia 44, Pa. Various types of radio—electronic hardware, sockets, plugs, terminal strips and boards. Binding posts fuse holders, etc. Connectors: rack, panel, and pressurized.	410
Eitel-McCullough, Inc., San Bruno, Calif. Diodes, triodes, tetrodes, pentodes, klystrons, cathode-ray, vacuum switches, vacuum capacitors, vacuum pumps, vacuum apparatus, tube hardware, vacuum tube materials and components.	36
Elastic Stop Nut Corp. of America, Union, N.J. Self-locking rollpins, and elastic stop nuts.	458
ELCO Corp., Philadelphia 40, Pa. Tube sockets, tube shields and plugs, and connectors. "Varicon" connectors, and plugs and receptacles. New connector systems, miniature in size but high current carrying capacity, flexible and polarity.	337
Electrical Industries, Inc., Newark 4, N.J. Hermetically sealed terminals and headers for use on relays, transformers, and other equipment requiring vacuum tight seals.	212
Electric Reactance Corp., Olean, N.Y. Ceramic capacitors, trimmers, wire wound resistors, printed circuits.	63
Electro Precision Products, Inc., College Point, L.I., N.Y. Coax Connectors, and Microwave Equipment.	431
Electro Motive Mfg. Co., Inc., Willimantic, Conn. See: Arco Electronics, Inc.	338
Electro-Tech Equipment Co., New York 13, N.Y. Electrical measuring instruments, and industrial control equipment, and constant and variable voltage transformers. Selenium rectifiers for industrial needs, Eagle signal timers, and photoswitch.	411, 412
Electronic Associates, Inc., Long Branch, N.J. Model 205 series Variplotter plotting board; automatic curve following equipment for use with analog computers, etc.; automatic digital curve plotting equipment; "EASE" (Electronic Associates Simulation Equipment)—a high-precision, general-purpose analog computer.	98, 99
Electronic Computer Corp., Brooklyn 17, N.Y. Small electronic digtail computer. Components for computers.	505
Electronic Devices, Inc., Brooklyn 15, N.Y. Rectifiers: Minisel-subminiature encased selenium rectifiers. Plastisel-miniature electronic selenium rectifiers. Powersel-commercial and power rectifiers. Minisel instrument rectifiers. Rectifier test equipment . . . power supplies. Resistors: carbon resistors.	408
Electronic Instrument Co., Inc., Brooklyn 11, N.Y. Test equipment instruments and kits, including oscilloscopes, vacuum tube voltmeter, signal tracers, signal generators, multitesters, sweep generators, tube testers, comparator bridge, battery eliminators, high voltage probes, high frequency probes, crystals, etc.	362

(Continued on page 88A)

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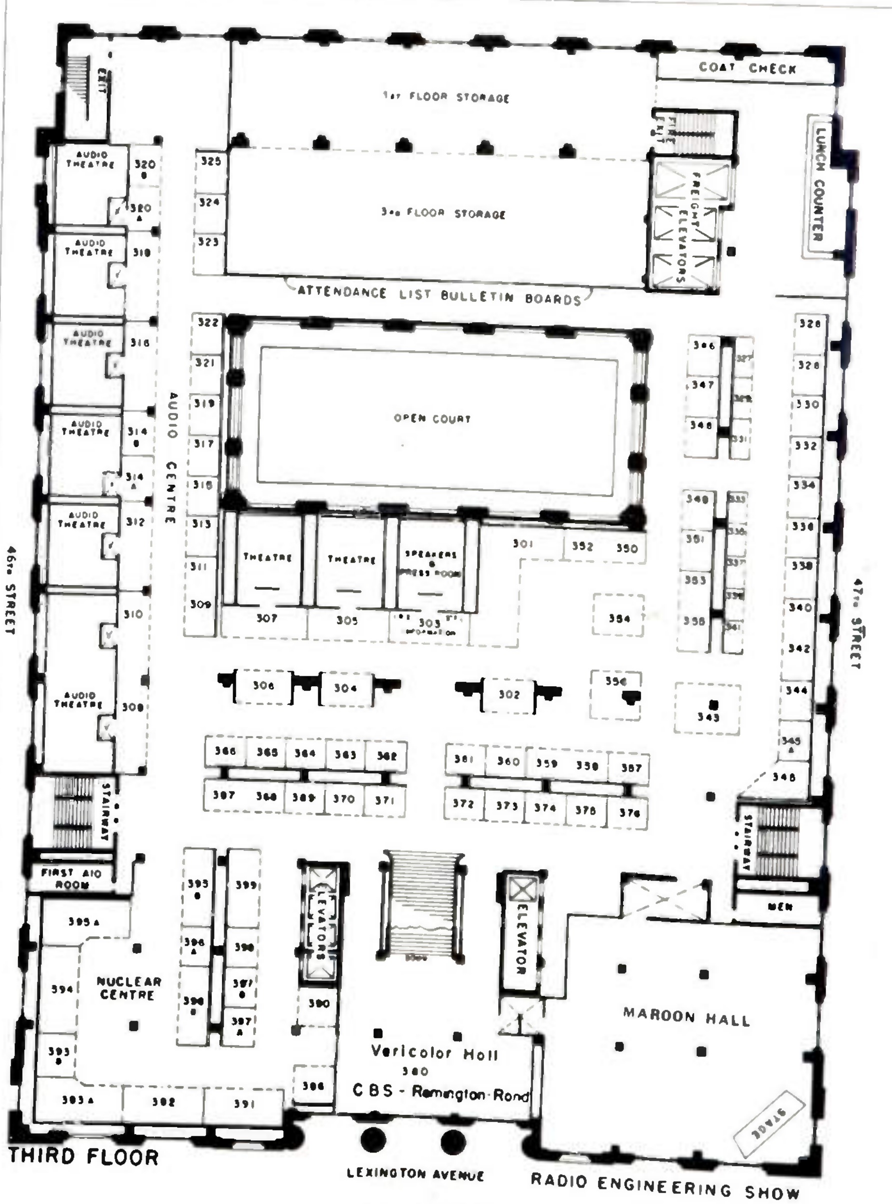
What to see at the Radio Engineering Show

(Continued from page 86A)

Firm	Booth
Electronic Mechanics, Inc., Clifton, N.J. Mykroy (glass bonded mica) Kel-F (Trifluorochloroethylene) Teflon (Tetrafluoroethylene).	336
Electronic Tube Corp., Philadelphia 18, Pa. Multi-channel oscilloscopes, standard, multi-gun and special cathode-ray tubes, dc amplifiers.	274, 275
Electronics, New York 18, N.Y. See: McGraw-Hill Pub. Co.	200
H. R. Ellis, New York 17, N.Y. Airmec Lab., test equipment. Enthoven, H. V., cored solder. Painton & Co. Ltd., Potentiometers-resistors Induction Motors Corp., miniature ac motors. Parmeko, transformers, cans, parts. Belling & Lee, sockets, connectors, filters, fuses. General Accessories Co., sockets, connectors, pilot lights, phone tips. Plastic Process Co., Electro Acoustic.	244
El-Tronics, Inc., Philadelphia 33, Pa. Binary and decade scaling units; portable Geiger-Mueller survey units; portable ionization chambers—cutie pie type; count rate meters and count rate meter attachments; Nuclear accessories for laboratories and hospitals; laboratory audio oscillator; square wave generator; insulation tester; wide band oscilloscope. Shielded manual sample changer, remote handling tongs. Noise and field intensity meter 20 to 400 mc, impulse generator.	386
Emeloid Co., Inc., Hillside 5, N.J. Printed circuits and fabricated and printed plastic articles such as nameplates and special parts. The printed circuits have finishing operations such as piercing, blanking, assembling, also patent pending flush type for switches.	472
Empire Devices, Inc., Bayside, L.I., N.Y. Broadband crystal mixers, noise and distortion analyzer 6 to 110 kc; coaxial attenuator pads. Resistive step attenuators, uhf.	349
Engineering Research Associates, Inc., St. Paul 4, Minn. Unit-Magnetic recorder using new boundry-displacement technique; shaft-monitor shaft position indicator: Magnetic recording heads; pulse transformers; pictures and descriptive brochures on ERA digital computers and magnetic drum storage systems.	361
Erie Resistor Corp., Erie, Pa. Miniaturized ceramic and button silver mica capacitors, printed circuits, stand-off and feed-thru capacitors. High temperature (250° C) button silver mica capacitors. Barium titanite piezo electric elements.	91

Continued on page 92A)



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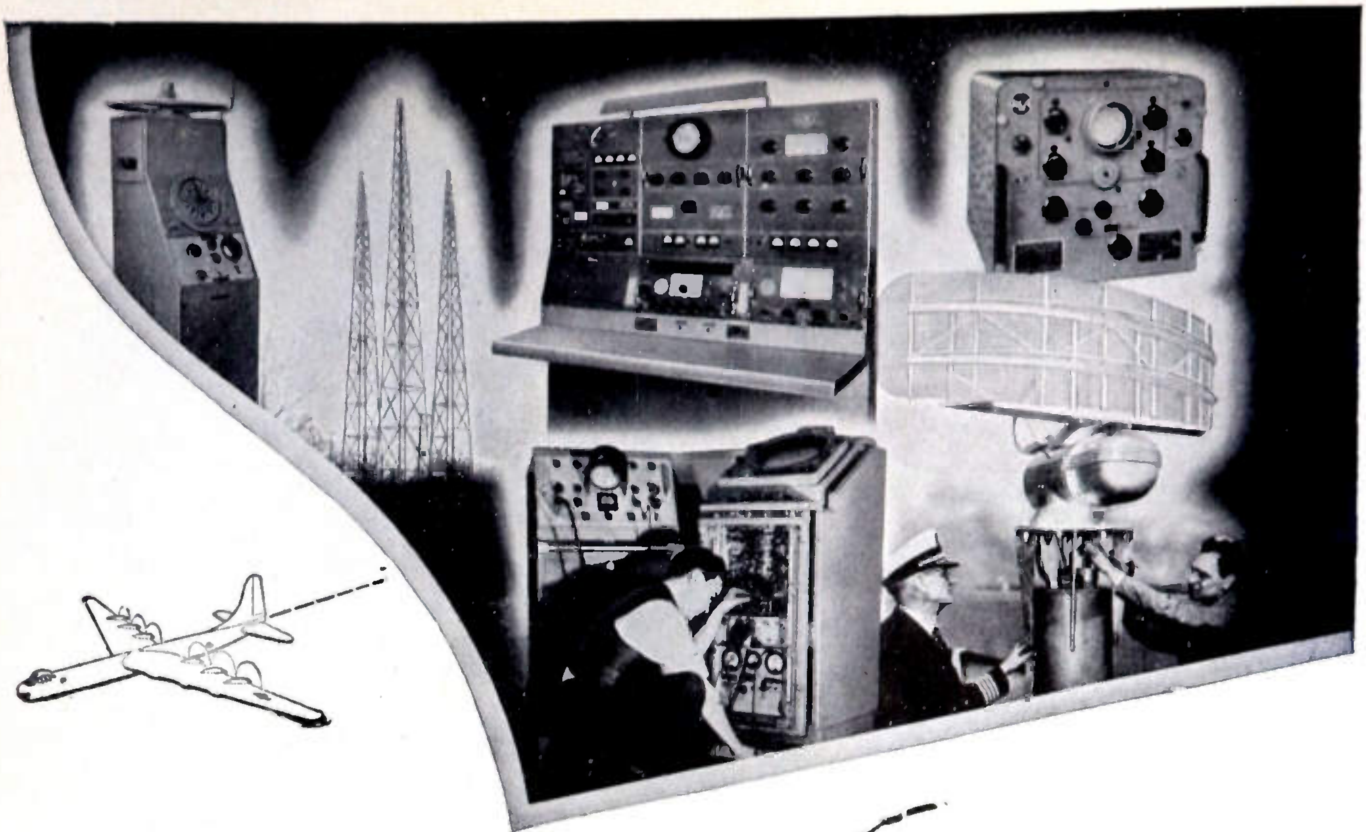

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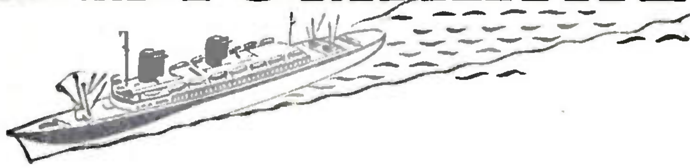

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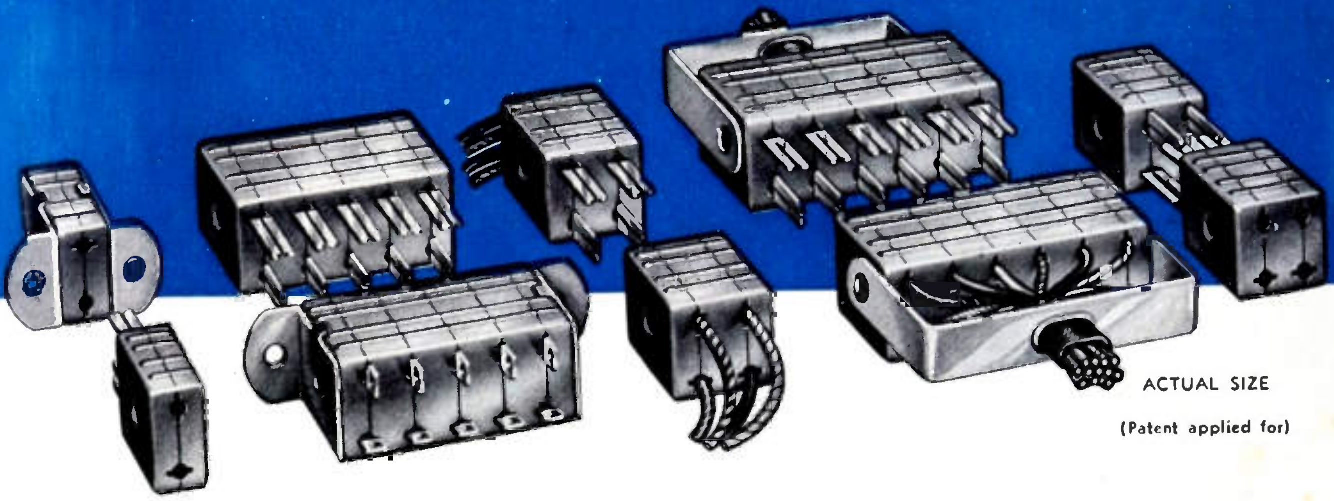

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WITH "KEYING CONTROL" VERSATILITY NEVER BEFORE ACHIEVED!

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HOW NEW IS NEW? You'll never realize until you see how Varicons' four basic parts give you the maximum number of "Keying Control" variations; plus contact combinations in any number demanded by your specific needs. For the new product you're designing, or the redesign of a present product, you'll want the full, specific Varicon story. We'll have it on your desk by return mail, upon the receipt of your inquiry.

General Specifications — Male and female elements identical :: Contacts always under pressure; cannot be overstressed or overstrained :: Current rating 30 amps, 115 volts :: Withstanding voltage maximum 4000 volts between closest terminals :: Low contact resistance :: Low capacitance, 300 ohm line spacing :: Easy insertion pressure :: Excellent retention pressure between contacts.

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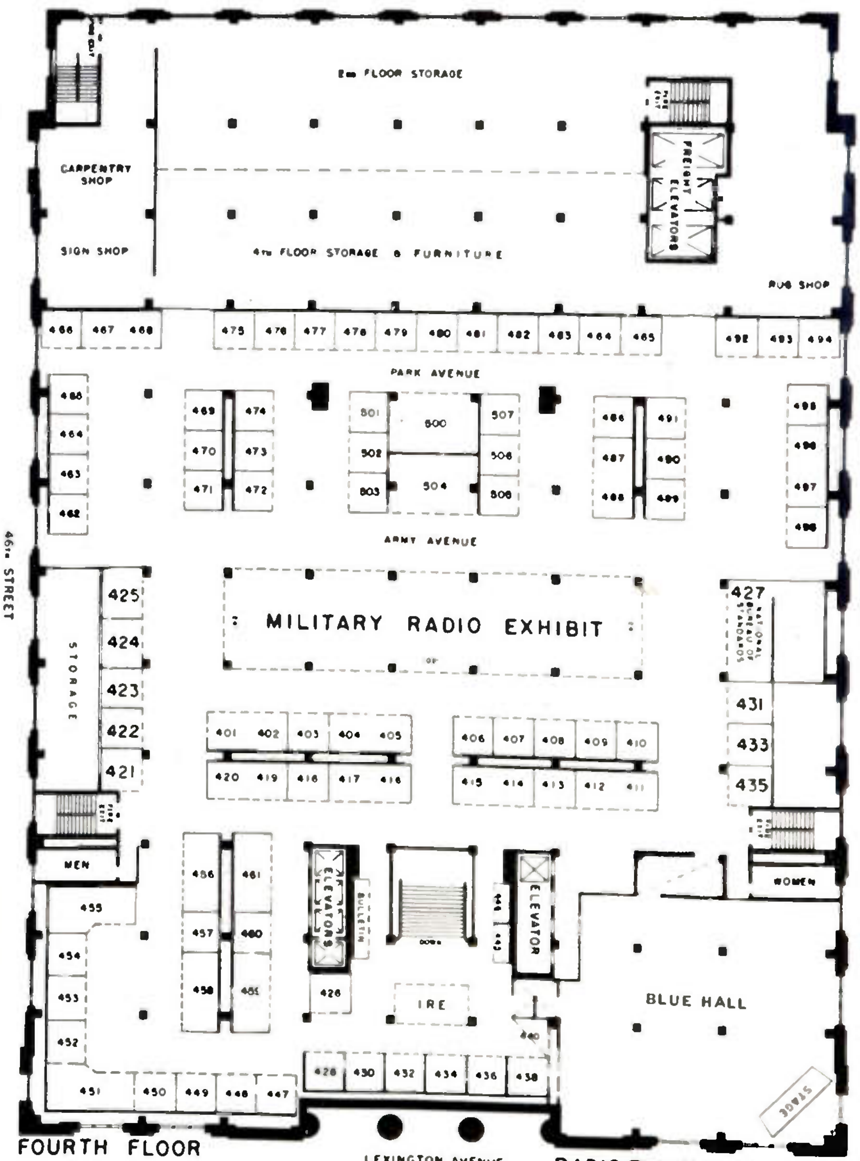
What to see at the Radio Engineering Show

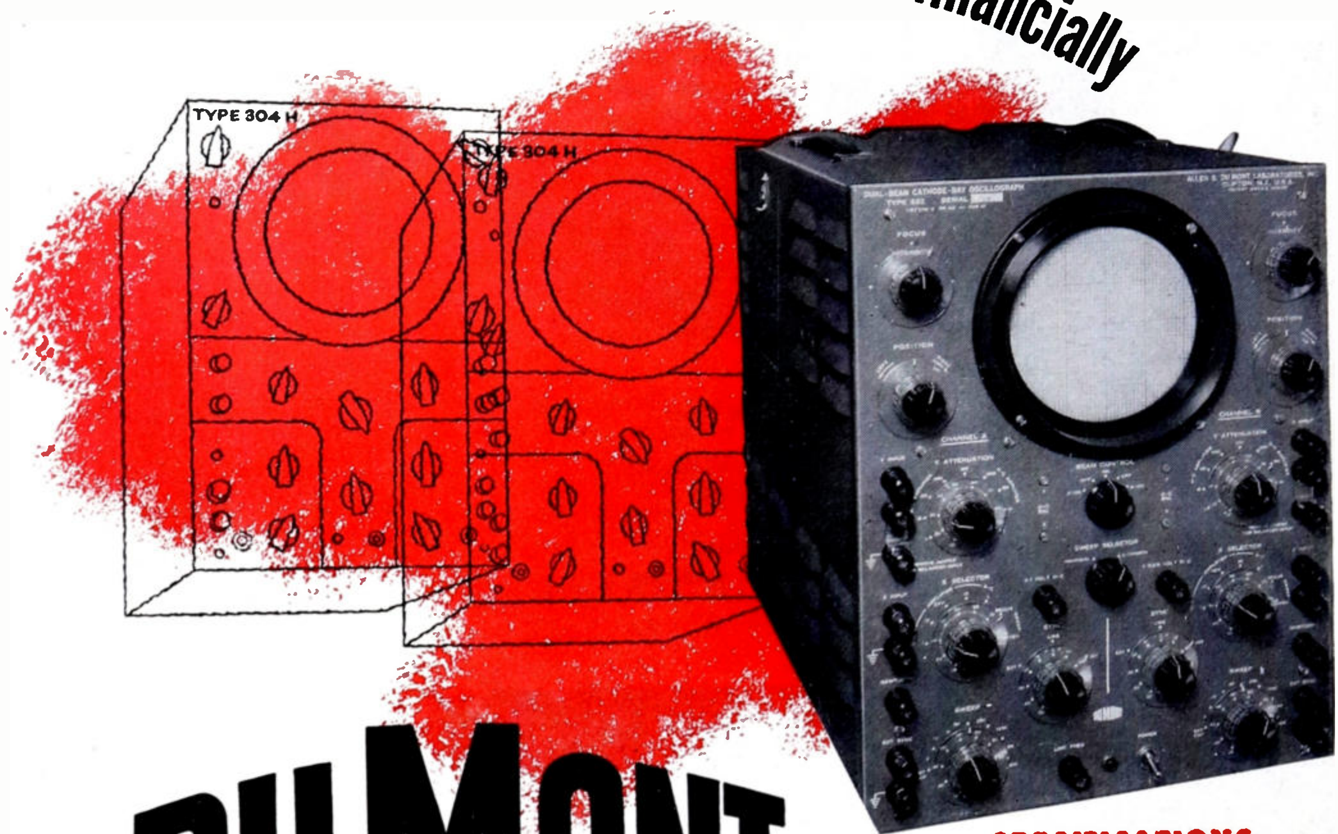
(Continued from page 88A)

Firm	Booth
Ethylene Chemical Corp., Summit, N.J. Teflon insulating material.	498
Fairchild Camera & Instrument Co., Jamaica 1, L.I., N.Y. Precision linear and non-linear potentiometers. Fairchild-Polaroid oscilloscope recording cameras. Oscillo-record (continuous-motion oscilloscope recording) camera.	238, 239
Falstrom Co., Passaic, N.J. Instrument and control panel boards. Aluminum chassis. Miscellaneous fabricated metal parts, including cold rolled steel, stainless steel, and aluminum for the communication, electronics and aviation fields.	423
Fairchild Recording Equipment Corp. Whitestone, L.I., N.Y. Professional synchronous tape recorders, disk recorders and transcription tables, multi purpose pickups, equalizers, preamplifiers, cuing amplifiers, thermostylus kits, control track generators, automatic framing devices, etc.	325A
Federal Telecommunications Labs., Inc., Nutley 10, N.J., Federal Hall VHF Omnidirectional radio range antenna, TV microwave relay link, and VHF airport radio direction finder, airborne distance measuring equipment, TV sound channel equipment. TV equipment: Flying spot scanner, master monitor, sync generator, high quality monitor, low voltage power supply, color sync generator, color flying spot scanner, and color master monitor. Impedometer, all metal traveling wave tubes, traveling wave tubes.	135
Federal Telephone & Radio Corp., Clifton, N.J., Federal Hall Pulse time modulation microwave equipment; transmitting, rectifier, and industrial vacuum tubes; selenium rectifiers and hf cable; mobile radio: vehicular and railroad.	135
Federated Metals Div., Amer. Smelting & Refining, New York 5, N.Y. Solders and other non-ferrous metals used in the radio and electronics industry.	414, 415
Federated Purchaser, Inc., Newark 4, N.J. R.C.A.—G.E.—Amphenol—Cornell-Dubilier—Hytron—Altec Lansing—University Loudspeakers—Triplett—Simpson—Weston—Federal Telephone & Radio—Radio Receptor Co.—Ward Leonard—Chicago Transformer—Sola Electric Radiart—Raytheon—Sprague—Centralab—Sangamo—Ohmite—Amperite—Alpha—Beldon—U. S. Engineering—Littlefuse.	473
Ferris Instrument Co., Boonton, N.J. Slotted measuring line, signal generators, microvolters, radio noise and field strength meters, and calibrators.	1-3

(Continued on page 94A)



A general-purpose DUAL-beam oscillograph to fit your needs technically and financially



the DU MONT TYPE 322

Not just another specialized dual-beam oscillograph, but a brand-new type designed for general development work but rugged enough for production testing and industrial applications as well. Compactness, lightweight, ruggedness and versatility mark the Du Mont Type 322 as another milestone in cathode-ray oscillography.

FEATURES

All the well-known features of the 304-H, and . . .

- Thoroughly field-tested.
- Individual and common time bases with driven or recurrent sweeps and sweep expansion on all sweeps.
- Conventional single-ended input with stepped and vernier attenuators, or balanced input with no attenuation, on both Y-axes.
- Concentric controls for easy-to-operate, compact control panel.
- High-gain D-C amplifiers on both channels.
- Amplitude calibration on either channel on both axes.
- Illuminated scale with dimmer control.

SPECIFICATIONS

Cathode-ray Tube — Type 5SP — Dual-beam Cathode-ray Tube. Accelerating potential, 3000 volts.

Y-Deflection Sensitivity — 0.028 peak-to-peak (0.01 rms) volts/inch from D-C to 300 KC (50% down at 300 KC); A-C coupling, 10% down at 5 c.p.s.

X-Deflection Sensitivity — 0.3 peak-to-peak (0.1 rms) volts/inch from D-C to 300 KC (down 50% at 300 KC); A-C coupling down 10% at 5 c.p.s.; common, D-C to 200 KC (down 50% at 200 KC).

Linear Time Base—Recurrent and driven sweeps variable in frequency from 2 to 30,000 c.p.s. Front panel connections provided for lower frequency by adding external capacitance.

Intensity Modulation — Input impedance 0.2 megohm, paralleled by 80 $\mu\mu$f. Negative signal of 15 volts peak blanks beam at normal intensity settings.

Beam Control Switch — On front panel to turn beams on or off independently or simultaneously.

Calibrator — Regulated potentials of 50 millivolts and 1 volt peak-to-peak squarewave at power line frequency available at front panel binding posts.

Power Source — 115/230 volts — 50-400 c.p.s. — 225 watts.

Dimensions — Height 15¾", width 12½", depth 22⅞", weight 75 lbs.

$835.00

Write for complete technical details: **Instrument Division Allen B. Du Mont Laboratories, Inc. 1500 Main Avenue, Clifton, N. J.**

SEE US AT THE RADIO ENGINEERING SHOW, BOOTHS 120 to 128

What to see at the Radio Engineering Show

(Continued from page 92A)

Firm	Booth
The Filtron Co., Inc., Flushing 55, N.Y. Miniaturized radio interference suppression filters.	45
Fisher Radio Corp., New York 17, N.Y. High quality receivers and amplifiers, preamplifier, the Concertone tape recorder, and the RJ speaker system.	314A
Ford Instrument Co., Div. Sperry Corp., Long Island City 1, N.Y. Synchro and servo motors, resolvers, mechanical differentials. Low inertia servo motors with high voltage control windings. Work directly into tubes thus eliminating need for transformer in amplifier.	213
Freed Transformer Co., Inc., Brooklyn 27, N.Y. Transformers, reactors, precision laboratory measuring equipment, electric wave filters, high "Q" toroid inductors, pulse transformers.	103
Furst Electronics Chicago 6, Ill. Wide band dc amplifier, wow meter, various regulated power supplies.	363
The Fusite Corp., Cincinnati 13, Ohio A complete line of Fusite glass-to-metal terminals for hermetic sealing of all types of electrical components. Exhibit will include Fusite terminals applied to actual products of our customers in the field of electronics, instruments, switches, transformers, and refrigerator compressors.	360
Gates Radio Co., Quincy, Ill. Broadcast and television speech input equipment, a vhf 250 watt FM relay transmitter, standard broadcast 1000 watt transmitter. Several assemblies and groups of switching and control apparatus for broadcast and communications systems of various types.	214
General Ceramics & Steatite Corp., Keasbey, N.J. Technical ceramics, including steatite, porcelain, titanates, zircon steatite, light duty refractories, end products consisting of insulators, coil forms, switch parts, sockets, connectors, and hermetic seal terminal bushings of the metalized and compression seal types, ferramics (magnetic ferrites). Dielectric materials.	47
General Electric Co., Schenectady 5, & Syracuse, N.Y. Second harmonic magnetic converter, photoelectric recorder. Capacitors, soldering irons, relays, rectifiers, switches phono cartridges and diamond styli, speakers tone arms, preamplifiers TV components—Germanium power rectifier, germanium diodes—TV test equip.—oscilloscopes—marker generator—sweep generator power supplies. Tubes. C-R—TV & radio receiving TV picture—uhf rf power, gas switching, uhf and vhf types.	113-119
General Electric Co., Apparatus Div., Schenectady 5, & Syracuse, N.Y. Transformers: MIL-T-27, Permafil, and silicone. Voltage regulators, magnetic amplifier, dry-type high voltage power pack, magnetic current reference.	397B
General Precision Laboratory, Inc., Pleasantville, N.Y. G.P.L. Camera chain, 35 mm TV projector by Simplex, G.P.L. 16 mm projection equipment, completely remotely controlled TV camera.	18-20
General Radio Co., Cambridge 39, Mass. UHF TV Station monitor. UHF Impedance Measuring Equipment. Unit Instruments. Vacuum-Tube bridge. Variac ® Autotransformers. Sound-Level Meter. Sound-Survey Meter, Sound-Level Calibrator. Octave-Band Analyzer. Sound Analyzer. Vibration meter, Vibration analyzer. Vacuum-Tube Voltmeters. Crystal Voltmeter. Megohmeters, light meter, output power meters, microvolter, standard-signal generators, coaxial elements.	92, 93
General Transformer Co., Homewood, Ill. Transformers—Power audio, reactors, for commercial and military applications. Rectifier power unit and battery charger—PP34. Perma power, ac-dc power supplies to convert battery operated radio sets to 110 v ac line operation.	449
John Gombos Co., Inc., Irvington 11, N.J. Tracking switches, dial light sockets, filters, terminator assemblies, beryllium copper connectors, and precision parts used in the electronic industry.	263

(Continued on page 95A)

What to see at the Radio Engineering Show

(Continued from page 94A)

Firm	Booth
Grant Pulley & Hardware Co., Flushing, N.Y. Electronic equipment slides—a complete line of slides which enable electronic apparatus (chassis, consoles, racks, etc.) to slide in or out of its casing. Mechanisms available which can tilt unit to any desired angle for servicing. Slides are made for loads up to 2,000 pounds. Complete engineering liaison service available.	462
Gray Research & Development Co., Inc., Hartford 1, Conn. Industrial color television; Telop, studio TV projector; Multiplexer, studio TV use. Camera turret, studio TV use. Transcription arms & equalizers.	S-9, S-10
Grayhill See: Wally Swank	221
Green Instrument Co., Inc., Cambridge 39, Mass. Pantograph engraver for name plates, dials and scales. Instrument panels up to 19" in height by any length. Rotary tables, self-centering vies, clamping fixtures, and cutter grinder. Special machinery for production engraving.	369
Guardian Electric Mfg. Co., Chicago 12, Ill. Hermetically sealed relays, AN approved electrical controls, radio relays, and television components.	367, 368
Gulton Mfg. Corp., Metuchen, N.J. Acoustic and electric delay lines, Glennite subminiature capacitors and components, cathode followers and amplifiers, capacitor bridge transducer display (microphones, hydrophones, vibration pickups, ultrasonic transducers, phono pickups) Glennite piezoelectric ceramics.	428
Halldorson Co., Chicago 40, Ill. Complete line of transformers and reactors for new construction and replacement purposes. High fidelity transformers for broadcast and music reproducing applications.	470
H. B. Hardman, Belleville, N.J. See: Robinson, Edward	479
Harrison Radio Corp., New York 7, N.Y. "Electronic Information Quotient" An automatic device loaned by the U.S.N. to rate a person as to accuracy and speed of response to questions.	323
A. W. Haydon Co., Waterbury 20, Conn. Chronometric governed dc timing motors, time delay relays, and sequence or repeat cycle timers.	54
Helland Research Corp., Denver 9, Colo. Galvanometers, 400 and 500 recorders, 708 table mounted recorder and 708 rack mounted recorder, and 800-1 single channel amplifier.	304
Heldor Mfg. Co., Bloomfield, N.J. Hermetic seal bushings, weld studs, and MIL-T-27 cans, blind inserts. Bushing assemblies, header assemblies.	S-17
The Helipot Corp., S. Pasadena, Calif. Precision linear single and multiturn potentiometers and servo controls, turns and position indicating dials, and miniaturized servo controls—linear.	84
Heminway & Bartlett Mfg. Co., New York 18, N.Y. Twisted nylon lacing cords, and flat braided nylon lacing cords.	327
Heppner Mfg. Co., Round Lake, Ill. Loud speakers, horizontal transformers, ion traps, beam centering controls, focus devices, vibration mounts, gun heaters, cable assemblies, rf chokes, ferrite antenna coils, correcting magnets, standpipe connectors.	321
Hermetic Seal Products Co., Newark 7, N.J. Hermetic seals, glass-metal. Hermetically sealed terminals, bushings, headers, plugs and bases, in single terminals or multiheaders for the electronic industry.	129
Hewlett-Packard Co., Palo Alto, Calif. Electronic test and measuring instruments. S24A test set—for gun fire radar testing.	40, 41

(Continued on page 122A)

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

PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 40 — *February, 1952* — NUMBER 2

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES

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

Harold L. Kirke

VICE-PRESIDENT, 1952-

Harold L. Kirke was born on January 7, 1895, in London, England. He received army wireless training, and during World War I, he served as a radio instructor to officers classes in the British Army Signal schools. From 1920–1924, he was with the Marconi Wireless Company experimental section, and was an important member of a group that built and operated the Writtle Broadcasting Station. In 1924, Mr. Kirke joined the British Broadcasting Corporation as a development engineer, and in 1925, he became the head of the development department. This section later became the research department. Mr. Kirke is now the Assistant Chief Engineer of BBC.

Mr. Kirke was the chairman of the radio section of the Institution of Electrical Engineers from 1944–1945, and was also a member of the council of that body. In June, 1947, he was made a Commander of the Order of the British Empire. He is the chairman of the Acoustics Group of the Physical Society, and a member of the Physical Society Council.

Becoming a Member of the IRE in 1925, Mr. Kirke was elevated to Senior Member in 1943, and was presented the IRE Fellow Award in 1945, for "services to broadcasting in the British Isles and in particular for his leadership in the research activities of the British Broadcasting Corporation."

What Are We Going to Use for Engineers?

R. D. BENNETT

In the early days of American life, the scope and functions of government were relatively restricted and simple. The more complex industrialized system of today has brought a vast expansion in the scope of governmental operations, and has correspondingly increased their individual complexity and their interrelationships. This change, in turn, has imposed great burdens and responsibilities on government officials and has made serious demands on their skill, experience, and ingenuity.

One of the resulting technical problems, which is of basic significance to the people of all modern countries, is clearly and constructively analyzed in the following guest editorial from the Technical Director of the United States Naval Ordnance Laboratory at Silver Spring, Maryland. The author is a Fellow of the IRE.—*The Editor.*

Those of us who have responsibilities, whose successful discharge depends on a continuing supply of young engineers, view with great concern the disparity between the engineering work load we expect and the supply of talent likely to be available.

Last year's crop, while not the highest on record, was a bumper one, yet the demand was at least twice the supply. The average for the next five years will be about half those available in 1951. While demand in research and development may taper off in a year or two, the increased need in production may easily offset any such decrease. Out of the relatively meager supply of engineers available, the armed forces may be expected to take the substantial fraction who have been trained under ROTC auspices, and unless we formulate policies to the contrary, some additional ones will be taken through selective service.

Therefore, we find ourselves in a position where our technical output is limited by our trained engineering manpower. Unless we husband this trained manpower with care, this output may be too low for us to meet successfully the challenge we face. What can we do?

It seems clear that there will be no easy solution to this problem; however, a co-operative effort on the part of all hands may yield an adequate answer.

First, we can choose between what we would like to have and what we really need.

Secondly, we can simplify some of the military "gadgetry" we are building.

Thirdly, we can curtail, temporarily, our demands for less essential goods and services.

Next, we can make better use of the engineers we do have. This means we must cease hoarding for hoped-for contracts; improve our management of engineering talent; extend our work week; fleet up subengineering talent; separate and segregate the less difficult parts of our engineering jobs and get them done by less extensively trained people; get engineers in the armed services into jobs which use their engineering talent.

Then, we can intensify our on-the-job training of engineers or potential engineers. We can expand opportunities for undergraduate as well as advanced academic training. We can teach relatively untrained people to do specific jobs, and we can train management to generate better teamwork.

Finally, we must broaden the base of our engineering manpower structure. To do this we need to get less shadow and more substance into the secondary school training of those who have the wit to absorb it. We must encourage more women to enter the profession. We must be more vocal in demonstrating that engineering training is a good basis for almost any calling, and we may have to adjust the relative financial rewards of engineers so that it does not take ten years for the engineer to catch up economically with his fellow high school graduate who took up a trade.

We can prosecute all these approaches to the solution of our problem. If we prosecute them with vigor, we can haul ourselves out of a tight situation. If we do not, we shall not only fail in our effort to have both guns and butter; we can also expect our troubles to multiply and our security to be seriously jeopardized.

Admissions and Transfers*

RAYMOND F. GUY†, FELLOW, IRE

THE BEST INTERESTS of our profession are served by having a strong, efficient, and smoothly functioning society such as our IRE. The high esprit de corps and the pride in membership in our Institute are a reflection of many tangible and intangible factors including, through regional representation, the opportunity to participate in the formulation of policies and practices, candid and frequent discussion of Institute problems, and the maintenance of high professional standards.

As in all professional societies, the IRE has various membership grades which reflect the professional competence, experience, and character of the individual, and his standing in his chosen profession. The grade of Student Member provides for membership by engineering undergraduates. The grade of Associate provides an avenue through which nonprofessional people may become members of the Institute, without voting and certain other privileges. The grade of Member denotes professional experience and competence, and provides for professional recognition of qualified young men early in their careers, or others of as yet limited professional ability. The grade of Senior Member recognizes and denotes high professional competence, extended experience, and the ability to assume major responsibility in the practice of one's profession. And the grade of Fellow is an honor conferred by the Board of Directors for outstanding contributions to the science or technology of radio and allied fields. All members are encouraged to aspire to the highest grade for which they are qualified, and to take pride in the professional competence which it denotes.

Our members are entitled to have the stature of the membership grades maintained, and applications for admissions and transfers handled smoothly, fairly, and promptly. Your officers, directors, committees, and headquarters staff make every effort toward that end. The Admissions Committee is a hardworking and conscientious group, presently under the chairmanship of Mr. H. P. Corwith. The Membership Committee, currently under the chairmanship of Mr. Ringland Krueger, is deeply concerned with encouraging membership among qualified nonmembers, upgrading qualified members, and analyzing the reason for resignations and minimizing them. And the Public Relations Committee, now under the chairmanship of Mr. Lewis Winner, is developing the program by which the Institute and the accomplishments of its members are brought to the attention of their employers, the industry, and the public.

These Committees have as their functions the maintenance of the stature of the membership grades; enhancement of knowledge and pride in what they represent; encouragement of qualified nonmembers to seek membership and of qualified members to advance their grades; prompt, impartial, fair, and helpful action on membership applications; the maintenance of staisfactory public relations; and enlightenment of the public, the industry, and others concerning the accomplishments and activities of our members, the Institute, and the profession.

It is believed particularly timely to familiarize members of the Institute with the manner in which admissions and transfers are handled, and to solicit their co-operation in overcoming some of the difficulties which add to the burden of the Admissions Committee, the headquarters staff, some of your officers, and new applicants.

Applications for admission or transfer go through a routine which has been modified at intervals in an attempt not only to reduce the time of processing, but also to insure the utmost fairness to the applicant. Until recently applications for admission or transfer, upon receipt, were processed in the following order:

1. They went to the File Department to determine if the applicant had formerly been a member of the Institute.
2. If not, they were sent to the Applications Department, where the reference forms were mailed to those listed by the applicant.
3. The application and the returned reference forms were sent to the File Department to determine whether the references were qualified under the bylaws.
4. The application, reference forms, and other pertinent information were transmitted to the Admissions Committee for consideration.
5. The Admissions Committee, meeting monthly, recommended either that the application be approved, that additional information or qualified sponsors be sought through the applicant, that no affirmative action be taken because the applicant was not qualified, or that a suggestion be made to the applicant that he accept a lower grade of membership consistent with his qualifications.
6. Applications recommended for approval are transmitted to the Executive Committee for final action. Applications not recommended for approval go to the Membership Relations Co-ordinator for review and any special action deemed advisable in collaboration with the Admissions Committee and the Executive Committee. The Membership Relations Co-ordinator is a member of the IRE Executive Committee, and is responsible to it for activities of the Membership, Admissions, and Public Relations Committees.

Although many applications were received without adequate qualified references, they nevertheless were forwarded to the Admissions Committee because, in some cases, it had been found that the applicants were so obviously qualified that the reference requirements could be waived in part. However, the number of such applications lacking the required number of qualified references has become so large that the processing routine has been modified. Under the new routine, applications from the United States and Canada having insufficient qualified references are held at headquarters while the applicant is requested to complete the quota. They are then processed as formerly. Applications outside of the United States and Canada are processed under the old routine. The reason for the difference lies in the fact that most of the applications are from the United States and Canada, little time is consumed in corresponding with applicants, and in most localities qualified references may be found. On the other hand, in foreign countries there are relatively few applicants, correspondence requires a relatively long time, and qualified references are relatively few in number.

Every practical step has been taken to expedite favorable action on applications when the applicant is qualified and when he has provided the information and list of qualified references as specified in the Institute bylaws. Consideration of an application must, of course, be based upon the information contained in the file of correspondence.

In the processing of applications and transfers, two difficulties recur which delay action and unnecessarily add to the work load of the office staff and your fellow members who contribute their time and effort in committee to serve you. One of the most troublesome delays in the admissions or transfer processing arises because too often the references given are either nonmembers of the Institute, or they are not of the required membership grade. It is found frequently that only one or two of the references comply with the membership-grade requirements for references, and at times none do.

The Institute bylaws and membership literature specify exactly, and in the most simple terms, that on an application for Associate Grade, at least three references of Associate or higher grade are required; for Member, at least four of Member Grade or higher are required; and for Senior Member, at least five of Senior Member or Fellow are required. Compliance with these simple requirements ordinarily is not difficult. But if assistance is desirable in seeking qualified references, it is recommended that the applicant attend one or more of his local IRE Section meetings and obtain the assistance of the officers, which will be given gladly. If it is not feasible to attend a Section meeting, one may obtain assistance from IRE headquarters, or from among those qualified references to whom one is known. If these avenues are not available, discretion will be exercised in accepting references from members of other scientific societies in related fields, provided that they hold comparable grade in those societies. It is also possible to exercise discretion in difficult cases where one of the references who is a nonmember of IRE is the applicant's supervisor and is familiar with his work.

The Admissions Committee seeks to be helpful and constructive and to aid appli-

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† National Broadcasting Co., RCA Building, New York 20, N. Y., and Membership Relations Co-ordinator of the I.R.E.

cants. But as your guardian in the maintenance of member grade standards and stature, they have no choice but to withhold favorable action if the applicant is not qualified or if he has neglected to comply with the requirements laid down in the bylaws.

In preparing applications, all too frequently the applicant neglects to submit sufficient information to substantiate the experience or other qualifications which he has claimed. It is helpful for an applicant to bear in mind in preparing applications that the Admissions Committee will know only what he submits to them, and must judge his work largely upon the description he he provides. The comments of his references are relied upon to substantiate his claims and not, in general, to add further details. An applicant should recognize that he has an obligation to comply with the bylaws of an organization in which he seeks membership or upgrading, and he should perceive that the processing of his application would be facilitated if he complied as completely as possible.

It would be helpful for applicants, insofar as possible, to select references to verify his entire experience, rather than having all of them selected to verify only the last year or two. This in itself may make it easier to find references, particularly if the applicant has recently transferred to new employment and has not yet established many new acquaintances.

The second main difficulty arises because persons receiving reference forms, for various reasons, do not reply, or do not do so promptly. Frequently, this is found when the person to whom it was referred feels that he does not know the applicant, does not know enough about him, or is reluctant to recommend a denial of the application. In such cases, it is necessary for the headquarters group to follow up the original reference with further correspondence, pointing out that the application remains dormant until the necessary number of references have been heard from.

It is urged that those receiving reference forms return them promptly with their recommendations or, if they prefer not to make recommendations, so to indicate in order that other references may be used or sought. In fairness to the applicant and others involved, those named as references should inform headquarters of their wishes in the matter. Indefinite delay in returning reference forms may prejudice unfairly the ultimate outcome of an application, whereas prompt return with a note that the applicant is unknown or not well enough known will not influence the action taken.

A large proportion of new applications for membership are stimulated by association with IRE members, and many of the applications for transfer are stimulated by the Sections Committee or officers. It is urged that, where possible, members of the Institute should aid new applicants in filling out their forms so as to insure that experience is properly specified, at least the minimum number of qualified references is named, and that the forms are complete and properly addressed. In so doing you can be of great assistance in eliminating a duplication of work and much loss of time, and you will also help prevent loss of membership on the part of applicants who become irritated or discouraged by extended correspondence, requests for more names, and the like. And it is urged that members applying for upgrading facilitate the prompt processing of their applications by specifying at least the minimum number of qualified references and providing complete information. Those charged with responsibility for approving or denying requests for admission or transfer have for their guidance a manual which has been developed and improved through the years. It has insured consistency and uniformity of action, and maintenance of the stature of the grades. You are urged, in preparing applications, to keep in mind that an applicant's qualifications must be judged by the information contained in the written material presented for consideration, which normally consists of the applicant's own statements and the recommendations and comments of his qualified references.

It is recommended that applicants provide their references with a summary of their technical training and experience because, frequently, those named as references have an incomplete recollection of an applicant's professional history. In the absence of such recollection, the persons named as references must request it, or restrict their comments, to the possible disadvantage of the applicant.

The Application of Damping to Phonograph Reproducer Arms*

WILLIAM S. BACHMAN†

This paper is published with the approval of the IRE Professional Group on Audio, and has been secured through the co-operation of that Group.—*The Editor.*

Summary—**Large forces are developed at the stylus tip of a conventional phonograph reproducer arm because of excitation of the resonance of the arm mass with the suspension compliance. This paper presents an analysis of the problem and describes a reproducer-arm design in which mechanical resistance is introduced in the pivots. By this means, control of the arm resonance is obtained without increasing the stylus-tip impedance of the reproducer.**

THE STYLUS-BEARING force of a disk reproducer upon the record must be small to limit the bearing pressure to reasonable values. This low force is called upon to hold the stylus in contact with the groove against the dynamic forces developed at the stylus point at arm resonance, and to move the reproducer arm about its vertical pivot along the spiral path which the record groove presents. By providing damping in both the horizontal and vertical pivots, this resonant force is greatly reduced, improved resistance to external shock is obtained, and protection against damage from accidental release of the reproducer head is achieved.

At low frequencies, say below 500 cps, the mechanical system of a phonograph reproducer arm and pickup cartridge may be represented by a mass suspended on a spring. In Fig. 1(a) (see following page), m represents the effective mass of the arm and cartridge assembly, referred to the stylus point. This is always less than

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† Columbia Records, Inc., 799 Seventh Avenue, New York, N. Y.

the weight of the assembly since it is partly supported at the far end by the pivot about which it rotates. The compliance (the reciprocal of stiffness) of the stylus suspension is represented by c.

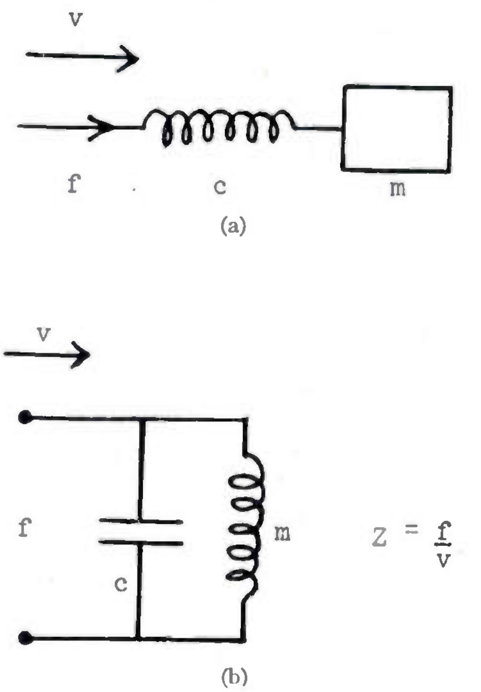

Fig. 1—(a) Mechanical schematic diagram of a conventional reproducer arm and cartridge at low frequencies.
c = the needle-suspension compliance of the reproducer cartridge
m = the effective mass of the arm and cartridge assembly, referred to the stylus tip
f = the force developed at the stylus tip
v = the velocity of motion of the stylus tip.
(b) Electrical equivalent of mechanical system of 1(a).
Z = the mechanical impedance at the stylus tip.

Below the resonant frequency of the system, which is often in the 30-to-60-cps region, the motion of the mass corresponds to the motion of the driving point. In other words, the arm and cartridge follow the slow progress of the spiral groove. Above the resonance, the system becomes mass controlled, and the arm does not follow the rapid undulations which the modulated groove imposes upon the reproducer stylus. This difference in motion between the stylus and the arm provides the stimulation of the pickup, to which the output voltage is proportional.

The operation of the system may also be described in electrical terms by referring to the analogous circuit in Fig. 1(b). The mass is analogous to inductance, compliance to capacitance, force to voltage, and velocity to current. The unidirectional motion which describes the radial motion of the arm in following the spiral groove corresponds to dc in the electrical system. The alternating velocity imparted to the stylus by the groove modulation corresponds to ac in the electrical network. Above the resonant frequency this velocity (or current) divides between the compliance, c, and mass, m, practically all of it admitted by the compliance in the useful range of the reproducer. The output voltage is proportional to the velocity to which the compliance is subjected in a magnetic type of reproducer, or to the integral of this velocity in a displacement-sensitive reproducer.

At the resonant frequency of the circuit of Fig. 1(b), the impedance, Z, reaches very high values, limited only by the Q of the system. The Q of the mass element is very high. There is usually some dissipation in the suspension compliance, however, but its value must be limited if the mid-range impedance of the system is not to be made too high.

To indicate schematically where the dissipative element appears, the mechanical system is shown in Fig. 2(a). This mechanical system schematic shows that deflection of the suspension spring is accompanied by work done in the friction or mechanical resistance element r.

The electrical equivalent of this mechanical system is shown in Fig. 2(b).

As a practical example, let us assume a suspension compliance of 10^{-6} cm/dyne and an effective arm mass, referred to the stylus tip, of 20 grams. These are in the range likely to be encountered in actual practice. Such an arm and cartridge assembly would have an arm resonance frequency of

$$f = \frac{1}{\sqrt{4\pi^2 mc}} \frac{1}{\sqrt{4\pi^2 \times 20 \times 10^{-6}}} = 35.6 \text{ cps.}$$

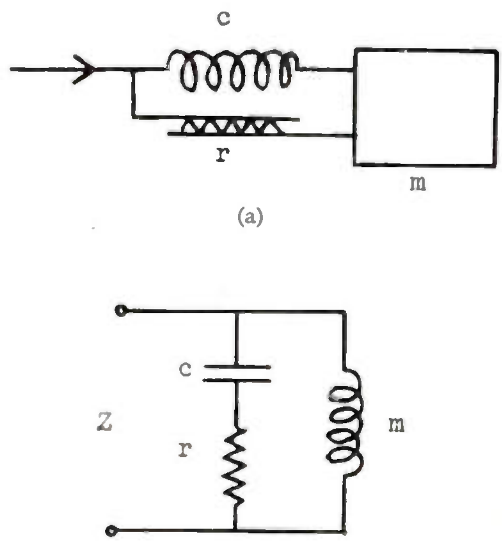

Fig. 2—(a) Mechanical schematic of phonograph reproducer arm and cartridge, with damping in the needle suspension.
r = the effective mechanical resistance resulting from the dissipation in the needle suspension. (b) Electrical equivalent of mechanical schematic in (a). Above resonance the impedance at the needle point Z can never become lower than r.

To critically damp this resonance, the value of r would have to be

$$r = \sqrt{\frac{4m}{c}} = \sqrt{\frac{4 \times 20}{10^{-6}}}$$

$$= 8{,}950 \ \frac{\text{dyne sec}}{\text{cm}} \text{ (mech. ohms).}$$

Now, inspection of the circuit in Fig. 2(b) shows that, above the resonant frequency, the impedance approaches r. (X_m gets very large and X_c gets very small.) Suppose that a velocity of 5 cm/sec rms were imposed upon the stylus. This velocity is considerably below the peak program level encountered in most records. The force at the needle point would be

$$f = rv = 9{,}850 \frac{\text{dyne sec}}{\text{cm}} \times 5 \frac{\text{cm}}{\text{sec}}$$

$$= 44{,}600 \text{ dynes or } 45.5 \text{ grams.}$$

If the included angle of the groove were 90°, the upward component of this lateral force would be equal to it, and the needle-bearing force would have to be in excess of $45.5x\sqrt{2}$ grams (45.5 is the rms value) to insure contact with both sidewalls of the groove. Such a value of needle force is absurd, particularly for microgroove records, so it becomes obvious that effective damping of the arm resonance cannot practically be obtained by this method.

It has been usual practice to increase arm mass to move the resonant frequency farther below the desired transmission band. This reduces the incidence of arm-resonance excitation by program material, but the resonant impedance is thereby increased, which further increases the susceptibility to jumping as a result of accidental mechanical shock.

If, in the circuit of Fig. 2(b), the resistive element were inserted in series with the mass element, as shown in Fig. 3(a), it would not affect the driving-point impedance of the system above the resonant frequency. In other words, the impedance above resonance of the circuit of Fig. 3(a) would be the same as that of Fig. 1(b), namely, approaching X_c.

Mechanically, Fig. 3(a) is represented by Fig. 3(b). The proper functioning of the system in the useful signal frequency range, as mentioned before, depends upon all of the imposed motion, v, being accommodated by the spring, c. Only below the resonant frequency is the motion of m significant. Putting the resistance in this position does mean that it will have to "carry the dc," or, in other words, to accommodate the motion imposed by the slow radial progress of the spiral. The velocity of this spiral is quite low. Suppose the record is turning at 78 rpm, with a spiral pitch of 100 lines to the inch. The radial velocity will be

$$v = \frac{1 \text{ in}}{100 \text{ rev}} \times \frac{78 \text{ rev}}{\text{min}} \times \frac{\text{min}}{60 \text{ sec}} = 1.3 \times 10^{-2} \text{ in/sec}$$

$$\text{or } 3.3 \times 10^{-2} \text{ cm/sec.}$$

The force necessary to move the arm at this velocity against resistance r is

$$f = rv = 8{,}950 \times 3.3 \times 10^{-2} = 295 \text{ dynes or } 0.3 \text{ gram.}$$

This is a satisfactorily low value, and furthermore, it is still lower with fine groove records turning at lower speeds.

From the mechanical schematic in Fig. 3(b), it is seen that the mechanical resistance element must be installed between the arm and the motorboard. It could be applied against any part of the arm, or at its pivots,

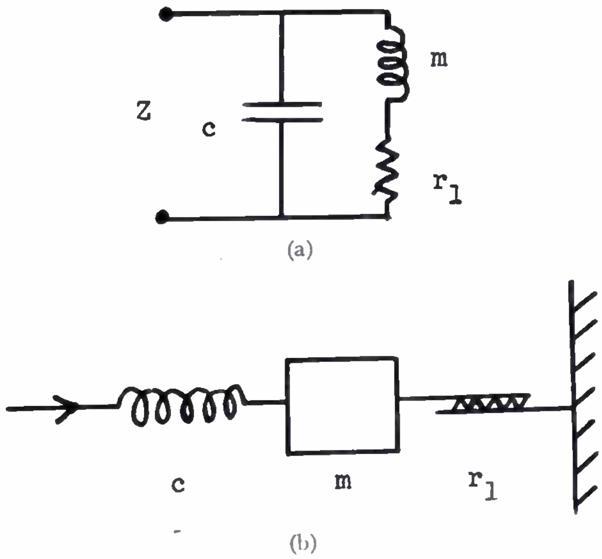

Fig. 3—(a) Modification of the circuit in section (b), which avoids raising the input impedance above resonance without sacrificing the effect of r upon damping. r_1 is a new circuit element, and not the resistance due to damping in the suspension compliance. (b) Mechanical schematic of electrical circuit of section (a). r must be inserted between the arm mass and the record or motorboard.

in a wide variety of forms. The idea of some friction connection in the pivots, or elsewhere, is violently opposed to the usual concept of a reproducer arm. In fact, failure of many arms to operate satisfactorily at the low bearing forces which LP microgroove records impose can be traced to excessive pivot friction. This friction is quite different from the desired mechanical resistance. It is true that friction causes energy to be dissipated as heat, as does mechanical resistance. The big difference is that pure mechanical resistance does not change in value as the velocity is varied. In ordinary rubbing friction, the resistance offered varies violently with the imposed velocity. A good familiar example is the large difference between starting friction and running friction, which is observed in bearings. As an example of linear mechanical resistance, consider a boat, floating in water. The resistance it offers is easily demonstrated by the work the engine has to do to move it at constant speed through still water and dead air. Yet this same boat may be moved, slowly, to be sure, with a surprisingly small force in still water and dead air. In this example, the friction of the boat moving through the water is linear, assuming that the velocity range does not extend into turbulence of the water. The friction behaves as a true mechanical resistance, in which the velocity of motion is proportional to the applied force.

While friction is one of the most severe restraints in the design of many mechanical systems, it is almost paradoxical that linear mechanical resistance is very difficult to obtain. Probably the purest form of mechanical resistance is that which obtains from the motion of a short-circuited conductor in a uniform magnetic field. This is "pure" only in the sense that it is linear. It cannot be separated from the mass of the conductor through which it is developed.

It is possible to compute the amount of mechanical resistance which a given conductor will develop in a magnetic field by using the two basic relations pertaining to a moving conductor in a uniform magnetic field, viz.,

$$f = Bli \tag{1}$$

$$e = Blv, \tag{2}$$

where

f = the mechanical force
b = the flux density
l = the length of the conductor
i = the current in the conductor
e = the voltage induced in the conductor
v = the velocity of motion of the conductor.

Now,

$$r = \frac{f}{v} = \text{mechanical resistance.} \tag{3}$$

Substituting (1) and (2) in (3),

$$r = \frac{B^2l^2i}{e} = \frac{B^2l^2}{R}, \tag{4}$$

where R is the electrical resistance of the conductor.

Now,

$$R = \frac{pl}{A}, \tag{5}$$

where

p = specific resistivity of the conductor
l = length of conductor
A = cross sectional area of conductor.

Substituting (3) in (4),

$$r = \frac{B^2lA}{p} = \frac{B^2V}{p} = \frac{B^2M}{pd}, \tag{6}$$

where

V = volume of conductor
M = mass of conductor
d = density of conductor material.

The ratio M/r has the dimension time and is analogous to the time constant L/R of an electrical circuit comprising these elements.

Solving (6) for the time constant:

$$\frac{M}{r} = \frac{pd}{B^2}.$$

Assuming a copper ring type of conductor in an annular air gap having a flux density of 10,000 lines per square centimeter,

$$\frac{M}{r} = \frac{1.6 \times 10^{-6}\ \text{ohm cm}}{1} \times \frac{8.9\ \text{gram}}{\text{cm}^3} \times \frac{\text{cm}^4}{10^{-8}\ \text{volt}^2\ \text{sec}^2}$$

$$\frac{M}{r} = 14.2 \times 10^2 \frac{\text{ohm cm}^2\ \text{gm}}{\text{volt}^2\ \text{sec}^2} \times \frac{\text{dyne sec}^2}{\text{gram cm}}$$

$$\times \frac{\text{volt}^2}{\text{watt ohm}} \times \frac{\text{watt sec}}{\text{joule}} \times \frac{\text{erg}}{\text{dyne cm}} \times \frac{\text{joule}}{10^7\ \text{erg}}$$

$$= 142 \times 10^{-6}\ \text{sec.}$$

If the size of conductor were chosen so that 100 mechanical ohms would be developed, the mass of the conducting ring would be

$$M = \frac{M}{r} \times r = 142 \times 10^{-6}\ \text{sec} \times \frac{100\ \text{dyne sec}}{\text{cm}}$$

$$= 14.2 \times 10^{-3}\ \text{gram.}$$

The presence of mass with the resistance in this mechanical system is analogous to the presence of inductance in an electrical resistor. A resistor of 100 ohms having 14.2 millihenries built in would be far from a noninductive resistor.

Another, more attractive, method of obtaining linear mechanical resistance is through the use of viscous liquids. The resistance may be obtained by moving an impeller through a liquid, forcing the liquid through an orifice, or utilizing the viscous fluid as a film in shear. One difficulty to which these methods are subject is the change of viscosity with temperature. While this effect is large with most petroleum oils, the effect is much smaller in silicone oils. Since the specific value of the mechanical resistance is not critical in this application, either type of oil may be used over a reasonable temperature range.

In Fig. 4, a cross-sectional view of a design using a fluid film in shear is shown. The two concentric surfaces which bound the film are those of a ball and socket. The ball is part of the arm and the socket part of the mounting base. The arm is suspended on the point of a stud which is part of the base. This point is

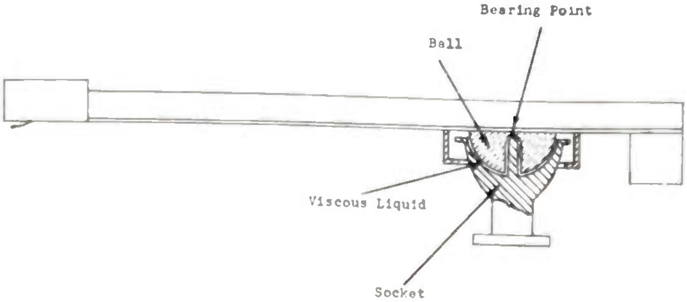

Fig. 4—Cross-sectional view of phonograph reproducer arm having viscous film damped pivot.

at the center of both the ball and socket, and above the center of gravity of the arm. A clearance of about 0.006 inch is maintained between the surfaces of the two spherical segments, and this space is filled with high-viscosity oil (having a viscosity of approximately 50,000 centistokes).

It is evident that this design provides mechanical resistance in both the lateral and vertical planes. While the theoretical analysis of the mechanical system at the beginned of this paper treated only the conditions obtaining in the horizontal plane, it is obvious that a resonance in the vertical plane will also be observed. This may or may not occur at the same frequency as the lateral resonance, depending upon the ratio of vertical to lateral compliance and the effective mass at the stylus tip in the two planes. Usually the vertical resonance is near the lateral resonance so that, substantially, the same value of mechanical resistance is required for its damping. Having damping in the vertical plane also provides protection from damage due to accidental release of the arm. The rate of fall is held to a low value so that the shock of contact is small and no bouncing results. For a reproducer having a needle force of 5 grams, and a damping resistance of 8,950 mechanical ohms, the rate of descent will be

$$V = \frac{f}{r} = \frac{5 \times 10^3 \text{ dyne cm}}{9 \times 10^3 \text{ dyne sec}} = 0.55 \text{ cm/sec.}$$

A similar retarding effect is obtained in the lateral plane which increases the resistance to lateral shock. To put it another way, the tendency of a reproducer arm to function as a seismograph is arrested.

The mechanical resistance obtained by fluid films is proportional to the area of the film and approximately inversely proportional to the film thickness. It has been observed that with thicker films and higher viscosity liquids a significant amount of compliance appears along with the resistance. If this compliance were too large, it would serve to uncouple the resistance element from the system. Small values of compliance, on the other hand, are helpful in that they permit the arm to follow severely warped or eccentric records readily.

It is interesting that the amount of mechanical resistance used in this reproducer arm is not a critical value. The upper limit of resistance is reached when it interferes with tracking records having reasonably small values of eccentricity or warpage. With the usual values of suspension compliance and mass, this would occur at several times the amount of resistance necessary to critically damp the arm resonance. At the other extreme, where the resistance approaches zero, the arm merely reverts to a conventional one, in which there is very low pivot friction. If the design is such that the resonance of the arm and suspension compliance occurs below the desired transmission band, the variation in resonant response, resulting from change in the mechanical resistance, is of little importance. Between these extremes, a wide range of improved performance exists. Even a resistance in the order of one-sixth of the critical value will cut the Q of the resonant system significantly, with a corresponding improvement in stability as a result.

Fig. 5 shows the performance of an experimental arm similar to the one shown in Fig. 4. A light arm, having approximately 20-grams effective mass at the stylus tip, was used with a crystal cartridge whose compliance was approximately $.5 \times 10^{-6}$ cm/dyne. The light arm and stiff cartridge were chosen so that the resonant

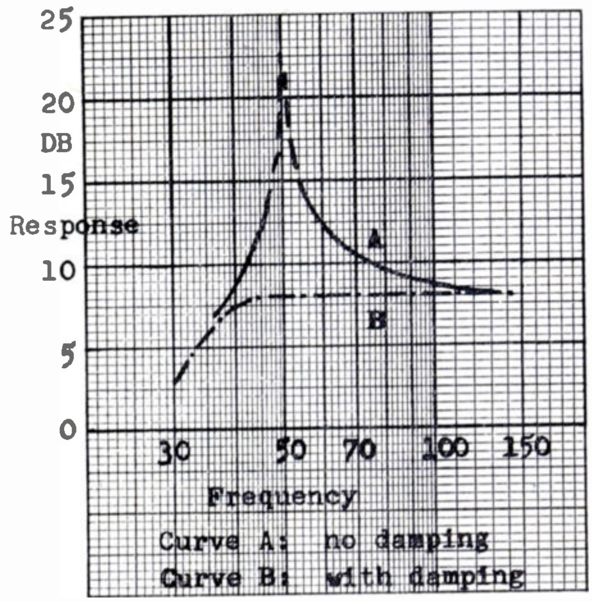

Fig. 5—Response of experimental arm with and without damping.

frequency would be high enough to avoid errors due to pointer vibration in the indicating instruments.

Without damping, the dynamic forces developed near resonance were so large that the stylus was forced out of the groove. For this reason it was not possible to measure the true resonant rise, and the curve therefore shows, with dashed lines, an estimated response in this region.

Fig. 6 is a photograph of a commercial design of a reproducer arm based upon the principle illustrated in

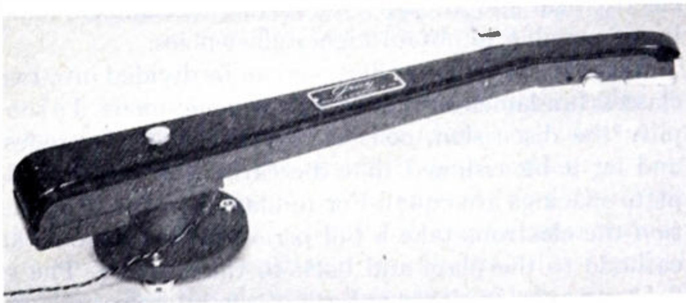

Fig. 6—Photograph of commercial design of a viscous damped reproducer arm.

Fig. 4. In this arm, the thumb screw over the pivot point permits adjustment of the film thickness to accommodate a wide range of temperature.

The application of linear mechanical resistance to the pivots of a phonograph reproducer arm provides effective means for damping the low-frequency arm resonance. When this resonance is brought under control, it is no longer necessary to have a large excess of needle force on the record over that required to follow the desired modulation of the groove.

Retarding-Field Oscillators*

J. J. EBERS†, ASSOCIATE, IRE

Summary—Pendulum and velocity-variation types of oscillations in retarding-field oscillators are discussed. The processes of bunching, drifting, and working of the electrons are explained, and a comparison is made to a reflex klystron. Recent results obtained from retarding-field oscillators are given.

Introduction

Positive-grid or retarding-field oscillators are among the oldest generators of radio-frequency power. Considerable confusion exists in the literature in regard to the energy conversion mechanism responsible for the oscillations. It is the purpose of this paper to clarify the electron mechanisms involved, as well as to investigate the possibility of the application of such mechanisms to wide-range, high-frequency generators in the 3,000 to 30,000 mc range.

Barkhausen-Kurz Oscillations

The original oscillations described by Barkhausen and Kurz[1] are dependent upon a pendulum-like oscillation of the electrons about the grid of a positive-grid triode. In order for this mechanism to function there must be a sorting of unfavorable electrons by the plate (or by the cathode in the event that the plate is operated quite negative with respect to the cathode). This type of oscillation mechanism is well described in the literature.[2] Barkhausen-Kurz oscillations have been observed to exist without a resonant circuit connected between the tube electrodes. The application of a tuned circuit results in higher interelectrode rf voltages which increase the energy transfer from the electrons to the electric field and intensify the sorting mechanism, making it possible to obtain higher efficiencies.

Barkhausen-Kurz oscillations can be divided into two classes: fundamental-mode, and harmonic-mode. To simplify the discussion, consider parallel-plane electrodes, and let it be assumed that the cathode-grid and grid-plate spacings are equal. For fundamental-mode operation the electrons take a full period to travel from the cathode to the plate and back to the cathode. The rf field can exist in either or both of the interelectrode regions; however, if it exists in both regions, then the two fields must be in phase. Under these conditions it is also possible to obtain odd-harmonic oscillations in which the transit time of the electrons from the cathode to the plate and back to the cathode is an odd number of periods of the rf voltage. If the rf field exists in only one interelectrode region, or if the two fields are of opposite phase, then it is possible to obtain even-harmonic oscillations.

Velocity Variation Oscillations

As soon as the rf voltage, built up in the interelectrode spaces, reaches a peak value on the order of one-tenth the dc voltage, a second phenomenon becomes appreciable. The electrons obtain large velocity variations which result in bunching and working of the electrons on the field, just as in klystrons and other velocity-variation tubes. Whereas the electron mechanism is quite simple, the mathematics describing it is not. For this reason, one of the simplest cases is analyzed here. Parallel plane geometry is assumed with a resonant circuit connected between grid and plate. This particular location of the tuned circuit was chosen because it is better adapted to high-frequency generators. The effects of space charge are neglected, and it is assumed that the electrons are intercepted by the grid after one transit of the grid-to-plate region. The assumptions of parallel planes and no electrons returning into the cathode-to-grid region can be met fairly well in practice by the application of electron optical principles. It has been observed[3] that a combination of space charge and the proper reflecting field can produce a ring-shaped refocus of the beam. The mathematical consideration of an oscillator in which the electrons are allowed to return into the cathode-to-grid region would be extremely difficult. Moreover, it is believed that such a tube would not be useful as a wide-tuning-range oscillator. Whereas the neglect of space charge would seem to be a very serious limitation on the usefulness of the theory, it must be remembered that the first-order effect is to shift the zero-potential plane in the direction of the grid. The field between the grid and the zero-potential plane is found to remain quite uniform. For small ac signals the debunching effect of space charge is not appreciable.

An approximate mathematical solution to this velocity-variation problem is given in the Appendix. Newtonian dynamics is applied to the electron motion to obtain the velocity and displacement as a function of time. For small ac signals an analytic expression for the conversion efficiency is obtained, based on the net gain or loss of energy of the electrons in traversing the grid-plate region. For the large ac signals a graphical

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† Electrical Engineering Department, The Ohio State University, Columbus, Ohio.

[1] H. Barkhausen and K. Kurz, "Die kurzesten mit vakuumröhren herstellboren wellen," *Phys. Zeits*, vol. 21, pp. 1–6; 1920.

[2] R. I. Sarbacher and W. A. Edson, "Hyper and Ultrahigh Frequency Engineering," John Wiley and Sons, pp. 521–549; 1943.

[3] O. Heil and J. J. Ebers, "*A new wide-range, high-frequency oscillator.*" Proc. I.R.E., vol. 38, pp. 645–650; June, 1950.

method is used to obtain the conversion efficiency. In order to obtain a better insight into the processes which compose the electronic mechanism, it is helpful to consider (12) of the Appendix. This equation gives the displacement of the electrons as a function of time and can be written as

$$\frac{\beta_0 \omega}{v_0} x = -(\omega t)^2 + \omega t(\beta_0 - 2K \cos \alpha) - 2K[\sin \alpha - \sin (\omega t + \alpha)] \quad (1)$$

where ω is the angular frequency of the rf voltage, β_0 is the dc transit angle of the electrons, K is the ratio of the peak rf to the total dc voltage between grid and plate, α is the phase angle of the rf field at the time the electron enters the grid-plate region, and v_0 is the initial velocity of the electrons corresponding to the grid voltage. This equation could be plotted directly to obtain a picture of the electron displacement curves for various values of α. However, the result would not give a clear picture because the curves would be parabolas with a sine wave superimposed. If the term $2K \sin (\omega t + \alpha)$ is transferred to the left side of the equation, there results

$$\frac{\beta_0 \omega}{v_0} x - 2K \sin (\omega t + \alpha) = -(\omega t)^2 + \omega t(\beta_0 - 2K \cos \alpha) - 2K \sin \alpha. \quad (2)$$

If the change of variable

$$X = \frac{\beta_0 \omega}{v_0} x - 2K \sin (\omega t + \alpha) \quad (3)$$

is made, (2) becomes

$$X = -(\omega t)^2 + \omega t(\beta_0 - 2K \cos \alpha) - 2K \sin \alpha. \quad (4)$$

The form of (4) indicates that in the new X and ωt co-ordinate system the displacement curves become true parabolas.

To plot the curves the X and ωt co-ordinate system is constructed according to (3). A parabolic template determined by the equation

$$X = -(\omega t)^2 + (\omega t)\beta_0$$

can then be constructed. The intersection of the $X=0$ axis, in the X and ωt co-ordinate system, and the parabola can be determined by a calibration of the parabola as a function of α according to (4). By this method curves are plotted in Figs. 1, 2, and 3 for $K=0.4$, 0.8, and 1.0, respectively.[4] Electrons are shown entering the grid-plate region at intervals of 22½ degrees. To simplify the picture, only returning electrons are shown during the last period. The dc transit angle β_0 is the same in all figures and is equal to 7.60 radians. It is shown in the Appendix that this dc transit angle is the optimum angle for first-mode operation.

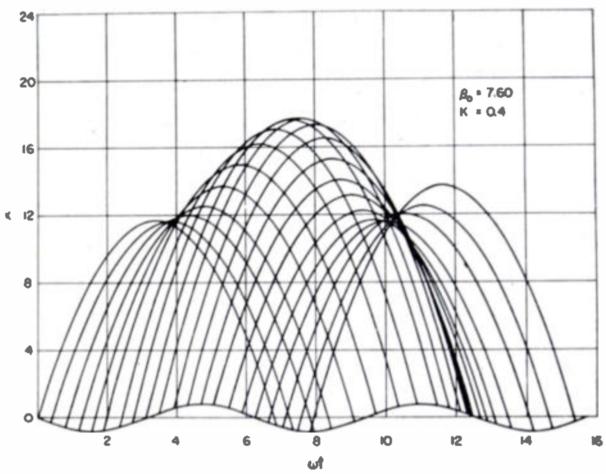

Fig. 1—Velocity modulation, bunching, and working in a retarding-field oscillator. $K=0.4$.

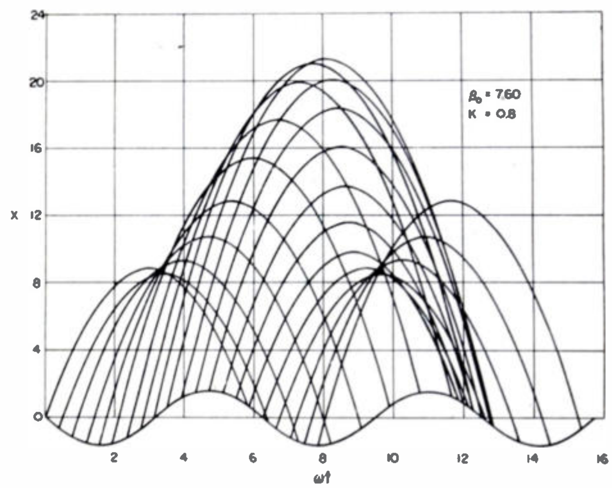

Fig. 2—Velocity modulation, bunching, and working in a retarding-field oscillator. $K=0.8$.

It is seen that upon entering the grid-plate region the electrons receive a velocity variation dependent upon the phase angle of the rf grid-to-plate voltage at their time of entry. Since the plate is negative with respect to the cathode, the electrons drift and become bunched in space. Under optimum conditions, the bunch is formed after the electrons have reversed direction, and the bunch returns to the grid at a time when the rf electric field opposes their motion so that the electrons deliver energy to the rf field. In Fig. 1 it is seen that for $K=0.4$ the bunch is poorly formed; many electrons return to the grid during a time when they continue to take energy from the rf field. In Fig. 2, for $K=0.8$, the bunch is well formed. By observing the phase of the sine wave which is used as a co-ordinate, it is seen that the bunch returns at an optimum time for delivering energy to the rf field. In Fig. 3 (see following page), for $K=1.0$, the electrons are over-bunched. A theoretical analysis (see Appendix) indicates a maximum efficiency

[4] This method of picturing electron motion was suggested by O. Heil and has been used by H. W. Heil in a German patent application: "Sekundarelektronemission," No. H 168 549 a/21 a4.

for $K=0.88$; thus there seems to be good correlation between bunching and efficiency.

For a reflex klystron in which the rf field is concentrated between two grids and in which the electrons drift and bunch in an rf field-free repeller region, it is found that the optimum drift angles in the repeller region are given by

$$\theta = 2\pi(n + 3/4), \; n \text{ an integer.} \tag{5}$$

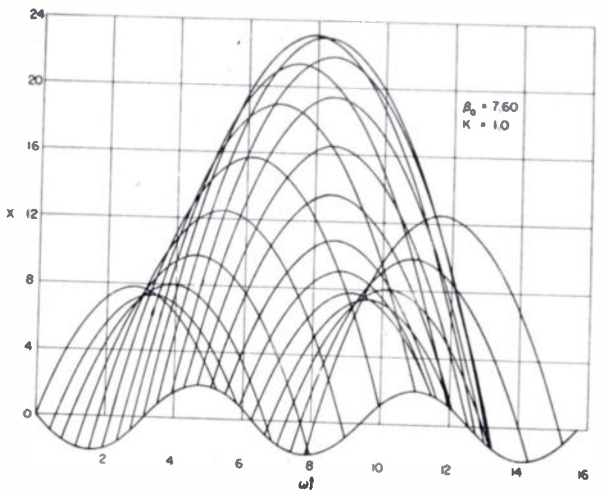

Fig. 3—Velocity modulation, bunching, and working in a retarding-field oscillator. $K=1.0$.

In the Appendix it is shown that the optimum drift angle for the first three modes of a parallel-plane retarding-field oscillator are 2.42π, 4.44π, and 6.45π. Thus the transit angles for optimum efficiency could be written approximately as

$$\beta_0 = 2\pi(n + 5/4). \tag{6}$$

Now if this transit time is divided into three parts as follows: (a) $\pi/2$ radians for velocity modulation, (b) $2\pi(n+3/4)$ radians for drifting and bunching, and (c) $\pi/2$ radians for working, an excellent correlation with reflex klystron behavior is obtained. A careful study of Fig. 2 indicates that the above division of the total transit time is indeed quite accurate. Since the energy transfer between the field and the electron is measured by $\int \bar{E}\cdot d\bar{l}$, most of the energy interchange occurs on the steep sections of the parabola (the modulating and working times) and very little energy interchange occurs at the top of the parabola (the drift or bunching time). In a reflex klystron, bunching occurs around an electron which crosses the gap when the rf field is zero and increasing in such a direction as to decelerate the electron. In the case of a plane retarding-field oscillator, bunching occurs around an electron which passes the grid when the rf field has its maximum accelerating value.

The small-signal theory, as developed in the Appendix, is of no value in determining the amplitude of the oscillations and the efficiency when the oscillator is working into a particular load. However, it provides a good criterion for oscillations to start. By equating the expressions for the power delivered by the beam and the power dissipated by the load, as derived in the Appendix, the following equations for the starting currents are obtained:

$$I_{0s} = 1.11 \times 10^{-12}\, \omega^2 d^2 G \text{ for mode 1,} \tag{7}$$

$$I_{0s} = 0.51 \times 10^{-12}\, \omega^2 d^2 G \text{ for mode 2,} \tag{8}$$

$$I_{0s} = 0.36 \times 10^{-12}\, \omega^2 d^2 G \text{ for mode 3.} \tag{9}$$

As a typical example, assume

$$f = 6{,}000 \times 10^6 \text{ cps,}$$

$$d = 10^{-4} \text{ meters.}$$

$$G = 2.5 \times 10^{-5} \text{ mho,}$$

then

$$I_{0s} = 23.6 \text{ ma for mode 1,}$$

$$I_{0s} = 10.9 \text{ ma for mode 2,}$$

$$I_{0s} = 7.7 \text{ ma for mode 3.}$$

Such starting currents are quite easily obtained.

From the description of the tubes and experimental results described in the literature, it is difficult to determine whether the Barkhausen-Kurz mechanism or the velocity-variation mechanism was responsible for the oscillations observed.[5] Because the two mechanisms can occur in the same tube under identical conditions, a tabulation of the optimum transit angles in the grid-plate region for different modes of operation furnishes a criterion for determining which mechanism is responsible for the oscillations, as in Table I. Since some of these transit angles differ by relatively small amounts, it seems probable that many of the Barkhausen-Kurz oscillations reported in literature were actually of the velocity-variation type, particularly in cases where optimum output was obtained with the plate quite negative with respect to the cathode.

TABLE I

Type of Oscillation	Transit Angle
Barkhausen-Kurz	π
Barkhausen-Kurz	2π
Velocity Variation	2.42π
Barkhausen-Kurz	3π
Barkhausen-Kurz	4π
Velocity Variation	4.44π
Barkhausen-Kurz	5π

Unfortunately, the sorting mechanism which is essential for Barkhausen-Kurz oscillations also can occur in velocity-variation oscillations. If the electrons which gain energy in crossing from grid to plate (see Fig. 2) are collected by the plate, the efficiency of the velocity-variation mechanism is definitely decreased. This results from the fact that these electrons are capable of delivering energy to the rf field. This would not be true if the dc transit angle were adjusted to 2π radians (for

[5] E. C. S. Megaw, "Electronic oscillations," *Jour. IEE*, (London), vol. 72, pp. 313–25; 1933.

Barkhausen-Kurz oscillations) instead of 2.42π radians. The entire effect of electrons striking the plate is complicated by secondary electrons. If the electrons strike the plate with more than 20 volts of energy, secondaries will be produced. With large rf voltages, it has been observed that the secondary emission ratio often is much larger than one. The effect of these secondaries on the oscillation mechanism would be difficult to determine analytically.

Experimental Results

In a recent paper[6] a new tube is described which is essentially a retarding field oscillator adapted to a resonant cavity. To see how this tube evolves from a reflex klystron, consider Fig. 4. At the top is shown a reflex klystron which is capable of generating high-frequency power with an efficiency of approximately 1.5 per cent. The electrons are velocity modulated in the concentrated rf field between the anode and grid; they then drift and bunch in the repeller region and return to do work in the rf field. In the center is shown a tube such as described by Heil and Ebers. The electrons are velocity modulated by the concentrated rf field near the nozzle. They then drift and bunch in a relatively weak rf field, and return to do work on the strong rf field.

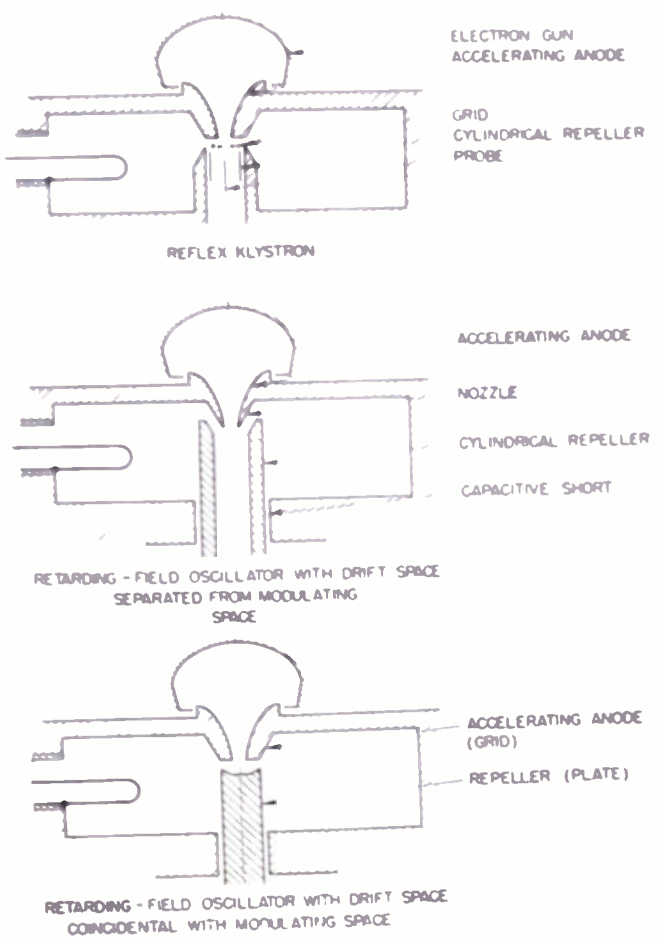

Fig. 4—Retarding-field oscillators.

Calculations of electron transit times for the oscillator shown in the center of Fig. 4 for the highest frequency oscillations show that the electrons only enter the repelling field a distance comparable to half the diameter of the anode aperture. In this region the ac and dc fields are quite uniform. Moreover, the transit time corresponds to approximately 1.1 periods, as compared to $\frac{3}{4}$ period for optimum efficiency of a reflex klystron. For these reasons it was decided to undertake the mathematical development given in the Appendix and to build a tube such as shown at the bottom of Fig. 4. In this case the fields are concentrated and quite uniform between the grid and repeller. The potential distribution for this electrode configuration is plotted in Fig. 5. Actually, the chronological order of the development of these tubes is correct, but it could be argued that the reverse order would have been more logical.

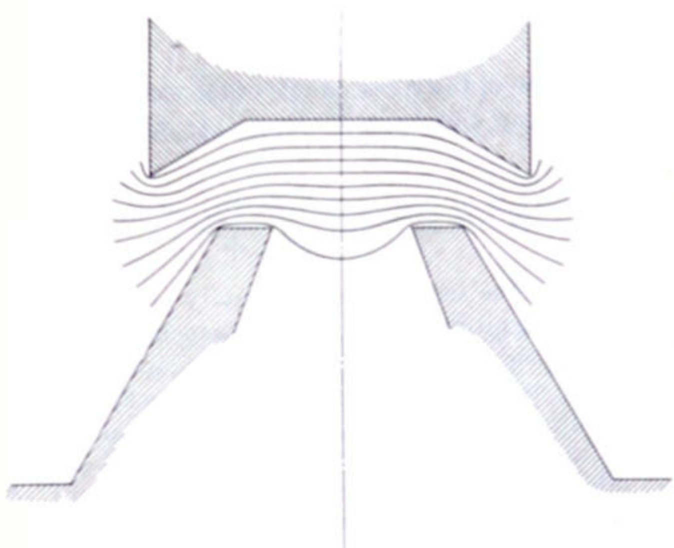

Fig. 5—Potential distribution in a planar, retarding-field oscillator.

A drawing of a tube designed to have a uniform repeller field is shown in Fig. 6. In order to obtain a check on the theory of retarding-field velocity-variation oscillations presented here, a design was chosen which would fit the assumptions of the theory as closely as possible. The field between the grid (hereafter called the anode) and negative plate (hereafter called the repeller) is quite uniform, as shown in Fig. 5. Very few electrons make more than one transit of the interaction gap, since the beam spreads on entering the cavity, and upon its return strikes the flat portion of the anode.

The actual tube was built on a demountable pump station. The accelerating anode was made in two parts so that the beam current could be varied with the cavity voltage held constant. The gun used in this model was developed by O. Heil and has a perveance of 4.33×10^{-6}.

[6] O. Heil and J. J. Ebers, *ibid.*

The power output was obtained by inserting a coupling loop into the cavity through a quartz tubing. A choke was provided to prevent loss of high-frequency energy on the repeller lead. The distance x of Fig. 6 was varied in different models.

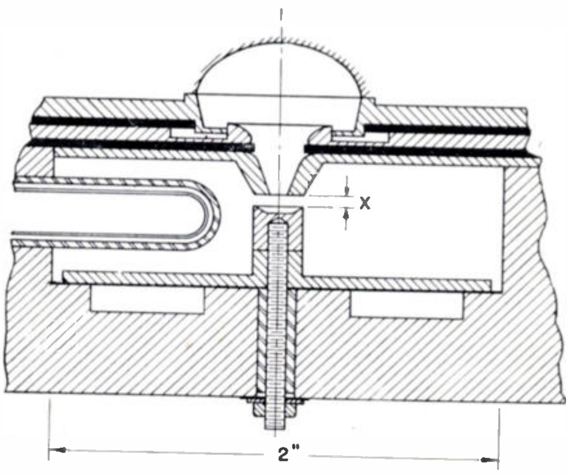

Fig. 6—Drawing of a planar, retarding-field oscillator.

Calculation of the transit angle of the electrons in the interaction gap may be made from (14), which may be written as

$$\beta_0 = 1.27 \times 10^4 \frac{d\sqrt{V_{A_2}}}{\lambda(V_{A_2} + V_R)}, \tag{14(a)}$$

where V_{A_2} is the second anode or cavity potential, and V_R is the magnitude of the negative repeller potential. Since the field extends into the anode aperture (see Fig. 5) the value of d is not exactly equal to the dimension x shown in Fig. 6. As a correction, the distance from the flat anode surface to the 0.9 equipotential surface was added to d in all cases (about 0.012 inch). A tabulation of β_0 for various values of x, and the corresponding measured resonant wavelength of the cavity are given in Table II. The theoretical optimum transit angles for modes 1 and 2 are 7.60 and 13.8 radians, respectively; however, it is recalled that space charge pushes the zero potential plane toward the grid. Thus it would be expected that transit angles, as calculated above, could be larger than the theoretical values. It has been observed that as the beam current is decreased, the calculated angles approach the theoretical values.

TABLE II

x	λ	β_0	
		Mode #1	*Mode #2*
0.0395"	10.4 cm	8.75 rad.	
0.063"	9.15 cm	8.55 rad.	15.0 rad.
0.083"	8.55 cm	8.90 rad.	14.9 rad.
0.073"	8.75 cm	8.25 rad.	15.2 rad.
0.070" (grids)	8.90 cm	7.80 rad.	

A curve of power output and efficiency as a function of anode voltage is shown in Fig. 7. Since the efficiency increases at a constant rate, it is probably true that space-charge debunching is not a limiting factor in efficiency. A curve of power output and efficiency as a function of beam current is shown in Fig. 8.

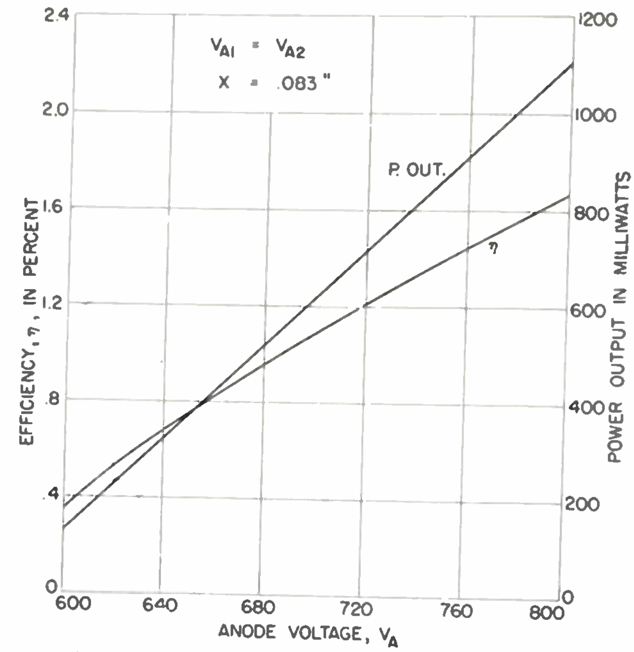

Fig. 7—Variation of power output and efficiency with anode voltage.

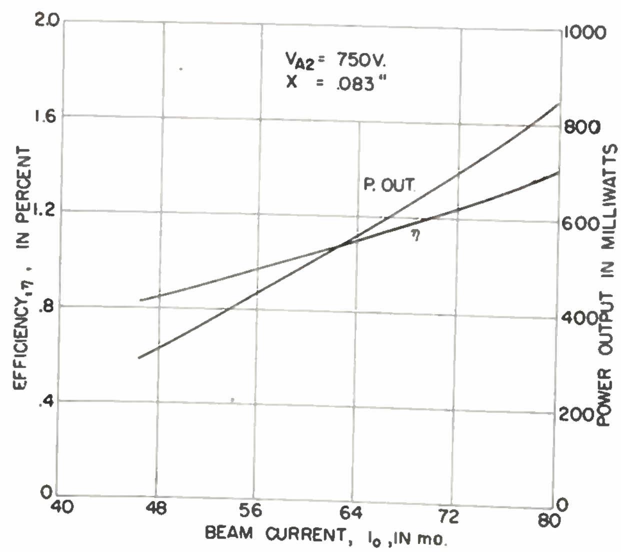

Fig. 8—Variation of power output and efficiency with beam current.

From additional data obtained on this tube, it was found that approximately twice as much energy was dissipated in the cavity as in the load. The largest portion of these losses occur in the repeller by-pass circuit. Recently, an improved-design tube was operated at

6,000 mc with an over-all efficiency of 4.5 per cent and a conversion efficiency of approximately 6.0 per cent.[7]

Conclusions

Of the two fundamentally different methods of generation of high-frequency energy by the use of a positive-grid or retarding-field oscillator, the velocity-variation type is the more adaptable to wide-range, centimeter-wave oscillations. The difference between the power produced in such oscillators and the theoretical values offers promise of greater efficiencies. The upper limit on the efficiency obtainable in practice is not known, but factors such as space-charge debunching and nonuniform transit times will probably make the theoretical value unobtainable. Work can be done toward determining the best electron optics and the best cavities for such an oscillator. To counteract the effects of space charge, it is possibly true that higher voltages and lower currents should be used. This would also have the advantage of increasing the cavity gap length and hence the cavity shunt impedance.

The principal advantage of these oscillators over reflex klystrons lies in their simplicity. From data which has been obtained to date, the electronic tuning range seems to be comparable to reflex klystrons; about one per cent. The cone-shaped anode (Fig. 4) probably lends itself better to wide-range tuning than one with a plane anode and repeller, since it concentrates the fields in the vicinity of the anode and allows the electrons room in which to drift without striking the repeller. If only a limited tuning range is desired, it is possible that a plane electrode system would give best results.

Acknowledgments

The author thanks his associates in the electron tube laboratory who have made suggestions concerning the theory, and who have aided in the construction of the tubes discussed in this paper. The author wishes to express his appreciation to E. M. Boone for his valuable criticism and suggestions in the capacity of advisor for doctoral research.

Appendix

Mathematical Analysis of Velocity-Variation Effects in the Grid-Plate Region of a Retarding-Field Oscillators[8]

Consider the parallel-plane electrode arrangement and electrical connections shown in Fig. 9. Electrons leave the cathode and are accelerated to a velocity v_0 corresponding to the grid voltage V_c. The plate is operated at a potential which is negative with respect to the cathode; thus the electrons are reflected in the grid-plate region. Upon their return to the grid it will be assumed that the electrons are collected. For simplicity it will be assumed that the electrodes are infinite, parallel planes. The effects of space charge will be neglected. The object of this analysis is to obtain an expression for the transfer of energy between the electron stream and the rf field in the grid-plate region.

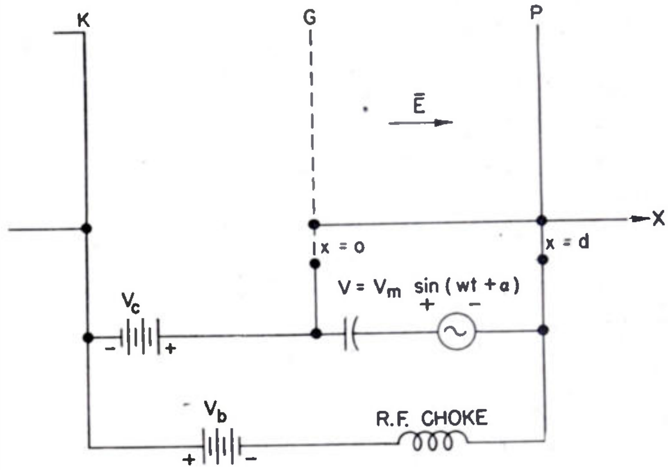

Fig. 9—Circuit for analyzing velocity-variation oscillations.

If the magnitude of the dc grid-to-plate voltage is defined as

$$V_1 = V_b + V_c, \tag{10}$$

and if the force on an electron is equated to the rate of change of momentum, an expression is obtained which can be integrated to give the velocity and displacement of an electron, which enters the grid-plate region when the phase angle of the rf field is α, as a function of time. The velocity is

$$v = v_0 - \frac{e}{md\omega}\left[V_1\omega t - V_m\{\cos(\omega t + \alpha) - \cos\alpha\}\right] \tag{11}$$

and the displacement measured from the grid plane is

$$x = v_0 t - \frac{e}{md\omega^2}\left[\frac{V_1(\omega t)^2}{2} - V_m\{\sin(\omega t + \alpha) - \omega t\cos\alpha - \sin\alpha\}\right]. \tag{12}$$

If τ is the total electron time of transit of the grid-plate region, then the transit angle will be given by

$$\beta = \omega\tau. \tag{13}$$

The transit angle in the absence of an rf field is defined as β_0;

$$\beta_0 = \frac{2mdv_0\omega}{eV_1}. \tag{14}$$

If the further definition is made that

[7] These results will be described in a forthcoming article by J. Moll and R. Wilmarth of the Ohio State University Tube Laboratory.

[8] An analysis of this problem has been done by W. Kleinsteuber, "Die bremsfeldanfachung bei grossen wechselspannungen," *Hochfreq. und Elecktroak.*, vol. 57, pp. 1–10; 1941, by a different method with similar results. The processes of velocity modulation, bunching, and working were not apparent nor was the reason why the different modes exist.

$$K = \frac{V_m}{V_1}, \tag{15}$$

then (11) and (12) become the following dimensionless equations:

$$\frac{v_r}{v_0} = 1 - \frac{2}{\beta_0}\left[\beta + K\{\cos\alpha - \cos(\alpha+\beta)\}\right] \tag{16}$$

$$0 = -\beta^2 + \beta(\beta_0 - 2K\cos\alpha) - 2K[\sin\alpha - \sin(\alpha+\beta)], \tag{17}$$

for electrons at $x = 0$.

Small-Signal Theory

Equation (17) is a transcendental equation in β, and if an analytical solution is to be obtained, the terms $\sin\beta$ and $\cos\beta$ must be expanded in power series. However β is large, in general, and an expansion about β_0 is used. If this is done and terms of the zero, first, and second order only are retained, a solution is obtained which, when substituted into (16), gives

$$\frac{v_r}{v_0} = -\left[1 + \frac{2K}{\beta_0^2}\{2\sin(\alpha+\beta_0) - 2\sin\alpha - \beta_0\cos(\alpha+\beta_0) - \beta_0\cos\alpha\} + \frac{2K^2}{\beta_0^2}\{1 + \cos^2\alpha - 2\cos\beta_0 - 2\beta_0\cos\alpha\sin(\alpha+\beta_0) + \sin^2(\alpha+\beta_0)\}\right]. \tag{18}$$

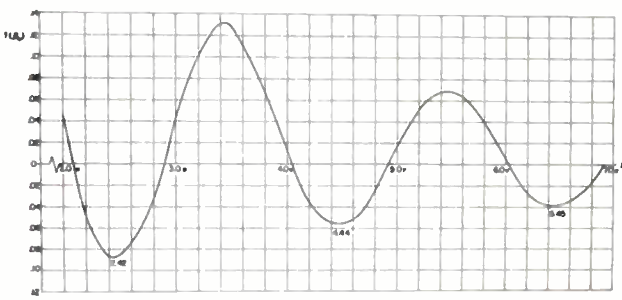

Fig. 10—Variation of $f(\beta_0)$ with β_0.

The conversion efficiency of the mechanism (transfer of electron beam energy to rf field energy) is given by the simple equation,

$$\eta_{conv} = 1 - \frac{1}{2\pi}\int_0^{2\pi}\frac{v_r^2}{v_0^2}\,d\alpha. \tag{19}$$

Because of the integral properties of sinusoidal functions, the square of (18) is easily integrated. The resulting efficiency is given by

$$\eta_{conv} = -4K^2\left[\frac{4 + 3\beta_0^2 - (4+\beta_0^2)(\cos\beta_0 + \beta_0\sin\beta_0)}{\beta_0^4}\right] \tag{20}$$

which can be written as

$$\eta_{conv} = -4K^2 f(\beta_0). \tag{21}$$

The function $f(\beta_0)$ is plotted in Fig. 10. The negative values of this function are of interest since they correspond to positive values of efficiency as given by (21), and hence they represent a net transfer of energy from the electron stream to the rf field. It is observed that negative maxima occur for β_0 equal to 7.60, 13.80, and 20.22 radians. These regions will hereafter be referred to as modes 1, 2, and 3, respectively.

If the dc voltages have been adjusted to give the optimum drift angles mentioned in the foregoing paragraph, the efficiency can be plotted as a function of K, the ratio of ac and dc voltage, for the three modes. These are the parabolic curves in Fig. 11. Now let it

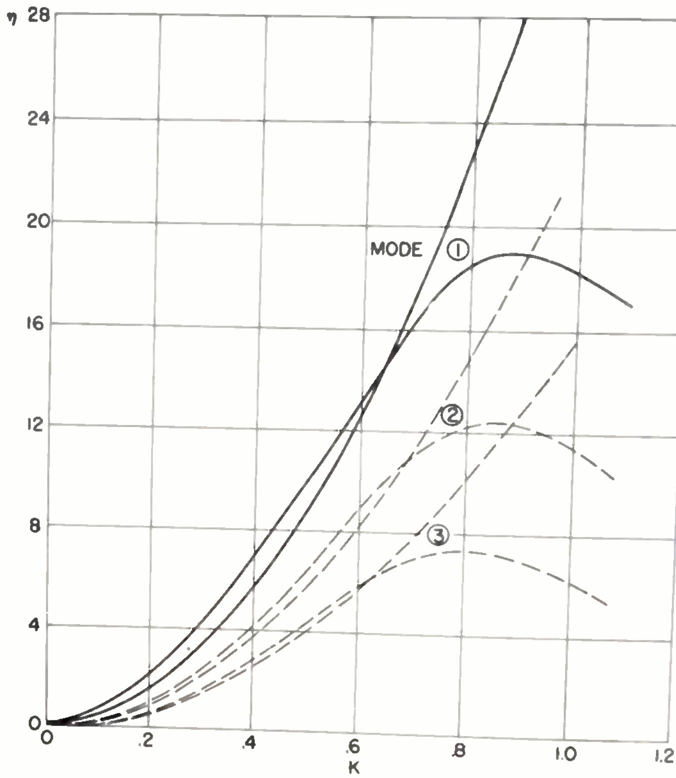

Fig. 11—Small-signal and large-signal efficiency as a function of K

be assumed that the mechanism described above is operating as an oscillator. A parallel resonant circuit with shunt conductance G is connected between grid and plate. A shunt conductance G is assumed to dissipate all the conversion energy. If the circuit is oscillating it must be true that the power delivered by the beam is equal to the power dissipated in the conductance G. Thus if P_{del} is the power delivered and P_{dis} is the power dissipated, then

$$P_{del} = P_{input}\eta_{conv} = I_0V_c\eta_{conv} = -4I_0V_cK^2f(\beta_0)$$

or

$$P_{del} = \frac{4I_0V_cf(\beta_0)}{V_1^2}V_m^2 \tag{22}$$

and

$$P_{\text{dis}} = \tfrac{1}{2} G V_m^2. \tag{23}$$

The power-delivered and the power-dissipated curves are both proportional to the square of the rf voltage, and hence will not intersect. This implies that oscillations would build up in intensity indefinitely provided they would start. Thus it is obvious that the efficiency curves obtained are inconsistent with experiment and whereas they may be quite accurate for small signal, they, by necessity, must exhibit a saturation effect for the condition of large ac signals.

Large-Signal Theory

In order to obtain the saturation effect mentioned above, it is necessary to obtain more accurate solutions of (17). A sufficiently accurate analytical solution could be obtained, but a solution by means of a series of approximations is probably easier and permits an accuracy only dependent on the care in making the computations.

Equation (17) can be written as

$$\beta = (\beta_0 - 2K \cos \alpha) - 2K \left[\frac{\sin \alpha - \sin (\alpha + \beta)}{\beta} \right]. \tag{24}$$

As a first approximation, β can be taken equal to $(\beta_0 - 2K \cos \alpha)$ and this value of β substituted in the right-hand side of (24). This gives a new value of β which can again be substituted in (24). If this procedure is repeated two or three times, a sufficiently accurate value of β will be obtained. A plot of the variation of β with K for different values of α is shown in Fig. 12. The dc transit angle is taken to be 7.60 radians which corresponds to the optimum angle for first-mode operation as determined by the small-signal theory. The dotted lines indicate the first approximation. The values of β obtained can then be substituted in (16). The ratio v_r^2/v_0^2 can be plotted as a function of α and the efficiency obtained by graphical integration. These results are also shown in Fig. 11, where efficiency is plotted as a function of K. Also plotted is a possible (not calculated) variation of efficiency for the other modes.

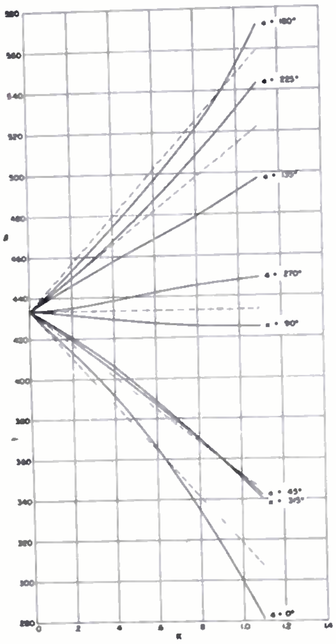

Fig. 12—Variation of transit angle with K for various α.

The question naturally arises as to whether the value of β_0 chosen for the first mode actually gives optimum efficiency for large signals. The efficiency was calculated for values of β_0 on either of 7.60 radians, and in both cases a decrease was obtained.

Correction

David Atlas, co-author with Ludwig Katz of the correspondence, "Optimum Vertical Resolution in Microwave Probing of the Atmosphere," which appeared on page 1341 of the October issue of the Proceedings of the I.R.E., has brought the following error to the attention of the editors:

Equations (1) and (2) should read as follows:

$$t = p \sin \theta + h\phi / \tan \theta. \tag{1}$$

and

$$\cos^3 \theta - \cos \theta + \frac{\phi h}{p} = 0. \tag{2}$$

Improvements in Image Iconoscopes by Pulsed Biasing the Storage Surface*

R. THEILE† AND F. H. TOWNSEND‡, SENIOR MEMBER, IRE

***Summary*—In storage-type television camera tubes using high-velocity electrons for scanning, such as the iconoscope and image iconoscope, the storage surface is stabilized to an equilibrium potential by secondary emission. A number of undesirable characteristics, such as spurious signals, absence of "black-level" information, and relatively low efficiency, are usually associated with tubes of this kind. However, these disadvantages are considerably reduced if the mean potential of the storage surface is shifted negatively. A method is investigated for obtaining such required potential shift by periodically irradiating the storage surface with high-velocity electrons while simultaneously reducing the collector potential, these periodic processes being carried out during suitable intervals in picture transmission, such as the frame-blanking period.**

Application of this principle has been most successful in cases where the picture is projected intermittently as in the "memory-scanning" method of film transmission, but for continuous pickup the advantages are only partly attainable.

Introduction

FOR STABLE OPERATION of storage-type camera tubes, the potential of the storage surface must be restored to an equilibrium value at each scanning. There are two stable potentials which the insulated storage-plate surface can assume under bombardment, depending on whether secondary electrons are emitted in excess of unity ratio or not. If the electrons are of relatively high velocity, e.g., of the order of 1,000 volts, the surface, because of secondary emission, assumes a potential close to that of the collecting electrode which is usually common with the tube anode. But if the electron velocity at the storage plate is close to zero, there are substantially no secondary electrons, and the surface accumulates charge until its potential approximately equals that of the cathode of the electron source.

In camera tubes of the iconoscope and image-iconoscope type[1,2] the surface-potential stabilization is obtained by the first method, that is, by secondary emission. This "anode-potential stabilization" has the great advantage of being completely stable at all light levels, but tubes operated in this manner show undesirable effects because of "secondary electron redistribution." In such tubes the storage surface is insulated, and an equilibrium must be maintained between the number of electrons arriving at and leaving this surface. This in turn requires that, on the average, all of the emitted secondary electrons in excess of one for each arriving primary electron must return to the surface, thus giving rise to "redistribution" effects. These effects become apparent as spurious signals and cause other defects in the pictures.

The second method, known as "cathode-potential stabilization," is employed, with considerable success, in the Orthicon and image-Orthicon camera tubes used in modern television.[3-6] The scanning beam does not liberate free electrons for redistribution, and hence there are no spurious signals from this source. Also, higher storage efficiency may be obtained, as well as preservation in some tubes of true black level in the signal. On the other hand, low-velocity scanned tubes are not without limitations.[7]

Consideration of the signal generation processes show that some of the major advantages of low-velocity scanning can be obtained in high-velocity scanned tubes by shifting the mean potential of the storage surface in a negative direction. There have been a number of proposals for obtaining this desired effect: by use of semiconducting storage plates,[8] by continuous or pulsed diffuse irradiation of the storage surface with low-velocity electrons,[9] by scanning with very high-velocity electrons,[10,11] and so on. However, apart from the use of bias-light and rim-light techniques in iconoscopes for film transmission[12-14] which obtain the required results

* Decimal classification: R583.6. Original manuscript received by the Institute, September 4, 1950; revised manuscript received July 11, 1951.

† Pye Ltd., Cambridge, England.
‡ Cathodeon Ltd., Cambridge, England.

[1] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, Inc., New York, N. Y., Chapts. 10 and 11; 1940.

[2] J. D. McGee, "Electronics," The Pilot Press, London, England. Edited by B. Lovell, Chapt. 4; 1947.

[3] A. Rose and H. Iams, "The Orthicon, a television pickup tube," *RCA Rev.*, No. 4, pp. 186–199, October, 1939 "Television pickup tubes using low-velocity electron-beam scanning," Proc. I.R.E., vol 27, pp. 547–555; September, 1939.

[4] A. Rose, P. K. Weimer, and H. B. Law, "The image Orthicon, a sensitive television pickup tube," Proc. I.R.E., vol. 34, pp. 424–432, July, 1946.

[5] J. D. McGee, "A review of some television pickup tubes," *Jour. IEE* (London), vol. 97, pt. III, pp 377–392; November, 1950. "Distant electric vision," Proc. I.R.E., vol 38, pp. 596–608; June, 1950.

[6] R. B. Janes, R. E. Johnson, and R. S. Moore, "Development and performance of television-camera tubes," *RCA Rev.*, vol. 10, no. 2, pp. 191–223; June, 1950.

[7] L. H. Bedford, "Television-camera tubes," *Wireless Eng.*, vol. 28, pp. 4–16; January, 1951.

[8] H. Salow, "Speichernde Bildfanger mit halbleitendem Dielektrikum," *Fernsehen und Tonfilm*, pp. 1–4; January, 1939.

[9] A. W. Vance and H. Branson, U. S. Patent 2,147,760, issued February 21, 1939. (British Patent 434,942.)

[10] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, Inc., New York, N. Y., p. 306; 1940.

[11] J. D. McGee and H. G. Lubszynski, "E. M. I.—cathode-ray television transmission tubes," *Jour. IEE* (London), vol. 84, p. 131; 1939.

[12] W. S. Percival, C. O. Browne, L. R. J. Johnson, and F. Blythen, British Patent 490,845, accepted August 22, 1938.

[13] O. H. Schade, U. S. Patent 2,368,884, issued February 6, 1945. (British Patent 569,436.)

[14] S. Helt, "Practical Television Engineering," Murray Hill Books, Inc., New York, N. Y., pp. 188–190, 1950.

to a partial degree, it is not known to the authors that any of the proposed methods have been successfully used in a practical television service.

The paper deals with investigations into a method of obtaining the required shift of the equilibrium potential of the storage surface by periodically charging the storage capacitance during the frame-blanking intervals,[15] a system which has made possible considerable improvements in film transmission using storage-type camera tubes.[16]

Operation of Image Iconoscope-Type Tubes under Normal Conditions

The investigations described have been carried out with tubes known commercially as the "Photicon,"[17] which is of the image-iconoscope type, but the proposals are applicable to some other camera tubes employing the storage principle.

Fig. 1 shows a schematic diagram of a tube of this type. As the principle of operation is well known, it will not be discussed in detail. However, it is necessary to draw attention to some of the peculiarities of operation in order to simplify the understanding of the material which follows.

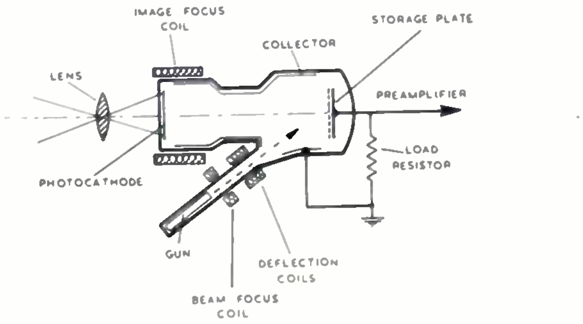

Fig. 1—Schematic diagram of an image iconoscope-type tube.

Up to the stage involving the storage surface, the translation of the picture signal is satisfactory and is at maximum efficiency as the photo current is always saturated. However, at the storage plate, where the picture charge pattern is formed, the signal translation is less satisfactory owing to the unfavorable potential conditions at the scanned surface, which are inherent in the method of surface-potential stabilization by secondary emission.

This will be explained in greater detail in conjunction with Fig. 2, which shows qualitatively the effects of secondary emission at an electron-bombarded target. Fig. 2(a) is a schematic diagram of the experimental setup used in obtaining the information displayed in Fig. 2(b). This setup consists of a vacuum tube having means of generating an electron beam which is accelerated towards a conductive target plate, with external means of varying the potential between the target plate and the collector electrode.

Fig. 2(b) shows the target current i_p (difference between primary and secondary currents) as a function of the target/collector potential difference e. This characteristic is mainly determined by the initial velocity distribution of the emitted secondary electrons. The number of secondary electrons collected decreases as the retarding field is increased until eventually all the secondaries are returned and the target current then equals the incident primary current. On the other hand, if the target potential is sufficiently negative to that of

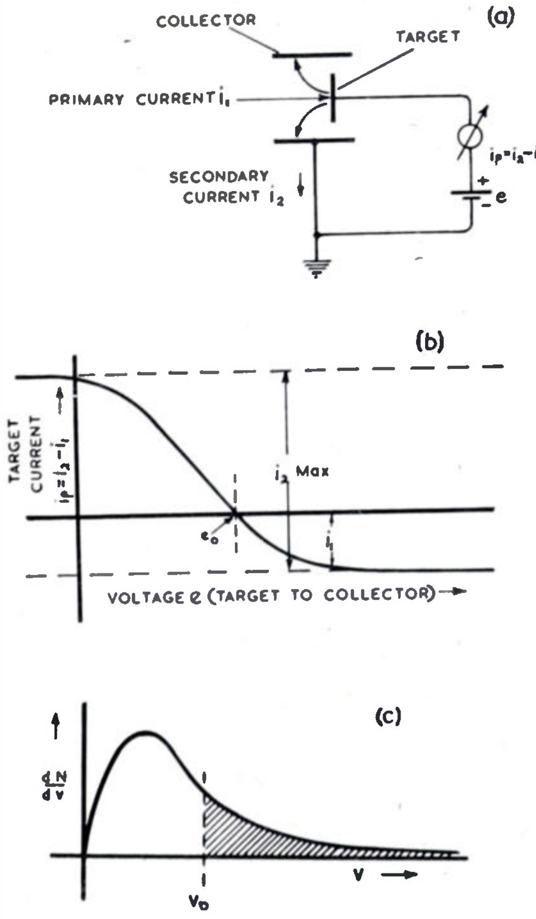

Fig. 2—Qualitative effects of secondary emission: a. Test setup. b. Signal-plate current characteristic. c. Velocity distribution.

the collector, all the emitted secondary electrons are collected and the target current is equal to the excess of secondary current over primary current. Even if no potential difference exists between the target and the collector, the emission velocities are such that complete

[15] Subsequent to the writing of this paper, the authors became aware of the patents: A. V. Bedford, R. D. Kell, U. S. Patent 2,108,097 issued February 15th, 1938, and T. L. Delvaux (Compagnie Francaise Thomson Houston), French Patent 901.393, issued July 25th, 1945, in which the application of such periodic pulsed charging is mentioned for the operation of normal iconoscopes, but obviously these proposals have not been developed to any known practical technique.

[16] W. R. Cheetham, N. Q. Lawrence, and R. Theile, "The design of 16-mm television film-chain employing pulsed Photicon pickup technique," *Jour. Soc. Mot. Pic. & Telev. Eng.*; to be published.

[17] Trade mark of Cathodeon Ltd., Cambridge, England. Registered in the United States and Great Britain.

collection of the secondary electrons can take place, provided that no appreciable modification of the space potential is caused by the electron charge itself, a condition which is closely approached during normal camera-tube operation. The range between no collection and full collection of the secondary electrons is determined by the distribution of the emission velocities, which usually have a greatest probability value of about 2 volts, the exact value depending on the properties of the bombarded surface, and so forth.

Of particular importance in the operation of camera tubes is the point e_0 in Fig. 2(b). Here the average number of secondary electrons collected equals that of the arriving primary electrons, that is $i_p=0$. At this point it is possible to disconnect the target lead without causing any change in the operating conditions. Alternatively, if by reason of the operating conditions the target current is essentially zero, for example, when the target consists of an insulated layer, then e_0 will automatically be established as the operating equilibrium value. The potential e_0 is determined by the velocity distribution shown in Fig. 2(c). In this curve, the number of electrons dN per velocity interval dv, (between v and $v+dv$) is plotted against v. The existence of the equilibrium potential e_0 means that only electrons having velocities greater than v_0 can surmount the retarding field, v_0 being defined as that value at which all emitted electrons with velocities from v_0 to infinity have an integrated number equal to that of the incident primary electrons.

Applying these results to image-iconoscope camera tubes in which a concentrated electron beam periodically scans an insulated storage surface, we find that the area immediately under the scanning beam assumes the positive equilibrium potential, and simultaneously the excess secondary electrons are redistributed over neighboring parts of the surface. This low-velocity electron redistribution forms a negative charge layer, which is essential and necessary for the picture charge-pattern development. There is a continuously proceeding interchange of electrons in such manner between the area being scanned, other areas of the storage plate and, of course, the collecting electrode, that the mean storage-plate current is zero over a complete cycle of the scanning process, provided the picture content is constant. The departures of this current from the mean value in positive and negative direction represent the signal modulation. Since, therefore, the black signal is not constant in relation to the interline (zero beam) level which is available as a reference during line-blanking time, it is not possible to use this reference for black-level restoration.

This is one of the serious defects of high-velocity scanned camera tubes, which makes it necessary for the dc level of the outgoing signal to be continually adjusted manually. It is extremely difficult, if not impossible, to carry out this procedure satisfactorily when frequent and violent changes of light level occur, as often happens in the transmission of cinematograph films.

Other major defects of high-velocity electron-scanned tubes are spurious signals and edge flare, which are due to nonuniformity of the redistribution process. It follows from the description of the surface-potential stabilization by secondary emission, that the area immediately behind the scanning beam is left as the most positive part of the storage surface. Consequently, there is a tendency for the free secondary electrons to migrate in the direction opposite that of the motion of the scanning beam, thus accumulating most negative charge at the areas where the scanning process starts, and least charge at those parts which are scanned at the end of each cycle. As a result of this nonuniform charge distribution, spurious signals are generated in the scanning process, and the modulation ratio due to the picture content is not constant over the picture area. These spurious signals ("tilt and bend" or "dark spot") appear in the transmitted picture as shading of the background level, the shading being most intense at the top left corner of the received picture and decreasing towards the bottom and right-hand edges. This shading can be compensated, but the compensation requires adjustment with changes in picture content.

Edge flare is also caused by the sense of the redistribution, and appears along the edges of the picture where the scanning process ends, particularly at the bottom edge in a conventional television system. As the scanning beam switches immediately to the top after having scanned the bottom part, there is no region of scan-induced secondary electrons following this area in close proximity. As a result of this, the last lines at the bottom of the picture receive very little charge by redistribution. Consequently, the surface potential of this area is less negative than that of the greater part of the storage surface. This results in the signal from this area appearing to be relatively white and having a reduced depth of modulation. The effect is most marked when the lowest part of the picture contains areas of black picture content which meet the bottom edge. But, if the picture area adjacent to this bottom edge is largely white, then there is redistribution of secondary electrons induced by the photoelectrons bombarding the neighborhood, and the effect is much reduced under these conditions.

Another unfavorable consequence of surface-potential stabilization by secondary emission is that the range of possible voltage change at the surface is limited between the positive equilibrium potential and the maximum negative potential due to accumulated redistribution electrons. This leads to low efficiency of storage and to a nonlinear transfer characteristic which has decreasing slope with increasing light levels. Consequently, the maximum possible signal to noise ratio is limited, but these properties of nonlinear transfer characteristic and incomplete storage efficiency are in some respects advantageous. They result in good half-tone rendition, "sharpness" of definition is retained in pictures including fast-moving subjects, and external gamma correction unit is not required.

Principle of Improvement by Pulsed Biasing

It is clear from the above considerations that improvements in the characteristics of high-velocity scanned tubes could be expected from the reduction or elimination of secondary electron redistribution. This means that collection of the secondary electrons must be improved, which necessitates a relative shift of the storage-surface potential in a negative direction (see Fig. 2, page 147.)

It is known that temporarily improved pictures may be obtained from an iconoscope or image iconoscope by suddenly increasing the collector potential in a positive direction. Within a short time, however, the storage surface loses its charge because of the increased collection of secondary electrons, the surface potential follows the changed (more positive) collector potential, and new equilibrium conditions are established, which result in the operating conditions reverting to those which existed before the application of the positive potential. In order to maintain the required potential difference between the storage surface and the collector, it is obviously necessary to replace, continuously or periodically, the excess of collected secondary electrons at the emitting surface.

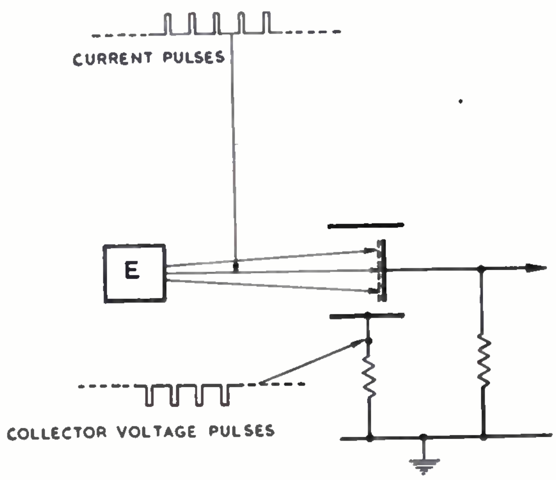

Fig. 3—Basic principle of "pulsed biasing" the storage surface.

This electron replacement may be achieved by a combination of electron pulses sent on to the storage plate and voltage pulses on the collector, as schematically illustrated in Fig. 3. E represents a pulsed electron source generating a diffuse beam with velocities of similar order to those of the electrons from the photocathode or the scanning beam. Simultaneously, negative voltage pulses are applied to the collector, thus establishing a retarding field in front of the storage plate. Consequently, during the pulse period, all secondaries from the storage surface must return to the surface, which becomes progressively more negatively charged, until either equilibrium with the collector pulse potential is established, or, should the amplitude of this potential be very high, until the cessation of the pulse. During the intervals between pulses, the collector reverts to its normal potential, leaving the storage plate charged negatively to it. As a result of this, the tube operates under improved conditions, and, because the charging process is periodically repeated, the improved conditions are maintained.

An estimate of the requirements for the pulsed electron-beam current i and the collector amplitude can easily be made. It is assumed that the duration of the blanking interval is 5 per cent of the whole frame period, that the pulsing period t occupies approximately half of this interval, and that the total storage-plate capacity C is 5,000 pf, these being typical values found in practice. We further assume that the electron source is of high internal impedance, and that the collector pulse amplitude is sufficient to retard substantially all emitted secondary electrons. The process is then time proportionate.

$$\text{Potential change } \Delta e = \frac{\text{additional stored charge}}{\text{capacity}} = \frac{i \times t}{C},$$

therefore, $i = C \times \Delta e / t$.

As the required potential shift Δe is of the order of only a few volts (see Fig. 2), the charging pulse current required is of the order of 10^{-5} amps.

Tubes of the image iconoscope type are most suited to the pulsed-bias mode of operation since it is possible to utilize the picture photocathode as the source of electrons E noted in Fig. 3. It is necessary only to illuminate the photocathode periodically from a suitably controlled light source, such as a miniature cathode-ray tube as shown in Fig. 4 (see following page), a gas-discharge tube, or an incandescent lamp in conjunction with a rotating or vibrating shutter.

Fig. 4 illustrates the procedure adopted for the transmission of cinematograph films, utilizing intermittent projection on to the photocathode of the pickup tube during the frame-blanking period (memory scanning). The mode of operation is as follows: The whole frame period is divided into three intervals. During the first interval the charging of the storage surface takes place according to the method described. During the following period the image of the stationary film is projected on to the tube photocathode, whereby a charge pattern corresponding to the picture content is developed on the surface of the storage plate under improved conditions. The total time of these two intervals must not exceed the duration of the frame-blanking interval. During the third period of the cycle the charge pattern is evaluated by the scanning beam under conditions of high efficiency as almost all the secondary electrons generated are collected. Picture projection and scanning

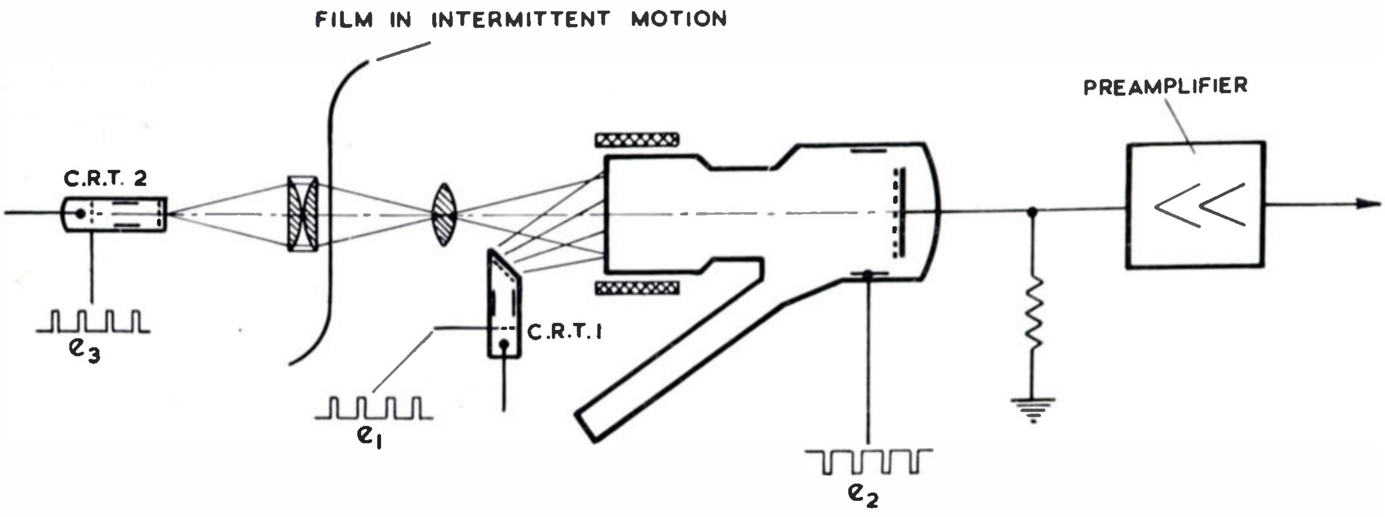

Fig. 4—Pulsed-bias operation of an image iconoscope for the television transmission of films.

restore the normal equilibrium condition, the negative potential shift being reinstated during the first period of the following cycle. The whole process is repeated at field frequency, e.g., 60 or 50 cps in normal television systems.

Fig. 4 also illustrates the use of a cathode-ray tube as a suitable light source for the intermittent picture projection. The control electrodes of the cathode-ray tubes CRT 1 and CRT 2 and the collector of the pickup tube are fed with voltage pulses in accordance with the sequence illustrated in Fig. 5. Between times 1 and 2 the charging process takes place, between 2 and 3 the picture projection, and between 3 and 1′ the evaluation by scanning. Waveform e_1 illustrates the pulses fed to the control electrode of CRT 1_1, e_2 the pulses fed to the tube collector, and e_3 the pulses fed to the control grid of CRT 2 to control the light-projection source.

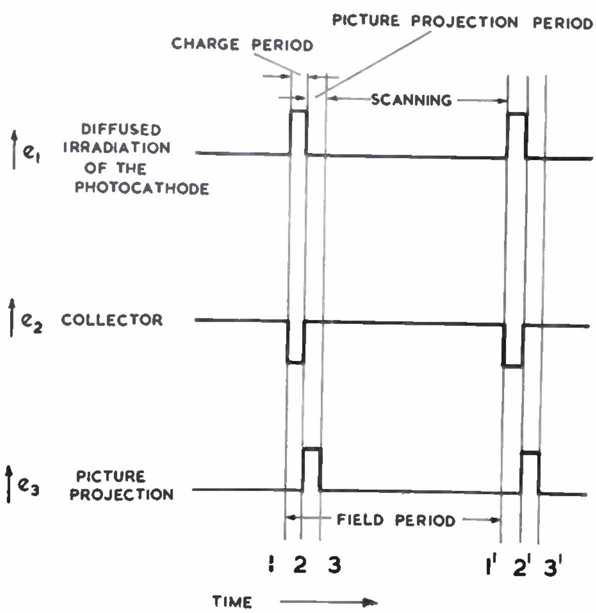

Fig. 5—Timing sequence of voltage pulses shown in Fig. 4.

Analysis of the Output Signal of Pulsed-Biased Image Iconoscopes

As only the general character of the produced signal is of interest at this stage, second-order effects, such as those due to the redistribution of residual secondary electrons, incomplete charge evaluation, and the like, are neglected during the first part of the following analysis, but are discussed at a later stage.

Consideration is first given to the application of the pulsed-bias technique for the transmission of cinema films, as outlined in Fig. 4. The signal-plate current waveforms, illustrated in Fig. 6, are obtained when transmitting an evenly illuminated field at different light levels: section (a) showing the current waveform obtained with no illumination (dark field), section (b) with medium-intensity illumination (gray field), and section (c) with high-level illumination (white field).

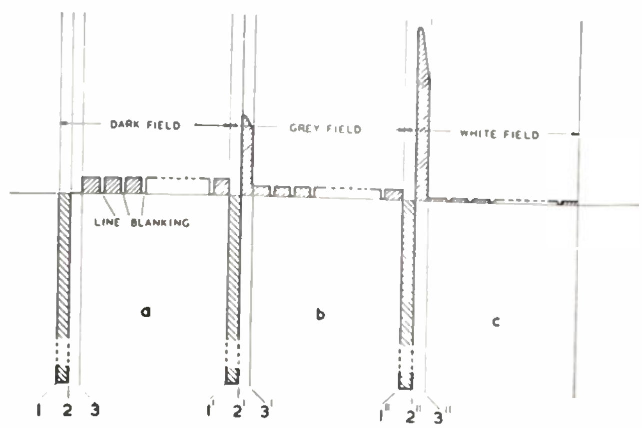

Fig. 6—Schematic-output waveforms from a tube operated as shown in Fig. 4 (intermittent picture projection).

During the charging intervals, 1–2, 1′–2′, 1″–2″, there is present only the primary current originating from the pulsed photocathode or other source as all secondary electrons are returned to the storage surface. This charge current is plotted in a negative direction since the resulting voltage developed at the amplifier input is negative relative to ground. During interval 2–3 (Fig. 6(a)) there is no signal-plate current owing to the absence of any photocathode illumination (dark field). It is only during the scanning interval 3–1′ that electrons can reach the collector by secondary emission, thus producing a positive-going signal-plate current, interrupted during the line-blanking intervals. Negative-going pulses similar to those occurring during interval 1–2, are also developed during intervals 1′–2′ and 1″–2″.

The conditions of grey field, illustrated in Fig. 6(b), are different, as during the picture-projection interval 2′–3′ a positive-going pulse is developed as a result of secondary emission due to photoelectron bombardment of the storage surface. In consequence of the partial discharge of the storage capacity, the amplitude of the signal developed during the subsequent scanning process 3′–1″ is lower than that developed under dark field conditions.

During transmission of the brightly illuminated field (Fig. 6(c)) the surface is almost completely discharged to the normal collector potential in the picture-projection period 2″–3″, and the amplitude of the signal developed during the subsequent scanning period 3″–1‴ is small.

In order to obtain the described results, it is important that the scanning-beam current is so adjusted that, in all cases, the total charge is completely evaluated. In other words, the mean signal current per picture element must have a direct relationship to the element charge.

The electron charges arriving at and leaving the storage surface must balance; therefore, as the charge is equal to the time integral of the current, it is necessary that the area equivalent to the negative-going pulses, as indicated in Fig. 6, shall, over a complete period, equal the total of the areas equivalent to the subsequent positive-going pulses. The distribution of the positive part of the signal between the picture projection and the scanning interval will vary according to the average light flux of the scene to be transmitted. As the amplitude of the signal during scanning decreases with increasing light level, the sense of the signal is negative, that is, it is of similar polarity to that obtained during normal operation of this type of tube. This makes it possible to introduce a continuous change from normal- to pulse-operated conditions.

It may be seen from Fig. 6 that information regarding the mean level of picture brightness is included in the picture signal when referred to the interline (zero beam) level. This unidirectional nature of the signal is due to the existence of the heavy counter-going pulse during the charging (nontransmission) period. The dc information is retained during the passage through an *RC* coupled amplifier if the well-known principle of dc restoration is used with reference to the line-blanking interval signal.

Although the mode of operation has been explained in conjunction with Fig. 6 for the transmission of pictures in the form of evenly illuminated fields only, it applies equally to fields of irregular light distribution. During the picture-projection interval, a charge pattern is developed on the storage surface and an average current flows corresponding to the integral of the elementary currents over the whole storage plate. The scanning process subsequently erases the remaining charge, thereby producing the varying signal current, periodically interrupted by the line blanking, the level during the blanking intervals corresponding to peak white.

As explained above, the beam-current density must be high enough to discharge every picture element completely. This means that some redistribution always exists, as, at least during the final part of the discharge process, some secondary electrons are returned to the storage surface. Fortunately, no serious disadvantages are caused by this residual redistribution, because the tendency of the secondaries to migrate in the direction opposite the direction of motion of the scanning beam is even greater under pulsed-bias operation than under normal operating conditions. Consequently, the secondary electrons do not disturb the charge pattern of those parts of the storage surface yet to be scanned, and any nonuniformity of the redistributed charge landing behind the beam is eliminated during the following pulse-charge interval, which precedes the next impression of the picture-charge pattern and scanning process.

There is also some redistribution during the picture-projection period, mainly towards the end of it, as areas which correspond to bright parts in the scene assume nearly the equilibrium potential. This can reduce slightly the storage efficiency, and may also cause some local variations in the picture contrast.

In order to obtain maximum efficiency and advantage of the pulsed-bias method of operation, it is desirable to keep the residual redistribution to a minimum. The surface-potential shift, established during the pulse period, should therefore be as high as possible. However, limitations are placed upon this requirement as the potential pattern amplitude on the storage surface has to be kept within a limited range in order to avoid second-order effects, such as those due to chromatic aberration in the focusing of the image and scanning electrons, deflection of the electrons by transverse fields, and the like. Fortunately, these limitations do not seriously restrict the application of pulsed-bias techniques.

With the restricted voltage range of each storage element, the amount of charge which can be stored depends on the element capacity. The suitable choice of the storage-plate capacity is now determined by the average light which is available in the scene, on which depend the average number of photoelectrons and, consequently, the average charge to be stored. For operation at low light levels, for example, small storage capacities are preferable in order to reach the desirable maximum potential change with the low photocurrent available. The optimum beam current is then also small. On the other hand, if high light levels are available, it is preferable to use relatively large storage capacities which can retain a larger charge in the same permissible voltage range. It then becomes possible to obtain very strong output signals with negligible noise content under these conditions, which is an important advantage of the proposed method of operation.

The introduction of pulsed-bias operation modifies the transfer characteristic (signal output against light input) from that which is normal in high-velocity scanned tubes. The initial portion of the transfer characteristic is made more straight and the level at which output limitation occurs extends to higher values of light level, with increasing degree of potential shift. The half-tone rendition, as seen on a normal television receiving tube, is consequently different from that obtained under normal operation. If a high degree of biasing is applied, it may be necessary to "de-gamma" the signal in the following amplifier, as is required for flying spot scanners and Orthicon and image-Orthicon pickup tubes.[18,19] The variations of the transfer characteristic with the operating conditions are very similar to those obtained from the image Orthicon, the shift of the storage surface potential corresponding to the change of the mesh potential.

The analysis of the signal output makes it clear that under pulsed-bias operation the average signal amplitude is considerably smaller than the charge and discharge pulse amplitudes (see Fig. 6). It is therefore necessary to eliminate the peak pulse signals by suitable clipper circuits before or in the first stage of the preamplifier following the camera tube, otherwise the system would be overloaded and the useful signal distorted. This can be easily carried out by known methods.[16]

As explained, the interline pulses of the picture signal current correspond to peak white. This level is therefore not very suitable as an interval in which the "clamp circuit" for dc restoration is effective, as the real black level, although in fixed relation to these interline pulses, (see Figs. 6 and 8), will vary with changes in bias light pulses, in beam current, and the like. It is preferable to provide a restoration reference level corresponding to true black anywhere in the blanking period where the clamp circuit operates. This can be done by inserting a black strip at the left-hand side of the picture, which, of course, must be scanned before the picture blanking is ended. This requires that the camera-tube line-deflection retrace time be shorter than the picture-blanking time. The clamp circuit, effective only for a short period immediately before the actual commencement of the line scanning, then holds the black level from which the picture signal is developed in a unidirectional sense (Fig. 6).

(a)

(b)

(c)

Fig. 7—Television pictures produced by the "Photicon" camera tube under intermittent projection conditions: a. With "pulsed bias" applied. b. As (a) with pulses removed, showing reduction of signal amplitude. c. As (b) with gain increased to obtain signal amplitude comparable with (a) showing marked presence of flare.

[18] T. C. Nuttal, "Some aspects of television-circuit technique: Phase correction and gamma correction," *Bull. schweiz elektrotech. ver.*, vol. 40, pp. 619–622; August, 1949.
[19] C. L. Townsend and E. D. Goodale, "The orthogam amplifier," *RCA Rev.*, vol. 11, pp. 399–410; September, 1950.

Fig. 7 illustrates some of the results obtainable by the application of pulsed-bias operation. In Fig. 7(a) is shown a television picture produced with intermittent picture projection under pulse-bias conditions. Fig. 7(b) shows the picture produced when the biasing pulses are removed, all other operating conditions remaining unchanged. From comparison of these two pictures it may be seen that, in addition to the other advantages claimed, pulsed-bias operation results in increased signal output from the camera tube. This gain is normally of the order of two to three times, considerably improving signal-to-noise ratio in the picture. Higher gains may be obtained by using a greater amplitude of biasing; however, this would probably make it necessary to "de-gamma" the picture, and may also introduce additional spurious signals, such as those arising from local variations in secondary-emission ratio. The results shown are obtained under the best compromise conditions. By increasing the amplifier gain so that the picture amplitude is similar to that shown in Fig. 7(a) the presence of edge flare, which exists under normal operating conditions, becomes more easily observable. This is shown in Fig. 7(c).

The conditions of operation during the transmission of continuously illuminated scenes, as required for live pickup, are more complicated than those during transmission of intermittently projected film, as considered so far. These complications are due to the fact that the charge-pattern development (picture projection) and scanning processes occur simultaneously. One complete frame period consists of two intervals only—pulse charge and scanning.

To analyze the picture signal under these conditions we first assume that during the pulse-charge interval the storage plate is charged to an even potential across its whole surface, any residual charge pattern being erased. The complete signal output of the tube may be analyzed into three separate components. Fig. 8(a) shows the charge current which occurs only during the pulse-charging interval, and consists of a succession of negative pulses at frame frequency. The second component, illustrated in Fig. 8(b), represents the integrated electron current from the storage surface due to secondary emission by photoelectron bombardment. The amplitude of this current decreases during the frame period owing to the time decay of each individual discharge process and owing to the fact that the process of scanning progressively reduces the area of the surface which has not been brought to the equilibrium potential. The shape of the waveform of this component is affected by the degree of potential shift that is, whether or not the secondary emission is saturated. The third component (Fig. 8(c)) is the signal due to the scanning process. This signal shows a change in depth of modulation during the frame period as the storage time is very short at the commencement of the frame-scanning period, but increasingly longer as the scanning proceeds. Therefore, the signal-plate current is high at the commencement of frame deflection whatever the picture content, but with increasing time the current may approach zero if areas corresponding to white areas in the picture are scanned.

Fig. 8(d), shows the actual signal current resulting from the combination of the three components Fig. 8(a–c). Two shortcomings are apparent, namely, the changing base line and the change in modulation depth. It is relatively easy to eliminate the first-mentioned additive component (Fig. 8(b)) as the signal has periodic interruptions during the line-blanking time, the level of which can be used as a reference level for a clamped dc restorer in the amplifier chain in the same manner as other low-frequency distortions, hum, and the like are suppressed. It is only necessary that those stages of the amplifier preceding the dc restoration have a linear transmission range sufficiently wide to avoid distortion of the signal.

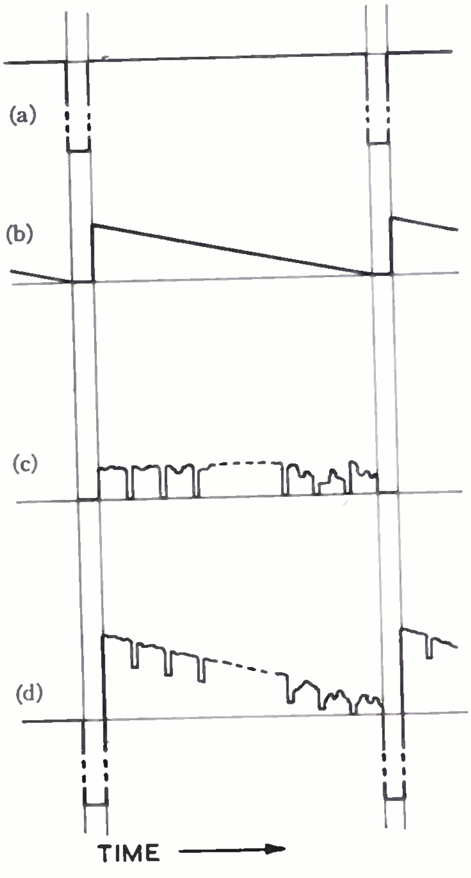

Fig. 8—Schematic-output waveforms from a pulsed-biased image iconoscope under continuous pickup conditions.

The second shortcoming, due to change in depth of modulation, is of much greater importance, and means that no satisfactory signal can be obtained under the assumptions made.

In spite of this, experiments showed that usefully improved pictures can be obtained provided that the potential shift is limited to a low value. This is found to be sufficient to reduce the spurious signals and flare, to give the signal a more unidirectional character, and to balance the slight decrease in the depth of modulation from top to bottom of the picture which occurs under normal operating conditions due to nonuniformity of redistribution.

These results indicate that the lack of charge-pattern development during the early part of the picture period is not as serious as was assumed. Investigations show that this is because the storage time in the tubes under consideration is usually shorter than the frame period. Therefore, by making the charge pulse as short as possible and timing it to occur at the beginning of the frame-blanking interval, the remaining part of this interval is sufficient to build up an appreciable charge pattern at the top of the picture before scanning commences. There still remains a modulation of the signal during the scanning period, but this can be equalized by an inverse modulation of the amplifier gain.

It is also found that an indirect method of charging the storage surface helps to preserve at the top of the picture the charge pattern developed during the preceding frame period. One method of performing this indirect charging is to mask the photocathode so that the pulse illumination does not irradiate the area on which the picture is projected. Instead it irradiates only the adjacent areas, those which mainly surround the bottom part of the picture where the disturbing flare shows up most under normal operation. The process of storage-plate charging occurs by redistributing secondary electrons from the neighboring bombarded areas. These results are not surprising in view of those obtained from bias light technique with memory scanning in iconoscope film scanners,[14] where a potential shift is obtained from the continuous landing of low-velocity photoelectrons at the storage surface which have no appreciable influence on the picture-charge pattern, but impart only an additive charge. A suitable method of carrying out the indirect method of pulse charging is by the insertion of a funnel between the camera lens and the photocathode. This is so formed that it frames the picture area and masks it from illumination by the pulse light source. Experiments with intermittent picture projection in which the phase of the projection relative to the scanning interval is adjustable, indicate that with such an indirect method of charging the shape of the picture-charge pattern is retained to a useful degree after the charge pulse has taken place.

Conclusions

The described "pulsed-bias" technique improves the operating characteristics of image-iconoscope camera tubes by simple means. Experimental results are in close agreement with the given analysis, and show that the major faults of these tubes are eliminated or considerably reduced: Flare and "dark spot" are brought to low levels, no adjustments to the shading controls are required during operation, a unidirectional sense of the output signal is obtained, and with suitable dc restoration in the amplifier, the constancy of black level is well maintained. A further important feature is that, owing to the improved efficiency of signal generation, the output signal amplitude, and consequently the signal-to-noise ratio, are increased by about two to three times.

The full advantages of the technique are obtainable on the transmission of intermittently projected pictures, such as "memory-scanned" cinema films. For continuous pickup, only partial application of the technique is possible. Although the improvements so obtained are appreciable, the practical use of this technique is doubtful as other methods of bias application, now under development, show promise of better results. It is hoped that a report on this work will be published in the near future.

Provision for the pulsed-bias method of operation can be easily incorporated in image-iconoscope camera equipment so that the same tube may be used biased for film transmission and unbiased or partially biased for direct pickup (studio) use. This meets the long-established requirement for a camera tube able to give satisfactory service for both purposes, and the possibility of more flexible equipment utilization is thereby increased, bearing in mind that modern versions of the image iconoscope, such as the Photicon, are giving highly satisfactory results as studio and outside broadcast pickups in Great Britain and other European countries.

Results obtained on tests and in service for film transmission compare favorably with those obtained from other high-quality devices, such as flying spot scanners and image dissector tubes.

Apart from the improved equipment utilization, the use of a camera tube in conjunction with standard (synchronized) cinema projectors for the televising of films allows stationary film to be viewed, which means that single frame captions may be transmitted or pictures may be previewed for adjustment or other purposes.

The pulsed-bias technique as described has been incorporated in telecine equipment manufactured for the British Broadcasting Corporation by Pye Ltd. of Cambridge, England, utilizing the "Photicon," a British-made camera tube of the image-iconoscope type.

Acknowledgments

The investigations were carried out in the Laboratories of Pye Ltd. and Cathodeon Ltd., both of Cambridge, England, and the authors wish to express their thanks to the directors of these companies for permission to publish these results. Thanks are also due to many of their colleagues for suggestions, criticisms, and assistance, and to R. W. Lee of General Precision Laboratories, Pleasantville, N. Y., for helpful discussions.

The Q of a Microwave Cavity by Comparison with a Calibrated High-Frequency Circuit*

HUGH LeCAINE†

Summary—A comparison method has been developed for the direct measurement of cavity Q at microwave frequencies. It is particularly useful for high values of Q, such as 5,000 to 15,000. The over-all error in the measurement is estimated to be less than ±3 per cent.

A two-channel superheterodyne technique is used, in which both channels are driven by the same frequency-swept oscillator, and both channels use the same local oscillator. The cavity is inserted in the radio-frequency stage of the first channel, and a comparison circuit is inserted in the intermediate-frequency stage of the parallel channel.

The two resonance curves are displayed on the same oscilloscope for alternate sweeps of the oscillator. When the resonance curves are made to coincide, the Q of the cavity is n times the Q of the comparison circuit, where n is the ratio of radio frequency to intermediate frequency.

Cavity shunt resistances can be measured on the same apparatus.

I. The Method

Most practical methods of measuring the unloaded Q of a cavity follow either one of two basic procedures: (1) The relative energy transmitted through the cavity is measured over a range of frequencies including the resonant frequency, using extremely loose coupling both in and out. (2) The impedance looking into the cavity is measured over a similar range of frequencies.

Sproull and Linder[1] followed the first procedure, sweeping through the range of frequencies automatically with a "frequency-swept" klystron oscillator, and recording the transmitted energy on an oscilloscope. Then the Q was determined from the resulting "resonance curve" by observing the spread in frequency between half-power points. The method described below is essentially that of Sproull and Linder, with the addition of a variable "comparison curve" displayed on the same oscilloscope and taken from a high-frequency resonant circuit of variable Q. The result is a considerable increase in the accuracy of the measurement, particularly when the Q of the cavity is high.

A two-channel superheterodyne technique is used in which both channels are driven from the same frequency-swept oscillator, and both channels use the same local oscillator as shown in the block diagram of Fig. 1. The cavity is inserted in the radio-frequency stage of the first channel, and the comparison circuit is inserted in the intermediate-frequency stage of the parallel channel. Then any increment in frequency at the cavity is identical with the resultant increment in frequency at the comparison circuit.

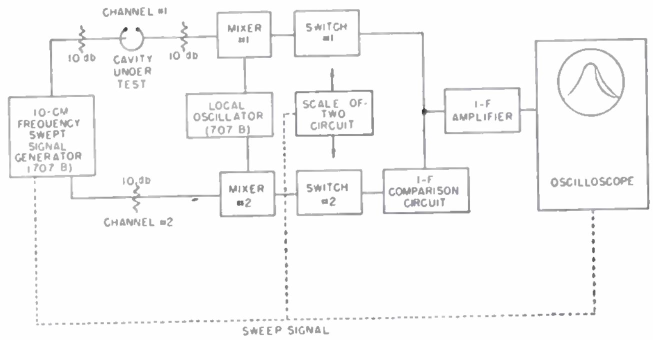

Fig. 1—A block diagram of the two-channel superheterodyne Q measuring setup.

The two resonance curves are displayed on the same oscilloscope for alternate sweeps of the oscillator, and the Q of the comparison circuit is varied until the two curves come into coincidence. Then the spread in frequency between the half-power points on one curve will be identical with that on the other, while the resonant frequencies will be in the ratio n, where n is the ratio of radio frequency to intermediate frequency. Hence the Q of the cavity will be n times the Q of the comparison circuit.

The equipment was required to measure the Q's of cavities resonant at about 2,800 mc, and it was decided to design for a range of Q's from 5,000 to 15,000. An intermediate frequency of 19.5 mc was selected so that n equalled 144. Thus the required range of Q's for the comparison circuit is about 35 to 105. Such a range of values is readily realized at 19.5 mc in an accurately calibrated variable Q circuit.

II. The Advantages

The outstanding advantage of the comparison method of measuring cavity Q's is a decided increase in the accuracy of interpreting the display on the cathode-ray tube, particularly when the Q is very high. In the single resonance curve method, the difference in

* Decimal classification: R119.33. Original manuscript received by the Institute, May 22, 1950. Presented, Joint Meeting, URSI-IRE, Washington, D. C., April, 1950.

† National Research Council of Canada, Ottawa, Canada.

[1] R. L. Sproull and E. G. Linder, "Resonant cavity measurements," Proc. I.R.E., vol. 34, pp. 305–312; May, 1946.

frequency at the half-power points must be read from the face of the tube, and when the Q is very high this difference may be less than a scale division on the wavemeter. For example, the difference in frequency is only 280 kc for a Q of 10,000 at a resonant frequency of 2,800 mc.

Montgomery[2] has described a heterodyne system of producing two markers or pips which may be set at the half-power points. The difference in frequency between the two pips can be determined with great accuracy, but the error in determining the half-power level is still considerable. In any case, the over-all error in the single resonance curve method is seldom less than ± 7 per cent.

In the comparison method the cathode-ray tube is used simply to bring the two resonance curves into coincidence between the half-power points. Complete coincidence below the half-power points is not always possible, since the two resonance curves are not quite identical in shape. The error introduced at this point is approximately the ratio of spot diameter to one-half the face diameter, or about ± 1 per cent.

In addition, the comparison method has several lesser advantages over the single resonance curve method, each contributing to the over-all accuracy of the measurements.

(1) Since the superheterodyne detector is highly sensitive, it is extremely easy to make the coupling both into and out of the cavity much smaller than necessary to avoid cavity loading effects.

(2) The response-law of a crystal used as a first detector in a superheterodyne circuit is much more reliable than that of a crystal used as a simple detector. The response-law of the second detector does not matter, since it is common to both systems.

(3) The two mixers are separate in the two channels, and must be similar; but all other components are common and affect both curves equally. Hence there are no close restrictions on distortion in the oscilloscope, distortion in the amplifier, or amplitude and frequency modulation of the swept oscillator.

The over-all error for the comparison method, measuring Q's of about 10,000, is estimated to be less than ± 3 per cent.

III. The Measurement of Shunt Resistance

The shunt resistance of a cavity can be measured on the same equipment by the "capacitance insertion" or "resistance insertion" method. In the first case, a small dielectric cylinder is placed in a suitable position in the cavity and the shunt resistance is calculated from the resultant change in resonant frequency. The method is essentially that of Sproull and Linder,[1] but uses a superheterodyne technique with one outstanding advantage.

The absolute difference in frequency is exactly the same in the radio- and intermediate-frequency stage, where its value relative to the mean frequency is very much greater. A small signal from a calibrated signal generator may be inserted in the intermediate-frequency stage, to appear in the output as a zero-beat marker superimposed upon the resonance curve. The marker may be set on the peak of the resonance curve, or on some point on the side of the resonance curve, both before and after insertion of the dielectric cylinder, and the difference in frequency noted.

The cavities under test at the National Research Council were for use with tunable magnetrons in an experimental electron accelerator. Thus, there was no need for a precise measurement of the resonant frequency, and this problem was not considered.

IV. The Component Parts of the Instrument

A. The Frequency-Swept Oscillator

Any oscillator capable of being swept over the required frequency range may be used. The frequency change need not be any particular function of the applied voltage, since in any case the same sweep is applied to both channels. In the 10-cm band, the 707-B klystron is a suitable tube.

B. The Mixers

The two mixers should be as nearly identical as possible, using simple untuned circuits, and matched crystals.

If the cavity is replaced in the first channel by a flat attenuator with the same insertion loss, and if the comparison circuit is removed from the other channel, then the two channels will be identical except for whatever differences there may be in the two mixers. If the two curves obtained on the oscilloscope under these conditions lie together over the required frequency band, then the mixers are satisfactory.

C. The Switches

The switches are used to present the resonance curve of the cavity and that of the comparison circuit on alternate sweeps of the same oscilloscope. They consist of two 6AC7 tubes biased to cutoff on alternate sweeps. The divider circuit is a standard "scale-of-two" circuit, triggered by a pulse derived from the sweep generator in the frequency-swept oscillator. A change in gain from "off" to "on" of zero to 40 or 60 db is realized readily, and this is more than adequate since the two curves are adjusted to coincide in any case.

[2] C. G. Montgomery, "Techniques of Microwave Measurements," Radiation Laboratory Series, vol. 11, pp. 396–407; McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

D. The Comparison Circuit

The comparison circuit is a simple coil and condenser combination with a variably coupled loading resistor in shunt, as shown in Fig. 2. The Q of the circuit without the loading resistor must, of course, be higher than the highest Q required for comparison purposes. Then any required Q may be realized by simply varying the coupling of the loading resistor.

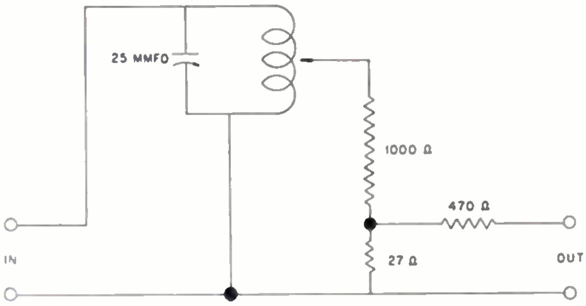

Fig. 2—The "comparison" circuit in the intermediate-frequency stage.

It is highly desirable for simple operation that the loading resistor should change the Q of the circuit without detuning it, and thus it should be relatively nonreactive. Hence the capacitance in the comparison circuit should be quite large so as to minimize the effect of the natural shunt capacitance across the resistor. A suitable value was realized for a range of Q's from 35 to 500 at 19.5 mc, when the capacitance was adjusted to give the circuit a resonant impedance of about 1,000 ohms with the coupling of the loading resistor set to give a Q of about 200. At the same time, the loading resistor itself should be selected for minimum inductance. Carbon resistors were found unsuitable because of their poor temperature characteristics. A specially built noninductive, wire-wound resistor was finally selected for this purpose.

It has been found that such a comparison circuit holds a calibration exceedingly well, chiefly because there is very little drift in the value of the loading resistor. Suitable calibrating gear has been built into the setup.

In calibrating the comparison circuit the Q is measured for a large number of separate couplings of the loading resistor, and the equivalent Q of the cavity is marked directly on the coupler scale, thus making the instrument direct reading. For any particular coupling, the Q is determined by measuring the difference in frequency between the half-power points on the resonance curve of the comparison circuit, using a frequency-calibrated signal generator with the frequency of the output controlled manually.

The output from the signal generator is fed through an automatic-gain-controlled 19.5 mc amplifier to the grid of the 6AC7 switching tube, as shown in the block diagram of Fig. 3. The output of the amplifier remains constant as frequency is varied, but may be set as required at either a "peak" (full) output, or a "0.707" (3 db down) output. First the resonance point is found with the amplifier output set at "0.707" by adjusting the frequency for a maximum signal from the diode across the comparison circuit. Then the amplifier output is switched up to "peak" and the signal from the diode across the comparison circuit is brought back down to its former value by detuning the signal generator, thus determining the first of the two half-power points. The other one is found by detuning in the other direction. The frequency at each of the two half-power points is read directly from the signal generator.

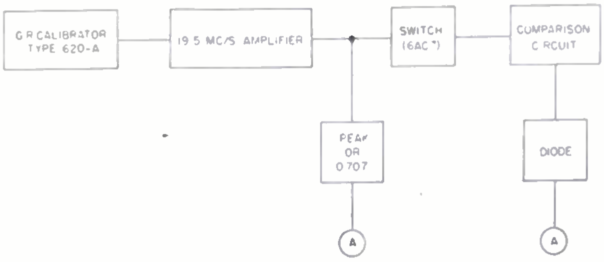

Fig. 3—A block diagram of the calibration circuit.

The response-law of the diode across the comparison circuit is of no importance, since the diode output is brought to the same value for all readings.

The measurement takes into account all frequency-sensitive elements from the plate of the switching tube to the junction point of the two channels. It does not take into account slight variations in impedance with frequency in the grid circuit, which must balance similar variations in the other channel, as explained above under "The Mixers."

The General Radio "Calibrator" Type 620-A was found to be an exceedingly satisfactory signal generator for calibrating the comparison circuit. It uses a 1 mc crystal for checking each of ten 1 mc bands, including the band required here.

E. The Intermediate-Frequency Amplifier

The intermediate-frequency amplifier is of standard construction. It was designed for a bandwidth of about 7 mc, so that a wide frequency sweep could be used in setting up a new cavity. Since this amplifier is common to both channels, its design is not critical.

An intermediate frequency of 19.5 mc was selected because it lies in the middle of one of the bands of the General Radio "Calibrator" Type 620-A used in calibrating the comparison circuit.

A Homopolar Tachometer for Servomechanism Application*

CRAIG C. JOHNSON†

Summary—**Certain design considerations, construction detail, and performance data are presented for a small disk-type homopolar tachometer which was developed as a low inertia, low noise, low friction rate signal source for incorporation into a high-performance servomechanism. Performance characteristics reveal that a unit of this type has certain advantageous features which recommend it over conventional tachometers, although a low signal output level is an inherent disadvantage.**

FOR CERTAIN high-performance servomechanism and electromechanical analog computer systems in which it is necessary to obtain a voltage proportional to a shaft rate (wherein this quantity is used for feedback and/or intelligence), it is desirable to have available a low inertia, low friction tachometer, the output signal of which is linear with the input shaft rate. It is also desirable that the electrical noise components in this output signal be held to a minimum. The attainment of these characteristics in a tachometer will enhance the performance of the servo system under both static and dynamic conditions. Commutator ripple and relatively high brush friction associated with permanent magnet and other dc generators are generally undesirable. AC drag-cup induction generators, though mechanically satisfactory, present problems because of electrical phasing difficulties, residual voltage effects, harmonic content in the generated signal, and rectification and filtering in such systems where a dc output is required. The homopolar tachometer described in this paper was developed as a component in a high-performance servo system in an attempt to circumvent some of the above-mentioned difficulties.

The principle of the homopolar generator is believed to have been discovered by Faraday, and may be stated simply, as follows: If a conducting disk is rotated between the poles of an annular magnetic field of constant strength, as indicated in Fig. 1(a), an emf is induced between the periphery of the disk and its center of rotation. The polarity of the induced signal is in accord with Lenz's law; thus by reversing the direction of disk rotation, the polarity of the induced signal is also reversed. No eddy-current damping will exist so long as the magnetic field is circumferentially homogeneous. In terms of the physical parameters of the conductor and magnetic field, the value of the induced voltage is given by the following expression:

$$E = \frac{\beta l \bar{v}}{10^8} \text{ volts,} \tag{1}$$

where

β = magnetic-field intensity in gauss

l = length in cm of conductor cutting magnetic lines of force

$\bar{v}$ = average velocity of cutting lines of force in cm/sec.

In as much as the output-voltage level for such a unit is characteristically low as compared to conventional tachometers of roughly the same physical size, it will usually be necessary to amplify this signal for system utilization. Because of this fact, and since a low disk moment of inertia will usually be desirable in a high-performance system, it follows that the output voltage/inertia ratio might be a useful criterion in selecting an optimum rotor-magnetic field configuration. Referring again to Fig. 1(a) and substituting appropriate values into (1), one finds the output voltage for this configuration to be

$$E = \frac{3}{8}\left(\frac{\beta\omega r_0^2}{10^8}\right) \text{ volts,} \tag{2}$$

where

ω = angular rate of disk rotation in rad/sec.

(It is assumed in (2) that $r_i = r_0/2$ for a small disk in order to provide clearance for a shaft and supporting

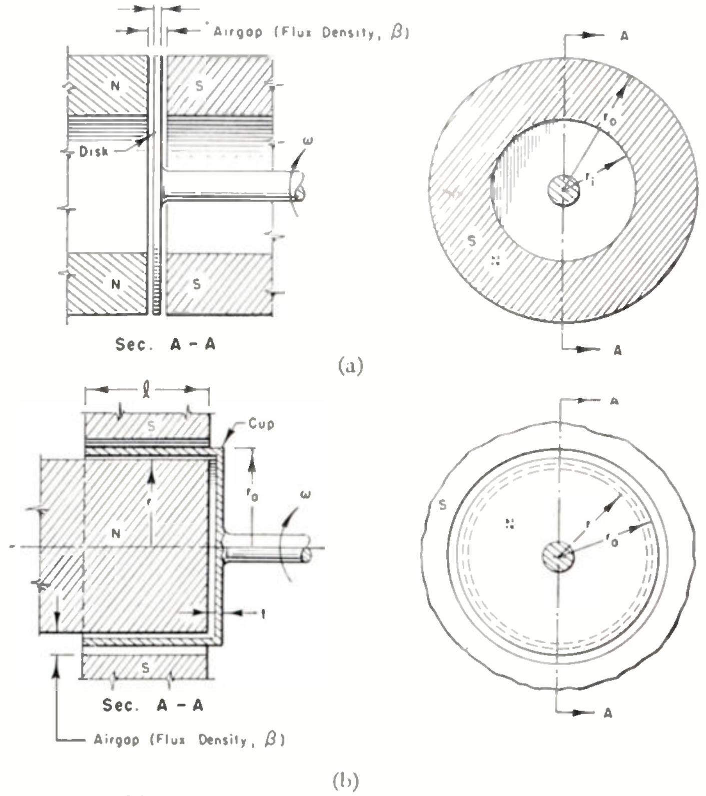

Fig. 1—Magnetic field—rotor configurations.

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† North American Aerophysics Laboratory, Downey, Calif.

bearings.) Then, assuming a solid disk of thickness t and density δ, and neglecting the moment of inertia of the shaft and bearings, the output voltage/inertia ratio is given in appropriate units by

$$\frac{E}{J} = \frac{3}{4}\left(\frac{\beta\omega}{\delta\pi t r_0^2}\right)\frac{1}{10^8}. \qquad (3)$$

For purposes of comparison, the configuration shown in Fig. 1(b) will also be considered. This unit consists of a cup rotating in a radial magnetic field similar to the voice coil magnetic circuit used in most loud speakers. From the standpoint of good magnetic design, it will be assumed that

$$\pi r^2 = 1.2(2\pi r l).$$

For a narrow air gap, it will be valid to let $r_0 \approx r$, and it will be further assumed that $t \ll r_0$. Under these conditions, and letting β, ω, δ, t, and the output voltage E be the same for the two configurations (i.e., the cup and the disk) it can be shown that

$$\left(\frac{E}{J}\right)_{\text{disk}} = 2.16\left(\frac{E}{J}\right)_{\text{cup}}. \qquad (4)$$

Further analyses along these lines will reveal that the disk has an E/J higher than any other single-rotor configuration. Moreover, it can be shown that E/J for cascaded disks may be improved over that of a single disk by a factor equal to the number of disks in series, this conclusion being based on the comparison of two units of equal output voltage. However, the construction of a unit utilizing smaller disks in cascade would be rather elaborate and, since each disk requires brushes, it could be expected that higher friction and noise levels would result.

In view of the above results, together with the fact that a disk-type unit would be easier to fabricate, it was decided that such a unit as shown in Fig. 2 should be constructed. In order to reduce the size of the exciting coil, the magnetic circuit was made of Allegheny 4,750 high-permeability alloy, which saturates around 10,000 to 11,000 gauss. With this value of β, the dimensions of the magnetic field were chosen so that the output voltage would be approximately 1 mv per rps. With an aluminum (24ST) disk 2 inches in diameter and 0.030 inch thick and a total air gap of 0.060 inch, it was found that a 1,500-amp turn coil would be required. A coil of 1,030 turns was layer wound of No. 20 Formex insulated wire. An exciting current of 1.5 amp dc essentially saturates the magnetic circuit and results in the dissipation of about 15 watts in the coil. The coil is electrically floating so that the polarity of the magnetic field may be reversed by switching the coil input leads.

As mentioned at the outset, low electrical noise and low mechanical friction are two important characteristics desired in a unit of this type. The quality of the electrical pick-off brushes will directly determine both the noise and friction levels. Generally, an increase in brush pressure will decrease noise and increase friction so that some compromise solution is required. For this unit, reasonably satisfactory performance for the peripheral contacts was achieved through the use of two brushes (shorted together) phased approximately 90 space degrees apart on the disk, a unique spacing which appeared to reduce chattering effects. The contacts consist of Paliney No. 7 bars soldered onto thin leaf springs which were over-damped with layers of scotch electrical tape. These contacts ride on a thin silver ring which was shrunk over the disk periphery. It was found necessary to lap the contacts and ring with fine abrasive in order to reduce output noise. For symmetrical performance with direction of disk rotation, it was found desirable to orient the leaf springs in opposition, as shown in Fig. 2. Brush pressure was adjusted for optimum results. Because of low rubbing velocities the center contact is not critical and consists merely of a spring bearing against a brass pin which is pressed into the disk center. It is apparent from Fig. 2 that the shaft

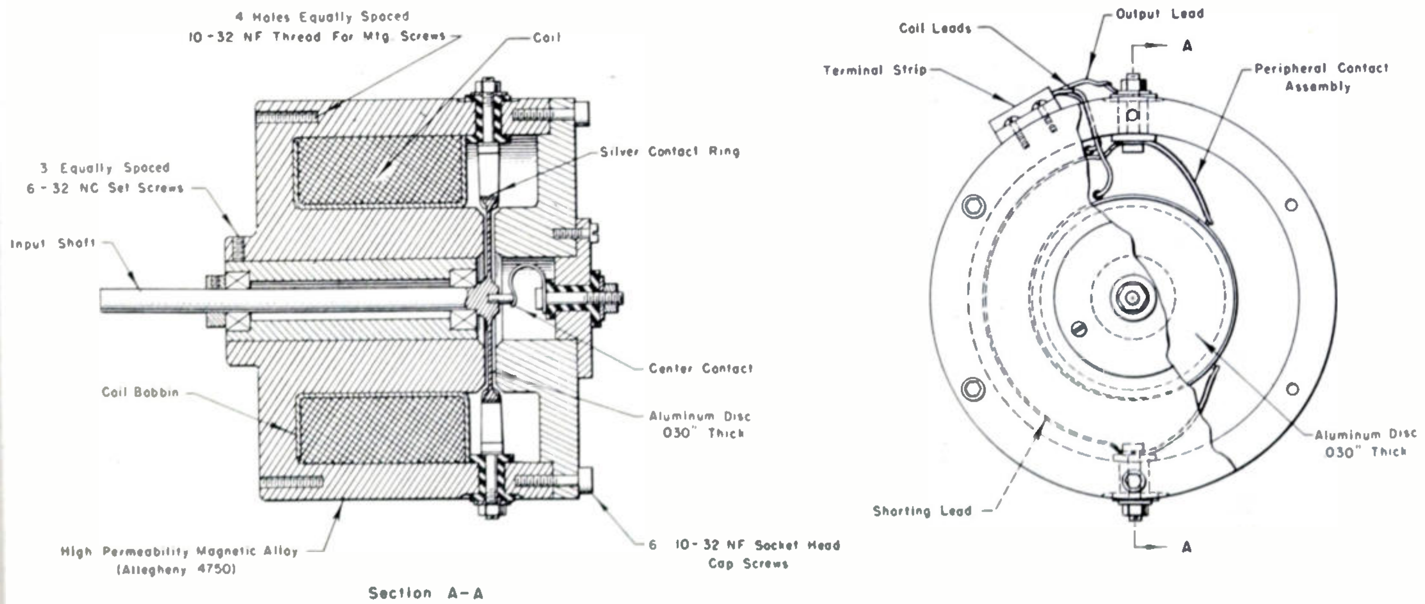

Fig. 2—Homopolar generator.

and disk are electrically grounded through the bearings to the case. Thus, the center contact serves only as a direct ground for eliminating bearing noise. It is possible, of course, through suitable insulation, to construct a unit with a double-ended output.

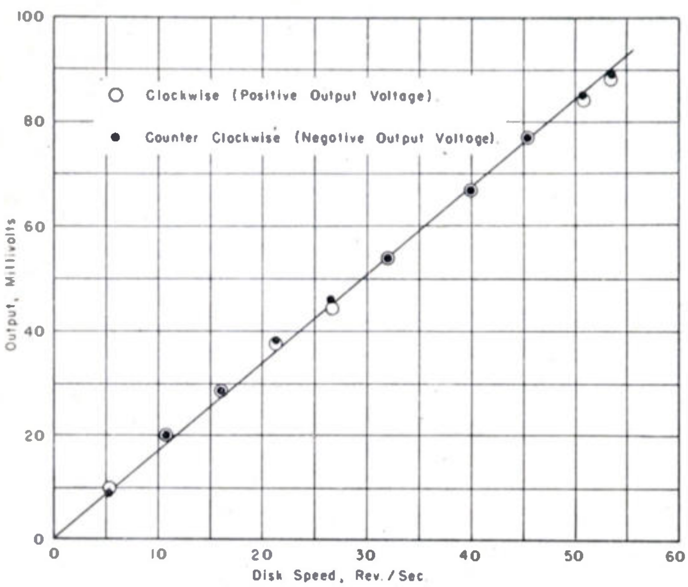

Fig. 3—Output volts versus speed at constant field current (1.5 amp dc).

Linearity of output voltage versus speed of rotation is indicated in Fig. 3. The output voltage is about 1.7 mv per rps, which is somewhat higher than design calculations. This may be attributed to the fact that magnetic fringing effects were neglected in these calculations. No attempt was made to measure the electrical time constant of the disk, but it is believed that it will not cause any effects below, for instance, 50 cps. The moment of inertia, J, of the disk equals 2.5×10^{-6} slug-ft^2 (where 50 per cent of this value is represented by the silver ring). Starting friction torque is about 0.4 oz. in.

The electrical noise present in the output is high frequency in nature, and its peak-to-peak value was found to be less than 1 per cent of the signal voltage for all values of speed up to 3,200 rpm. The noise level was observed with the generator output feeding into the grid of an amplifier tube and filtered with a 0.01 μf condenser. Current loading of the output materially increases the noise level so that a moderately high impedance load appears essential to satisfactory operation. The generation of thermoelectric potentials between the Paliney No. 7 brush contacts and the silver ring, due to frictional and coil dissipation heating, appears negligible since the output voltage versus speed curve (Fig. 3) is essentially symmetrical with direction of rotation. Appreciable thermoelectric potentials would bias the curve and destroy such symmetry.

The actual unit described above has been incorporated into the system for which it was designed. Performance of the unit in this system has proven qualitatively superior to the performance of other types of tachometers which were previously utilized. Operation has been essentially trouble free, except that it appears desirable to clean and relubricate the contact surfaces occasionally.

It is believed that the homopolar tachometer may be improved considerably for use with servomechanisms and related systems. A permanent magnet field, a reduced over-all size, and a mercury-pool electrical pick-off present intriguing possibilities. The major disadvantage of a low output signal level remains, however, thus necessitating the rather stringent requirements for low-drift, high-gain amplification.

ACKNOWLEDGMENT

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A Portable, Direct-Reading Microwave Noise Generator*

E. L. CHINNOCK†

Summary—**This paper discusses the factors which influenced the design of a directly calibrated portable microwave noise source, utilizing a fluorescent lamp. The variation of the noise power output and the impedance match as a function of the operating temperature are considered, and the portable unit is described.**

THE USE OF the gaseous discharge in an ordinary fluorescent lamp as a source of microwave noise power has been suggested by Mumford. The uniformity and stability of these lamps make them attractive for use as a tool for the measurement of noise figures of microwave circuits.[1] There is, however, a slight temperature correction to be applied when the greatest accuracy is demanded. The unit to be described includes a convenient means of allowing for this correction.

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† Bell Telephone Laboratories, Inc., Holmdel, N. J.

[1] W. W. Mumford, "A broad-band microwave noise source," *Bell Sys. Tech. Jour.*, vol. 28, pp. 608–618; October, 1949.

A plot of the data taken on an early model shows the magnitude of this correction in Fig. 1. These data cover the operating temperature range from 30 to 50°C, as measured by a mercury thermometer placed in contact with the waveguide circuit surrounding the lamp. It is seen that the excess noise power increases by 1.1 db in this 20-degree range, corresponding to a negative temperature coefficient of −0.055 db per degree C.

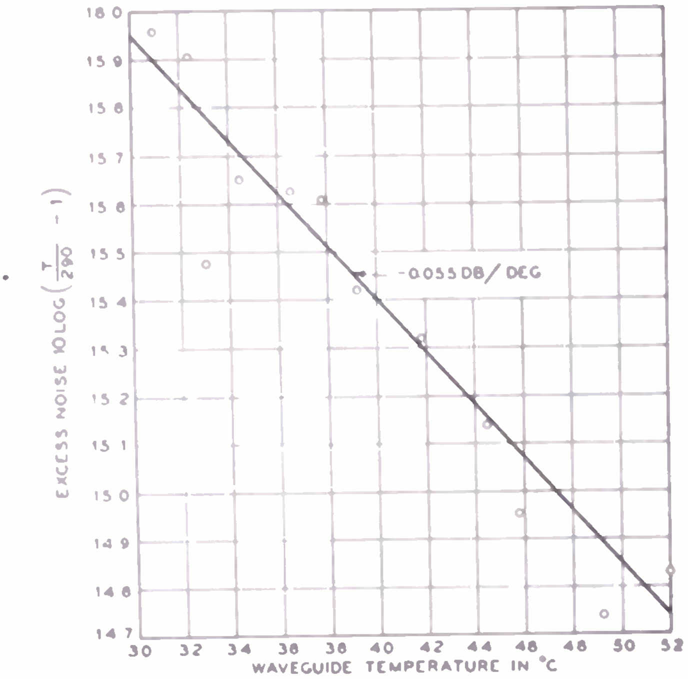


Fig. 1—Temperature coefficient of microwave noise source.

While this is a small coefficient, it was felt that we might do even better by operating at a lower temperature, since when the mercury is frozen out, the discharge would be characteristic of the remaining argon whose electron temperature (and hence microwave noise power output) might be less than that of the mixture of gases.[2] At some intermediate temperature, the coefficient should be zero.

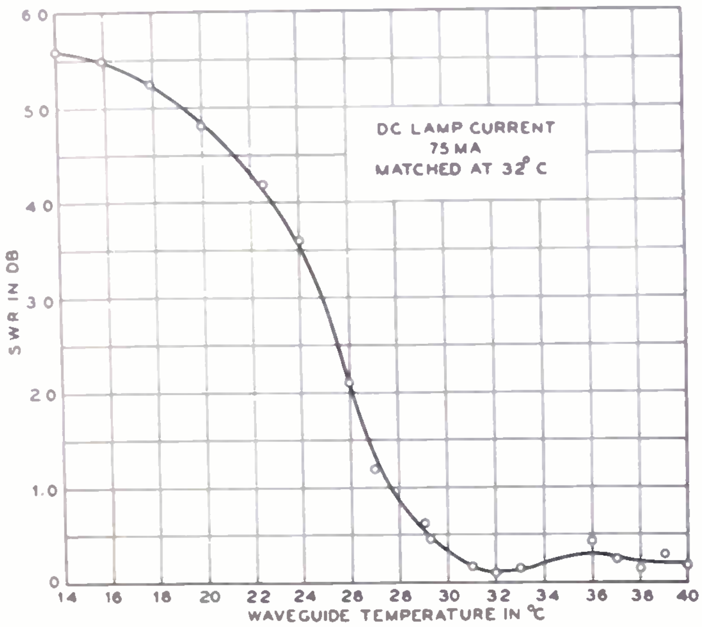
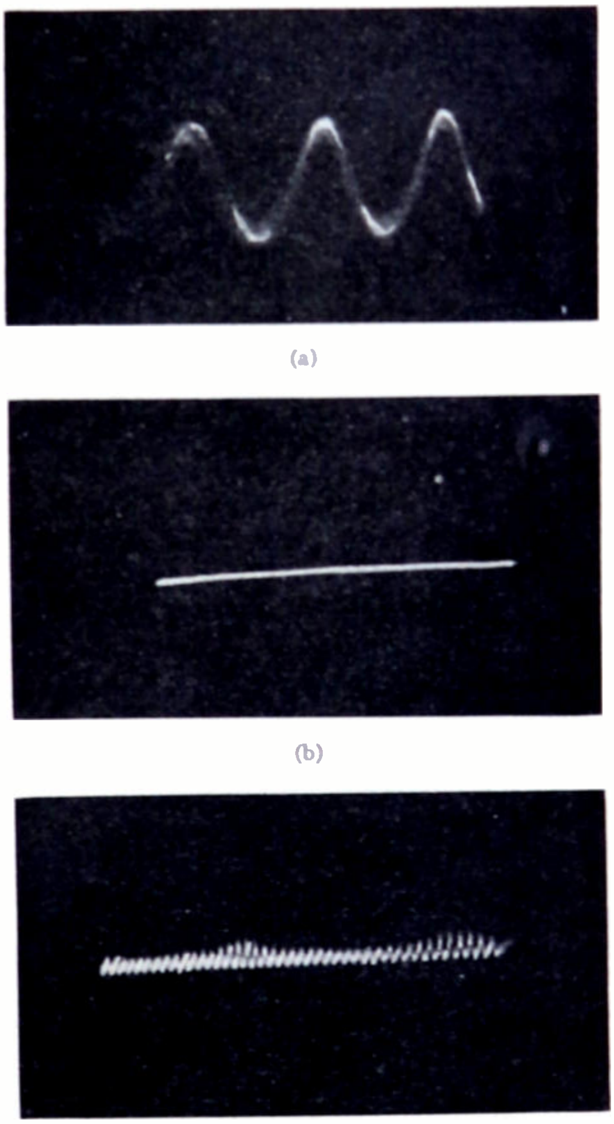


Fig. 2—Impedance match versus temperature.

Such a region was sought and found, but unfortunately the microwave impedance match to the waveguide became worse as the temperature was lowered, as shown in Fig. 2. For these data, the lamp was matched to the waveguide at 32°C and the average reflection coefficient was noted, using a directional coupler in a match meter. This was interpreted in terms of standing-wave ratio for plotting along the ordinate. Efforts to rematch the lamp at the lower temperatures met with no success, which suggested that the impedance was not remaining constant. Using an oscilloscope to observe the reflection coefficient as a function of time, the patterns shown in the photographs of Fig. 3 were obtained. Fig.



(a)

(b)

(c)

Fig. 3—Reflection coefficient versus time displayed on an oscilloscope. (a) temp = 14°C, 1,500∼sweep, avg w 10.3db, avg SWR—5.7db, IF atten 25db. (b) temp = 31°C, 1,500∼sweep, avg w 42.0db, avg SWR—0.14db, IF atten 25db. (c) temp = 31°C, 60∼ sweep, avg w 42.0db, avg SWR—0.14db, IF atten 0db.

[2] H. Johnson and K. R. De Remer, "Gaseous discharge super-high-frequency noise sources," Proc. I.R.E., vol. 39, pp. 908–914; August, 1951.

3(a) shows that at 14°C the reflection coefficient varied with time cyclicly at a frequency of a few kilocycles. Figs. 3(b) and (c) show that at 31°C these effects were reduced. In Fig. 3(c) the receiver gain was 25 db greater than in Figs. 3(b) and (a), revealing that the oscillations were still present, though negligible. From the data of Figs. 2 and 3, it was concluded that if a good match was desired, operation of the lamp should be confined to ambient temperatures above 28°C, measured at the waveguide.

In the course of these measurements, evidence of considerable time lag was observed in the mercury thermometer indication, and it was thought that this could be reduced if the temperature at the lamp itself could be measured. Several means for measuring temperatures are possible. A thermocouple would involve the maintenance of some kind of constant temperature bath. Mechanically indicating thermometers would be too bulky and would lack the speed and accuracy desired. Since a resistance thermometer has many attractive features, it was tried. Fig. 4 shows a temperature

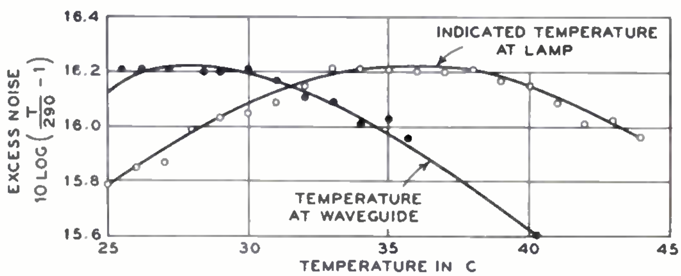


Fig. 4—Excess noise output versus temperature.

run on one lamp as the ambient temperature was changed. One curve is for the mercury thermometer mounted on the exterior waveguide; the other is for the resistance thermometer mounted directly on the lamp. The predicted maximum in the excess noise occurred when the waveguide temperature was about 28°C, at which time the lamp temperature was 36°C. Evidently the temperature difference between the lamp bulb and the waveguide was about 8 degrees.

The resistance thermometer consisted of a bridge using WE106A 60-ohm resistors in three arms and a temperature-sensitive element in the fourth. The 106A resistors are relatively insensitive to temperature, the coefficient being 0.00017 per cent per degree C at 20°C, whereas the Driver Harris number 99 alloy used in the fourth arm had a coefficient of 0.006 per cent per degree C at 20° C. This was wound on a thin paper tube which could be slipped over the T5 lamp at the anode end, and placed as near the active waveguide as possible without disturbing the microwave field. The resistance of this winding was trimmed to balance the bridge at 20°C, and the upper limit of 60°C was set by the voltage applied across the bridge. Calibration of the bridge and the indicator, a 0 to 200 μa 50-ohm meter, Fig. 5.

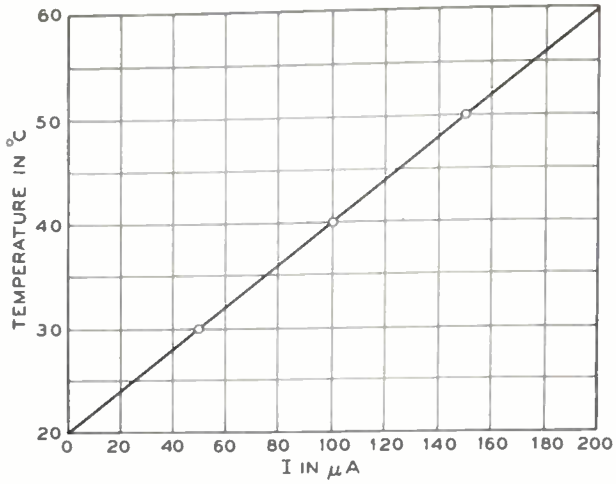


Fig. 5—Calibration of resistance thermometer.

Feeling that means were now available for measuring the lamp temperature quickly and accurately, ten lamps were measured in the high-frequency head. The excess noise versus the indicated lamp temperature for each of these is plotted in Fig. 6, which shows that each of the lamps exhibited evidence of the existence of an operating temperature where the coefficient was substantially zero, and that the spread among the lamps increased as the temperature increased, being ± 0.15 db at 26 degrees and ± 0.45 db at 45°C. The region near a lamp temperature of 32°C is thus interesting for

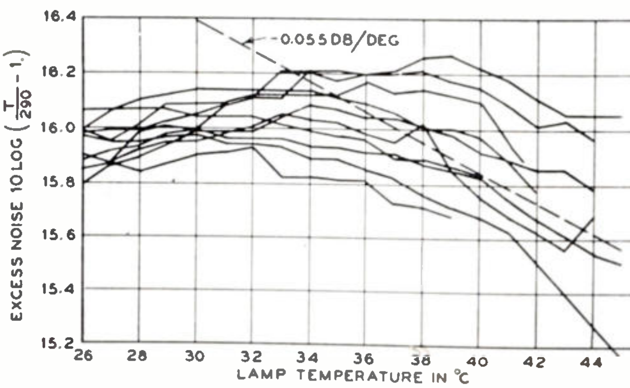


Fig. 6—Excess noise versus temperature for ten lamps.

two reasons; not only is the temperature coefficient negligible, but also the spread among different lamps is small. Unfortunately, this occurs at a temperature where the oscillation in the match may be bothersome, i.e., when the lamp temperature is 32°C (waveguide temperature around 24°C). According to the data of Fig. 2, the average SWR might be about 3.5 db, an intolerable value in some instances.

Thus it appears that the useful region should be confined to lamp temperatures above about 36°C, as indicated in Fig. 6. This corresponded to a waveguide temperature of about 28°C on Fig. 2, above which the SWR was less than a db. The temperature coefficient of -0.055 db per degree C seems to fit fairly well most of the lamps tested, and the spread among the lamps was about ±0.3 db, as found previously. Accordingly, the temperature-measuring bridge was designed to include the region above 36°C.

A representative lamp was chosen from the lot and installed in a circuit. Its excess noise was measured over a temperature range of 29 to 46°C, and is given by the data of Fig. 7. A hand-calibrated scale was pasted on the face of the indicating meter of the resistance thermometer, thereby giving a direct-reading scale for excess noise power output in db.

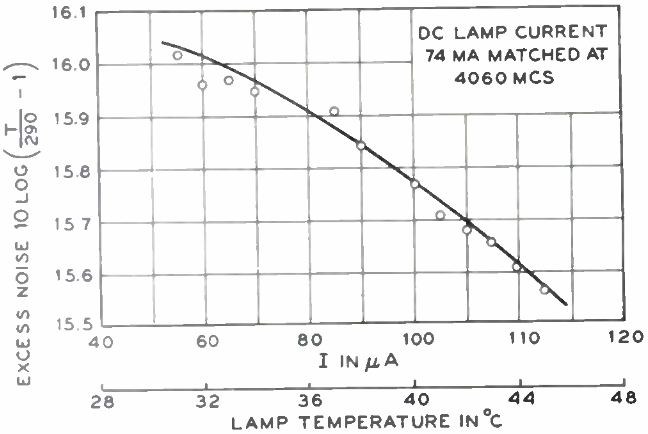


Fig. 7—Excess noise output versus temperature of the completed unit.

A conventional power supply mounting the resistance thermometer indicating meter, and a dc lamp current indicating meter, completed the unit. Switches were provided (SW 3, 4, 5, 6) to include any or all of a bank of fixed resistors in series with a fine trimming resistance R_8, so that the lamp current could be set at any value from 50 to 100 ma, the limits required to match the lamp to the waveguide over the range from 3,700 to 4,500 mc. It might be pointed out here that if the lamp is matched at the mid-band, the SWR will not exceed 3 db over the band, and this is sufficiently good for many measurements. The resistance thermometer

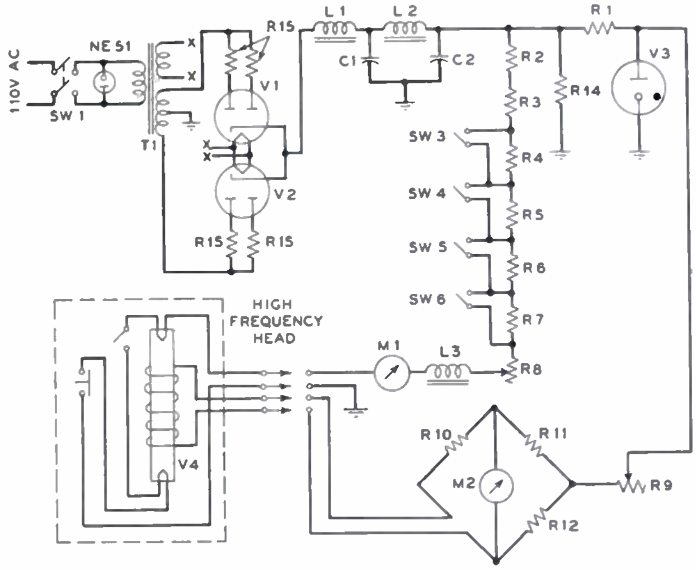


Fig. 8—Schematic of power supply and resistance thermometer

C_1C_2—Dual 40 μf 450-v electrolytic cond.
R_1—Two 3K ohm 2 w resistors
R_2R_3—630 ohm 10 w resistors
$R_{4,5,6,7}$—400 ohm 10 w resistors
R_8—500 ohm 25 w potentiometer
R_9—20K ohm potentiometer
$R_{10,11,12}$—W.E. Co. 60-ohm type 106A resistors
R_{13}—Driver Harris alloy #99 0.002" dia. wound on paper tube to approx. 60 ohms (see text).
R_{14}—100K ohm 1 w resistor
R_{15}—10 ohm 2 w resistors
$SW_{1,3,4,5,6}$—SPST toggle switches
SW_2—Fluorescent push switch 1 NC contact momentary break, one NO momentary make contact.
L_1L_2—18 henry 100 ma chokes
L_3—6 w fluorescent lamp ballast
M_1—0–100 ma meter
M_2—0–200 microampere 50 ohm meter
PJ—Plug and jack connector
V_1V_2—6×4 vacuum tubes
V_3—OB2 regulator tube
V_4—6 w daylight fluorescent lamp
T_1—Power transformer 700v CT 90 ma 5v 3a 6.3v 3.5a

bridge derives its power from the same supply, with a gas tube, V_3, acting as a regulator. Fig. 8 is a schematic diagram of the unit and a listing of the parts. Fig. 9 is a photograph of the completed unit.

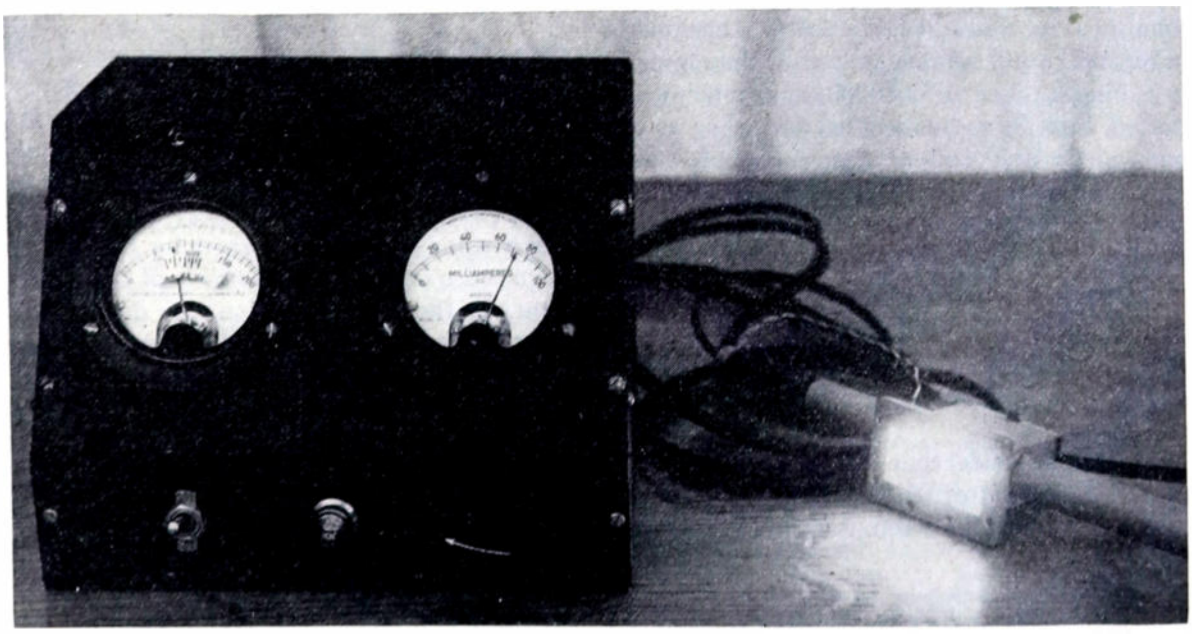


Fig. 9—Front view of completed unit with lamp lighted.

The completed direct-reading-microwave noise generator unit has been in use in the laboratory for some time, and has proven to be a useful tool for the measurement of microwave noise figures on various equipments. The ease of operation and stability of this portable unit have been both time saving and gratifying.

Discussion on "The Permittivity of Air at a Wavelength of 10 Centimeters"*

W. ERIC PHILLIPS

C. M. Crain:[1] The article on permittivity of air and water vapor at 10 centimeters by Mr. Phillips, published in the July, 1950, PROCEEDINGS, was indeed interesting as it illustrates the world-wide interest in determining these important quantities.

There are a few points in the article, however, which I feel warrant comment. Mr. Phillips states that the published results of permittivity of dry air are confined to frequencies of the order of 50 megacycles. In the last few years at least two articles have been published on measured values of dielectric constants of dry air or its constituents in the neighborhood of 3.2 centimeters. Birnbaum at the Bureau of Standards in Washington, (1.) published measured values of dielectric constants of gases in this region. The Electrical Engineering Research Laboratory at The University of Texas (2.) has published measured values for the dielectric constants of both dry air and water vapor at 9,340 megacycles. Their measurements were obtained by heterodyning two stabilized Pound oscillators (3.).

Equation (3) in Mr. Phillips' article is apparently correct only if the radius of the resonant cavity remains constant. If I interpret Fig. 4 correctly, the outside wall of the resonant cavity was still subjected to atmospheric pressure when the cavity was evacuated. If the radius of the copper tubing is taken as 5.080000 centimeters when it is evacuated, then its radius will be, according to my calculations, neglecting end effects, 5.080068 centimeters when the pressure inside is one atmosphere. Then equation (3) for air should be

$$k_e = \frac{\dfrac{1}{\lambda_{ta}^2} + \left(\dfrac{r_{mn}}{2\pi a_a}\right)^2}{\dfrac{1}{\lambda_{tv}^2} + \left(\dfrac{r_{mn}}{2\pi a_v}\right)^2} = \frac{\dfrac{1}{\lambda_{ta}^2} + 0.005677137}{\dfrac{1}{\lambda_{tv}^2} + 0.005677344}.$$

It is important to note that the constants in the numerator and denominator of the above equation are *different*. The difference between 0.005677344 in the above equation, and 0.00567695, as used by Mr. Phillips, is of negligible consequence. If the above relation is used, for example, to calculate k_e for dry air in series 11a and 11b, page 789, one gets $k_e = 1.000552$, instead of 1.000573. Hence, it appears that the value for k_e of dry air at 0°C and 760 mm Hg, on page 790, should be $1.00058_6 \pm 0.00002_3$. This value compares with 1.000577 in reference (1.) and 1.000572 in reference (2.).

There is apparently a misprint in equation (4). The value 2π is obviously not correct. Using k_e, as above, equation (4) would read

$$k_e - 1 = \frac{210}{T}\left(P + \frac{48P_e}{T} H\right)10^{-6}.$$

It is not clear how Mr. Phillips was able to quote a value for k_e for water vapor at 100° C on the basis of his measurements at 23° C and the Clausius-Mosotti relation. Water vapor is a polar molecule, and hence has a dielectric constant which is approximately related to temperature as follows:

$$k_e - 1 = \frac{A}{T} + \frac{B}{T^2},$$

where A and B are constants which may be determined by measuring k_e at two different temperatures. It is possible, however, using the Debye equation (4), to calculate the value of k_e at 100° C from the measured value at 23° C if one assumes either values quoted by other observers for the dipole moment of the water molecule, or uses a value from the literature for k_e of water vapor at frequencies in the infrared region of the frequency spectrum. In the infrared region, the dipole moment does not contribute to the dielectric constant; hence, $k_e - 1$ can be taken as simply A/T.

* W. E. Phillips, "The permittivity of air at a wavelength of 10 centimeters," PROC. I.R.E., vol. 38, pp. 786–790; July, 1950.

[1] Department of Electrical Engineering, The University of Texas, Austin 12, Texas.

BIBLIOGRAPHY

1. H. Lyons, G. Birnbaum, and R. Kryder, "Measurement of complex dielectric constant of gases at microwaves," *Phys. Rev.*, vol. 74, p. 104; 1948.
2. C. M. Crain, "The dielectric constant of several gases at a wavelength of 3.2 centimeters," *Phys. Rev.*, vol. 74, p. 691; September, 1948.
3. R. V. Pound, "Electronic frequency stabilization of microwave oscillators," *Rev. Sci. Instr.*, vol. 17, p. 490; 1946.
4. S. Glasstone, "Textbook of Physical Chemistry," D. Van Nostrand Co., Inc., New York, N. Y., p. 545; 1946.

A General-Purpose Electronic Wattmeter*

DON E. GARRETT†, ASSOCIATE, IRE AND FRANK G. COLE†, ASSOCIATE, IRE

Summary—**A direct reading wattmeter is described which will read either positive or negative peak or average power of complex voltage and current waveforms containing components in the frequency range from dc to 71 kc. The instrument will measure up to 50 watts, covered in three ranges with an accuracy of 3 per cent. The scale may be extended by suitable resistance changes. The voltage and current inputs are mutually dc isolated to allow measurement of the current at any point between the ground and the voltage source.**

Multiplication of the current and voltage is accomplished by modulating a 10-mc carrier with the current, and then modulating the resultant with the voltage in a cascade of two suppressed-carrier modulators. The instantaneous product is recovered in a phase-sensitive demodulator, and then passed through an integrating network for average power, or through a peak detector for peak power.

The wattmeter was designed with particular reference to the measurement of the plate dissipation in television horizontal output tubes. It was designed, nevertheless, to be a general-purpose instrument capable of accurately measuring power in any type of circuit where the voltage and current involved can be measured.

INTRODUCTION

THE MEASUREMENT of the plate dissipation of the horizontal-sweep output tube in television receivers has, in the past, been a nebulous problem. Economics require receiver manufacturers to employ small tubes operating near maximum capacity. Consequently, it is necessary to determine the actual plate dissipation of power-amplifier tubes in some manner to avoid epidemics of tube failure. Also it is desirable to have the method compatible with design procedures where the effect of a circuit change or adjustment can be noted immediately. The power measurement is made difficult by the complexity of plate voltage and current waveforms. It is further complicated by the voltage crest of four kv or more during the retrace period, while the average voltage during current conduction is perhaps only a hundred volts.

Many methods have been devised to measure the plate dissipation of the horizontal output tube. These can be divided roughly into tube-envelope temperature measurements and voltage-and-current measurements.

Power measurements made by the temperature methods consist of noting the bulb temperature during the actual operating conditions and then reproducing the same temperature later with a dc plate voltage and current. Plate dissipation is then given by the product of these dc values. Screen and filament dissipation, of course, must be held constant.

A relatively crude method that has been used consists of painting stripes of wax, each with different melting points, on the bulb and then noting which of these melt during operation. Other methods have included a resistance wire bridge, a thermocouple, and an oil-bath calorimeter. Of these, the oil-bath calorimeter is perhaps the most accurate but the most inconvenient. Temperature methods are not popular, however, because of the time required to make temperature measurements.

Heretofore, power measurements made by the voltage and current methods have been indirect and less accurate than the temperature methods because of the complex nature of the waveforms involved. Independent readings of average plate voltage and current cannot be employed to yield the average plate dissipation because true average power is defined as the time average of the instantaneous product of current and voltage. It is possible to form the product graphically from oscilloscope traces of the voltage and current, but this procedure is slow and not very accurate because the high surge voltage during retrace obscures the voltage during actual plate-current conduction.

An indirect method utilizing voltage and current measurements has been devised by C. E. Torsch[1] and had been used by the Receiver Department of the General Electric Company in the design of horizontal output stages. By this method, a pessimistic though approximate average plate voltage during conduction is found by making certain measurements at the secondary of the output transformer. Admittedly, the method is not theoretically accurate, but it does give a slightly conservative comparative dissipation figure in a relatively short time.

This review indicates that direct-reading accurate wattmeter would greatly facilitate power measurements in television circuits. The requirements for the instrument are

(a) complex waveforms must be handled,
(b) dc components of the inputs must be preserved,
(c) separate dc reference levels for the current and voltage inputs must be allowed so that the current can be measured in the plate circuit of tubes to avoid the screen current, and
(d) frequencies from dc to several hundred kc should be passed so that power contained in the harmonics may be measured.

None of the electronic wattmeters discussed in the literature[2] satisfy all the requirements simultaneously.

* Decimal classification: R245.3. Original manuscript received by the Institute, February 8, 1951; revised manuscript received, August 6, 1951.

† Receiver Department, General Electric Company, Electronics Park, Syracuse, New York.

[1] C. E. Torsch, "A universal-application cathode ray sweep transformer with ceramic iron core," *Tele-Tech*, vol. 9; January 1950.

[2] J. R. Pierce, "A proposed wattmeter using multielectrode tubes," PROC. I.R.E., p. 1743; October, 1930. G. Dexter, "Universal af-rf wattmeter," *Radio News*, p. 66; May, 1948. L. R. Malling, "Electronic wattmeter," *Electronics*, p. 133; November, 1945. E. Mittelmany, "High-frequency wattmeter," *Electronics*, p. 324; March, 1945. R. Bauch, "Hot-wire wattmeter," *Elektrotech. Z.*, p. 530; July 9, 1903. G. H. Brown, J. Epstein, and D. W. Peterson, "Direct-reading wattmeters for use at radio frequencies," PROC. I.R.E., p. 403; August, 1943. H. M. Turner and F. T. McNamara,

Their major shortcoming is that they cannot operate with separate dc reference or ground points for the voltage and current inputs and still retain the dc component. The difficulty may be overcome if the voltage and current information is completely contained in the envelope and phase of a carrier signal. In this case, capacitor or transformer coupling can be used, thereby allowing dc isolation with the dc component of the inputs preserved. The wattmeter described in this paper is based upon this principle.

Circuit Description

The block diagram of Fig. 1 shows how the watt-meter makes use of carrier modulation to form the current and voltage product.

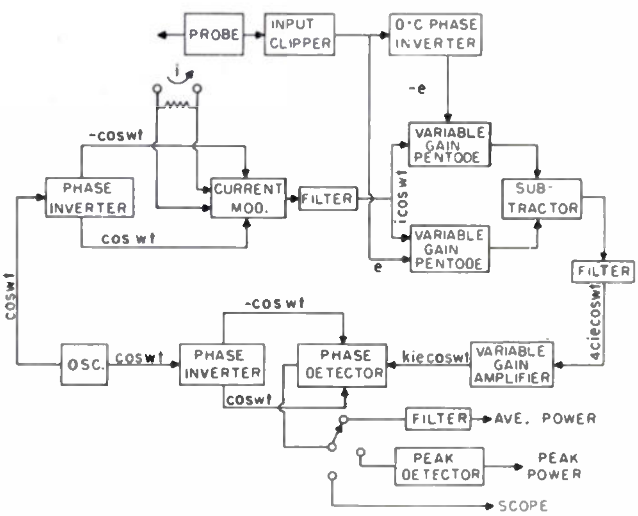


Fig. 1—Block diagram of electronic wattmeter.

Fundamentally, in the wattmeter described, a carrier is modulated with the current waveform and then again modulated with the voltage waveform. The result is demodulated to yield the instantaneous voltage and current product. The average power is found by passing the instantaneous product through a low-pass filter, and the peak power is found by passing the product through a peak detector.

Current Modulator

The object of the current modulator is to modulate the carrier frequency with a voltage proportional to the instantaneous current. If, for example, the carrier is defined by ($\cos wt$) and the current by ($I \cos qt$), then the output from the current modulator should be ($I \cos qt \cos wt$). This, by identity, is equal to $1/2I[\cos (w+q)t + \cos (w-q)t]$, and may be seen to contain only the side-bands with no carrier component. Consequently, the current and voltage modulators must be balanced or suppressed-carrier modulators.

One of the basic requirements for the wattmeter is that an arbitrary dc voltage may exist between the current and voltage inputs. This requirement necessitates dc isolation between the modulators. The type of suppressed-carrier modulator used as the current modulator has diodes as the nonlinear element,[3] and may be dc isolated to the extent of the coupling capacitor ratings if crystals are used, or of the filament-cathode ratings if vacuum diodes are used.

The current modulator is, then, a sampling device for measuring the amplitude and polarity of the current input at the carrier-frequency rate. The circuit consists of two biased diodes, shown as V-4 and V-5 in the complete schematic Fig. 2, with the plate of one connected to the cathode of the other, and with the carrier being applied push-pull to the remaining cathode and plate. With the current terminals shorted and with equal amplitude push-pull inputs, each diode will conduct the same, giving a zero net voltage across the tuned-circuit filter L_3. A balance adjustment is provided so that the carrier inputs may be matched, thereby insuring complete voltage cancellation across L_3. A voltage, determined by the current passing through the current-sampling resistor, is applied across both diodes. This voltage biases the diodes so that more of one phase of the carrier reaches the common connection that arrives from the other phase. The net carrier at this point is, therefore, proportional to the current with its phase being determined by the polarity of the current bias voltage. The peak carrier amplitude must always be greater than the bias voltage if its output amplitude is to be determined by the current. If the current frequency is much lower than that of the carrier, the net voltage appearing at the common terminal of the diodes is approximately a square wave defined by the expression

$$e = Ki[1/2 + 4/\pi(\cos wt - 1/2 \cos 3wt + 1/5 \cos 5wt + \cdots)], \quad (1)$$

where i is the instantaneous current value and w the carrier angular frequency. The tuned filter, L_3, removes the harmonics, leaving

$$e = Ki \cos wt, \quad (2)$$

where K is the constant of proportionality.

Crystal diodes are used to avoid filament isolation problems and subsequent capacitive coupling between the cathode and filament circuits. In addition, with the use of crystals, the entire current modulator is immune to supply-voltage changes, thereby providing maximum stability. It has been found, however, that crystal diodes are sensitive to temperature changes, and occasional balance checks must be made.

The size of the current-sampling resistor determines the input voltage for a given current and can, therefore, be used to determine the wattage range. In the watt-

"An electron tube wattmeter and voltmeter, and a phase-shifting bridge," Proc. I.R.E., p. 1743; October, 1930. E. Mallett, "A valve wattmeter," *Jour. IEE*, vol. 73, p. 295; 1933. M. A. H. El-Said, "Novel multiplying circuits with application to electronic wattmeters," Proc. I.R.E., p. 1003; September, 1949.

[3] B. Chance, V. Hughes, E. F. MacNichol, D. Sayre, and F. C. Williams, "Waveforms," McGraw-Hill Book Co., New York, N. Y. Chapt. 11, p. 389; 1949.

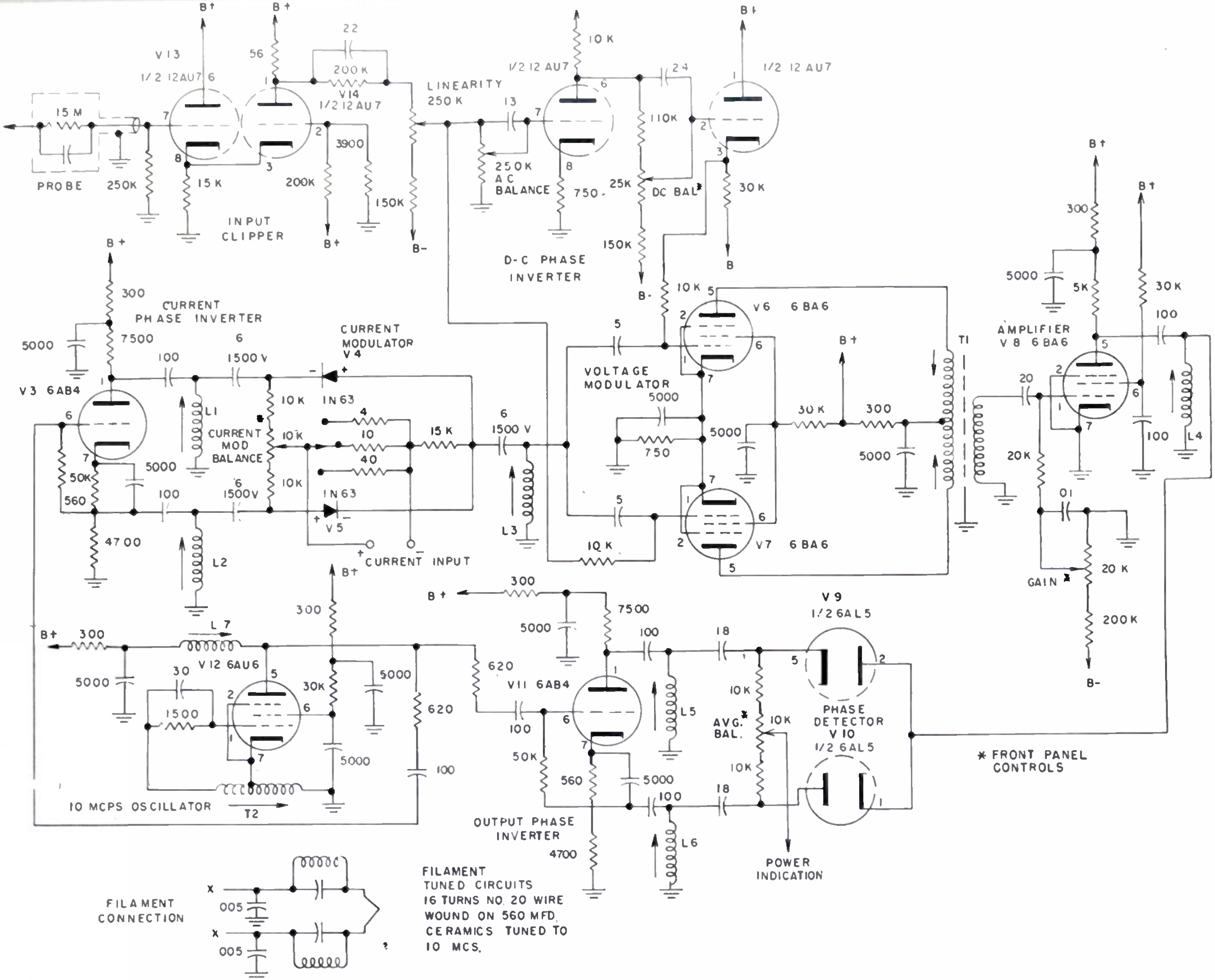


Fig. 2—Circuit schematic of electronic wattmeter.

meter described, three ranges are provided by switching in either a 4-ohm, 10-ohm, or 40-ohm resistor.

Voltage Modulator

The next problem is to remodulate the above signal with the voltage input. The use of a second diode modulator is precluded since the output amplitude of this type of modulator is dependent upon the smaller input voltage. However, a satisfactory circuit is obtained with two variable-gain tubes connected so that their plate current subtracts and their grid voltages are, respectively,

$$e_{g1} = i \cos wt + e \tag{3}$$

and

$$e_{g2} = i \cos wt - e, \tag{4}$$

where e is the voltage waveform.[4] The plate current of these tubes, over a small range, may be expressed by the power series

$$ip = a + be_g + ce_g^2 + de_g^3 + \cdots . \tag{5}$$

Substituting the grid voltages of (3) and (4) into the power series of (5) and then subtracting, one obtains

$$i_{p1} - i_{p2} = 2be + 2de^3 + 4cie \cos wt + 2di^2e \cos^2 wt \tag{6}$$

for the first four terms of the power series. The subsequent tuned circuits remove all components except the carrier-frequency term, leaving

$$i_{p1} - i_{p2} = 4cie \cos wt. \tag{7}$$

The constants a, b, c, and d may be matched over the operating range by adjustment of the relative dc bias and the ac grid drive on the variable-gain pentodes, V-6 and V-7, Fig. 2. It was originally thought that the voltage-modulator pentodes would have to be carefully matched. Under actual operation, several randomly selected tubes operated as linearly as matched tubes when the individual quiescent bias was properly adjusted.

Voltage-Input Circuit

Push-pull voltage inputs are required by the voltages modulator. A direct-coupled phase inverter is used to provide the push-pull voltages since the instrument must preserve the dc component. The voltage is sampled through a capacitively compensated high-impedance probe having an input impedance of 15 megohms and 1.5 $\mu\mu$f. Other probes having a different impedance in conjunction with various terminating circuits can extend the voltage range of the wattmeter to any desired value.

For accurate measurement of the plate dissipation of television horizontal output tubes, the high surge during retrace is removed by a clipper. Removal of the surge allows increased sensitivity over the active portion of the cycle. The validity of surge voltage clipping is dependent upon no plate-current flow during the high-voltage peak. To eliminate measurement errors here, the grid drive on the horizontal output tube must cut off the plate current instantaneously. If the cut-off voltage is sluggish or of insufficient amplitude, then certainly some current will flow during the plate voltage pulse, contributing a small error to the wattage reading.

Amplifier

The voltage modulator output is passed through an adjustable-gain amplifier, V-8 Fig. 2, so that the wattage indication may be varied for calibration purposes, and to compensate for tube aging and drift. Although little error has been found to be associated with this variable-mu tube amplifier, let us determine the spurious modulation products generated. The first four terms of the power series of (5) indicate an output voltage from V_8, when filtered by the plate circuit, of

$$e = K\left[bie + 3/4dk^2(ie)^3\right] \cos wt. \tag{8}$$

The spurious term introduced by intermodulation is $3/4\, dk^2(ie)^3$, where d is the third-order constant defined by (5). The third-order curvature of the tubes characteristic is very small compared to the linear term, and, in addition, the signal input to this stage is only a few mv. Consequently, the error introduced should not be the limiting feature of the wattmeter.

Demodulator

The remaining problem is to demodulate or to recover the carrier modulation defined by (7). The demodulator must be more than an amplitude detector since the phase of the carrier contains the power polarity information. Therefore, the demodulator must be both amplitude and phase sensitive. This characteristic is exhibited by the common diode-type phase detector.[5] The modification of the conventional phase detector used allows ground reference for both the carrier and signal inputs. The circuit configuration is exactly the same as the current modulator

The diode demodulator can best be explained by examining the voltages involved. The diodes, V-9 and V-10, Fig. 3, in combination with their associated RC circuits, comprise separate peak detectors. Consequently, the capacitors will charge up to the peak voltage across V-9 and V-10, respectively. Let the peak amplitude of the push-pull carrier inputs be E, and the peak amplitude of the modulated input be proportional to the instantaneous modulation (ie). The latter stipulation will be true if the modulating frequency is much lower than the carrier frequency. The peak voltage across V-9 and V-10 will be $(E+ie)$ and $(E-ie)$, respectively. The voltage at the output terminal will be one-half the difference of the voltage on each charg-

[4] W. L. Everitt, "Communication Engineering," McGraw-Hill Book Co., New York, N. Y., 2nd ed., p. 390; 1937.

[5] W. L. Emery, "Ultra-High-Frequency Radio Engineering," Macmillan Co., New York, N. Y., p. 41; 1944.

ing capacitor, giving a demodulated output of the instantaneous product (*ie*). With no input from the amplifier *V*-8, the voltage at the output terminal should be zero. A balance adjustment is provided so that the output will be zero in this circumstance.

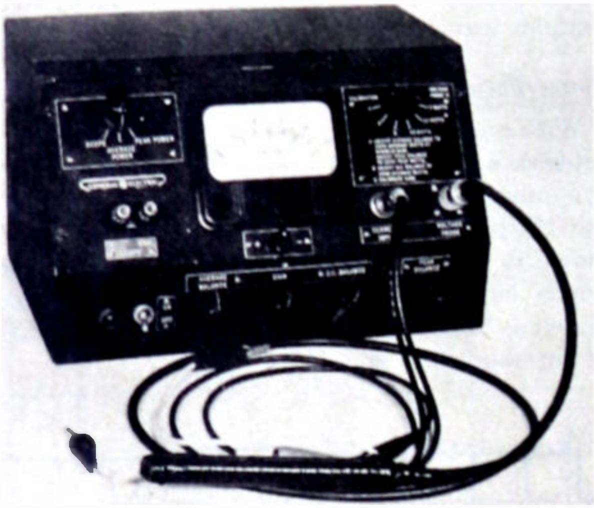


Fig. 3—Direct reading electronic wattmeter for complex waveforms, front view.

Output Filters

Three circuits are provided at the output of the phase detector. One is simply a resistance termination through which the power waveform may be observed on an oscilloscope. The second is an integrating circuit, or a low-pass filter, which averages the power over the cycle, thus giving the average power. The third is a peak detector which measures the peak power during cycle. The meter at the output is calibrated directly in watts.

Calibration

Calibration of the meter is accomplished by applying a known voltage to the current and voltage terminals. The resulting power indication may be calculated and the gain control adjusted for this reading on the meter at the output. In the instrument described, a fixed voltage is applied internally and the gain control is adjusted until the wattage indication coincides with a mark on the meter scale.

Stability

It is interesting to note that the stability of the carrier-frequency oscillator is not important providing the instrument is calibrated prior to using. Should the oscillator drift, the tuned circuit response results in reduced amplitude and introduces a phase shift in the carrier. The reduced amplitude may be directly compensated for by a suitable gain adjustment. The effects of the phase shift are not as obvious. However, since the output stage is a phase detector, the power indication will vary as the cosine of the phase-shift angle. This again merely produces a reduced sensitivity, and may be compensated for by a suitable gain adjustment. It was necessary to load down the tuned circuits so that the high-frequency components of the voltage and current inputs would not be attenuated; therefore, the tuned-circuit Q is relatively low. Consequently, a frequency shift in the oscillator will have only a small effect.

Stability is also affected by power-supply variations since the voltage input-circuit is direct coupled. To avoid as much instability as possible from this source, a well regulated power supply is required.

Shielding and Layout

Perhaps the most important consideration in the construction of the wattmeter is the circuit layout. When either the voltage or current input is zero, the wattage indication must be zero for all values of the nonzero input. To realize this requirement, either absolutely no feed through of the carrier can be tolerated, or the feed through must be equal but out of phase. For best results an attempt was made to both minimize and balance all cross talk. This was accomplished by symmetrical layout and extensive shielding. The shielding is apparent in the bottom view of the chassis, Fig. 4.

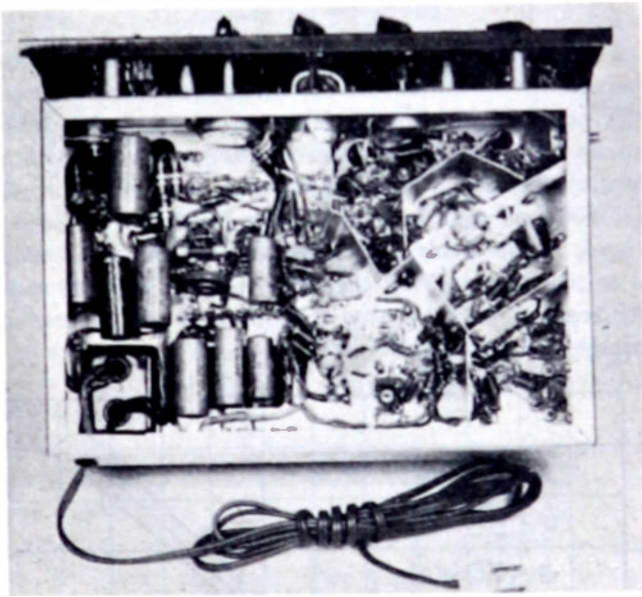


Fig. 4—Bottom view of electronic wattmeter.

Performance

The salient performance characteristics of the wattmeter that are of interest are the calibration, error, frequency response, and stability. Of primary interest, of course, is actual measurements on television sets.

DC Calibration

The dc calibration curves are shown on Fig. 5 (see following page) for various values of current and voltage input. The indication error may be seen to be within

3 per cent of full scale for all inputs within the range of the instrument. When the polarity of either input is reversed, the reading is duplicated in the negative direction.

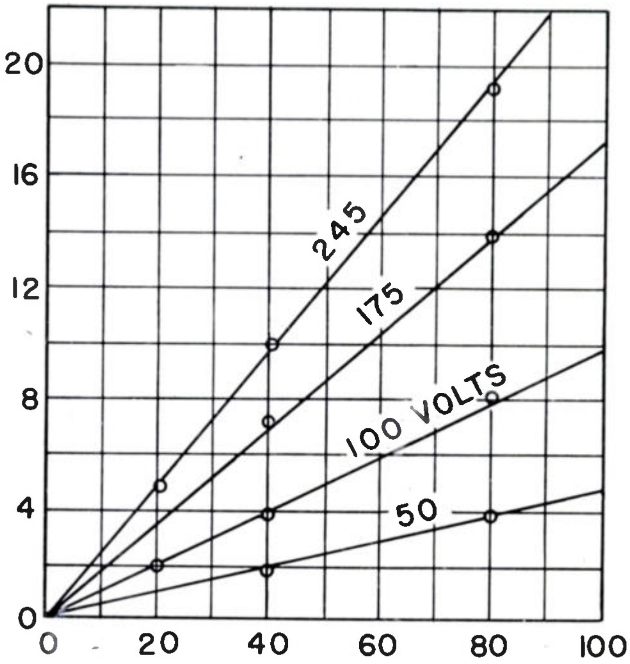


Fig. 5—DC performance.

A desirable feature of this wattmeter is that the unit operates at the same level on all ranges since the range selector is at the input. The maximum voltage that can be handled without overloading the voltage input circuit is 245 volts. The maximum current inputs are 100 ma, 400 ma, and 1 amp for the 5-, 20-, and 50-watt ranges, respectively. Perhaps an improvement would be to provide external terminals for the current sampling resistor so that any combination of current and voltage could be handled.

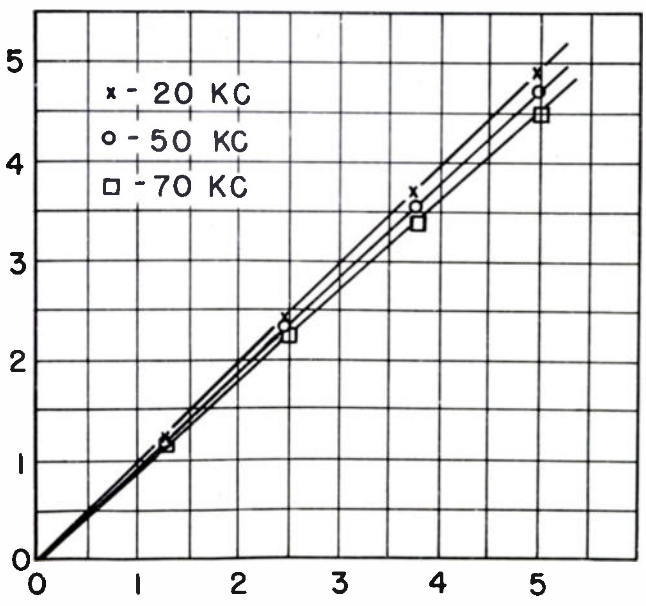


Fig. 6—Square-wave inputs.

Square-Wave Test

Fig. 6 shows the power indication resulting from a negative-going, square-wave voltage input and a dc current input. The calculated power, as shown in the figure, agrees well with that measured over the applicable frequency range of the wattmeter. This test supplements the dc calibration curves and, in addition, demonstrates the ability of the instrument to handle complex waveforms.

Power-Factor Test

With a 60-cps source, the power consumed by an *RC* network was measured with both this wattmeter and a dynamometer-type wattmeter. For two values of voltage, the power factor was varied from unity to 0.014, as shown in Fig. 7. The resulting indication agreed very closely, as may be seen from the curve. It was necessary, of course, to use a high-quality dynamometer wattmeter as the standard.

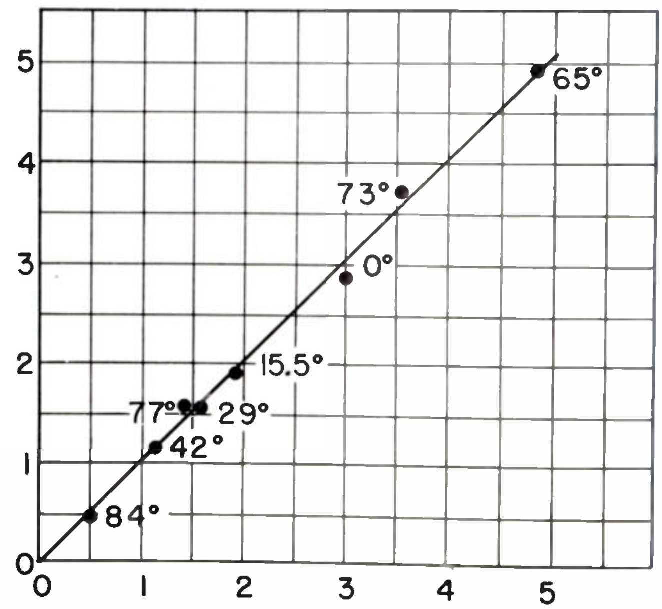


Fig. 7—Power-factor test.

Frequency Responses

The frequency response curves for the wattmeter are shown in Fig. 8. The response was measured with dc current and ac voltage inputs. Since the power was zero, the power waveform was observed on an oscilloscope and the amplitude as a function of frequency was noted. The 0.707 point may be seen to fall at 71 kc. Even though this is below the goal set, it is thought to be satisfactory since the lower-frequency power predominates in the horizontal output tube of television sets. This is true because the plate voltage during current conduction remains practically constant, with the higher-frequency components associated with rise time and high-frequency response contributing very little.

Actually, there are some high-frequency harmonics in the voltage waveform, and these must be passed if a true power reading is to be obtained. Since the higher frequencies are not passed, it is expected that the wattage indication, in television applications, will be slightly low.

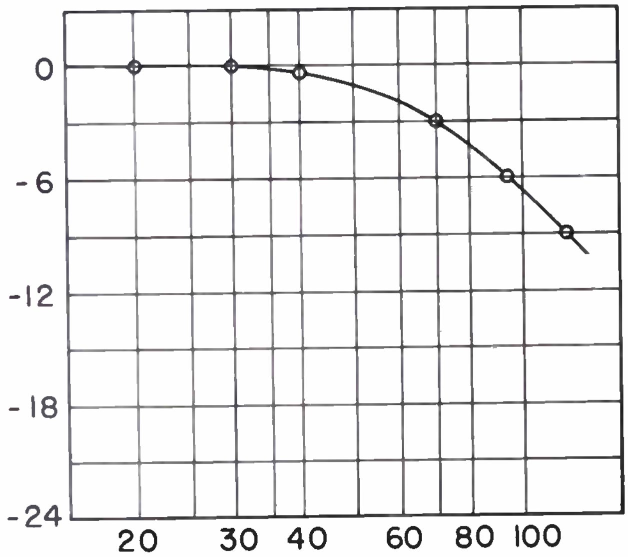


Fig. 8—Frequency response.

Television-Set Testing

The wattmeter has performed satisfactorily with regard to television-set measurements. A comparative check made on a General Electric 16T1 set using the wattmeter, the temperature method, and the approximate method of reference one gave 8.8 watts, 8.92 watts, and 10.5 watts, respectively, of horizontal output-tube plate dissipation. The wattmeter reads slightly less than the temperature method, which is assumed to be the most nearly correct, as would be expected since the wattmeter attenuates some of the higher-frequency components. Actually, the two readings are not significantly different since the wattmeter allowable error is 3 per cent. The approximate method of reference one results in a somewhat larger reading, as may be predicted from an analysis of the method.

A calibrated source simulating the voltage and current of a television horizontal output stage is used to check the probe and wattmeter calibration for pulsed inputs. The calibrator, therefore, insures accurate readings for a television signal.

Drift

There is a gradual drift in the zero setting and calibration during the first hour or so of operation. After a suitable warm-up period the unit stabilizes, and recalibration is necessary again only after a prolonged period of use. Sufficient controls for drift adjustment are brought out on the front panel. These include the voltage modulator dc balance, phase-detector balance, and gain control. A calibration switch is incorporated which terminates the voltage and current inputs so that it is not necessary to disconnect the test leads from the unit being measured when the wattmeter is recalibrated. In addition, the switch applies a calibration voltage to each modulator for zero adjustment, and then applies the voltage to both inputs for gain adjustment.

Conclusions

Let us examine the performance of the electronic wattmeter to see if the initial requirements have been met. The performance stipulations for the instrument were: Complex waveforms must be handled including the dc component; an arbitrary dc voltage must be allowed to exist between the voltage and current inputs; and frequencies from dc to several hundred kc should be passed.

Complex waveform performance was established by square-wave tests, power-factor tests, and actual measurement of the television horizontal output-tube plate dissipation. By square-wave tests and by the dc performance curves, dc operation was also established. The very nature of the circuitry of the wattmeter is sufficient to indicate that a dc voltage may exist between the voltage and current inputs. Even so, while checking the instrument calibration, the current-terminal potential was varied over wide limits with no effect whatsoever on the wattage indication.

The frequency response is below that desired, being only 71 kc. This response is adequate for TV applications, as has been explained. There is no basic reason, however, why this figure cannot be revised upwards to at least 1 mc if a particular application should require this response.

The accuracy is within 3 per cent, which is acceptable in view of the purpose of the wattmeter of providing an instantaneous wattage reading for developmental work on television sets.

Although the unit has been designed to measure television horizontal output tube losses, versatility has been maintained. The wattage range may be varied to suit any individual measurement problem by suitable resistance changes. The wattmeter can be used equally well for measuring miscellaneous circuit losses, such as in the high-voltage transformer, yoke, and vertical circuits. Its peak-reading feature makes it valuable for measuring peak power in pulsed circuits, such as those encountered in radar. In short, it can be used to measure power where any sort of complex waveforms are encountered.

Acknowledgment

The authors wish to acknowledge their indebtedness to Charles E. Torsch of the General Electric Company for his constructive criticisms and for his continued interest, which made this development possible.

Distortion of a Frequency-Modulated Signal by Small Loss and Phase Variations*

F. ASSADOURIAN†

Summary—General formulas are developed for harmonic and total distortion in the frequency of the outputs of linear transmission systems with pure frequency-modulated inputs and with amplitude and phase characteristics involving wiggles that can be represented approximately by single sinusoidal functions of small amplitude. The amplitude wiggles represent departure from flatness, and the phase wiggles departure from linearity. This first-order analysis yields a result for total distortion D, which varies linearly with the amplitude of either wiggle if the amplitude of the other is made zero. It will be seen that D is periodic in the frequencies of the wiggles and in the audio frequency p and carrier frequency ω_c of the frequency-modulated input. Of course, D is also a function of its index of modulation m. The formulas for D can be applied to amplifiers, filters, and the like in a communication system that satisfies the above assumptions.

General distortion formulas are applied to waveguides loaded by pure resistances. Among other things, the so-called "long-line" effect in distortion is discussed. Graphs show the dependence of distortion on various parameters.

I. Sinusoidal Representation of Amplitude and Phase Wiggles

THE ANALYSIS starts with the general linear four-terminal network indicated in Fig. 1. If E_1 is a constant-voltage generator of frequency ω, then the solution for E_2 in terms of E_1 has the form

$$E_2 = G(j\omega)E_1 = A(\omega)e^{-j\phi(\omega)}E_1, \tag{1}$$

where $G(j\omega)$ is dimensionless and is written in polar form. As is well known, nonflatness of $A(\omega)$ and nonlinearity of $\phi(\omega)$ will generally produce harmonic distortion in the frequency of E_2 if E_1 is a pure frequency-modulated wave.

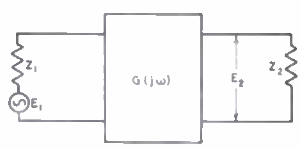


Fig. 1—Linear four-terminal network.

It will be assumed now that $A(\omega)$ and $\phi(\omega)$ have the forms indicated in (2).

$$\left.\begin{aligned} A(\omega) &= 1 + \mathcal{E}_1 \cos 2b(\omega - \omega_1), \\ \phi(\omega) &= c(\omega - \omega_0) + \mathcal{E}_2 \sin 2d(\omega - \omega_2). \end{aligned}\right\} \tag{2}$$

The amplitude and phase characteristics described by (2) are illustrated in Fig. 2. It is assumed that $\mathcal{E}_1$ and $\mathcal{E}_2$ are small and that higher powers of $\mathcal{E}_1$ and $\mathcal{E}_2$ can be disregarded in all steps of the subsequent analysis. It should be noted that (2) applies not only to waveguides, but to other types of networks as well.

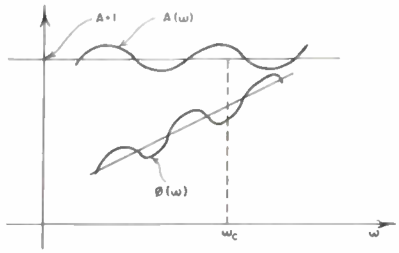


Fig. 2—Amplitude and phase characteristics with sinusoidal variation.

II. Derivation and Discussion of Distortion Formulas

To obtain the distortion in the frequency of E_2 (see Fig. 1) when E_1 is a pure frequency-modulated input, the sideband analysis is followed. E_1 is defined by

$$\begin{aligned} E_1 &= E_0 \sin (\omega_c t + m \sin pt) \\ &= E_0 \sum_{k=-\infty}^{\infty} J_k(m) \sin (\omega_c + kp)t, \end{aligned} \tag{3}$$

where E_0 is constant, ω_c is the carrier frequency, p is the modulating frequency, and m is the index of modulation. The instantaneous frequency Ω in (3) is the derivative of the phase and has the form

$$\Omega = \omega_c + mp \cos pt. \tag{4}$$

Note that mp represents the maximum deviation of the modulating from the carrier frequency.

If each term of the input E_1 is altered suitably with the use of $A(\omega)$ and $\phi(\omega)$ according to the steady-state theory, the output E_2 is given by

$$\left.\begin{aligned} &\frac{E_2}{E_0} = \sum_{-\infty}^{\infty} [1 + \mathcal{E}_1 \cos 2b(\omega_c' + kp)] \\ &J_k(m) \sin [(\omega_c + kp)(t - c) - \mathcal{E}_2 \sin 2d(\omega_c'' + kp) + \omega_0 c], \\ &\omega_c' = \omega_c - \omega_1, \qquad \omega_c'' = \omega_c - \omega_2. \end{aligned}\right\} \tag{5}$$

As expected, the linear term $c\omega$ in the expression for ϕ leads to a delay, but does not produce distortion. It is convenient to replace $t-c$ by τ.

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† Federal Telecommunication Laboratories, Inc., Nutley, N. J.

If (5) is expanded and higher powers of $\mathcal{E}_1$ and $\mathcal{E}_2$ are neglected, the result is

$$E_2/E_0 = (A^2 + B^2)^{1/2} \sin (\omega_c\tau + \omega_0 c + \psi), \quad \tau = t - c, \tag{6}$$

where A, B, and Ψ are defined by

$$\left.\begin{aligned} A &= \sum_{-\infty}^{\infty} J_k(m)\left[\cos kp\tau + \mathcal{E}_1 \cos 2b(\omega_c' + kp) \cos kp\tau \right. \\ &\quad \left. + \mathcal{E}_2 \sin 2d(\omega_c'' + kp) \sin kp\tau\right] \\ B &= \sum_{-\infty}^{\infty} J_k(m)\left[\sin kp\tau + \mathcal{E}_1 \cos 2b(\omega_c' + kp) \sin kp\tau \right. \\ &\quad \left. - \mathcal{E}_2 \sin 2d(\omega_c'' + kp) \cos kp\tau\right] \end{aligned}\right\} \tag{7}$$

$\tan \Psi = B/A$.

Equation (6) indicates that the output voltage E_2 is generally no longer a pure frequency-modulated wave. It now has a new instantaneous frequency

$$\Omega = \omega_c + d\psi/d\tau. \tag{8}$$

It turns out that both ψ and $d\psi/d\tau$ are periodic functions of time with fundamental period $2\pi/p$ and can be expanded in Fourier series; $d\psi/d\tau$ can be expressed in the form

$$\left.\begin{aligned} \frac{1}{p}\frac{d\psi}{d\tau} &= m \cos p\tau + \mathcal{E}_1 \sum_{k=-\infty}^{\infty} kJ_k(2m \sin pb) \cos \\ &\quad \left[k(pb + \pi/2) + 2b\omega_c'\right] \cos kp\tau + \mathcal{E}_2 \sum_{k=-\infty}^{\infty} kJ_k \\ &\quad (2m \sin pd) \sin \left[k(pd + \pi/2) + 2d\omega_c''\right] \sin kp\tau. \end{aligned}\right\} \tag{9}$$

The last result is basic in distortion calculations. Higher-harmonic terms beyond the first represent distortion in the original modulation. It should be recalled that (9) is a first-order result that is applicable to any network for which $A(\omega)$ and $\phi(\omega)$ have the forms indicated in (2) and Fig. 2. Even if $A(\omega)$ and $\phi(\omega)$ are approximately given by (2) for a finite frequency range that includes the essential band of E_1, (9) may still be used.

For distortion calculations, (9) may be written in the form

$$\left.\begin{aligned} \frac{1}{p}\frac{d\psi}{d\tau} &= m \cos \theta + \sum_{k=1}^{\infty} k(\mathcal{E}_1 A_k \cos k\theta \\ &\qquad + \mathcal{E}_2 B_k \sin k\theta), \quad \theta = p\tau, \\ A_k &= J_k(u)\left[\cos (kb_1 + b_2) \right. \\ &\qquad \left. + (-1)^{k+1} \cos (-kb_1 + b_2)\right], \\ B_k &= J_k(v)\left[\sin (kd_1 + d_2) \right. \\ &\qquad \left. + (-1)^k \sin (-kd_1 + d_2)\right], \\ u &= 2m \sin pb, \qquad v = 2m \sin pd, \\ b_1 &= pb + (\pi/2), \qquad b_2 = 2b(\omega_c - \omega_1), \\ d_1 &= pd + (\pi/2), \qquad d_2 = 2d(\omega_c - \omega_2). \end{aligned}\right\} \tag{10}$$

One can put (10) in the final form

$$\begin{aligned} \frac{1}{p}\frac{d\psi}{d\tau} &= m \cos \theta + \sum_{2}^{\infty} kC_k \sin (k\theta + \phi_k), \\ C_k^2 &= \mathcal{E}_1^2 A_k^2 + \mathcal{E}_2^2 B_k^2, \end{aligned} \tag{11}$$

where the first term of the sum in (10) has been omitted because it leads to higher-order terms in the distortion.

The definitions of distortion used in this report are given by

$$\begin{aligned} &k\text{th-harmonic distortion} = D_k = kC_k/m, \\ &\text{total distortion } D = \left(\sum_{2}^{\infty} k^2 C_k^2\right)^{1/2} \Big/ m. \end{aligned} \tag{12}$$

An expression for harmonic distortion D_k can be obtained from (10) and (11).

If D_A is the total distortion due to amplitude wiggles alone ($\mathcal{E}_2 = 0$) and D_ϕ is the total distortion due to phase wiggles alone ($\mathcal{E}_1 = 0$), then one can write

$$D^2 = D_A^2 + D_\phi^2, \qquad D_k^2 = D_{kA}^2 + D_{k\phi}^2. \tag{13}$$

It is evident from (13) and previous equations that, to a first order, there is no combination of amplitude and phase wiggles that will improve the distortion, harmonic or total, due to either alone.

The summation in (12) leads to

$$\left.\begin{aligned} 4m^2\frac{D_A^2}{\mathcal{E}_1^2} &= 4m^2 \sin^2 pb \\ &\quad + 2m \sin pb \cos 4b(\omega_c - \omega_1)\left[J_1(4m \sin pb)\right. \\ &\quad \left. + 4mJ_0(2m \sin 2pb) \sin^2 pb - \frac{J_1(2m \sin 2pb)}{\cos pb}\right] \\ &\quad - 8m^2 J_0(4m \sin^2 pb) \sin^2 pb \cos^2 pb \\ &\quad + 2mJ_1(4m \sin^2 pb) \\ &\quad - 16J_1^2(2m \sin pb) \sin^2 pb \cos^2 2b(\omega_c - \omega_1), \\ 4m^2\frac{D_\phi^2}{\mathcal{E}_2^2} &= 4m^2 \sin^2 pd \\ &\quad + 2m \sin pd \cos 4d(\omega_c - \omega_2)\left[J_1(4m \sin pd)\right. \\ &\quad \left. - 4mJ_0(2m \sin 2pd) \sin^2 pd + \frac{J_1(2m \sin 2pd)}{\cos pd}\right] \\ &\quad + 8m^2 J_0(4m \sin^2 pd) \sin^2 pd \cos^2 pd \\ &\quad - 2mJ_1(4m \sin^2 pd) \\ &\quad - 16J_1^2(2m \sin pd) \cos^2 pd \cos^2 2d(\omega_c - \omega_2). \end{aligned}\right\} \tag{14}$$

It can be seen from (14) that D_A has period π in pb and $2b(\omega_c - \omega_1)$, and that D_ϕ has period π in pd and $2d(\omega_c - \omega_2)$. Note that $p = 2\pi f_a$, where f_a is audio frequency, and b or d is related to the number of amplitude or phase wiggles per unit frequency (see Fig. 2). Fig. 2 can be used to give a crude partial check on the results in (10) and (14), if it is recalled that the amplitude and phase wiggles have period π in pb and pd, respectively. For example, if $pb = \pi$, then ω_c and all the sidebands of the frequency-modulated input fall on a flat line as far as amplitude is concerned. Hence, there should be no distortion D_A due to amplitude alone. From (14), it is seen that $D_A=0$ for $pb=\pi$. If $pd=\pi$, then ω_c and the sidebands of the frequency-modulated input fall along a line through the phase characteristic, and D_ϕ should be zero. From (14) it is seen that $D_\phi=0$ for $pd=\pi$.

Fig. 2, with the linear term in the phase removed, and (10) yield information about harmonic distortion. If $2b(\omega_c-\omega_1)=n\pi$, n integral, then ω_c is at the peak of an amplitude wiggle, and the sidebands fall at points that are symmetrical with respect to this peak. In this case, (10) shows that $D_{kA}=0$ for k even. If $2b(\omega_c-\omega_1)=(2n+1)\pi/2$, then ω_c is at a node and the points for the sidebands are odd symmetrical about this node. In this case, $D_{kA}=0$ for k odd. If $2d(\omega_c-\omega_2)=n\pi$ then ω_c is at a node, and the sideband points are distributed with odd symmetry about this node. For this case, $D_{k\phi}=0$ for k even. Finally, if $2d(\omega_c-\omega_2)=(2n+1)\pi/2$, then ω_c is at a maximum or minimum, and there is even symmetry for the sideband points. For this case, $D_{k\phi}=0$ for k odd.

A thorough analytical and numerical discussion of the distortion results in (10) and (14) will not be given in this paper. Since such a discussion is very complicated and probably of no practical value in general form, it will be made later for the special case of the waveguide. However, one interesting statement can be made about D in general.

If neither pb nor pd is close to $n\pi/2$ and m is sufficiently large, then approximately

$$\left.\begin{aligned} D_A &\approx \mathcal{E}_1 \sin pb, \\ D_\phi &\approx \mathcal{E}_2 \sin pd, \\ D^2 &\approx \mathcal{E}_1^2 \sin^2 pb + \mathcal{E}_2^2 \sin^2 pd, \end{aligned}\right\} \tag{15}$$

for pb and pd unequal to $n\pi/2$ for any integral n and for large m.

It will be seen later in the case of the waveguide that D in (15) is close to its maximum value for a given value of pb ($b=d$ in this special case) and that the over-all maximum value of D is close to its value in (15) for $pb=\pi/2$. It is possible that similar conclusions may apply to the general case. If this is true, then (15) indicates over-all maxima of $D_A=\mathcal{E}_1$, $D_\phi=\mathcal{E}_2$, and $D^2=\mathcal{E}_1^2+\mathcal{E}_2^2$. For example, if $\mathcal{E}_1=\mathcal{E}_2=0.05$, then $D_A=D_\phi=0.05$ and $D=0.07$. It can readily be shown from (14) that, for $pb\neq n\pi/2$ and $pd\neq n\pi/2$, D_A and D_ϕ have maxima and minima with respect to $2b(\omega_c-\omega_1)$ and $2d(\omega_c-\omega_2)$ at $4b(\omega_c-\omega_1)=n\pi$ and $4d(\omega_c-\omega_2)=n\pi$. The two n's need not refer to the same integer.

This concludes the general discussion of the first-order analysis of distortion caused by sinusoidal wiggles in amplitude and phase of transmission systems. If the mathematical description of such wiggles in the essential band of frequencies for the input frequency-modulated wave requires more than one sinusoid, the derivation of distortion results becomes more complicated, but remains possible along the lines of the previous analysis.

III. Derivation of Waveguide Amplitude and Phase Characteristics

We now apply previously obtained distortion results to lossless waveguides. Fig. 3 shows their equivalent transmission-line representation. It is possible to derive the special forms that the distortion expressions take for the waveguide by regarding the output voltage E_2 as being composed of voltages due to the main incident wave and the first re-reflection. The first-order analysis of this report would imply that the remaining re-reflections can be neglected. This point of view can be justified by using the paired-echo method or by superposing the main incident wave and the first re-reflection for each frequency in the input frequency-modulation spectrum. Harmonic-distortion formulas derived from this point of view appear as sideproducts in articles on multipath transmission by Crosby[1] and Corrington.[2]

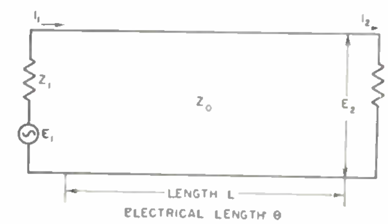

Fig. 3—Terminated waveguide.

Now consider the waveguide of Fig. 3. To apply previous distortion results to the waveguide, the quantities b, c, d, ω_1, ω_2, $\mathcal{E}_1$, and $\mathcal{E}_2$ have to be expressed in terms of waveguide parameters. From Fig. 3 and standard transmission-line equations, one can express E_2 in terms of a constant-voltage generator E_1 by the relation

$$\begin{aligned} E_2 &= \frac{Z_1Z_2}{(Z_1+Z_2)\cos\theta + j(1+Z_1Z_2)\sin\theta} E_1 \\ &= A(\theta)e^{-j\phi(\theta)}E_1. \end{aligned} \tag{16}$$

The impedances Z_1 and Z_2 in (24) have been normalized with respect to the characteristic impedance Z_0 of the waveguide.

For real Z_1 and Z_2, one can write

$$\left.\begin{aligned} A(\theta) &= \frac{Z_1Z_2}{[(Z_1+Z_2)^2+(1-Z_1^2)(1-Z_2^2)\sin^2\theta]^{1/2}} \\ \tan\phi(\theta) &= \frac{1+Z_1Z_2}{Z_1+Z_2}\tan\theta. \end{aligned}\right\} \tag{17}$$

[1] M. G. Crosby, "Frequency-modulation propagation characteristics," Proc. I.R.E., vol. 24, pp. 898–913; June, 1936.

[2] M. S. Corrington, "Frequency-modulation distortion caused by multipath transmission," Proc. I.R.E., vol. 33, pp. 878–891; December, 1945.

It is convenient to rewrite (17) in terms of reflection factors r_1 and r_2 defined by

$$\left.\begin{aligned} r_1 &= \frac{1 - Z_1}{1 + Z_1}, \qquad Z_1 < 1 \\ r_2 &= \frac{1 - Z_2}{1 + Z_2}, \qquad Z_2 < 1. \end{aligned}\right\} \tag{18}$$

If Z_1 or Z_2 is greater than one (or both are), then the numerators in (18) are reversed, but the final results are not affected.

With the use of (18), (17) becomes

$$\left.\begin{aligned} A(\theta) &= \frac{(1 - r_1)(1 - r_2)}{2[(1 - r)^2 + 4r \sin^2 \theta]^{1/2}}, \qquad r = r_1 r_2 \\ \tan \phi(\theta) &= \frac{1 + r}{1 - r} \tan \theta \end{aligned}\right\} \tag{19}$$

Before proceeding, consider the picture of distortion if either load in Fig. 3 is matched, i.e., if $r_1 = 0$ or $r_2 = 0$. In either case, $A(\theta)$ reduces to a constant and ϕ becomes θ. Since θ is approximately linear in frequency ω, ϕ becomes linear in ω. In other words, for $r = 0$, the system has a flat amplitude characteristic and linear phase characteristic. Hence, as is well known for this case, the system will not distort the frequency of a frequency-modulated input. If $r \neq 0$, then energy from the generator travels back and forth between Z_1 and Z_2 and undergoes partial reflections at each end. The energy at the load end comes from the main incident wave and re-reflections. It is the latter that cause distortion trouble in the case of a frequency-modulated input.

If now r is assumed to be small, it can be shown that (19) becomes

$$\left.\begin{aligned} A(\theta) &\approx 1 + r \cos 2\theta, \\ \phi(\theta) &\approx \theta + r \sin 2\theta. \end{aligned}\right\} \tag{20}$$

There remains the problem of expressing θ in terms of ω. Waveguide theory yields the formula

$$\theta = \frac{L\omega}{v}\left[1 - \frac{(\omega_1)^2}{(\omega)^2}\right]^{1/2} = a\omega, \tag{21}$$

where L is the physical length of the guide, v is velocity of light in unbounded space of the waveguide medium, and ω_1 is the cutoff frequency of the propagated mode. If ω is restricted to a small band, then the radical can be assumed to have a constant value taken at the center of the band, so that a is essentially constant. The use of (21) in (20) leads to the expressions

$$\left.\begin{aligned} A(\omega) &= 1 + r \cos 2a\omega, \\ \phi(\omega) &= a\omega + r \sin 2a\omega. \end{aligned}\right\} \tag{22}$$

IV. Distortion Formulas for Waveguide

If (22) is compared to (2), it is seen that $\mathcal{E}_1 = \mathcal{E}_2 = r$, $b = c = d = a$, and $\omega_1 = \omega_2 = 0$. The variable part of the output frequency as given by (11) thus becomes

$$\left.\begin{aligned} \frac{1}{p}\frac{d\psi}{d\tau} &= m \cos p\tau \\ &\quad + 2r \sin 2a\omega_c \sum_{k=2,4,\cdots}^{\infty} kJ_k(2m \sin ap) \\ &\quad \sin k(p\tau - pa - \pi/2) \\ &\quad + 2r \cos 2a\omega_c \sum_{k=3,5,\cdots}^{\infty} kJ_k(2m \sin ap) \\ &\quad \cos k(p\tau - pa - \pi/2). \end{aligned}\right\} \tag{23}$$

The general distortion formulas in (12) now become

$$\left.\begin{aligned} D/2^{1/2}r &= \{\sin^2 ap\{1 + [J_0(4m \sin ap) \\ &\quad + J_2(4m \sin ap)] \cos 4a\omega_c\} \\ &\quad - (2/m^2)J_1^2(2m \sin ap) \cos^2 2a\omega_c\}^{1/2}, \\ D_2 &= (4r/m)J_2(2m \sin pa) \sin 2a\omega_c, \\ D_3 &= (6r/m)J_3(2m \sin pa) \cos 2a\omega_c \cdots \end{aligned}\right\} \tag{24}$$

In interpreting the distortion results in (24), it is instructive to consider Fig. 4. The wiggles in Fig. 4 are phased 90 degrees apart, have amplitude r, and have period π/a in ω. The discussion following (14) can be applied here. It leads to the conclusions that there is odd-harmonic distortion if $2a\omega_c = n\pi$, even-harmonic distortion if $4a\omega_c = (2n+1)\pi$, and no distortion if $ap = n\pi$.

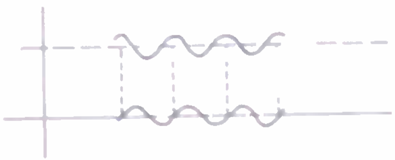

Fig. 4—Amplitude and phase characteristics of a terminated waveguide. The upper curve corresponds to $A(\omega) = 1 + r \cos 2a\omega$ and the lower curve to $\phi(\omega) = r \sin 2a\omega$ with the linear term omitted.

It is interesting to examine (24) in limiting cases. If $m \sin ap$ is very small, either because of small m or because of small ap, and if $\sin 2a\omega_c \neq 0$, then it can be seen that (24) reduces to

$$D \approx 2rm \sin^2 ap \sin 2a\omega_c \approx D_2, \tag{25}$$

for $m \sin ap$ small and $\sin 2a\omega_c \neq 0$.

In this case, distortion is chiefly of the second-harmonic type. If $m \sin ap$ is small and $\sin 2a\omega_c = 0$, then (24) reduces to

$$D \approx rm^2 \sin^3 ap \approx D_3 \text{ for } m \sin ap \text{ small and } \sin 2a\omega_c = 0. \tag{26}$$

Here, distortion is primarily of the third-harmonic type. Note that, for small $m \sin ap$ and $\sin 2a\omega_c = 0$, distortion drops much more rapidly with $m \sin ap$ than for the case $\sin 2a\omega_c \neq 0$.

If $m \sin ap$ is large, then (24) becomes

$$D \approx 2^{1/2} r \sin ap. \tag{27}$$

This result is essentially independent of m and $a\omega_c$. It cannot be attributed primarily to any single harmonic.

A brief discussion will now be given of the total distortion D. It depends on the dimensionless parameters r, ap, $a\omega_c$, and m. Since D varies linearly with r, no further discussion of its dependence on r is required. Perhaps the most striking feature about (24) is that D is periodic with fundamental period π in either ap or $2a\omega_c$ for fixed m. In other words, D cannot increase indefinitely with ap or $2a\omega_c$, but passes periodically through zeros and maxima. Since a is linear in waveguide length L, increasing L indefinitely does not produce an indefinitely increasing D.

It appears at first glance from the foregoing that the description of distortion in the waveguide as a long-line effect is inaccurate. However, as will be seen later, the first maximum of D with respect to L for fixed p, ω_c, and m may occur for a large value of L, so that increasing L from zero to this value will produce monotonically increasing D. In this case, increasing distortion can be regarded as a long-line effect so far as physical length L is concerned, but is actually confined to electrical lengths $ap<\pi$.

V. Construction of Distortion Graphs for Waveguides

A picture of the distortion introduced by loaded waveguides can be obtained by plotting distortion D against physical length L of waveguide or audio frequency f_a. Assume that the carrier frequency is fixed at 5,000 mc. Assume also that the maximum instantaneous deviation in the frequency of the frequency-modulated input is fixed. For this case, the index m varies with audio frequency $f_a=p/2\pi$. Assume that $m=1$ when $f_a=10$ mc, so that $m=10^7/f_a$ for any other audio frequency f_a.

Now consider the expression given for a in (21). The cutoff frequency ω_1 for the lowest mode in a standard 2-inch by 1-inch waveguide is about $2\pi\times3.16\times10^9$. The value of the radical in (21) is about 0.775 for $\omega=\omega_c$. The value of a is therefore given by $a=7.87\times10^{-10}\ L$, seconds, for L in feet. The expressions for ap, $a\omega_c$, and m to be used in the distortion formulas (24) for the present case become

$$\left.\begin{aligned} 2a\omega_c &= 2\pi\times 7.87L,\\ ap &= 2\pi\times 7.87\times 10^{-10} f_aL,\\ m &= 10^7/f_a. \end{aligned}\right\} \qquad (28)$$

Equation (28) shows that the functions of $2a\omega_c$ and ap in (24) have respective periods of 0.0635 feet and $0.0635\times10^{10}/f_a$ feet in L. In other words, the period of the ap terms is $10^{10}/f_a$ times the period of the $2a\omega_c$ terms. Since $f_a\leq10^7$ in the present application, the terms involving $2a\omega_c$ in (24) have a high frequency compared to those involving ap. Equation (28) also shows that the functions of ap in (24) have period $6.35\times10^8/L$ cycles in f_a. Note that the functions of $2a\omega_c$ in (24) are independent of f_a.

Graphs are provided in Figs. 5 and 6 of $D/2^{1/2}r$ against waveguide length L for different values of audio frequency f_a and against f_a for different values of L. Consider first the plots in Fig. 5 of $D/2^{1/2}r$ against L for different values of f_a. For large values of f_a, these

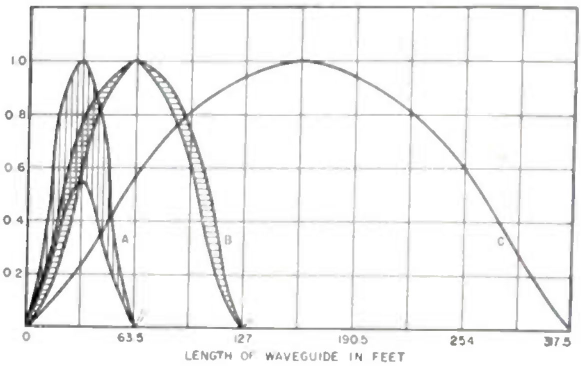

Fig. 5—Total distortion D plotted against waveguide length L. Curves A are for $f_a=10$ mc, and $m=1$; B, $f_a=5$ mc, and $m=2$; C, $f_a=2$ mc, and $m=5$.

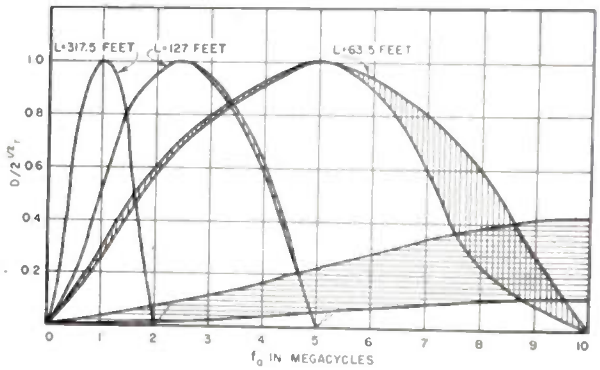

Fig. 6—Total distortion D plotted against audio frequency f_a with $m=10/f_a$.

graphs show a shaded area bounded by two curves. The actual curve lies in this shaded area and has high-frequency wiggles that have not been indicated. Hence, for large f_a, D changes very rapidly for a slight change in L. As noted above, these high-frequency wiggles have a period of 0.0635 feet in L independently of f_a. In the graph for $f_a=10$ mc, for example, there are 1,000 wiggles in the shaded area. The amplitude of these wiggles tends to zero as f_a is decreased, and hence m is increased. Only one period of each curve is drawn in Fig. 5.

Next consider the plots in Fig. 6 of $D/2^{1/2}r$ against f_a for various values of L. For small values of L, these graphs show shaded areas that are interpreted differently from the shaded areas of Fig. 5. In the present case, any actual curve again lies within the corresponding shaded area but no longer has any wiggles. A slight change in the parameter L from 0 to 0.0635 feet will yield a distortion curve without wiggles anywhere in a shaded area. Notice that these shaded areas shrink vertically as L is increased. As in Fig. 5, only one period of each curve is drawn in Fig. 6.

Measurements of Wavelengths and Attenuation in Dielectric Waveguides for Lower Modes*

C. W. HORTON† AND C. M. MCKINNEY‡, MEMBER, IRE

Summary—The wavelength of a wave guided by a dielectric rod has been measured for the TE_{0n}, TM_{0n}, HE_{1n}, and EH_{1n} modes for $n=1$ and 2. The measurements were made on polystyrene, lucite, textolite, and paraffin rods, as well as on plastic tubes filled with Nujol and dioxane. The measurements agree very well with the solutions of the characteristic equation. The solution of the characteristic equation corresponding to the second radial mode and the angular mode of order one has two branches, both of which are found experimentally. Measurements are made of the attenuation in a dielectric rod due to losses in the dielectric and due to bending of the dielectric guide.

I. Introduction

Many workers, including Zahn,[1] Schriever,[2] Southworth,[3] Mallach,[4] and Chandler[5] have measured the properties of dielectric waveguides. However, most of these measurements were made on small rods which could support only the first-order radial modes. Hondros and Debye,[6] Carson, Mead, and Schelkunoff,[7] Wegener,[8] Bondi and Pryce,[9] Whitmer,[10] Abele,[11] Elsasser,[12] Stratton,[13] and others have studied the theoretical aspects of the dielectric waveguide, but no comprehensive discussion of the theory has been published.

In the following discussion the wavelength of the guided wave will be described[14] by an "apparent index of refraction," which is defined as λ guide/λ external medium and denoted by n_a. The value of n_a may be measured or it may be computed from the characteristic equation given by Stratton.[15]

II. Experimental Technique

The desired modes were first generated in metallic waveguides by means of mode converters, and the dielectric rod was tapered to fit the metallic guide. Three different methods were used to measure the guided wavelength.

In the first method the waveguide was terminated by a large copper sheet perpendicular to the axis of the guide. This sheet produced a large standing-wave ratio that could be measured easily with a moving probe. Small rods yielded a simple standing wave which could be measured accurately. In the case of large rods, the tapered section that served as a transition between the metallic and the dielectric guide was short enough to give rise to a second, and in some cases a third, radial mode. When this happened, the standing-wave pattern had a modulation envelope, as shown in Fig. 1, from which one can determine the guided wavelength for both modes.

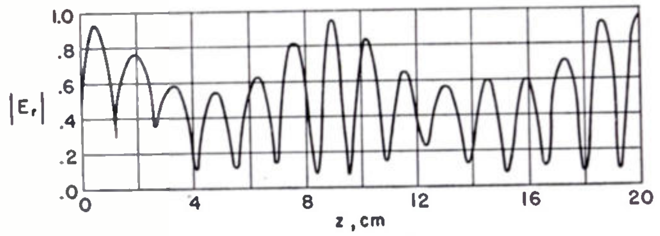

Fig. 1—The dependence of $|E_r|$ on the distance along a dielectric rod which has two modes, TM_{01} and TM_{02}, of propagation. The rod is made of textolite, and has a diameter of 2 5/16 inches. The free-space wavelength is 3.20 cm.

The second method is a variation of a procedure used by Chandler.[16] In this method a hole is cut in each of two copper plates which are mounted perpendicularly to the dielectric rod that passes through the holes. When the distance between the plates is properly adjusted, a resonant cavity results. This method will be referred to as the "resonator method (moving plates)." It works well on rods of small diameter, for which most

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† Defense Research Laboratory, The University of Texas, Austin, Texas.

‡ Formerly, Defense Research Laboratory, the University of Texas, Austin, Texas; now, Texas Technological College, Lubbock, Texas.

[1] H. Zahn, "Über den Nachweis elecktromagnetischer Wellen an dielektrischen Drähten," *Phys. Zeitsch.*, vol. 16, pp. 414–416; November 15, 1915.

[2] O. Schriever, "Elektromagnetische Wellen an dielektrischen Drahten," *Ann. d. Phys.*, vol. 63, pp. 645–673; 1920.

[3] G. C. Southworth, "Hyper-frequency wave guides—general considerations and experimental results," *Bell Sys. Tech. Jour.*, vol. 15, pp. 284–309; April, 1936.

[4] Peter Mallach, "Air material command report, F-TS-2223-RE," February, 1948. Translated by P. L. Harbury.

[5] C. H. Chandler, "An investigation of dielectric rod as wave guide," *Jour. Appl. Phys.*, vol. 20, pp. 1188–1196; December, 1949.

[6] D. Hondros, and P. Debye, "Elektromagnetische Wellen an dielektrischen Drähten," *Ann. d. Phys.*, vol. 32, pp. 465–476; 1910.

[7] J. R. Carson, S. P. Mead, and S. A. Schelkunoff, "Hyper-frequency wave guides—mathematical theory," *Bell Sys. Tech. Jour.*, vol. 15, pp. 310–333; April, 1936.

[8] H. J. Wegener, "Propagation velocity, wave resistance, and attenuation of electromagnetic waves in dielectric cylinders," Air Material Command Microfilm ZWB/FB/Re/2018, R. 8117 P 831. The first and third parts have been translated by M. M. Astrahan and W. C. Jakes, Jr., of the Microwave Laboratory, Northwestern University.

[9] H. Bondi and M. H. L. Pryce, "Dielectric cylinders as waveguides," Admiralty Signal Establishment Report No. M.434, August, 1942.

[10] R. Whitmer, "Fields in nonmetallic waveguides," Proc. I.R.E., vol. 36, pp. 1105–1109; September, 1948.

[11] M. Abele, "Teoria della propagazione di un campo elettromagnetico lungo una guida dielettrica a sezione circolare," *Il Nuovo Cimento*, vol. 5, pp. 3–13; August, 1948.

[12] Walter M. Elsasser, "Attenuation in a dielectric circular rod," *Jour. Appl. Phys.*, vol. 20, pp. 1193–1196; December, 1949.

[13] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 524–527; 1941.

[14] Gilbert Wilkes, "Wavelength lenses," Proc. I.R E., vol. 36, pp. 206–212; February, 1948.

[15] See Eq. (9), p. 526, of footnote reference 13.

[16] See Fig. 3, p. 1190, of footnote reference 5.

of the energy is outside of the rod, but it is not suitable for use on large rods.

In order to apply the resonator method to large rods, it is necessary to make the holes in the copper sheets smaller than the dielectric rod so that the sheets could reflect a significant amount of the energy. The dimensions of the dielectric rod that was placed between the two plates were calculated in advance from the characteristic equation. This made it necessary to tune the klystron oscillator through a small frequency range only. This method will be referred to as the "resonator method (fixed plates)."

In all of the measurements the source of energy was a type TS-13 microwave signal generator operating near a frequency of 9,275 mc, and the detector was a crystal detector in conjunction with a high-gain amplifier. The index of refraction of the samples used was measured at the operating frequency by transmission and reflection methods.[17]

III. The TE_{0n} Mode

Fig. 2 is a graph of some of the roots of the characteristic equation that governs the propagation of the TE_{0n} mode along a rod whose index of refraction is 1.5. The TE_{01} mode transducer did not function well, and only three reliable measurements were obtained. These measurements gave values of 1.03, 1.13, and 1.23 for n_a corresponding to values of 0.596, 0.796, and 1.00 for d/λ_0, respectively. Here d is the diameter of the rod which was made of lucite with $n=1.60$.

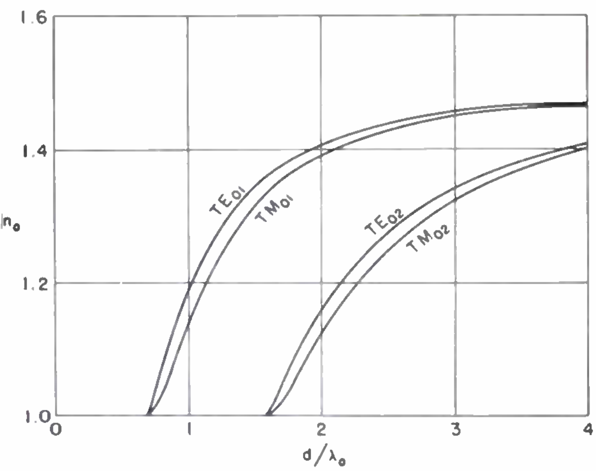

Fig. 2—Theoretical curves of n_a versus d/λ_0 for a rod whose index of refraction is 1.5. The TE_{0n} and the TM_{0n} modes.

IV. The TM_{0n} Mode

Fig. 2 also shows some of the roots of the characteristic equation that governs the propagation of the TM_{0n} mode along a rod whose index of refraction is 1.5. Fig. 3 shows some experimental data for lucite and textolite rods and for a polystyrene tube containing Nujol. The walls of the polystyrene tube were 1/16 inch thick. Nujol was used because it has an index of refraction of 1.5 corresponding to the theoretical curves shown in Fig. 2. The indices of refraction of the textolite and lucite rods were 1.58 and 1.60, respectively. The TM_{01} mode in the circular metallic waveguide was transformed from a TE_{10} mode in a rectangular waveguide by means of a suitable junction.

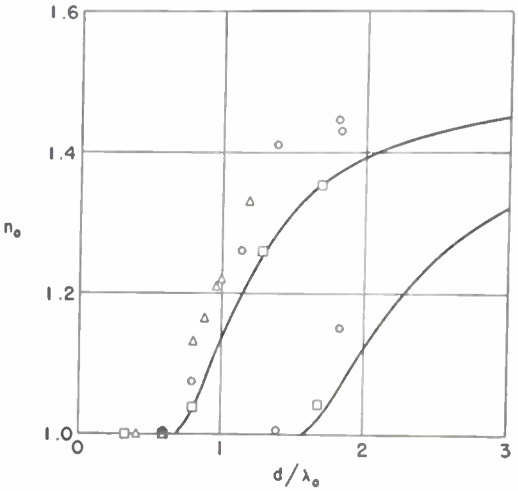

Fig. 3—Measurements of n_a versus d/λ_0 for the TM_{0n} mode. ○ textolite ($n=1.58$). △ lucite ($n=1.60$). □ polystyrene tubes filled with Nujol ($n=1.50$). The solid curve is a theoretical curve for $n=1.5$ and should be compared with the squares only. The data for textolite were measured by the resonator method (moving plates), while the other data were obtained with the probe method.

V. The EH_{1n} and the HE_{1n} Modes

The mode with angular order one is of more practical importance than the modes with no angular variation; consequently, it has been discussed more fully in the literature. Bondi and Pryce[18] have pointed out that the angular mode $n=1$ is unique in that it is the only mode for which a rod of arbitrarily small diameter can propagate a guided wave. Bondi and Pryce[19] call this mode the "fundamental mode," while Wegener[20] refers to it as the "HE_{10} mode." This designation is changed to HE_{11} in the present paper. The higher radial modes occur in pairs, as shown in Fig. 4, which is a plot of some of the roots of the characteristic equation. Wegener[20] says that these higher modes will be called "*HE* waves if the field structure of a cross section resembles that of an *H* wave and *EH* waves if it resembles that of an *E* wave." Although this description of the difference does not seem adequate, a comparison of Figs. 2 and 4 indicates that it is justifiable to characterize the two radial modes as Wegener suggests. This distinction is placed on a firmer basis by Abele,[21] who shows that the coefficients of the components of the transverse electric field have different signs on the two branches associated with each radial mode.

[17] Carol G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., Chapt. 10; 1947.

[18] See p. 2 of footnote reference 9.
[19] See p. 1 of footnote reference 9.
[20] See p. 9 of the translation of footnote reference 8.
[21] See pp. 9–10 of footnote reference 11.

Fig. 4 is a graph of some of the roots of the characteristic equation for the angular variation of order one and for the first two radial modes. The detailed structure of the second radial mode does not appear to have been illustrated before. The curves are computed for $n = 1.5$. In order to check the theoretical solution of the characteristic equation that is shown in Fig. 4, a series of paraffin blocks were molded and measured by the resonator method (fixed plates). The resulting data are shown in Fig. 4.

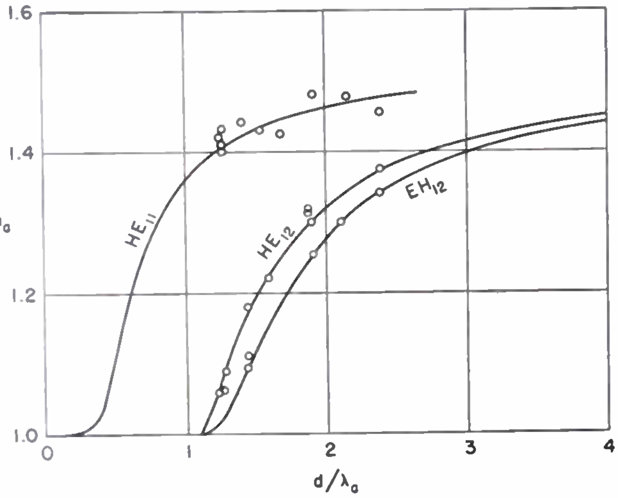

Fig. 4—Measurements of n_a versus d/λ_0 for paraffin ($n = 1.50$) by means of the resonator method (fixed plates). The solid curves are theoretical.

Fig. 5 shows further data on the HE_{11} mode for lucite and textolite rods and for polystyrene tubes filled with Nujol and dioxane. The dielectric constant of the dioxane solution was 1.60. Since the behavior of the HE_{11} mode should be independent of the source, the lucite rods were excited by using both a TM_{11} mode and a TE_{11} wave in the metallic waveguide.

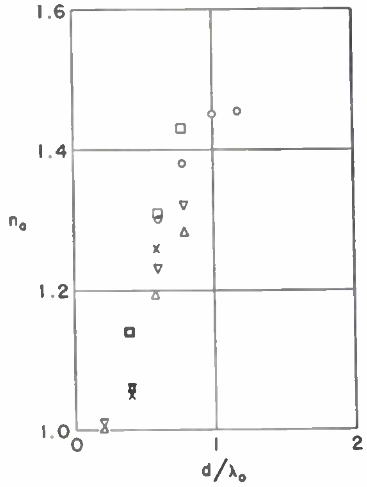

Fig. 5—Measurements of n_a versus d/λ_0 for the HE_{11} modes for various rods. ○ lucite ($n = 1.60$) measured by the probe method. □ lucite ($n = 1.60$) measured by the probe method. × textolite ($n = 1.58$) measured by the resonator method (moving plates). Δ polystyrene tubes filled with Nujol ($n = 1.50$) measured by the probe method. Δ polystyrene tubes filled with dioxane ($n = 1.60$) measured by the probe method. The source was a circular metallic waveguide with the TE_{11} mode in every case except that of the squares which were obtained from a TM_{11} mode.

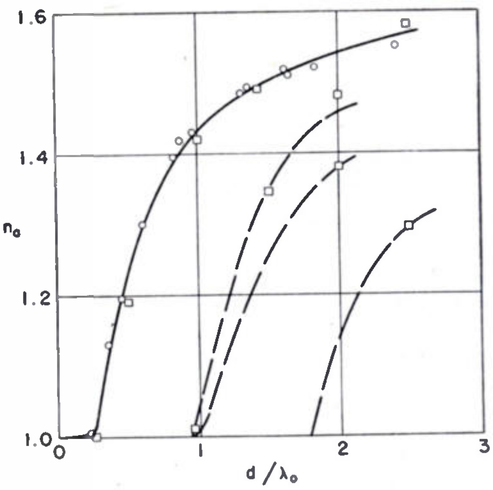

Fig. 6—Measurements of n_a versus d/λ_0 for polystyrene ($n = 1.60$). ○ resonator method (fixed plates) □ probe method. The curves are based on the experimental data.

Fig. 6 represents a series of measurements on polystyrene rods by the resonator (fixed plates) and the probe methods. The cutoff frequencies for the first three radial modes occur at $d/\lambda_0 = 0$, 0.97, and 1.79.

The TE_{11} modes were generated in the circular metallic waveguides by a simple tapered transition from a TE_{10} mode in a rectangular waveguide. The TM_{11} mode was generated in the circular metallic guide by means of two axial probe antennas excited 180° out of phase.

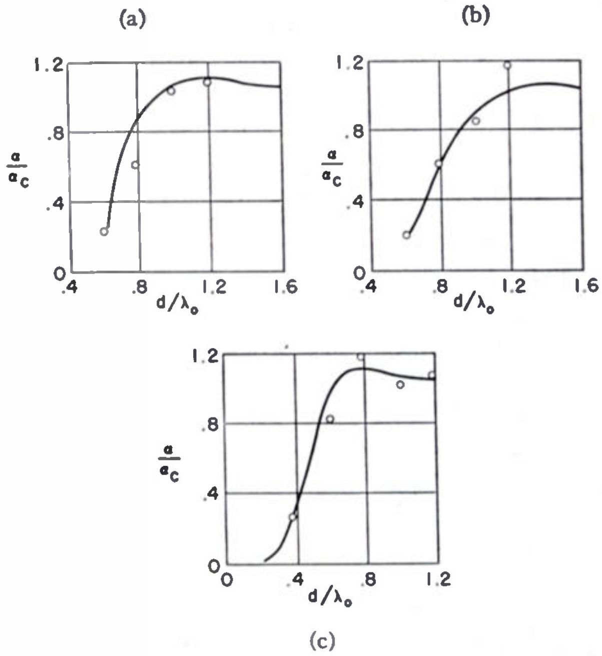

Fig. 7—The dependence of the attenuation of a lucite rod on the diameter. (a) TE_{01} mode. (b) TM_{01} mode. (c) HE_{11} mode. α_c is the attenuation in coaxial cable filled with lucite. The solid curves are from Wegener's theoretical work.

VI. Attenuation

Wegener[22] has computed the attenuation in a dielectric waveguide caused by the losses in the dielectric. Fig. 7 shows his theoretical curves and experimental data for a series of lucite rods. The lucite used in these experiments has a loss factor given by $\tan\delta = 0.01$.

It is frequently suggested that flexible dielectric guides can be used as connections to join sections of metallic guides. Fig. 8 shows the attenuation that results when two parallel waveguides separating a vertical distance h are connected with a flexible vinyl tube filled with Nujol. The tube is 2 feet long and has an outside diameter of 9/16 inch.

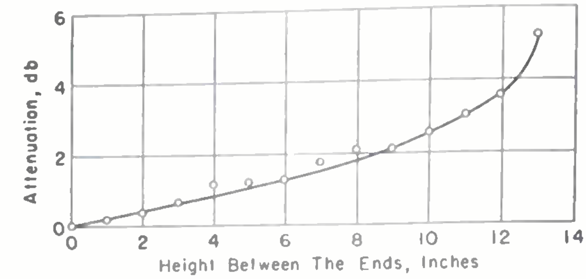

Fig. 8—The attenuation of a vinyl tube (od 9/16 inch) filled with Nujol ($n = 1.50$) as a function of the height between the ends. The data are for the HE_{11} mode

[22] See Figs. 13, 14, and 15 of footnote reference 8

The Short-Slot Hybrid Junction*

HENRY J. RIBLET†, ASSOCIATE, IRE

Summary—This paper describes a novel high-performance x-band hybrid junction. Its over-all dimensions are $1\frac{3}{4}'' \times \frac{1}{2}'' \times 2''$. It consists of a suitably loaded gap in the narrow common wall between two $\frac{1}{2}'' \times 1''$ waveguides. Over the frequency range 8,500- to 9,600-mc per second power equality within ± 0.25 decibels, isolation in excess of 30 decibels and a standing-wave ratio less than 1.07 may be obtained. The theory of the device is explained, and the particular advantages of this hybrid junction for a number of applications are outlined.

Introduction

The waveguide hybrid junction[1] plays an important part in a number of specialized waveguide circuits. In addition to its application as a power splitter, it is useful in the construction of balanced duplexers,[2] balanced mixers,[3] and broad-band switches.[4] Although special forms of waveguide hybrids have been used on occasion, the most common are the "magic tee" and the "hybrid ring." Both of these have in common the characteristic that when power enters one of the terminals it divides between two of the others so that the outgoing voltages at equally distant terminals are either in phase or exactly out of phase. There exists, however, another large class of waveguide hybrid junctions at whose equidistant output terminals the voltages are always in quadrature. One of the earliest of these has been called a right-angle hybrid.[5] The possibility of quadrature hybrid junctions having broad-band characteristics has been pointed out by N. I. Korman, in an unpublished work, and by Riblet and Saad.[6] For many applications, however, these junctions are unduly large. It is the object of this article to describe a compact broad-band hybrid junction of the quadrature type which lends itself, for many applications, to more efficient use of space than is possible with conventional hybrids.

Although the short-slot hybrid is closely related to the family of directional couplers, it is not, strictly speaking, a member according to the definition given by Mumford.[7] Nevertheless, a structure which has the same general appearance but which is a directional coupler has been described by Surdin.[8] Moreover, the feasibility of obtaining hybrid performance from paral-

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† Microwave Development Laboratories, Inc., Waltham, Mass.

[1] W. A. Tyrrell, "Hybrid circuits for microwave," Proc. I.R.E., vol. 35, pp. 1307–1313; November, 1947.

[2] J. Reed, "Rat Race Duplexing," M.I.T. Radiation Laboratory Report 885; February, 1946.

[3] W. A. Tyrrell, *ibid.*

[4] W. D. Lewis, and L. C. Tillotson, "A non-reflecting branching filter for microwaves," *Bell Sys. Tech. Jour.*, pp. 83–84; January, 1948.

[5] C. W. Zabel, "Balanced Duplexers," Microwave Duplexers, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N.Y., vol. 14, p. 367; 1948.

[6] H. J. Riblet and T. S. Saad, "A new type of waveguide directional coupler," Proc. I.R.E. vol. 36, p. 64; January, 1948.

[7] W. W. Mumford, "Directional couplers," Proc. I.R.E., vol. 35, pp. 160–166; February, 1947.

[8] M. J. Surdin, "Directional couplers in waveguides," *Jour. IEE* (London), pt. III, A 931 (no. 4) pp. 735–736; 1946.

lel waveguides apertured by a single large hole is suggested by Dicke.[9]

Simple General Theory[10]

To date all directional coupler-like hybrid junctions have had a plane of symmetry running their full length. Such an arrangement is shown schematically in Fig. 1. When power is incident on the main guide 2 at terminal 2_i, it proceeds along that waveguide until it encounters the coupling section. Under suitable conditions, by the time the energy reaches the end of the coupling section, it will have divided so that the energy leaving at terminal 1_o just equals that leaving at 2_o. If in addition no power leaves at terminal 1_i and none is reflected at terminal 2_i, assuming perfectly matched terminations at 1_o and 2_o, the structure is an ideal waveguide hybrid unction.

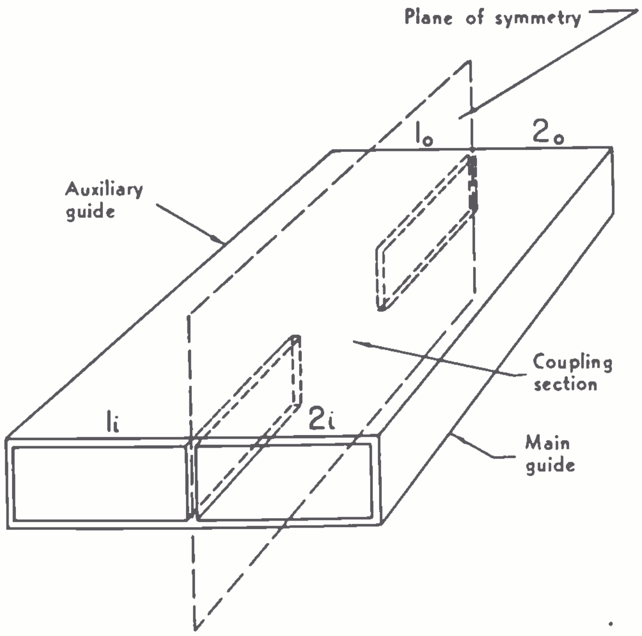

Fig. 1—Schematic hybrid junction.

It is now rather easy to derive several of the important characteristics of this type of hybrid junction and to state certain fundamental conditions which must be satisfied.

Since the reader should have no difficulty in reconstructing the arguments with the help of the Lippmann and Kyhl references, the results will be presented without details.

Conditions for complete isolation: *The reflected voltages in the even and odd modes shall both be zero.* When this condition is satisfied, the condition for power division becomes the following:

Condition for hybrid performance: *The transmitted voltages for the even and odd modes must differ from each other by 90 degrees.* This may be restated as

$$L\left(\frac{1}{\lambda_g^e} - \frac{1}{\lambda_g^0}\right) + \phi_r = 1/4, \tag{1}$$

where L is the length of the coupling section, λ_g^e and λ_g^o are the guide wave lengths for the even and odd modes, respectively, and ϕ_r is the phase shift in the even mode contributed by the reflections from the ends of the coupling section. Two useful additional facts may be inferred also from simple vector arguments concerning the voltages in the even and odd modes. They are:

(a) The output voltages at terminal 1_o leads the voltage at 2_o by 90 degrees.

(b) The output voltage at 2_o leads by 45 degrees what it would be if there were no slot at all in the waveguide.

The first conclusion is a consequence of high isolation, and is independent of the power-division characteristics of the hybrid junction. The second conclusion requires high isolation and the additional assumption that the odd mode does not see the slot.

Special Theory

Since the slot has very little effect on the odd mode, further analysis of the performance of the short-slot hybrid requires a solution of the Maxwell equations which satisfies the boundary conditions for the structure of Fig. 1 excited in the even mode. Fortunately, the solution of this problem is completely known in terms of convergent series, thanks to the work of Carlson and Heins.[11] This results from the fact that the even mode in the coupling section can be expressed in terms of plane waves so that the problem reduces to that of determining the reflection and transmission coefficients for a pair of plane waves suitably incident on an infinite array of metal plates.

The condition for complete isolation requires that the admittance of the structure in both directions as seen from the center of the coupling section shall be real. Accordingly, we determine R_g, the reflection coefficient, when two in-phase plane waves are incident on a semi-infinite set of metal plates ($\alpha = \pi/2$) along the directions $\pm\theta$. Of course θ is determined by the requirement that the electric field be zero at the outer wall of the coupling section. In the notation of Carlson and Heins

$$R_g = e^{i(\theta_1 - \theta_2)} \frac{K \cos i - k}{K \cos i + k}. \tag{2}$$

[9] R. H. Dicke, "Principles of Microwave Circuits," M.I.T. Radiation Laboratory Series, McGraw-Hill Book Co., New York, N.Y., vol. 8, pp. 447–448; 1948.

[10] This discussion constitutes a very slight extension of that given by B. A. Lippmann, "Theory of Directional Couplers," M.I.T. Radiation Laboratory Report 860, pp. 33–35; December 28, 1945, and R. L. Kyhl, "Directional Couplers," M.I.T. Radiation Laboratory Series, McGraw-Hill Book Co., New York, N.Y., vol. 11, pp. 889–890; 1947.

[11] J. F. Carlson and A. E. Heins, "The reflection of an electromagnetic plane wave by an infinite set of plates, 1," *Quart. Appl. Math.*, vol. 4, no. 4, pp. 313–329; January, 1947.

Equation (1), which states the condition for equal power division, contains the term ϕ_r which must be evaluated. According to the reciprocity theorem, the phase shifts at the beginning and at the end of the coupling section are equal and given by the phase of T_g, the transmission coefficient for the two in-phase plane waves above. It may be shown that

$$T_g = \frac{2K}{K \cos \theta + k} e^{i(\theta_1 - \theta_2)}. \quad (3)$$

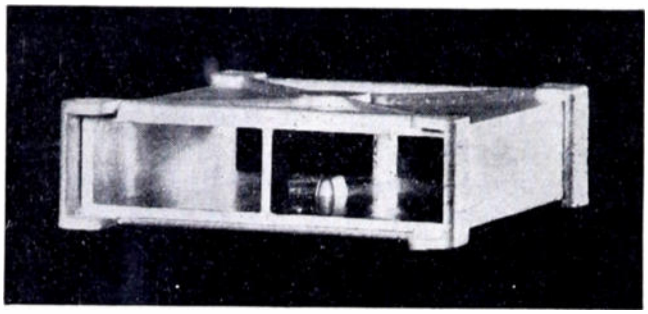

Fig. 2—Short-slot hybrid.

The actual short-slot hybrid shown in Fig. 2 differs from the schematic hybrid of Fig. 1 in two important respects: In the first place, capacitive domes have been provided in the coupling section, and secondly, the width of the coupling section has been reduced, by means of convex indentations, to below that of the twowaveguides leading up to it.

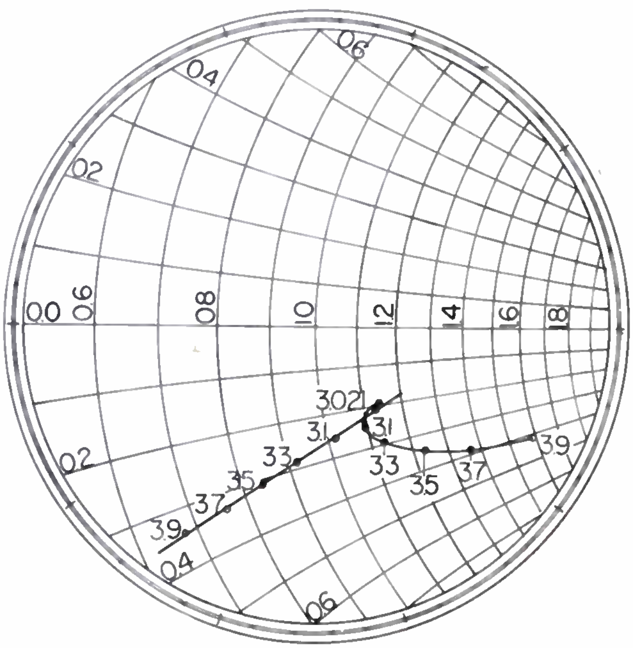

Fig. 3—Admittance of septum termination.

In early experiments, it was found that a slot length of 1.25 inches and a coupling section width corresponding to a=0.892 inch gave satisfactory hybrid performance over the frequency range from 8,500 to 9,600 mc per second. As a theoretical check, the admittance of the septated termination seen from the coupling section was calculated and has been plotted on a circle diagram as shown in Fig. 3. These points fall closely along a straight line over the frequency range from 7,700 to 9,900 mc per second. When these admittances are plotted relative to the center of the coupling section, the other curve of the figure is obtained. It is clear that a centrally located capacity may be provided which will give high isolation over relatively broad bands of frequencies.

Fig. 4 shows the relative phase shift due to the length of the aperture and also that due to the ends of the aperture. The compensating tendencies of these phase shifts explains the broad-band power-division characteristics obtained. It will be noticed that the sum of these phase shifts differs from the required 90 degrees. Of course, the central capacity and multiple reflections (ignored in this discussion) supply the remaining phase shift.

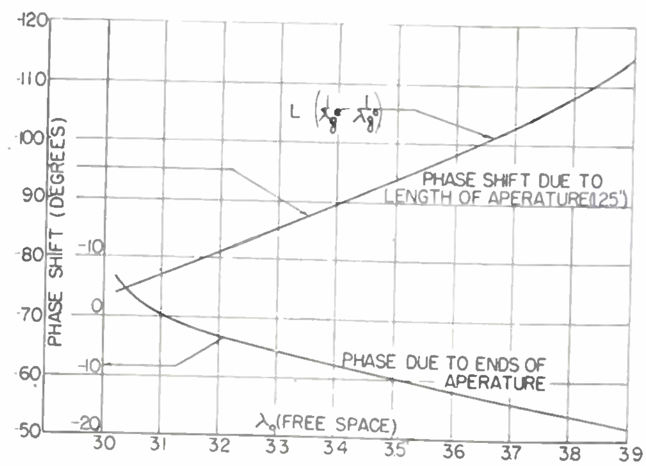

Fig. 4—Relative phase shift.

Present Performance

The present design of the short-slot hybrid junction was arrived at by a process of iterated selections from a small number of alternatives. As far as is known, it is not the optimum design and no effort has been made to obtain precise agreement between theory and practice. The theory is quite useful, however, in predicting the general effect of small changes in dimensions. Fig. 5 shows the best performance which has been obtained with one of the short-slot hybrids.

Applications

A review of all the applications of the hybrid junction would cover much of the waveguide art. It will be suf-

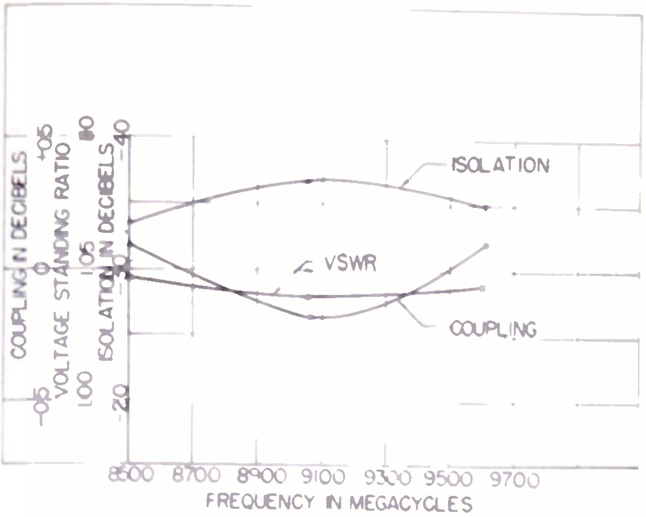

Fig. 5—Best performance.

ficient for our purposes to limit this discussion to two of the most important—namely the balanced mixer and the balanced duplexer. Fig. 6 shows a balanced mixer which uses this type of hybrid junction. Careful examination of the phase relationships shows that the 90-degree phase-shift characteristic of the hybrid junction in no way affects the ability of the mixer to discriminate against local oscillator noise. Pound[12] has gone through this argument in detail and indicated certain advantages for this arrangement in so far as image frequency power is concerned.

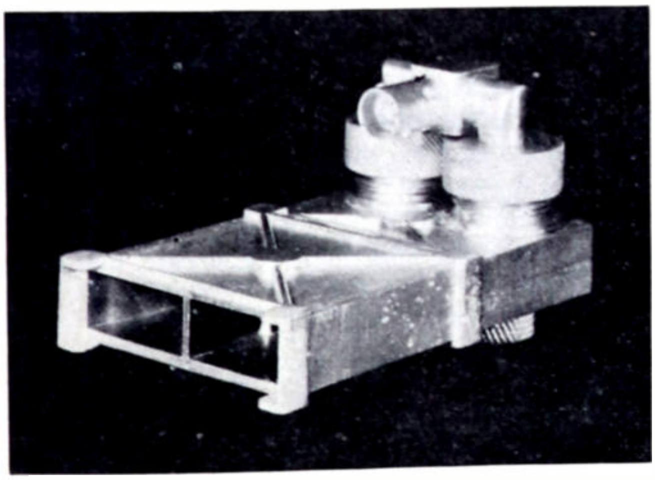

Fig. 6—Balanced mixer.

The principal advantage of the short-slot hybrid for mixer applications results from the close spacing between crystals which is made possible. Not only does it lend itself to more compact IF amplifier construction using the conventional balanced input transformer design, but it becomes a very simple matter to operate the crystals with a common dc bias. Fig. 6 shows such a mixer.[13] This circuit requires that one of the crystals be inverted with respect to the other. The possible advantages of this circuit over the conventional arrangement using a balanced input transformer appear to be a simpler mechanical arrangement, a perceptible tendency for the rf impedances of the crystal to balance more closely with a common dc bias, and a better over-all noise figure due possibly to the elimination of the closely coupled coils and their inherent losses and unbalances. Unfortunately, the mixer measurements were made with a small number of crystals, so the conclusions regarding noise figure and impedance balance must be considered tentative.

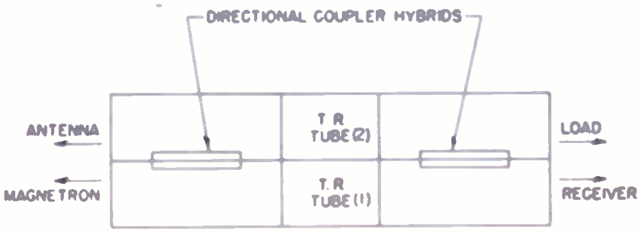

Fig. 7—Schematic balanced duplexer.

Fig. 7 gives a schematic drawing of a balanced duplexer using this type of hybrid. The inherent compactness of this arrangement as compared with the "model city" duplexer described by John Reed is apparent. The 90-degree phase-shift characteristic of the hybrid makes it possible to construct a perfectly symmetrical circuit without a frequency-critical quarter-wave-line length difference. This comes about as follows: Energy from the magnetron splits at the hybrid. The voltage crossing over leads the voltage going straight through by 90 degrees. At the TR tubes, which fire and thus act as short circuits, the energy is reflected without relative phase shift. Voltages in the magnetron branch arising from reflection at TR tube (2) experience an additional 90-degree phase advance, and thus destructively interfere with those which are reflected from TR tube (1). On the other hand, they reinforce in the antenna branch. On reception, the signal level is insufficient to fire the TR tubes, and the energy passes through the second hybrid when it recombines in the receiver branch, with little signal lost in the load.

[12] R. V. Pound, "Microwave Mixers," M.I.T. Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 16, pp. 277–279; 1948.

[13] The dc series features of this mixer were investigated by the author while employed at the Submarine Signal Co., Boston, Mass. The work was done under a subcontract supported by the Electronics Div., Glenn L. Martin Co., Baltimore, Md.

Fig. 8 is a photograph of a portion of a balanced duplexer which has been designed by the Reeves Instrument Company using these hybrids. One of the short-slot hybrid junctions is shown in the figure together with the two TR tubes and an ingenious mechanism for clamping the tubes in place.

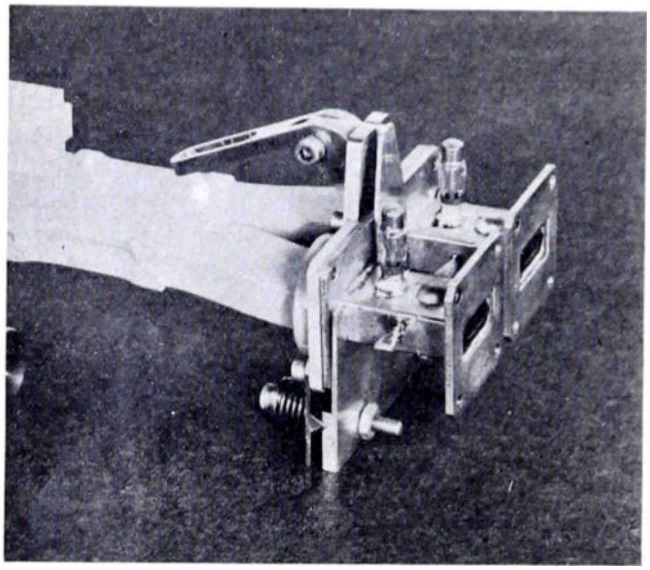

Fig. 8—Balanced duplexer half.

It is clear that the TR tubes may be replaced by shutters to obtain a broad-band mechanical switch or by band-rejection elements in which case one has the elements of a rejection filter. In addition, if the TR tubes are replaced by a pair of ganged movable pistons, a broad-band phase shifter or line stretcher results.[14]

Acknowledgments

I am indebted to Miss Eileen Quigley for the calculation on which the graphs of Figs. 3 and 4 are based. The numerical values were obtained by straightforward summation of the series involved. Dr. Bela Lingyel of the Naval Research Laboratory provided me with unpublished calculations based on very careful summation of these series. Fortunately, the two calculations agree (wherever they overlap) within the accuracy with which the graphs are plotted.

The photograph in Fig. 8 was provided by Mr. John Guarrera of the Reeves Instrument Company.

Appendix

It is useful to examine qualitatively some of the consequences when the hybrid junctions do not have perfect performance. The voltages B_i reflected out of the four terminals when voltages A_j are incident on them are given by the matrix equation

$$(B_i) = (S_{ij})(A_j),$$

where (S_{ij}) is the scattering matrix of the network. Since the network is lossless, the matrix (S_{ij}) is unitary.[15] This and symmetry imply the following relationships between S_{11}, S_{12}, S_{13}, and S_{14}:

$$\begin{aligned} |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 + |S_{14}|^2 &= 1 \\ S_{11}S_{12}^* + S_{12}S_{11}^* + S_{13}S_{14}^* + S_{14}S_{13}^* &= 0 \\ S_{11}S_{13}^* + S_{12}S_{14}^* + S_{13}S_{11}^* + S_{14}S_{12}^* &= 0 \\ S_{11}S_{14}^* + S_{12}S_{13}^* + S_{13}S_{12}^* + S_{14}S_{11}^* &= 0. \end{aligned} \quad \text{(a)}$$

From these equations, it may be concluded readily that

$$S_{14}S_{13}^* - S_{13}S_{14}^* = -2S_{13}S_{14}^* - S_{11}S_{12}^* - S_{12}S_{11}^*, \quad \text{(b)}$$

and

$$\begin{aligned} |S_{11}|^2\{S_{14}S_{13}^* - S_{13}S_{14}^*\} &= |S_{12}|^2\{S_{14}S_{13}^* - S_{13}S_{14}^*\} \\ &\quad + \{|S_{14}|^2 - |S_{13}|^2\}\{S_{11}S_{12}^* - S_{12}S_{11}^*\}. \end{aligned} \quad \text{(c)}$$

Since S_{12} is small compared with S_{13} and S_{14}, the coefficient of $|S_{11}|^2$ in (c) cannot vanish. Thus the vanishing of S_{12} implies that $S_{11}=0$. In words, complete isolation implies perfect match. Moreover, if $|S_{14}| = |S_{13}|$ we conclude immediately that $|S_{11}| = |S_{12}|$. If $|S_{14}|$ and $|S_{13}|$ differ by 0.25 db, which is the maximum inequality of output for these hybrids, the S_{11} will differ from S_{12} by an error of at most 2.2 per cent. Thus, for all practical purposes, we may immediately infer the maximum SWR from the measurement of isolation. In fact, the isolation measurement can be made with more accuracy than the SWR measurement.

If we write $S_{1j} = A_j e^{i\phi_j}$, we may conclude from the second equation of (a) that

$$A_1A_2 \cos(\phi_1 - \phi_2) = -A_3A_4 \cos(\phi_3 - \phi_4). \quad \text{(d)}$$

Assuming, in the most unfavorable case, that $\phi_3 = \phi_4$, we have

$$|\cos(\phi_3 - \phi_4)| \leq \frac{A_1A_2}{A_3A_4}.$$

Now the voltage isolation is given by

$$\frac{A_1}{\sqrt{2}A_3} \approx \frac{A_2}{\sqrt{2}A_4}.$$

Hence the isolation immediately determines the maximum amount by which the output phases may differ from 90 degrees. For 20 db of isolation the maximum phase error is of the order of 1 degree; for 30 db of isolation the possible error has dropped to 0.1 degree.

[14] This very interesting application was suggested to the author by Mr. Werner Koppl of the Glenn L. Martin Co.

[15] R. H. Dicke, ibid., pp. 148–149.

A Statistical Approach to the Measurement of Atmospheric Noise*

ROBERT S. HOFF†, MEMBER, IRE AND RAYMOND C. JOHNSON‡, ASSOCIATE, IRE

Summary—**A method of measuring and describing atmospheric noise based on statistical considerations is presented, and apparatus for making the measurement is described. Results obtained are compared on a statistical basis with those obtained by methods of noise measurements in common use. Data obtained at low frequencies are discussed.**

Introduction

THE PROBLEM of measuring atmospheric noise has been under study at the University of Florida since May, 1946 as a part of a low-frequency noise and wave-propagation investigation sponsored by the Watson Laboratories of the Air Force (Contract W28-099-ac-152). One of the objectives of the program is to establish a correlation between a measure of atmospheric noise and error in matching low-frequency loran pulses. In searching for a measurable characteristic of noise that would serve as a reliable index of its interference effect, the conclusion was reached that measurement of something more definitive than quasi-peak and average values are desirable so that correlations with interference effect can be improved under a wide range of noise conditions. The amplitude-probability-distribution method of measurement presented here was adopted and developed to meet the requirement.[1,2,3]

On Measuring Atmospheric Noise

The term "atmospheric noise" is used broadly to define any interfering radio waves generated by electrical disturbances of the atmosphere. According to the currently accepted theory,[4] these disturbing electromagnetic waves originate primarily from lightning flashes, and have energy components throughout the radio-frequency spectrum. Considering the chance occurrence of lightning flashes in time and the variability of flash orientation, current waveform, and conditions over the propagation path, the instantaneous noise voltage induced in a receiving antenna will not depend in any regular way on time as a variable. The same may be said of the envelope of the noise voltage even after the latter has been amplified by a receiver with a restricted pass band and further modified by limiting and other nonlinear processes in the receiver. This random-like voltage envelope appearing at the detector output is the noise voltage, the measurement of which is considered here. As an irregular function of time or time series $y(t)$, it may be treated by statistical methods.

Atmospheric noise is generally specified in terms of its average, rms, or quasi-peak value since these parameters can be measured conveniently with conventional equipment. Actually, however, the average and the mean-square values of a sample of noise of duration Δt are simply the first and second statistical moments of the time series for the period.[5] The nth moment is given by

$$M_n = \int_{-\infty}^{\infty} y^n f(y) dy, \tag{1}$$

where M_n is the nth-order moment, y is the amplitude variable, and $f(y)$ is the probability density function defining the probability of occurrence of various values of y. In the form for the moment given, the noise $y(t)$ is taken as a stationary time series during the period Δt, which makes $f(y)$ a one-dimension distribution function. Measurements made with equipment to be described later substantiate the assumption that the series is essentially stationary if the period of observation is of the order of one to ten minutes.

To describe a stationary time series by the method of moments, it is necessary to specify not only the first and second moments but the higher order moments as well.[6] The one exception is a density function following the Rayleigh Law, in which case it has been shown that either the first or second moment is sufficient.[7] Unfortunately, practical methods of measuring the higher order moments of atmospheric noise have not been brought to light. As a matter of record, the possibility of describing the noise condition by a measure of its quasi-peak value is even more remote, for such a measure bears no simple relation to a complete statistical description.

The alternate approach is, of course, to consider the direct measurement of the density function. The follow-

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† National Bureau of Standards, Washington, D. C.
‡ Assistant Research Engineer, Engineering and Industrial Experiment Station, University of Florida, Gainesville, Fla.

[1] The statistical approach was considered at the suggestion of the late Dr. G. W. Kenrick of the University of Puerto Rico.

[2] K. Maeda and H. Yokoyama, "Automatic Measurement of Atmospheric Noise," Report of Radio Research in Japan, Vol. 7, pp. 153–157; 1937. (Summary only was available to the author.)

[3] H. O. Peterson, "A method of measuring noise levels on short-wave radio-telegraph circuits," Proc. I.R.E., vol. 23, pp. 129–134; February, 1935.

[4] H. A. Thomas and R. E. Burgess, "Survey of Existing Information and Data on Radio Noise over the Frequency Range 1–30 mc/s," Department of Scientific and Industrial Research, Radio Research, Special Report No. 15, H. M. Stationery Office, London, England, pp. 24–25. (Contains a 186-item bibliography.)

[5] "Statistical Theory of Noise," Report No. 102, Research Division, Philco Radio Corp., p. 36.

[6] H. Cramer, "Mathematical Methods of Statistics," Princeton University Press, Princeton, N. J., p. 176; 1947.

[7] "Statistical Theory of Noise," *op. cit.*, pp. 59–60.

ing two probability functions are directly related to the probability density function, and lend themselves to direct measurement.

(a) The probability that the amplitude will fall within the limits y_0 and $y_0+\Delta y$ during the time interval Δt, where the probability is given by

$$P(y_0, y_0 + \Delta y) = \int_{y_0}^{y_0+\Delta y} f(y)dy. \qquad (2)$$

(b) The probability that the amplitude will be greater than y_0 during the period Δt, where the probability is given by

$$P(y > y_0) = \int_{-\infty}^{y_0} f(y)dy. \qquad (3)$$

Of the two forms given, the latter has been chosen, for reasons of circuit simplicity, as the basis for a practical measuring instrument.

The Measuring Method and Apparatus

Solution of (3) by electrical methods is accomplished by measurement of the per cent-of-time or probability that the noise envelope exceeds a reference level y_0 during the period Δt. To establish the probability during the period for values of y_0 throughout the range of amplitude variation, simultaneous measurements should theoretically be made at a number of reference levels. The period Δt should be sufficiently long so as to encompass a representative cross section of the noise in order to obtain reasonable statistical equilibrium and associated statistical significance. Practically, a Δt of the order of seconds is used, and measurements at the various reference levels are made consecutively, with from one to ten minutes being spent at each level to insure that a representative reading has been obtained. The result is a record of the time variation of probability of a sample of noise examined during the previous Δt seconds as the period advances in time. Measurement of probability at each level in turn introduces no error only if the function $f(y)$ does not change during the complete measuring cycle. Experience shows that the curves change slowly with time, except during sudden thunderstorms, sunrise, and sunset. It is noted here that more complex apparatus for making measurements simultaneously at all levels is a straightforward design problem, and would undoubtedly contribute to the accuracy of measurements during periods of rapidly changing noise conditions.

Operation of the practical measuring apparatus, shown in block-schematic form in Fig. 1, is based on the consideration that the average value over a period Δt of a series of rectangular pulses of equal amplitude, the width and spacing of which are determined by chance, is proportional to the per cent-of-time in which pulses occurred. This consideration is related to the per cent-of-time that a noise-envelope voltage exceeds a particular reference level by causing a current of fixed value to be interrupted suddenly whenever the reference level is exceeded, thus forming pulses which can be averaged as outlined above. The time constant of an RC network establishes the period Δt over which pulses are averaged. Per cent-of-time, in this case, is probability as it applies to past events.

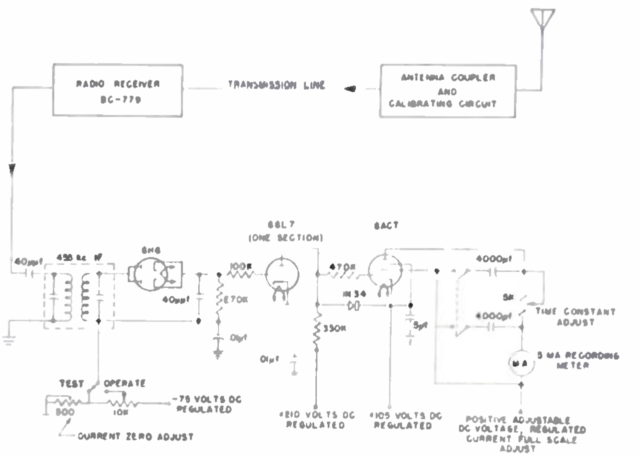

Fig. 1—Functional diagram of measuring system.

A dummy antenna capacitor is included in the antenna coupling unit for calibration of the system at any frequency in the range 100 to 400 kc. Operation of the receiver on the linear portion of its input-output characteristic is insured by using a step attenuator in the antenna circuit. Errors arising from the use of automatic gain control are thereby eliminated.[8]

In the measurement section of the system, current pulses are formed in the plate circuit of the 6AC7 tube by the action of the crystal diode and a series grid resistor. Adjustment is accomplished by setting the *current-zero adjust* resistor to maximum with the *test-operate* switch in the *test* position, adjusting the *current full-scale adjust* control for a 5-ma deflection of the recording meter and then resetting the *current-zero adjust* control for zero current in the recording meter.

Accuracy and operating flexibility of the system described can be improved by providing more gain following the diode detector, by the addition of an automatic attenuator and bias-level switching system, by providing for an up-scale reading meter and by incorporating a servosystem for automatic setting of the bias level when recording on a constant-percentage basis is desired. Many of these improvements are being considered for inclusion in improved models.

Presentation and Interpretation of Data

Raw data produced by the instrument described can be presented in their most fundamental form as a probability distribution curve for each period over which a cycle of measurement was made. Fig. 2 shows a typical sample of a recorder chart containing three cycles of measurement. Examples of plotted data are given in

[8] H. E. Dinger and H. G. Paine, "Factors affecting the accuracy of radio noise meters," Proc. I.R.E., vol. 35, pp. 75–81; January, 1947.

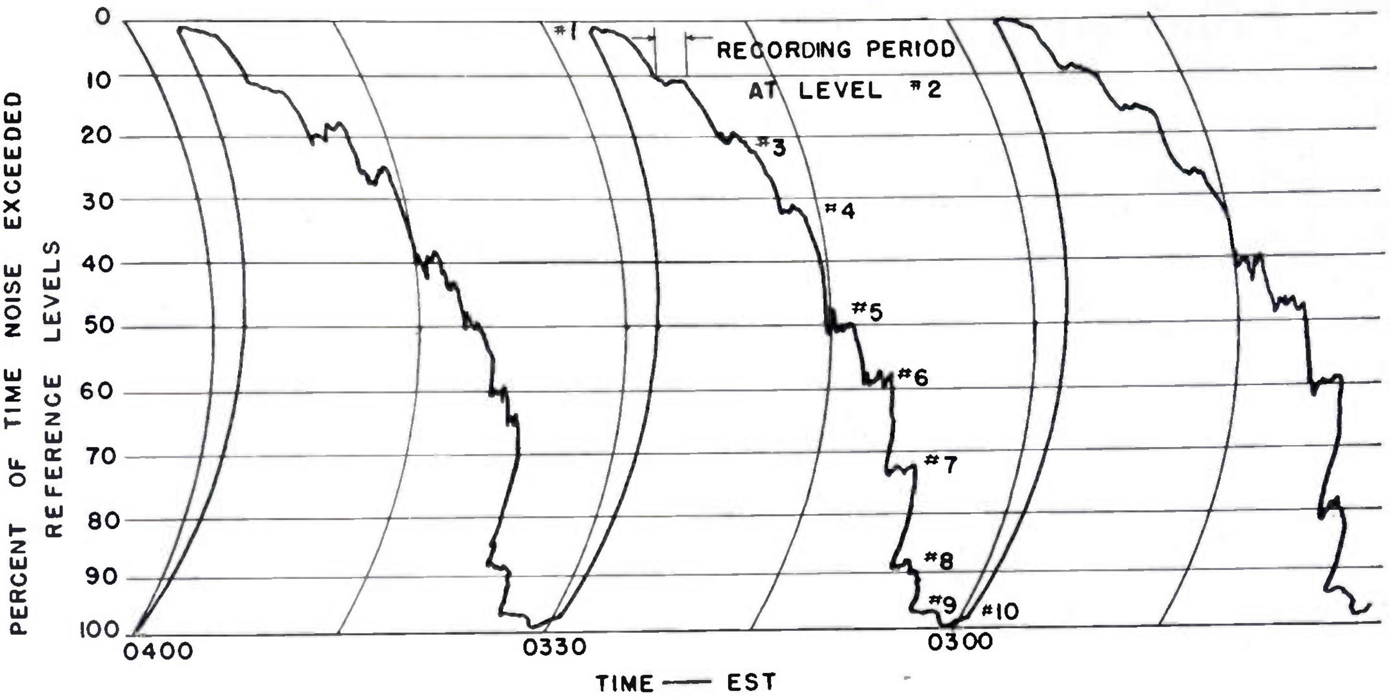

Fig. 2—Typical sample of recorder chart.

Fig. 3 for fluctuation noise and for atmospheric noise at three different times. All have been normalized to reduce the effect of general noise-amplitude level changes. The fluctuation noise curve follows the integral of the Rayleigh distribution closely, and provides a check on the proper operation of the system. Differences in the curves for atmospherics clearly point up the variations which may be expected.

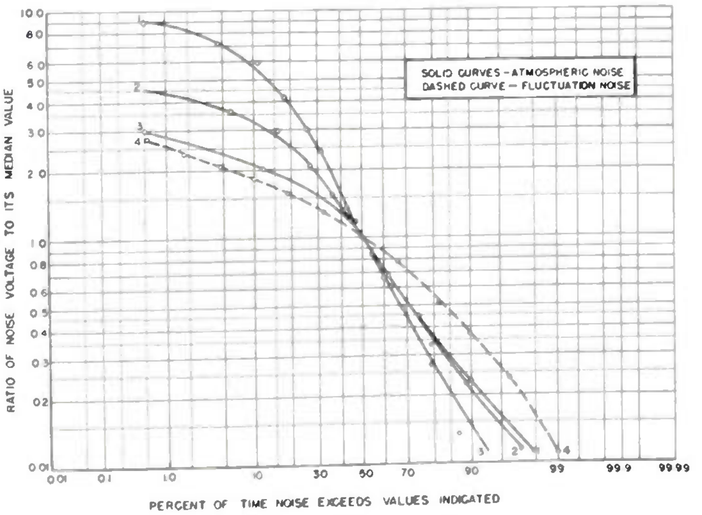

Fig. 3—Distribution curves of atmospheric and fluctuation noise.

Admittedly this and other obvious graphical presentations of the statistical measures of noise are time consuming to prepare and test for correlation with the interference effect of the noise. This problem is recognized, and efforts are being made to develop a mathematical expression that will fit the measured curves. In any case, the graphical presentations should serve as a basis for determining just which characteristics of noise play the major part in interfering with particular radio services whenever the interference effect can be evaluated on a quantitative basis.

Conclusions

The primary purpose of this report has been to point out the basic shortcomings of the conventional methods of measuring atmospheric noise and to present a more fundamental statistical method of measurement which shows promise of giving a much more accurate description of the noise condition. The amplitude-probability-distribution measurement provides this improved description, and should, therefore, prove to be a useful tool for the investigation of noise-interference effect and for the study of the thunderstorms from which atmospherics originate.

As for the apparatus described, it served the special purpose of demonstrating the practicability of the statistical measurement and furnished quantitative information as to the nature of and variations in the statistical characteristics of atmospheric noise at low frequencies. The operating principle is applicable to any probability distribution measuring problem in which data, either smooth or with discrete values, can be converted into a voltage with a corresponding amplitude variation.

There is ample evidence to show that objective atmospheric-noise measurements made in the past have not been used in communication-system planning without the application of a large factor of safety to compensate for the inadequacies of the noise data available. It is hoped, therefore, that the information presented here will provide the means whereby the interference effect of noise may be more accurately assessed and taken into account.

Shorted Stubs of High Resonant Impedance*

JOSEPH M. DIAMOND†, ASSOCIATE, IRE

Summary—**Curves are developed which specify the parameters of a coaxial or twin-line shorted stub of maximum resonant impedance when the frequency, outer dimension, and lumped capacitance at the open end are given.**

We are concerned with obtaining the maximum resonant impedance from a shorted length of coaxial or twin transmission line when the open end is loaded with lumped capacitance. Thus the line will be less than a quarter wavelength long. The quantity held constant will be the outside diameter for the coaxial line and the center-center spacing for the twin line.

When the lumped capacitance is zero, a quarter wavelength line is required; the optimum dimensions of this line for maximum impedance have been derived by Terman.[1] However, it is often impossible to avoid some lumped capacitance at the open end (if, for example, capacitive tuning is desired). The present investigation was, in fact, prompted by the necessity of obtaining a high-impedance tuned circuit at 400 mc for the measurement of high resistance by the susceptance-variation method. The curves presented here allow a line of maximum impedance to be calculated when the lumped capacitive reactance and outside line dimensions are given.

The curves are based on an extension of a paper by Nergaard and Salzberg,[2] who develop the following equation for resonant impedance from the fundamental transmission-line equations:

$$r = \frac{Z_0}{k}\left(\frac{1 - \cos 2\theta}{2\theta + \sin 2\theta}\right), \tag{1}$$

where

r = resonant impedance
Z_0 = characteristic impedance
θ = line length (degrees)
$k = R_0/2\omega L_0$
R_0 = line resistance per unit length
L_0 = line inductance per unit length
ω = angular frequency
C = lumped capacitance at open end.

To produce resonance the following condition must be satisfied:

$$\omega C Z_0 = \cot \theta. \tag{2}$$

Thus, if the line constants were known, the required θ and resulting r could be found. Since R_0 (including skin effect) and L_0 depend upon line dimensions, the design problem is more involved. The relations given below in Table I are well known.

TABLE I

Coaxial Line	Twin Line
$Z_0 = 138 \log b/a$ ohms	$276 \log d/a$ ohms
$L_0 = 4.60(10^{-7}) \log b/a$ henries per meter	$9.21(10^{-7}) \log d/a$ henries/meter
$R_0 = 4.16(10^{-8})\sqrt{f}(1/a - 1/b)$ ohms per meter (for copper)	$8.33(10^{-8})\ \sqrt{f}/a$ ohms/meter (for copper)
a = inner radius (cm)	a = conductor radius (cm)
b = outer radius (cm)	d = center-center spacing (cm)
f = frequency in cps all logs to base 10	

Combining these equations with (1) and (2), we obtain

$$\text{Coaxial} \quad \frac{r}{b\sqrt{F}} = \frac{219 X_c^2}{1 + 10^{X_c/138 \tan\theta}} \cdot \phi(\theta) \tag{3}$$

$$\text{Twin} \quad \frac{r}{d\sqrt{F}} = \frac{100 X_c^2}{10^{X_c/276 \tan\theta}} \cdot \phi(\theta), \tag{4}$$

where

$$\phi(\theta) = \frac{0.458}{\tan^2\theta}\left(\frac{1 - \cos 2\theta}{2\theta + \sin 2\theta}\right)$$

$$= 0.458\left(\frac{1 + \cos 2\theta}{2\theta + \sin 2\theta}\right) \tag{5}$$

$$X_c = \frac{1}{\omega C}$$

F = frequency in hundreds of mc.

Equations (3) and (4) give the resonant impedance obtained with a given X_c, F, outside line dimension, and θ, when Z_0 is adjusted to produce resonance. The equations are plotted in Figs. 1 and 2 as functions of θ, with X_c as parameter. For a given X_c, the peak of the curve shows that there is an optimum value of θ. The corresponding Z_0 is given by (2). Fig. 3 shows the optimum impedance obtained as a function of X_c, as well as the optimum values of θ and Z_0; these curves were obtained from the peak points of Figs. 1 and 2. As X_c becomes infinite, the values approach those given by Terman for the optimum self-resonant 90-degree line.

If the conductor is other than copper, the impedance obtained will be proportional to the square root of the conductivity, but otherwise all curves apply. Radiation, dielectric, and shielding losses are not included in this analysis. Thus, better agreement with calculated re-

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† Burroughs Adding Machine Co., Research Division, 511 N. Broad St., Philadelphia 23, Pa.

[1] For the original paper, see F. E. Terman, "Radio Engineers' Handbook," p. 191; also, F. E. Terman, "Resonant lines in radio circuits," *Trans. AIEE*, vol. 53, p. 1046.

[2] Nergaard and Salzberg, "Resonant impedance of transmission lines," Proc. I.R.E., vol. 27, p. 579.

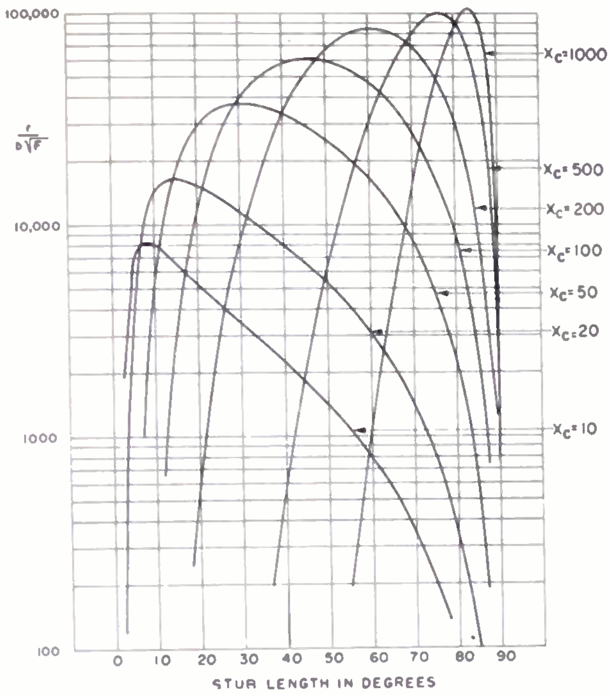

Fig. 1—Resonant impedance of a shorted coaxial stub. For resonance, $Zo = X_c/\tan\theta$
b = Outer radius in cm
F = Frequency in hundreds of mc
r = Resonant impedance in ohms
X_c = Lumped capacitive reactance at open end

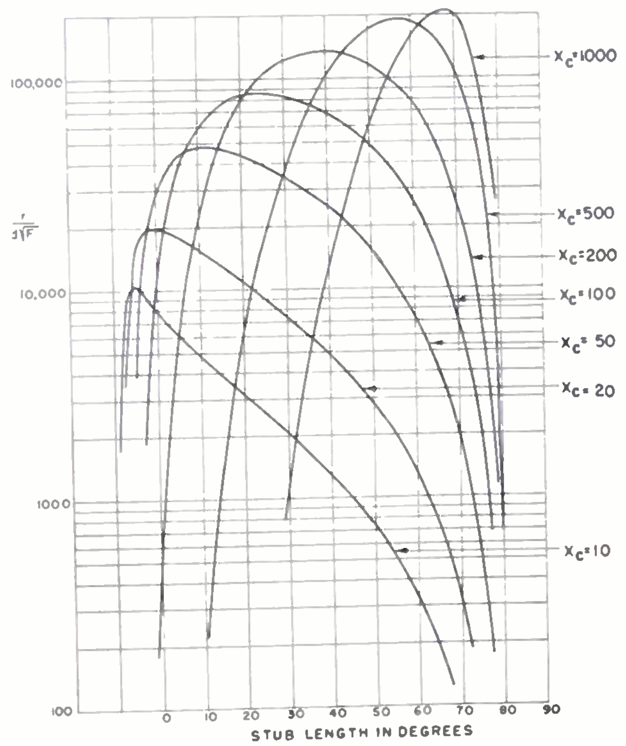

Fig. 2—Resonant impedance of a shorted twin-line stub. For resonance, $Zo = X_c/\tan\theta$
d = Spacing in cm
F = Frequency in hundreds of mc
r = Resonant impedance in ohms
X_c = Lumped capacitive reactance at open end

sults can be expected for the coaxial line, which is self-shielding, than for the twin line, which is affected by surrounding materials.

For example, a high-impedance coaxial stub is desired at 200 mc; the inner diameter of the outer conductor = 2 inches and the lumped capacitance = 10 $\mu\mu$f. We have, therefore, $F=2$, $b=2.54$ cm and $X_s=79.6$ ohms. Fig. 3 gives the parameters of the optimum line as $Z_0=97$ ohms, $\theta=39$ degrees, and $r/b\sqrt{F}=53$ kilohms. Therefore, the diameter of the inner conductor $=2/10^{97/138}$ $=0.396$ inches, the stub length $=(39/360)(300/200)(39.37)=6.40$ inches, and the resonant impedance $=(53{,}000)(2.54)\sqrt{2}=190{,}000$ ohms. These dimensions are not critical, however. The following table shows the impedance obtained and stub length requires when the inner conductor diameter is varied from $\frac{1}{4}$ to $\frac{3}{4}$ of an inch. Since it is difficult to interpolate between the curves of Figs. 1 and 2 in the neighborhood of the maxima, the values in Table II were calculated from (2) and (3):

TABLE II

Diameter of inner conductor, inches	$\frac{1}{4}$	$\frac{3}{8}$	$\frac{1}{2}$	$\frac{5}{8}$	$\frac{3}{4}$
Stub length required, inches	5.32	6.32	7.19	7.98	8.80
Resonant impedance, kilohms	177	190	190	176	154

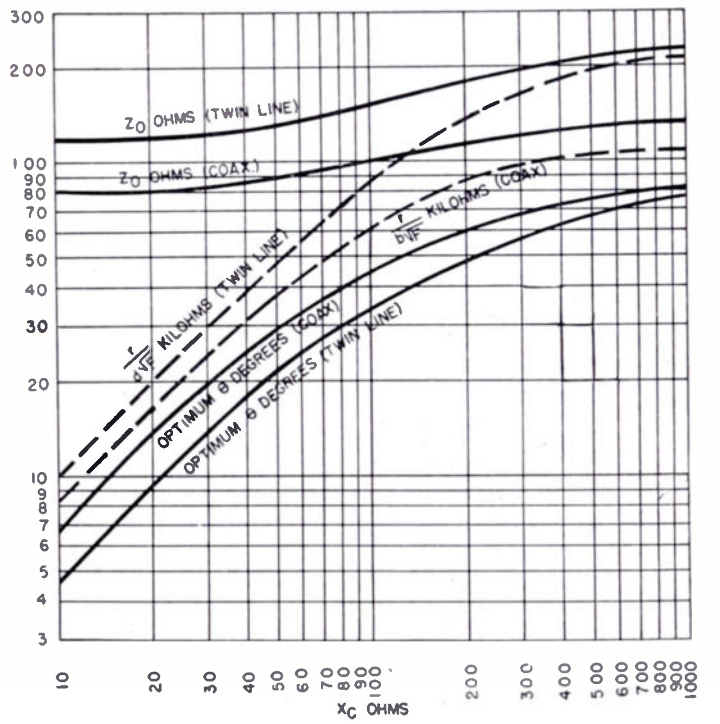

Fig. 3—Parameters and performance of the optimum line as a function of X_c.

The Influence of the Core Material on the Thermionic Emission of Oxide Cathodes*

H. A. POEHLER†, ASSOCIATE, IRE

Summary—**The influence of the core material on the thermionic emission of oxide cathodes was investigated. Alloys of nickel with 4.8 per cent Mn, 4.0 per cent Al, 0.38 per cent Mg, and 3.5 per cent W were used as cores, with pure electrolytic nickel as a core being used as a control. The experiments showed that both the dc and pulsed emission of oxide cathodes are dependent on the core to a marked degree.**

Introduction

THE EFFECT of the core material on the thermionic emission of oxide cathodes was early discounted by the work of Deinninger.[1] Lowry[2] and Beese[3] reopened the question in the 1930's, and Benjamin[4] contributed to it in 1935. Since 1939, however, the problem of the influence of the core material has received increasing attention.[5–12]

To date, the only work on the effect of the more common impurities found in commercial, cathode-type nickel on the thermionic emission of nickel-base oxide cathodes is that of Benjamin.[4]

Among other materials, the effect of which was studied by Benjamin, were 0.07-per cent Mg-, 0.34-per cent Mn-, and 2-per cent Al-nickel alloys, used as core material for oxide cathodes. His work, however, is open to several criticisms. The most serious of these is his measurement of saturation currents at such high temperatures as 1,020° K under dc conditions. Such measurements are subject to the following errors:

(a) Poisoning by evolution of gases from the anode.
(b) Poisoning by evolution of gases produced by the decomposition of anode deposits.[13,14]
(c) Heating of the coating by the I^2R loss in the coating itself, caused by the resistance of the coating. This is a serious criticism of Benjamin's work since he made the assumption that the "brightness temperature would depend only on the watts supplied to the filament." Thus, it is difficult to say what part of the differences in emission noted by Benjamin were due to, or were masked by, differences in I^2R loss in the coating.

Moreover, the use of two cathodes in one envelope is open to question because of the possibilities of affecting the emission of one cathode by the gases given off by the other cathode.

Discussion of Methods

A. Vacuum Systems

The system was pumped with a three-stage, glass-fractionation pump, using Octoil-S. The vacuum was measured with a distillation products, VG-IA ionization tube. A trap was located between the diffusion pump and the manifold. This trip was baked to 400° C each time the manifold was baked to 450° C. With this system, the pressure as measured at the gauge was consistently lower than 1×10^{-7} mm Hg, and generally 6×10^{-8} mm at tipoff.

B. Description of the Tube

The tube used in these experiments was the "Coomes Diode,"[15] which was developed by the Radiation Laboratory at M.I.T. The tube is illustrated in Fig. 1. The anode and leads are made of kovar. The tube is so designed as to permit the anode to be water cooled. The anode can then be operated at low temperatures, even for a relatively high anode dissipation. The design is also such as to allow the tube to be sealed with a minimum oxidation of the tube parts.

In order to measure cathode temperatures below 800° C, a thermocouple was added to the tube. The

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† General Precision Laboratory, Inc., Pleasantville, N. Y.

[1] F. Deinninger, "On the emission of oxides," *Ann. Phys.* (Lpz), vol. 25, p. 285; 1908.

[2] E. Lowry, "The role of core metal in oxide cathodes," *Phys. Rev.*, vol. 35, p. 1367; 1930.

[3] N. Beese, "Ba-Ni alloy core for oxide cathodes," *Phys. Rev.*, vol. 36, p. 1309; 1930.

[4] M. Benjamin, "The influence of metallic impurities in the core of oxide cathodes," *Phil. Mag.*, vol. 20, p. 1; 1935.

[5] W. Liebold, "Dissertation," University of Berlin, Berlin, Germany; 1941.

[6] A. Fineman and A. Eisenstein, "Studies of the interface of oxide cathodes," *Jour. Appl. Phys.*, vol. 17, p. 663; August, 1946.

[7] H. Jacobs and G. Hees, "Variations in the constants of Richardson's equation as a function of life for the case of oxide coated cathodes on nickel," *Phys. Rev.*, vol. 72, p. 174; July, 1947.

[8] W. Mutter, "Rectification characteristics of an oxide cathode interface," *Phys. Rev.*, vol. 72, p. 531; September, 1947.

[9] D. Wright, "Oxide cathodes, the effect of coating-core interface on conductivity and emission," *Proc. Roy. Soc. A*, vol. 190, p. 394; 1947.

[10] J. Acker, "A.S.T.M. committee work—factory tests on cathode Ni," *Proc. I.R.E.*, vol. 36, p. 376; March, 1948.

[11] H. Jacobs, G. Hees, and W. Crossley, "Thermionic emission of oxide coated filaments," *Bull. Amer. Phys. Soc.*, vol. 23, p. 12; 1948.

[12] R. McCormack, "A standard diode for radio-tube-cathode core material approval tests," *Proc. I.R.E.*, vol. 36, p. 376; March, 1948.

[13] H. Jacobs, "Dissociation energies of surface films of various oxides as determined by emission measurements of oxide cathodes," *Phys. Rev.*, vol. 69, p. 692; 1946. Also, *Jour. Appl. Phys.*, vol. 17, p. 596; July, 1946.

[14] H. Bruining, H. Hamaker, and A. Alten, Jr., "On the activation of oxide cathodes," *Philips Res. Rep.*, vol. 2, p. 171; 1947.

[15] E. Coomes, J. Buck, A. Eisenstein, and A. Fineman, "Alkaline Earth Oxide Cathodes for Pulsed Tubes," N.D.R.C., Div. 14, OEMsr 262, Report 933; March 30, 1946. (Publ. No. PB 55739, Office of Technical Services, U. S. Dept. of Commerce.)

couple was composed of an extremely fine, 0.002-inch diameter molybdenum wire, which was welded onto the cathode sleeve immediately adjacent to the coating.

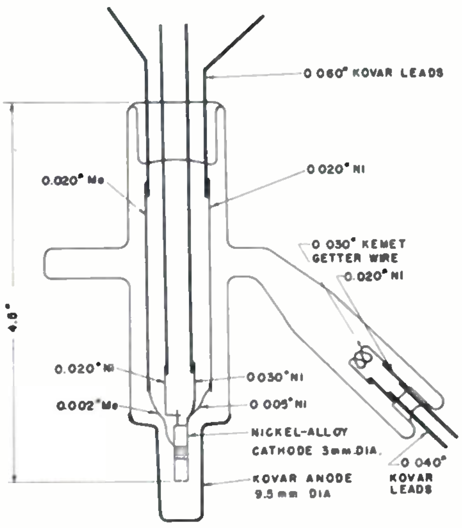

Fig. 1.—Diode construction.

To minimize end effects, only the center 4 mm of the 14-mm long cathode were coated. Before tipoff, all tubes were gettered by flashing an iron-clad barium getter, which had been carefully degassed before cathode activation.

C. Techniques and Processing

The utmost attention was paid to the adherence of a uniform processing technique for all the tubes. The processing techniques[15] developed by the M.I.T. Radiation Laboratory were taken as the basis for this work, with the following exceptions: Al-, Mg-, and Mn- nickel-alloy cathodes were only vacuum fired, not hydrogen fired; Mn-nickel alloy cathodes were vacuum flashed at 950° C instead of 1,000° C because of the volatility of the manganese; and all the tubes were continuously flushed with dry nitrogen during sealing to minimize oxidation of the leads.

The salient points of the processing technique may be briefly enumerated. The kovar anode is outgassed by electron bombardment at 1,500 volts from a dummy tungsten filament. The central portion of the anode is heated to 800° C (brightness temperature) until the pressure is 5×10^{-7} mm. The cathodes are vacuum fired in a separate envelope. Each is flashed at 1,000° C for 5 minutes, and held at 850° C until the pressure is 2×10^{-7} mm. The fired cathodes are sprayed with an equimolecular mixture of barium and strontium carbonates in a nitrocellulose binder to a controlled coating weight of from 9 to 12 mg/cm.^{2} When the sprayed cathodes are ready for mounting, the diodes are carefully opened with a hot wire and the sprayed cathode is mounted in place of the tungsten filament that was used to bombard the anode. The technique is such that the processed parts are exposed to the air only for the irreducible time necessary for spraying, mounting, and sealing.

D. Emission Measurement

The most direct method of approach is to use direct-current measurements at the normal operating temperatures of from 700 to 900° C, and to increase the plate voltage until the plate current begins to saturate. To consider this current as the temperature-limited emission of the cathode for that temperature, and hence as a measure of the cathode emission, is a serious error.

The primary faults are the following:

1. The Anode Effect—It is extremely difficult, if not impossible, to construct and sufficiently outgas an anode so that it will not liberate gases when saturation currents are drawn from the cathode. This is primarily because the saturation currents of oxide cathodes are so high and oxide cathodes are readily poisoned by minute amounts of gases that will react with free barium.

2. I^2R Heating and Coating Changes—In drawing currents in the neighborhood of saturation, the I^2R loss in the coating itself becomes comparable to the heater input.

For these reasons, in order to measure reliably the the cathode-emission ability, it is necessary to take measurements at much reduced temperatures and currents, i.e., at from 150 to 400° C, 10^{-10}–10^{-6} a/cm^2. Under these conditions the anode dissipation, and hence the gas liberation from the anode, is kept at an absolute minimum; furthermore, the passage of current through the cathode is kept exceedingly small so as to cause a minimum of disturbance to the existing physico-chemical makeup of the cathode.

In an effort to overcome the anode limitation, efforts were made as early as 1930 by Thomson,[16] and later by Maddock,[17] Heinze,[18] and Patai,[19] to draw emission current in pulses. In this manner, high peak currents could be drawn at operating temperatures of from 700 to 900° C at low anode temperatures, and hence at low poisoning levels. Thomson was able to reach peak currents as high as 3a/cm^2 in this manner.

Within recent years, by use of microsecond pulses, it was found that oxide cathodes could deliver up to 100 a/cm^2, space-charge limited,[15,20] and up to 150 a/cm^2,

[16] B. Thomson, "High efficiency emission from oxide-coated filaments," *Phys. Rev.*, vol. 36, p. 1415; 1930.

[17] A. Maddock, "Activation of oxide cathodes," *Phil. Mag.*, vol. 19, p. 422; 1935.

[18] W. Heinze, "Calorimetric determination of work function for oxide cathodes," *Ann. Phys.* (Lpz), vol. 16, p. 41; 1933.

[19] E. Patai and G. Frank, "Emission constants of oxide cathodes," *Z. techn. Phys.*, vol. 16, p. 254; 1935.

[20] E. Coomes, "Pulsed properties of oxide cathodes," *Jour. Appl. Phys.*, vol. 17, p. 647; August, 1946.

produced by a dielectric breakdown or by an I^2R heating of the interface is not as yet clear.

A typical set of data illustrating the effect of core material on pulsed emission is shown in Fig. 5. The data

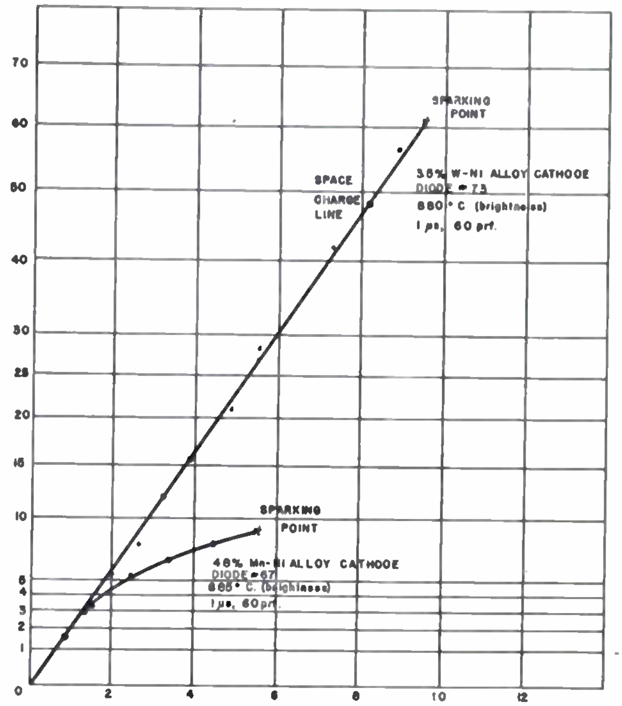

Fig. 5—Effect of the core material on the pulsed emission characteristics of experimental diodes. Ordinate is current in amperes per square centimeter; abscissa is applied anode potential in kilovolts. Draw on $\frac{3}{2}$-power paper.

was taken at 880° C., 1 μs, 60 prf. The emission characteristic of the 3.5-per cent W-Ni core tube is typical of that of a good emitter. No falling off of emission from the space-charge limited line could be determined, the emission being limited by sparking. The emission characteristic of the 4.8-per cent Mn-Ni tube is characteristic of a poor emitter. The emission falls away from the space-charge curve at low values, and is ultimately limited by sparking.

The accuracy of the low-temperature dc measurements is limited by the accuracy with which the cathode temperature could be measured. Uncertainty in determining the temperature of the cold junction of the thermocouple limits temperature measurements to ± 4 per cent. The emission-current measurements were made to ± 3 per cent, and the voltage measurements to within ± 2 per cent. Pulse measurements are within ± 5 per cent. The temperature, when expressed as "brightness" temperature, could be measured to ± 2 per cent.

Discussion of Results

The data of Table I show that the core material exerts a marked influence on the thermionic emission of oxide cathodes. The observed influence is in conformity with already published theories, as will be shown.

The work of Becker[27] and other[28] has shown that the oxide coating behaves as an impurity semiconductor, small amounts (in the order of 0.2 per cent) of free alkaline-earth metals furnishing the impurity. Furthermore, Prescott and Morrison[29] have shown that the emission of an oxide cathode is dependent on the amount of free barium present in the coating. They found that the thermionic emission increases with the barium content up to a concentration of 30 μg[30] of barium per cm² of superficial area. Free barium, however, can be produced by reaction of the core material with the oxide coating, such as

$$a\text{Ke} + b\text{BaO} \rightarrow \text{Ke}_a\text{O}_b + b\text{Ba},$$

where Ke represents the core metal.

Benjamin[4] advanced the thesis that the emission of nickel-alloy base oxide cathodes was related to the reducing power of the additive alloyed with the nickel. He used the heats of formation of the most likely formed oxides as a measure of the reducing power. In this manner he was able to explain the emission of 2-per cent Al-Ni, 0.07-per cent Mg-Ni, and 0.34-per cent Mn-Ni alloys. The emission of pure nickel, however, was out of place in this scheme, and Benjamin attributed this to small traces of reducing elements in the pure nickel, in spite of the fact that the other alloys were made using the same pure nickel as base.

In an investigation of the thermionic emission of oxide cathodes using eleven different core materials, Liebold found[5,28] that his results could not all be explained by the theory of Benjamin. He proposed a modified theory, into which the experimental data obtained here fits.

Benjamin divided the core materials into two groups: (a) those core materials for which the most probably formed interface oxides have a dissociation pressure, or a vapor pressure, higher than that existing in the tube (about 10^{-6} mm), at the operating temperatures; and (b) those core materials for which the dissociation pressure and vapor pressure of the most probably formed interface oxides are lower than the pressure existing in the tube.

In the first group, the interface compounds that are formed do not persist. As soon as they are formed, they decompose, giving off oxygen, or they are evaporated away. Hence, no interface is formed. For these cores, basing Ba formation on the reduction of BaO by the core, Liebold concludes that the thermionic emission should increase with the reducing power of the core. To

[27] J. Becker, "Phenomena in oxide-coated filaments," *Phys. Rev.*, vol. 34, p. 1323; 1929.
[28] G. Herrmann and S. Wagener, "Die Oxydkathode," Johann Barth, Leipzig, Germany; 1943.
[29] C. Prescott and J. Morrison, "The oxide-coated filament," *Jour. Amer. Chem. Soc.*, vol. 60, p. 3047; 1938.
[30] Per 1 mg BaO, 1 mg SrO of coating.

this extent Liebold's theory is the same as Benjamin's. Indeed, Liebold found that the thermionic emission of Au, Pt, Pd, Cu, and Ni increased with the heats of formation[31] of the oxides most probably formed at the interface. These oxides all have dissociation pressures above 3×10^{-6} mm.

For the second group, an interface compound is formed that persists. This interface, by its interposition, retards the reduction by the core metal of BaO to Ba, and does so to a larger degree as it grows in thickness. Indeed, Liebold found that for W, Mo, Ta, Cr, and Zr, the emission decreased as the heats of formation of the most probably formed oxides increased. The dissociation and vapor pressures of all these oxides are considerably below 10^{-6} mm.

Finally, Liebold explains the high thermionic activity of pure, nickel-base oxide cathodes. According to Liebold, the heat of formation of the nickel oxide is just large enough to supply a sufficient reduction of BaO to Ba, without, at the same time, the dissociation pressure of NiO (3×10^{-6} mm at 950° C) being so low as to permit an appreciable interface formation. Finally, the dissociation pressure is not so large that the oxygen liberated by the dissociation of NiO is too much for the getter to handle; otherwise, the cathode would soon be poisoned by oxygen.

The results that were obtained for the 3.5-per cent W-Ni alloy may at first seem to be a contradiction because Liebold, in testing the emission of pure tungsten core oxide cathodes, found their emission to be inferior, and noted the presence of an interface formation, which he concluded was probably WO_2. However, this contradiction is resolved when we remember that a large part of the interface formation takes place during the breakdown of the carbonate to the oxide, as a result of the thermal dissociation of $2CO_2$ to O_2 and $2CO$, and with the subsequent oxidation of the W core by the O_2. However, since the core is a 96.5-per cent Ni-3.5-per cent W alloy, instead of pure tungsten, not as much tungsten is accessible to the oxygen during the brief (3 to 4 minutes) carbonate-to-oxide breakdown procedure. As a result, a relatively smaller interface is formed. Furthermore, if it can be assumed that the tungsten alloyed in the core gradually diffuses to the surface where it is removed by combination with the oxygen of the BaO to form Ba (noting that the heats of formation indicate that W is more active in reducing BaO to Ba than Ni), it becomes understandable that the 3.5-per cent W-Ni alloy, when used as the base of an oxide cathode, yields a more efficient emitter than pure nickel. The sparking-current data for the 3.5-per cent W-Ni alloy supports this view.

The results obtained with a 0.38-per cent Mg-Ni, alloy, when used as core, show thermionic emissions inferior to that of pure nickel. Rooksby[32] has noted that an interface compound is formed when nickel containing small amounts of magnesium is used as the core for oxide cathodes, and he analyzed this interface to be MgO. It will be assumed that an MgO interface also is formed in the Mg-Ni cathodes tested here. The low values of sparking current obtained for 0.38-per cent Mg-Ni give further indication of the formation of an interface. Decreased emission of the 0.38-per cent Mg-Ni alloy core oxide cathodes, therefore, is probably due to the formation of an MgO interface which persists and retards the action of the Mg of the core in its reduction of the BaO.

The improved emission reported by Benjamin[4] with 0.07-per cent Mg-Ni as core, and the general use by the Germans of 0.07-per cent Mg-Ni for oxide cathodes during World War II,[33] can be explained by reasoning similar to that applied to the 3.5-per cent W-Ni core. As a result of the small amount of magnesium in the core, the interface formation during conversion of the carbonate to oxide is reduced. The interface that is formed is not sufficient to impede the reducing action of the magnesium in the core.

The results with 4.0-per cent Al-Ni show a reduced emission relative to the use of pure nickel as core. Rooksby[32] has analyzed the interface formed on 2-per cent Al-Ni core oxide cathodes, and found it to be $BaO \cdot Al_2O_3$. Unfortunately, data on the dissociation and vapor pressures of this compound are not available. The presence of an interface in the cathodes analyzed by Rooksby, however, indicates that the vapor pressure and the dissociation pressure are below that of normal tube operation. It will be assumed that $BaO \cdot Al_2O_3$ is the interface formed in our 4.0-per cent Al-Ni cathodes. That an interface is formed also is indicated by the low sparking currents obtained for 4.0-per cent Al-Ni cathodes (Table I).

The thermionic emission, therefore, is reduced for the 4.0-per cent Al-Ni cathodes for the reasons already outlined for the 0.38-per cent Mg-Ni alloy. The emission in this case is inferior to that obtained by Benjamin because of the increased aluminum content, the discussion being similar to that given for 0.38-per cent Mg.

Finally, the emission of 4.8-per cent Mn-Ni alloy was found to be considerably inferior to that of pure nickel. Benjamin[4] in his work with 0.34-per cent Mn also found it to be the worst of 0.07-per cent Mg-Ni, 2-per cent Fe-Ni, 2-per cent Al-Ni, and 0.2-per cent Th-Ni. No data on the interface compound of an Mn-Ni alloy oxide cathodes is available. In a manner

[31] It is to be noted that, strictly speaking, the heats of formation cannot be used as a guide to the reducing power of an element, but rather it is the free energy of the reaction that is the determining factor. Heats of formation, when used as a guide of reducing power, will give the correct result only when the entropies involved are the same for all the reactions compared. Furthermore, the values of the thermodynamic constants at the temperatures of the reaction should be taken, and not the values for room temperatures, as is so commonly done. Finally, it should be pointed out that the free energies in a reaction do not determine the rate of the reaction.

[32] H. Rooksby, "Application of X-ray technique to individual laboratory problems," *Jour. Roy. Soc. A.*, vol. 88, p. 308; 1940.

[33] T. Briggs, "European practices in the manufacture of cathodes," Proc. I.R.E., vol. 36, p. 376; March, 1948.

space-charge limited,[21,22] when an auxiliary dc current was simultaneously drawn.

Our emission measurements were made by measurement at low temperatures and low current drain, and by pulse testing at 1 microsecond and at a low repetition rate, such as 60 cycles.

In addition to the gases given off from the anode as a result of excessive heating, it has been shown, by Jacobs[13] and Bruining,[14] that cathodes may be poisoned by the dissociation of oxide deposits on the anode. Where cathodes contain appreciable quantities of impurities, for example, magnesium, silicon, manganese, and the like, we may also expect to find their oxides on the anode. These compounds will be decomposed, as Jacobs has shown, at critical potentials, corresponding to the heats of formation. If the anode voltages are kept below this level, no dissociation of the anode deposit occurs, and hence there is no gas liberation and consequent emission decay.

For this reason the anode potentials on all the dc, low-temperature measurements have been restricted below 6 volts. Since a contact potential of at least 1 volt is generally present, in effect, only 5 volts remain to produce dissociation. This potential is below that corresponding to the critical potentials for any of the compounds likely to be deposited on the anode. Proof of this statement is the fact that in our measurements, up to 6 volts, absolutely no emission decay could be noted.

1. Low-Temperature, dc Measurements: Currents in the range of 10^{-6}–10^{-11} amperes were measured with a dc amplifier, and cathode temperatures below 700° C were measured by means of a thermocouple, formed by spot welding 0.002-inch molybdenum wire to the cathode immediately adjacent to the coating.

2. Pulsed Measurements: The pulsed measurements were made with a standard, Link Model-4 modulator, which was triggered at 60 cycles by a standard P4 Browning synchroscope.

The emission current was determined by measuring the drop of potential across a noninductive resistor, and the potential was measured by means of a capacity potential divider.

E. Calibration

For the cathode-thermocouple calibration, special vacuum tubes, 2 feet in length, were constructed so that the one end containing the cathode could be kept in an oven, while the other end could be maintained at a fixed temperature. The temperature in the oven was measured by a chromel-alumel thermocouple immediately adjacent to the outside of the 2-foot tube and at the level of the cathode. The chromel-alumel thermocouple, in turn, was calibrated against three fixed points: the boiling point of water, the melting point of tin, and the melting point of zinc.

[21] W. Ramsey, "Sparking of Oxide Cathodes," N.D.R.C., Div. 14, OEMsr 358, Report 294; July 15, 1944. (Publ. No. PB12136, Office of Technical Services, U. S. Dept. of Commerce.)

[22] W. Ramsey, "General Survey of Sparking in H. V. Thermionic Tubes," N.D.R.C., Div. 14, OEMsr 358, Report 516, Oct. 31, 1945, 91p PB12128. Office of Technical Services, U. S. Dept. of Commerce.

RESULTS

By measuring the dc emission at low temperatures and at low current levels, and by measuring pulsed emission at normal operating temperatures, significant differences in emission, which can be attributed to the core,[23] were found. The results are tabulated in Table I.

TABLE I

TABULATED RESULTS

Core	*Tube No.*	*Zero-Field Current at 230° C* $\times 10^{-9}$ *a/cm²*		*Sparking Current at 880° C a/cm²*	
3.5% W-Ni alloy	38	385	460±66*	51	51±7
	27	395		40	
	25	520		52	
	73	525		61	
Electrolytic nickel	56	195	230±22	31	38±5
	31	245		37	
	51	250		38	
	66	240		45	
0.38% Mg-Ni alloy	37	46	65±14	28	30±3
	33	57		—	
	48	79		35	
	70	78		26	
4.0% Al-Ni alloy	35	42	37±7	27	25±2
	36	26		—	
	58	42		23	
4.8% Mn-Ni alloy	26	1.8	4±3	14	17±4
	67	1.4		11	
	34	8.0		19	
	71	4.1		22	

* Standard deviation.

The dc, low-temperature emission readings were taken in the range from 150 to 400° C and 10^{-10}–10^{-6} a/cm². A typical set of data, illustrating the effect of the core material, is shown in Fig. 2. Tubes #73 and #34 are identical in every respect, except the core material. At a 10-degree higher temperature, however, the emission of the 4.8-per cent Mn-Ni cathode is only one-hundredth that of the 3.5-per cent W-Ni alloy.

To obtain the zero-field currents tabulated in Table I, the saturation currents were extrapolated to zero-field by the use of the Schottky equation,[24] $\log I/I_0 = K/T\sqrt{E}$ where I is the saturation current at the anode potential E, I_0 is the saturation current when the field at the

[23] Chemical analysis of the cores is given in Tables II and III, Appendix.

[24] W. Schottky, "Ueber den Einfluss der Bildkraefte," *Phys. Z.*, vol. 15, p. 872; 1914.

The Schottky equation is derived from, and strictly applies only to, the emission of homogeneous metal surfaces. It is used for composite emitters, such as oxide cathodes, in lieu of an equation strictly applicable to composite cathodes. When so used, it is found that the slope is different from that given by the Schottky equation.

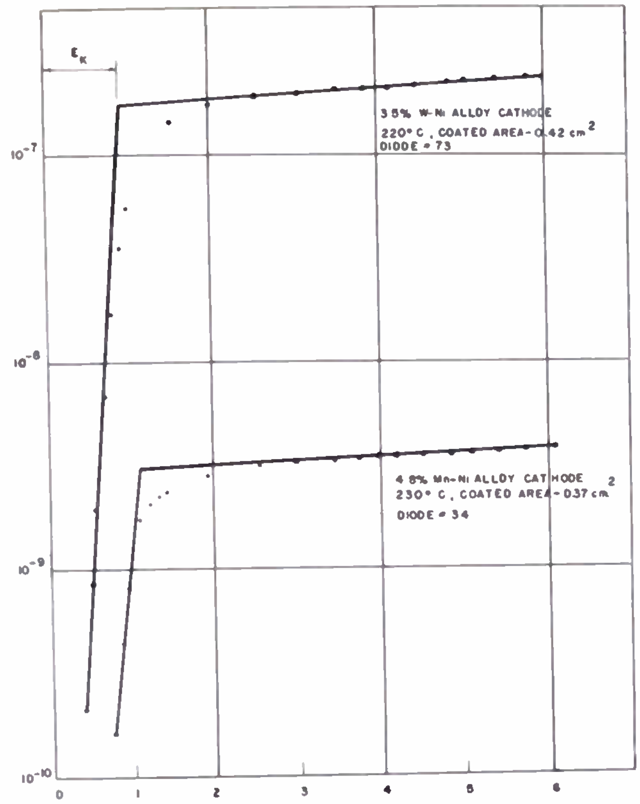

Fig. 2—Effect of the core material on the dc-emission characteristics of diodes at low temperatures. Ordinate is current in amperes; abscissa is applied anode potential in volts. E_k is the contact potential.

cathode is zero, T is the temperature, and K is a constant. Fig. 3 shows the extrapolation of the saturation currents to zero-field, using the Schottky relation. Account has been taken here of the contact potential, which must be added algebraically to the applied voltage to obtain the voltage that appears in Schottky's relation. The contact potential is given to a sufficient degree of accuracy by the interaction[25,26] of the initial and the saturation-current lines (Fig. 2).

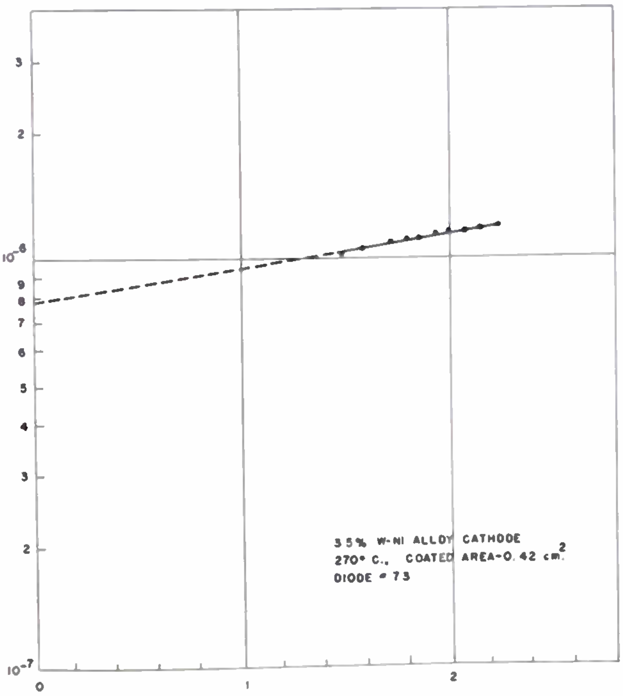

Fig. 3—Determination of the zero-field emission current extrapolation from the Schottky $\sqrt{E}$ characteristic. Ordinate is current in amperes. Abscissa is $\sqrt{E_A - E_K}$, where E_A is the applied anode potential and E_K is the contact potential; unit is $\sqrt{\text{volts}}$.

The saturation currents for each cathode were measured for at least three different temperatures. The zero-field currents were determined, and were plotted against temperature. A typical set of data is shown in Fig. 4.

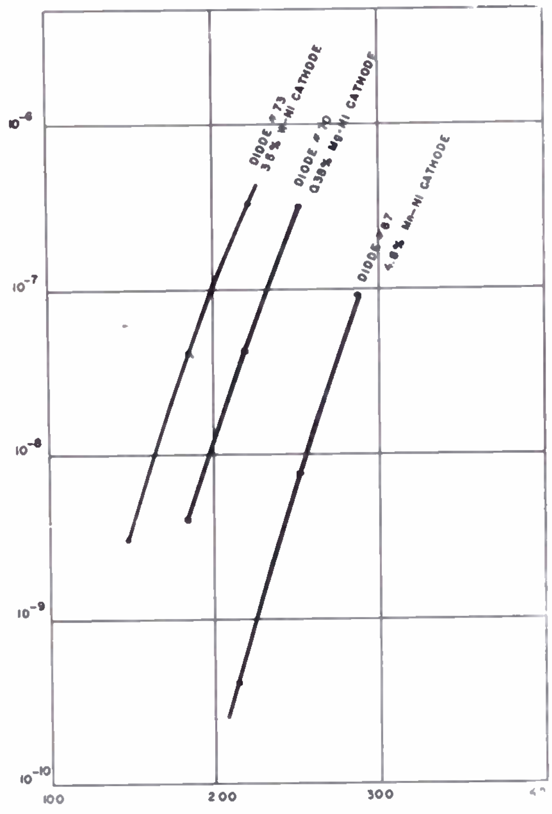

Fig. 4—Effect of the core material on the emission current of oxide cathodes. Ordinate is zero-field emission current in amperes per square centimeter; abscissa is temperature in degrees centigrade.

The comparative zero-field emission currents, shown in Table I, were taken from these graphs at the arbitrary temperature of 230° C. As noted earlier, the tubes were seasoned on the pumps. The emission measurements were made shortly after the tubes were taken off the vacuum system. The dc test was made first, the tubes being aged 10 minutes at 850° C (brightness) and 250 ma/cm² before the test. The tubes were retested after one hour dc operation at 850° C (brightness) and 250 ma/cm², with satisfactory agreement.

The sparking currents are significant because they limit the maximum current that can be drawn from a cathode. There is evidence[15,21,22] that sparking is initiated by a metallic vapor, but whether this vapor is

[25] H. Rothe, "Austrittsarbeit and Kontaktpotential," *Z. techn. Phys.*, vol. 6, p. 633; 1925.
[26] W. Heinze and S. Wagener, "Variations of emission constants of oxide cathodes during activation," *Z. Phys.*, vol. 110, p. 164; 1938.

produced by a dielectric breakdown or by an I^2R heating of the interface is not as yet clear.

A typical set of data illustrating the effect of core material on pulsed emission is shown in Fig. 5. The data

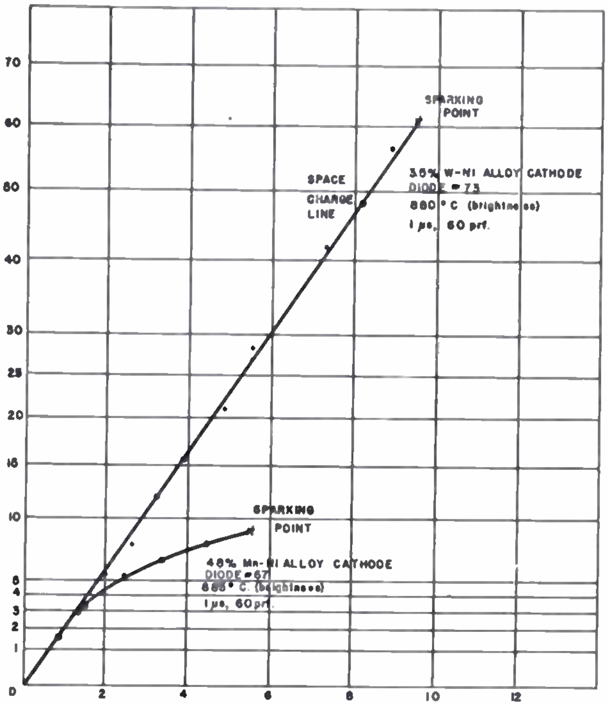

Fig. 5—Effect of the core material on the pulsed emission characteristics of experimental diodes. Ordinate is current in amperes per square centimeter; abscissa is applied anode potential in kilovolts. Draw on $\frac{2}{3}$-power paper.

was taken at 880° C., 1 μs, 60 prf. The emission characteristic of the 3.5-per cent W-Ni core tube is typical of that of a good emitter. No falling off of emission from the space-charge limited line could be determined, the emission being limited by sparking. The emission characteristic of the 4.8-per cent Mn-Ni tube is characteristic of a poor emitter. The emission falls away from the space-charge curve at low values, and is ultimately limited by sparking.

The accuracy of the low-temperature dc measurements is limited by the accuracy with which the cathode temperature could be measured. Uncertainty in determining the temperature of the cold junction of the thermocouple limits temperature measurements to ± 4 per cent. The emission-current measurements were made to ± 3 per cent, and the voltage measurements to within ± 2 per cent. Pulse measurements are within ± 5 per cent. The temperature, when expressed as "brightness" temperature, could be measured to ± 2 per cent.

Discussion of Results

The data of Table I show that the core material exerts a marked influence on the thermionic emission of oxide cathodes. The observed influence is in conformity with already published theories, as will be shown.

The work of Becker[27] and other[28] has shown that the oxide coating behaves as an impurity semiconductor, small amounts (in the order of 0.2 per cent) of free alkaline-earth metals furnishing the impurity. Furthermore, Prescott and Morrison[29] have shown that the emission of an oxide cathode is dependent on the amount of free barium present in the coating. They found that the thermionic emission increases with the barium content up to a concentration of 30 μg[30] of barium per cm^2 of superficial area. Free barium, however, can be produced by reaction of the core material with the oxide coating, such as

$$a\text{Ke} + b\text{BaO} \rightarrow \text{Ke}_a\text{O}_b + b\text{Ba},$$

where Ke represents the core metal.

Benjamin[4] advanced the thesis that the emission of nickel-alloy base oxide cathodes was related to the reducing power of the additive alloyed with the nickel. He used the heats of formation of the most likely formed oxides as a measure of the reducing power. In this manner he was able to explain the emission of 2-per cent Al-Ni, 0.07-per cent Mg-Ni, and 0.34-per cent Mn-Ni alloys. The emission of pure nickel, however, was out of place in this scheme, and Benjamin attributed this to small traces of reducing elements in the pure nickel, in spite of the fact that the other alloys were made using the same pure nickel as base.

In an investigation of the thermionic emission of oxide cathodes using eleven different core materials, Liebold found[5,28] that his results could not all be explained by the theory of Benjamin. He proposed a modified theory, into which the experimental data obtained here fits.

Benjamin divided the core materials into two groups: (a) those core materials for which the most probably formed interface oxides have a dissociation pressure, or a vapor pressure, higher than that existing in the tube (about 10^{-6} mm), at the operating temperatures; and (b) those core materials for which the dissociation pressure and vapor pressure of the most probably formed interface oxides are lower than the pressure existing in the tube.

In the first group, the interface compounds that are formed do not persist. As soon as they are formed, they decompose, giving off oxygen, or they are evaporated away. Hence, no interface is formed. For these cores, basing Ba formation on the reduction of BaO by the core, Liebold concludes that the thermionic emission should increase with the reducing power of the core. To

[27] J. Becker, "Phenomena in oxide-coated filaments," *Phys. Rev.*, vol. 34, p. 1323; 1929.
[28] G. Herrmann and S. Wagener, "Die Oxydkathode," Johann Barth, Leipzig, Germany; 1943.
[29] C. Prescott and J. Morrison, "The oxide-coated filament," *Jour. Amer. Chem. Soc.*, vol. 60, p. 3047; 1938.
[30] Per 1 mg BaO, 1 mg SrO of coating.

this extent Liebold's theory is the same as Benjamin's. Indeed, Liebold found that the thermionic emission of Au, Pt, Pd, Cu, and Ni increased with the heats of formation[21] of the oxides most probably formed at the interface. These oxides all have dissociation pressures above 3×10^{-6} mm.

For the second group, an interface compound is formed that persists. This interface by its interposition, retards the reduction by the core metal of BaO to Ba, and does so to a larger degree as it grows in thickness. Indeed, Liebold found that for W, Mo, La, Cr, and Zr, the emission decreased as the heats of formation of the most probably formed oxides increased. The dissociation and vapor pressures of all these oxides are considerably below 10^{-6} mm.

Finally, Liebold explains the high thermionic activity of pure, nickel base oxide cathodes. According to Liebold, the heat of formation of the nickel oxide is just large enough to supply a sufficient reduction of BaO to Ba, without, at the same time, the dissociation pressure of NiO (3×10^{-6} mm at 950° C) being so low as to permit an appreciable interface formation. Finally, the dissociation pressure is not so large that the oxygen liberated by the dissociation of NiO is too much for the getter to handle; otherwise, the cathode would soon be poisoned by oxygen.

The results that were obtained for the 3.5-per cent W-Ni alloy may at first seem to be a contradiction because Liebold, in testing the emission of pure tungsten core oxide cathodes, found their emission to be inferior, and noted the presence of an interface formation, which he concluded was probably WO_2. However, this contradiction is resolved when we remember that a large part of the interface formation takes place during the breakdown of the carbonate to the oxide, as a result of the thermal dissociation of $2CO_2$ to O_2 and 2CO, and with the subsequent oxidation of the W core by the O_2. However, since the core is a 96.5-per cent Ni-3.5-per cent W alloy, instead of pure tungsten, not as much tungsten is accessible to the oxygen during the brief (3 to 4 minutes) carbonate-to-oxide breakdown procedure. As a result, a relatively smaller interface is formed. Furthermore, if it can be assumed that the tungsten alloyed in the core gradually diffuses to the surface where it is removed by combination with the oxygen of the BaO to form Ba (noting that the heats of formation indicate that W is more active in reducing BaO to Ba than Ni), it becomes understandable that the 3.5-per cent W-Ni alloy, when used as the base of an oxide cathode, yields a more efficient emitter than pure nickel. The sparking-current data for the 3.5-per cent W-Ni alloy supports this view.

The results obtained with a 0.38-per cent Mg-Ni, alloy, when used as core, show thermionic emissions inferior to that of pure nickel. Rooksby[22] has noted that an interface compound is formed when nickel containing small amounts of magnesium is used as the core for oxide cathodes, and he analyzed this interface to be MgO. It will be assumed that an MgO interface also is formed in the Mg-Ni cathodes tested here. The low values of sparking current obtained for 0.38-per cent Mg-Ni give further indication of the formation of an interface. Decreased emission of the 0.38-per cent Mg-Ni alloy core oxide cathodes, therefore, is probably due to the formation of an MgO interface which persists and retards the action of the Mg of the core in its reduction of the BaO.

The improved emission reported by Benjamin[4] with 0.07-per cent Mg-Ni as core, and the general use by the Germans of 0.07-per cent Mg-Ni for oxide cathodes during World War II,[23] can be explained by reasoning similar to that applied to the 3.5-per cent W-Ni core. As a result of the small amount of magnesium in the core, the interface formation during conversion of the carbonate to oxide is reduced. The interface that is formed is not sufficient to impede the reducing action of the magnesium in the core.

The results with 4.0-per cent Al-Ni show a reduced emission relative to the use of pure nickel as core. Rooksby[22] has analyzed the interface formed on 2-per cent Al-Ni core oxide cathodes, and found it to be $BaO\ Al_2O_3$. Unfortunately, data on the dissociation and vapor pressures of this compound are not available. The presence of an interface in the cathodes analyzed by Rooksby, however, indicates that the vapor pressure and the dissociation pressure are below that of normal tube operation. It will be assumed that $BaO\ Al_2O_3$ is the interface formed in our 4.0-per cent Al-Ni cathodes. That an interface is formed also is indicated by the low sparking currents obtained for 4.0-per cent Al-Ni cathodes (Table I).

The thermionic emission, therefore, is reduced for the 4.0-per cent Al-Ni cathodes for the reasons already outlined for the 0.38-per cent Mg-Ni alloy. The emission in this case is inferior to that obtained by Benjamin because of the increased aluminum content, the discussion being similar to that given for 0.38-per cent Mg.

Finally, the emission of 4.8-per cent Mn-Ni alloy was found to be considerably inferior to that of pure nickel. Benjamin[4] in his work with 0.34-per cent Mn also found it to be the worst of 0.07-per cent Mg-Ni, 2-per cent Fe-Ni, 2-per cent Al-Ni, and 0.2-per cent Th-Ni. No data on the interface compound of an Mn-Ni alloy oxide cathodes is available. In a manner

[21] It is to be noted that, strictly speaking, the heats of formation cannot be used as a guide to the reducing power of an element, but rather it is the free energy of the reaction that is the determining factor. Heats of formation, when used as a guide of reducing power, will give the correct result only when the entropies involved are the same for all the reactions compared. Furthermore, the values of the thermodynamic constants at the temperatures of the reaction should be taken, and not the values for room temperatures, as is so commonly done. Finally, it should be pointed out that the free energies in a reaction do not determine the rate of the reaction.

[22] H. Rooksby, "Application of X-ray technique to individual laboratory problems," *Jour. Roy. Soc. A.*, vol. 88, p. 308; 1940.

[23] T. Briggs, "European practices in the manufacture of cathodes," Proc. I.R.E., vol. 36, p. 376; March, 1948.

similar to the other alloys, an interface compound probably is formed, reducing the emission for the reasons discussed for the 0.38-per cent Mg-Ni and 4.0-per cent Al-Ni alloys. The pulsed-emission data for the 4.8-per cent Mn-Ni alloy given in Table I are the lowest for all the alloys tested. The emission of the 4.8-per cent Mn-Ni alloy base oxide cathodes, however, is so reduced that it seems likely that another factor is at work. One such factor might be the formation of an interface compound that can oxidize barium.

Acknowledgment

The author wishes to express his appreciation to the Radio Receptor Company for the grant of a Radio Receptor Fellowship, to the Sylvania Electric Products Company, and to the Bell Telephone Laboratories for specialized services, such as hydrogen firing, to the International Nickel Company for the grant of the nickel alloys used as cores, and to Professor J. B. Russell of the Electrical Engineering Department of Columbia University.

Appendix

TABLE II

Quantitative Chemical Analysis of the Nickel Alloys*

Type	*INCO melt No.*	*Si*	*Mg*	*C*	*Mn*	*Fe*	*Cu*	*S*	*Ni+Co*	*Other*	*Ni*	*Co*
Grade A	—	0.20% max.	—	0.20% max.	0.35% max.	0.30% max.	0.25% max.	0.008% max.	99.00 min.	—	—	—
Electrolytic Ni	12808	0.03%	0.027%	0.012%	—	—	—	—	—		Remainder	0.76% †
0.38% Mg-Ni	12811	0.03%	—	0.012%	—	—	—	—	—	0.38% Mg	Remainder	0.76% †
4.0% Al-Ni	12827	0.03%	0.027%	0.012%	—	—	—	—	—	4.01% Al	Remainder	0.76% †
4.8% Mn-Ni	12888	0.03%	0.027%	0.012%	—	—	—	—	—	4.83% Mn	Remainder	0.76% †
3.5% W-Ni	12890	0.03%	0.027%	0.012%	—	—	—	—	—	3.48% W	Remainder	0.76% †

* A qualitative spectrochemical analysis is given in Table III.

† This special series of nickel melts, together with the analyses, was kindly supplied by the International Nickel Company, through the courtesy of Mr. E. M. Wise. Melt No. 12808 is the base melt to which Al, Mn, Mg, and W are added to obtain the other nickel alloys. The analysis is for the ingot.

TABLE III

Spectrochemical Qualitative Analysis* of the Nickel Alloys and of the Emission Mixture

	Estimated range	*Electrolytic nickel*	*0.38% Mg-Ni alloy*	*4.0% Al-Ni alloy*	*4.8% Mn-Ni alloy*	*3.5% W-Ni alloy*	*Raytheon emission mixture C-51-2*
Principal component	>10%	Ni	Ni	Ni	Ni	Ni	Ba, Sr
Major component	> 1%	—	—	Al	Mn	W	—
Minor component	0.1–3%	Co	Co, Mg	Co	Co	Co	Ca, Na
Impurities	0.01–0.3%	Fe, Mn, Si, Mg	Fe, Mn, Si, Ca	Fe, Mn, Si, Mg, Pb, Na	Fe, Cr, Si	Fe, Mn, Si, Zn	Mg
Traces	<0.03%	Ca, Cu, Na	Cu, Na, Pb	Ca, Cu, Cr, Zn	Ca, Cu, Na, Pb, Mg	Cu, Na, Pb, Mg, Ca, Cr	Pb
Faint traces	—	—	—	—	—	—	Al, Bi, Cu, Fe, Mn, Ni, Si

* Furnished through the courtesy of Dr. L. A. Wooten; analysis by Mr. W. Hartmann, Bell Telephone Laboratories. Analysis is for the rolled nickel sheet from which the cathodes were formed.

A Generalized Method for Analyzing Servomechanisms*

ARTHUR A. HAUSER, JR.†, MEMBER, IRE

Summary—**Through the use of the algebra of dyadics a unification of dc and ac servo theory is obtained. The effects arising from the asymmetry of the sidebands in an ac (carrier-frequency) servo are automatically taken into account. In the case of a symmetrical side-band carrier-frequency servo or a dc servo, the generalized operators introduced reduce to the ordinary transfer functions of the system components.**

I. Introduction

The problem of obtaining an expression for the loop transfer function of a carrier-frequency servomechanism cannot be handled by the usual methods of servo analysis. That is to say, the concept of transfer function is not applicable unless one is willing to assume that the frequency function of the components involved is symmetrical about the carrier frequency. Often such an assumption is valid, and leads to a good engineering design of a carrier-frequency servo system. On the other hand, this assumption, if made without due consideration, can lead to a poorly designed system.

Further, if the servo contains a frequency-sensitive network such as a parallel tee, the assumption of symmetrical sidebands precludes the possibility of obtaining information concerning the off-frequency operation of the servo system.

Once the loop-transfer function of the servo has been obtained, be it an ac or a dc servo, the usual methods of analysis for determining the transient response, or for testing the stability of the servo, may be employed. The central problem is that of obtaining the loop-transfer function of the servo, and that is the problem which is considered here.

Sobczyk[1] recently presented a method by which error lead-stabilized carrier-frequency servomechanisms may be analyzed. The present paper is, in a sense, a generalization of Sobczyk's work since it presents a method by which any type of carrier-frequency servomechanism may be analyzed. The analysis of dc servomechanisms, which has been exhaustively treated in recent literature, appears as a degenerate case of the technique presented here. The technique employs the theory of linear spaces in the particular form of the algebra of dyadics.[2] Through the use of this algebra, the concept of transfer function is generalized to that of transfer-operator. The transfer operators so defined behave with respect to carrier-frequency servos in much the same manner that transfer functions behave with respect to dc servos. Further, these transfer operators reduce to transfer functions in case that concept is applicable, as is the case with symmetrical sideband or dc servos.

II. The Vector Representation of Carrier-Frequency Voltages

In dc servo theory, it is usual and useful to represent a voltage of the form

$$v = K \cos (mt + \phi) \tag{1}$$

by means of the complex function

$$V = Ke^{j(mt+\phi)}, \tag{2}$$

where

$$j = \sqrt{-1}.$$

In this paper, voltages of the form

$$v = K \cos (mt + \phi) \cos (\omega_c t + \theta) \tag{3}$$

will be encountered where the factor $\cos (\omega_c t+\theta)$ is called the carrier, and ω_c is called the carrier (angular) frequency. The quantity m is called the modulating (angular) frequency and ϕ and θ are respectively the modulation and carrier phase shifts. The entire expression v will be recognized as a suppressed-carrier modulated voltage.

The voltage v of equation (3) may be represented uniquely as a vector, $\boldsymbol{V}$, having a complex argument. Thus,

$$\boldsymbol{V} = \boldsymbol{i}_\theta Ke^{j(mt+\phi)}. \tag{4}$$

Here $\boldsymbol{i}_\theta$ is a unit vector which makes an angle θ with some reference line defined by the base vector $\boldsymbol{i}_0$, and represents the carrier $\cos (\omega_c t+\theta)$.

In (4), the frequency ω_c of the carrier is nowhere noted, being understood. In systems in which more than a single frequency appears as, for example, in systems using demodulators or frequency changers, it is advisable to indicate the frequency, ω_c, to which the unit vector refers. In such cases ω_c may be used as a superscript and (4) may be written

$$\boldsymbol{V} = \boldsymbol{i}_\theta^{\omega_c} Ke^{j(mt+\phi)}. \tag{4a}$$

To unify the notation for dc and for carrier-frequency systems, it becomes necessary to represent the voltage v given by (1) as a vector. Thus, in the system of repre-

* Decimal classification: 621.375.13. Original manuscript received by the Institute, January 10, 1951; revised manuscript received August 8, 1951.

† Sperry Gyroscope Co., Great Neck, N. Y.

[1] A. Sobczyk, "Stabilization of carrier frequency servomechanisms," *Jour. Frank. Inst.*, vol. 246; July, 1948.

[2] L. Page, "Introduction to Theoretical Physics," D. Van Nostrand Co., Inc., New York, N. Y.; 1935.

sentation adopted here, the voltage $v=K\cos(mt+\phi)$ is written

$$v = K\cos(mt+\phi)\cos(0\cdot t+0) \tag{1a}$$

and is represented by the vector

$$V = i_0{}^0 K e^{j(mt+\phi)}. \tag{2a}$$

III. Representation of a Linear Network

A linear network is completely characterized by its (complex) transfer function $G(j\omega)$. This function, it will be recalled, specifies the amplitude ratio and phase shift to which an input sinusoid is subjected in the steady state, when the sinusoid is impressed on the network. Thus if a voltage

$$v_1 = A\cos(mt+\phi) \tag{5}$$

is impressed on a network with transfer function $G(j\omega)$ the steady state output v_0 is given by

$$v_0 = A\left|G(j\omega)\right|\cos\left[mt+\phi+\operatorname{arc}G(j\omega)\right]. \tag{6}$$

It is the fact that form of the voltage (5) is left invariant in the steady state by the linear network, which makes it possible to utilize the concept of transfer function.

If a voltage of the form

$$v_1 = K\cos(mt+\theta_1)\cos(\omega_c t+\theta_2) \tag{7}$$

is impressed on a linear network, its form is not left invariant. In fact, the steady state output voltage v_0 is

$$v_0 = K_1\cos(mt+\phi_1)\cos(\omega_c t+\phi_2) + K_2\sin(mt+\phi_1)\sin(\omega_c t+\phi_2). \tag{8}$$

However, if a more general voltage, namely,

$$v_1 = K_1\cos(mt+\theta_1)\cos(\omega_c t+\theta_2) + K_2\sin(mt+\theta_1)\sin(\omega_c t+\theta_2) \tag{9}$$

is impressed on a linear network, the form of the output is the same as the form of the input, the output voltage being

$$v_0 = C_1\cos(mt+\psi_1)\cos(\omega_c t+\psi_2) + C_2\sin(mt+\psi_1)\sin(\omega_c t+\psi_2). \tag{10}$$

The voltages v_1, and v_0 are representable in the forms

$$V_1 = i_{\theta_2}K_1e^{j(mt+\theta_1)} + i_{\theta_2-\pi/2}K_2e^{j(mt+\theta_1-\pi/2)} \tag{11}$$

and

$$V_0 = i_{\psi_2}C_1e^{j(mt+\psi_1)} + i_{\psi_2-\pi/2}C_2e^{j(mt+\psi_1-\pi/2)}. \tag{12}$$

The linear network may be thought of as a transformation, n which carries v_1 into v_0 and as such may be represented by another transformation N which carries the vector V_1 into the vector V_0. This transformation is a tensor of order 2 and the generalization of the transfer function $G(j\omega)$. In what follows, dyadic notation will be used, and hence the representation N of the network transformation, n, will be called the network transfer dyadic, or transfer operator of the network. Its form will now be derived explicitly.

Let the voltage

$$v_1 = K_1\cos(mt+\theta_1)\cos(\omega_c t+\theta_2) + K_2\sin(mt+\theta_1)\sin(\omega_c t+\theta_2) \tag{13}$$

be impressed on a network with transfer function $G(j\omega)$. Define

$$G\left[j(\omega_c+m)\right] = G^+e^{j\gamma^+}$$
$$G\left[j(\omega_c-m)\right] = G^-e^{j\gamma^-}$$
$$\phi_1 = \tfrac{1}{2}(\gamma^+-\gamma^-)$$
$$\phi_2 = \tfrac{1}{2}(\gamma^++\gamma^-)$$
$$A_1 = \tfrac{1}{2}(G^-+G^+)$$
$$A_2 = \tfrac{1}{2}(G^--G^+).$$

The steady state output voltage of the network may then be written

$$v_0=(K_1A_1+K_2A_2)\cos(mt+\theta_1+\phi_1)\cos(\omega_ct+\theta_2+\phi_2) + (K_1A_2+K_2A_1)\sin(mt+\theta_1+\phi_1)\sin(\omega_ct+\theta_2+\phi_2). \tag{14}$$

The network transformation, n, then transforms the voltage v_1 of (13) into the voltage v_0 of (14). Since the voltages v_1 and v_0 may be represented in vector form by the vectors V_1 and V_0, respectively, where

$$V_1 = i_{\theta_2}K_1e^{j(mt+\theta_1)} + i_{\theta_2-\pi/2}K_2e^{j(mt+\theta_1-\pi/2)} \tag{15}$$

and

$$V_0 = i_{\theta_2+\phi_2}(K_1A_1+K_2A_2)e^{j(mt+\theta_1+\phi_1)} + i_{\theta_2+\phi_2-\pi/2}(K_1A_2+K_2A_1)e^{j(mt+\theta_1+\phi_1-\pi/2)}. \tag{16}$$

The network transformation n may be represented by the dyadic

$$N = i_{\theta_2+\phi_2}i_{\theta_2}A_1e^{j\phi_1} + i_{\theta_2+\phi_2}i_{\theta_2-\pi/2}A_2e^{j(\pi/2)}e^{j\phi_1} + i_{\theta_2+\phi_2-\pi/2}i_{\theta_2}A_2e^{-j(\pi/2)}e^{j\phi_1} + i_{\theta_2+\phi_2-\pi/2}i_{\theta_2-\pi/2}A_1e^{j\phi_1}. \tag{17}$$

The vector representing the output of the network is then related to the vector representing the input by the relation

$$V_0 = N\cdot V_1. \tag{18}$$

The similarity of form between (18) and the expression

$$E_0 = G(j\omega)E_1, \tag{19}$$

which holds between the input and output when a simple sinusoidal voltage E_1 is impressed on the network, is striking. The vectors V_0 and V_1 replace the voltages E_0 and E_1 respectively, and the transfer dyadic N replaces transfer function $G(j\omega)$.

IV. Symmetrical Sideband Networks

A linear network is said to be a symmetrical sideband network with respect to ω_0 if[3]

[3] G^* is the complex conjugate of G.

$$G[j(\omega_0 + m)] = G^*[j(\omega_0 - m)] \tag{20}$$

for all $m \geqq 0$.

If the network under consideration is a linear passive network with a finite number of real positive circuit elements consisting of resistors, condensers and inductors, then it is always true that[4]

$$G(jm) \equiv G^*(-jm) \tag{21}$$

so that all such networks are symmetrical sideband networks with respect to $\omega_0 = 0$.

Since for the networks under consideration, the transfer functions $G(j\omega)$ are rational functions of $j\omega$, these networks cannot in general be symmetrical sideband networks with respect to a frequency ω_0 other than 0.

Even though this is the case, it is often true that real networks very nearly approximate the condition for symmetrical sidebands, and for the purpose of synthesizing carrier-frequency servomechanisms, it is often adequate to assume that the network under consideration is a symmetrical sideband network. This assumption usually simplifies the system equations considerably. It is because of this that these networks are of particular interest. Another reason that these networks are of interest is because they always arise in the theory of dc servos.

The main properties of symmetrical sideband networks are readily obtained.

Since

$$G[j(\omega_c + m)] \equiv G^*[j(\omega_c - m)]$$

$$G^+ \equiv G^- \qquad \gamma^+ \equiv \gamma^-,$$

hence

$$A_1 \equiv G^+ \qquad A_2 \equiv 0$$

$$\phi_1 \equiv \gamma^+ \qquad \phi_2 \equiv 0.$$

In this case the network transfer operator is

$$N = \mathbf{i}_{\theta_2}\mathbf{i}_{\theta_2}A_1e^{j\phi_1} + \mathbf{i}_{\theta_2-\pi/2}\mathbf{i}_{\theta_2-\pi/2}A_1e^{j\phi_1} \tag{22}$$

or

$$N = SG^+e^{j\gamma+},$$

where S is the idemfactor or unit dyadic.[2] Thus, in the case of symmetrical sideband networks, the network transfer operator has the simple form given by (22).

If $\omega_c = 0$ so that the servo under consideration is a dc servo, it is seen that

$$N = S^\circ G(jm), \tag{23}$$

where the superscript is included to indicate that $\omega_c = 0$, which is the usual transfer function employed in the treatment of dc servos if the idemfactor is disregarded. This shows that the network transfer dyadic as defined here is a logical generalization of the transfer function of a network, in the sense that it reduces to the transfer function in the simpler case of a dc or of any symmetrical sideband servo.

From (18) it follows that if an input

$$v_1 = K_1 \cos(mt + \theta_1) \cos(\omega_c t + \theta_2) \tag{24}$$

is applied to a symmetrical sideband network with respect to ω_c,

$$\mathbf{V}_0 = SG^+e^{j\gamma^+} \cdot \mathbf{i}_{\theta_2}K_1e^{j(mt+\theta_1)} = \mathbf{i}_{\theta_2}K_1G^+e^{j(mt+\theta_1+\gamma^+)} \tag{25}$$

$$v_0 = K_1 |G[j(\omega_c + m)]| \cos(mt + \theta_1 + \gamma^+) \cos(\omega_c t + \theta_2), \tag{26}$$

which shows that a network which has sidebands symmetrical with respect to the carrier frequency, does not phase-shift the carrier. In fact, the output is found by allowing the network transfer function taken relative to ω_c, i.e., $G[j(\omega_c + m)]$, to operate on the modulation alone. The carrier is preserved by the network. Thus, symmetrical sideband networks may be analyzed by the usual techniques employed in dealing with dc systems.

V. The Two-Phase Induction Motor

The two-phase induction motor is the power device used in most carrier-frequency servo systems of the instrument type. Hence, it is necessary to characterize this device before a discussion of carrier-frequency servos can be given.

It is possible beforehand to guess the nature of the transfer entity which must be associated with the two-phase induction motor. The motor accepts as input a suppressed-carrier modulated voltage, and gives as output a shaft rotation. For a complete unification of ac and dc servos, shaft rotations and dc voltages should be treated as vector quantities. The notation by which this may be done was illustrated in (2a). However, while such a notation leads to a complete unification, it is more cumbersome than the notation in which shaft rotations and dc voltages are treated as scalar quantities, as was done in (2). Agreeing then that shaft rotations and dc voltages are to be regarded as scalar quantities, the two-phase induction motor must be represented by an entity which transforms a vector to a scalar. Hence, the motor-transfer operator will be a vector whose scalar product with the vector representing the input, will give the scalar representing the output.

If Kv_1 represents the voltage applied to the control winding of the motor, and if $-K \cos(\omega_c t + \phi_0 - \pi/2)$ is the voltage applied to the reference winding of the motor, then the differential equation governing the position of the output, θ_0, of the motor will be taken to be[1]

$$\tau\ddot{\theta}_0 + \dot{\theta}_0 = 2\omega_c v_1 \cos(\omega_c t + \phi_0) - 2\dot{v}_1 \cos\left(\omega_c t + \phi_0 - \frac{\pi}{2}\right), \tag{27}$$

[4] James, Nichols, and Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co., Inc., New York, N. Y., p. 66; 1947.

where τ is a time constant associated with the motor and the load. It is convenient to rewrite (27) as two equations:

$$\tau\ddot{\theta}_0 + \dot{\theta}_0 = \rho(t)$$

$$\rho(t) = 2\omega_c v_1 \cos(\omega_c t + \phi_0) - 2\dot{v}_1 \cos\left(\omega_c t + \phi_0 - \frac{\pi}{2}\right). \quad (28)$$

If v_1 is taken to be the suppressed-carrier modulated voltage,

$$v_1(t) = K_1 \cos(mt + \theta_1) \cos(\omega_c t + \phi_0 + \lambda) + K_2 \sin(mt + \theta_1) \sin(\omega_c t + \phi_0 + \lambda), \quad (29)$$

then except for terms of frequency, $2\omega_c$, $\rho(t)$ is given by

$$\rho(t) = (2\omega_c K_1 - mK_2) \cos\lambda \cos(mt + \theta_1) + (2\omega_c K_2 - mK_1) \sin\lambda \sin(mt + \theta_1), \quad (30)$$

and it is thus seen that if a suppressed-carrier modulated voltage, $v_1(t)$ is applied, the motor acts like a dc motor to which a voltage $\rho(t)$ is applied. It is further seen that $\rho(t)$ is a demodulated version of the voltage $v_1(t)$.

Now if the voltage $v_1(t)$ is represented in vector form,

$$\boldsymbol{V}_1 = \boldsymbol{i}_{\phi_0+\lambda} K_1 e^{j(mt+\theta_1)} + \boldsymbol{i}_{\phi_0+\lambda-\pi/2} K_2 e^{j(mt+\theta_1-\pi/2)}, \quad (31)$$

and the response $\rho(t)$ is written in the complex form,

$$P = (2\omega_c K_1 - mK_2) \cos\lambda e^{j(mt+\theta_1)} + (2\omega_c K_2 - mK_1) \sin\lambda e^{j(mt+\theta_1-\pi/2)}, \quad (32)$$

it is seen that the demodulating portion of the two-phase induction motor can be described by the transfer operator $\boldsymbol{D}\cdot$ where

$$\boldsymbol{D} = \boldsymbol{i}_{\phi_0} 2\omega_c + \boldsymbol{i}_{\phi_0-\pi/2} m e^{-j(\pi/2)}. \quad (33)$$

The complex function P is then given by

$$P = \boldsymbol{D} \cdot \boldsymbol{V}_1. \quad (34)$$

Letting Θ_0 be the complex function representing the output, θ_0, of the motor, it follows from the first of (28) that

$$\Theta_0 = \frac{P}{jm(1 + j\tau m)} \quad (35)$$

or, using (34),

$$\Theta_0 = \frac{\boldsymbol{D}}{jm(1 + j\tau m)} \cdot \boldsymbol{V}_1 = \boldsymbol{M} \cdot \boldsymbol{V}_1, \quad (36)$$

and it is seen that the two-phase induction motor is described by the transfer vector $\boldsymbol{M}\cdot$ where

$$\boldsymbol{M} = \frac{\boldsymbol{D}}{jm(1 + j\tau m)} = \frac{\boldsymbol{i}_{\phi_0} 2\omega_c + \boldsymbol{i}_{\phi_0-\pi/2} m e^{-j(\pi/2)}}{jm(1 + j\tau m)}. \quad (37)$$

VI. The Induction Tachometer

The induction tachometer is a device which is often used to stabilize servomechanisms. In most applications, the tachometer is driven by the servo output shaft, and its output voltage is fed back to the input through appropriate networks to accomplish the stabilization. Besides the input shaft and the output winding, the tachometer also has another winding, the reference winding, which is usually supplied with an appropriately phased constant-amplitude voltage of frequency equal to that of the carrier.

If the voltage applied to the reference winding is proportional to v_c and if θ is the position of the input shaft, then the output voltage of the tachometer is given by

$$v_0 = -K \frac{d}{dt} (v_c \dot{\theta}). \quad (38)$$

Of particular interest is the case where the reference voltage is sinusoidal. Suppose that the reference voltage is proportional to $v_1 \cos(\omega_c t + \alpha)$. Then,

$$v_0 = K_1[\omega_c \dot{\theta} \sin(\omega_c t + \alpha) - \ddot{\theta} \cos(\omega_c t + \alpha)]. \quad (39)$$

If now the motion of the input shaft is also sinusoidal, so that θ is proportional to $\cos(mt + \phi)$, then

$$v_0 = K_1[-\omega_c m \sin(mt + \phi) \sin(\omega_c t + \alpha) + m^2 \cos(mt + \phi) \cos(\omega_c t + \alpha)]. \quad (40)$$

It has therefore been shown that if the input rotation of the tachometer is representable by

$$\Theta = K e^{j(mt+\phi)}, \quad (41)$$

and if the voltage applied to the reference winding is also sinusoidal, being proportional to $v_1 \cos(\omega_c t + \alpha)$, then the output voltage, v_0, of the tachometer may be represented by the vector

$$\boldsymbol{V}_0 = K_1[-\boldsymbol{i}_{\alpha-\pi/2} \omega_c m e^{-j(\pi/2)} + \boldsymbol{i}_\alpha m^2] e^{j(mt+\phi)}. \quad (42)$$

As might have readily been anticipated, the transfer operator of the tachometer is a vector which, when multiplying the scalar representing the input rotation, gives the vector representing the output. From (41) and (42), the transfer operator of the tachometer is seen to be the vector,

$$\boldsymbol{T} = -K_r\left[\boldsymbol{i}_{\alpha-\pi/2} m e^{-j(\pi/2)} - \boldsymbol{i}_\alpha \frac{1}{\omega_c} m^2\right]. \quad (43)$$

The output voltage of the tachometer is represented by

$$\boldsymbol{V}_0 = \boldsymbol{T}\Theta. \quad (44)$$

VII. Loop Transfer Functions of Servo Configurations

The two carrier-frequency servo configurations most commonly used in practice are shown in Figs. 1 and 2. These are, respectively, an error lead-stabilized servo, and a servo stabilized by tachometric feedback.

Referring to Fig. 1 on page 201, the following expression may be written

$$\Theta_0 = (N \cdot V_\epsilon) K_A \cdot M. \tag{45}$$

Writing V_ϵ and Θ_0 in the forms

$$V_\epsilon = i_0 \epsilon e^{jmt} \tag{46}$$

$$\Theta_0 = \theta_0 e^{jmt}, \tag{47}$$

it follows that

$$\frac{\theta_0}{\epsilon} = K_A (N \cdot i_0) \cdot M, \tag{48}$$

which is the required loop transfer function.

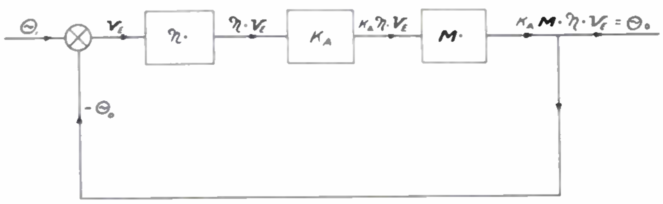

Fig. 1—Error lead stabilized servo.

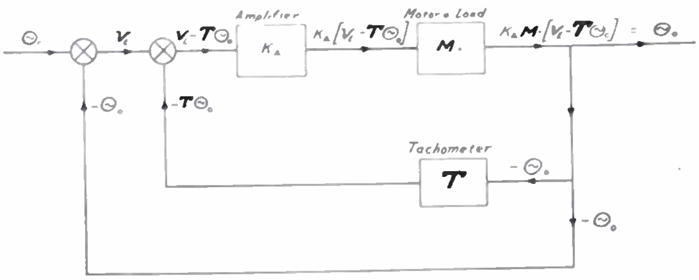

Fig. 2—Servo with tachometric feedback.

Referring to Table I, where the transfer operators of various servo components are tabulated, (48) may be expanded.

Carrying through this expansion,

$$\frac{\theta_0}{K_v \epsilon} = \frac{[(2\omega_c A_1 - mA_2)\cos\phi - j(2\omega_c A_2 - mA_1)\sin\phi]e^{j\phi_1}}{2\omega_c |G(j\omega_c)| jm(1 + j\tau m)}, \tag{49}$$

where

$$\phi = \phi_2 - \phi_0. \tag{50}$$

and

$$K_v = \lim_{jm \to 0} jm\left(\frac{\theta_0}{\epsilon} + 1\right) = 2K_A K_M \omega_c |G(j\omega_c)|. \tag{51}$$

Defining the phase margin ϕ_M by the equation

$$\phi_M = \angle\left(\frac{\theta_0}{K_v \epsilon}\right) - \pi, \tag{52}$$

the amplitude $|\theta_0 / K_v \epsilon|$ of the transfer function and the phase margin become:

$$\left|\frac{\theta_0}{K_v \epsilon}\right| = \frac{[(2\omega_c A_1 - mA_2)^2 \cos^2\phi + (2\omega_c A_2 - mA_1)^2 \sin^2\phi]^{1/2}}{2\omega_c |G(j\omega_c)| m(1 + \tau^2 m^2)^{1/2}} \tag{53}$$

$$\phi_M = \frac{\pi}{2} + \phi_1 - \tan^{-1}\frac{(2\omega_c A_2 - mA_1)\sin\phi}{(2\omega_c A_1 - mA_2)\cos\phi} - \tan^{-1}\tau m. \tag{54}$$

These results are in complete agreement with those of Sobczyk.[1]

One further comment should be made. The angle, ϕ, given by $\phi_2 - \phi_0$ can be calculated when ϕ_0 has been specified (ϕ_2 is known). The specification of ϕ_0 corresponds to selecting a phase for the reference winding of the motor. As stated in Table I, ϕ_0 should be taken to be the phase of the carrier portion of the motor input when the modulating frequency is zero.

TABLE I

Component	Input	Representation of Input	Representation of Output	Transfer Operator	Notes and Definitions
Linear Network with Transfer Function $G(j\omega)$	$v_1 = K_1 \cos(mt + \theta_1) \times \cos(\omega_c t + \theta_2) + K_2 \sin(mt + \theta_1) \times \sin(\omega_c t + \theta_2)$	$V_1 = i_{\theta_2} K_1 e^{j(mt + \theta_1)} + i_{\theta_2 - \pi/2} K_2 e^{j(mt + \theta_1 - \pi/2)}$	$V_0 = N \cdot V_1$	$N = i_{\theta_2 + \phi_2} i_{\theta_2} A_1 e^{j\phi_1} + i_{\theta_2 + \phi_2} i_{\theta_2 - \pi/2} A_2 e^{j(\pi/2)} e^{j\phi_1} + i_{\theta_2 + \phi_2 - \pi/2} i_{\theta_2} A_2 e^{-j(\pi/2)} e^{j\phi_1} + i_{\theta_2 + \phi_2 - \pi/2} i_{\theta_2 - \pi/2} A_1 e^{j\phi_1}$	$A_1 = \frac{1}{2}[G^- + G^+]$ $A_2 = \frac{1}{2}[G^- + G^+]$ $\phi_1 = \frac{1}{2}[\gamma^+ - \gamma^-]$ $\phi_2 = \frac{1}{2}[\gamma^+ + \gamma^-]$ $G^+ e^{j\gamma^+} = G[j(\omega_c + m)]$ $G^- e^{j\gamma^-} = G[j(\omega_c - m)]$
Two Phase Induction Motor with Voltage $-\cos[\omega_c t + \phi_0 - (\pi/2)]$ On the Reference Winding	$v_1 = K_1 \cos(mt + \theta_1) \times \cos \omega_c t + \phi_0 + \lambda) + K_2 \sin(mt + \theta_1) \times \sin(\omega_c t + \phi_0 + \lambda)$	$V_1 = i_{\phi_0 + \lambda} K_1 e^{j(mt + \theta_1)} + i_{\phi_0 + \lambda - \pi/2} K_2 e^{j(mt + \theta_1 - \pi/2)}$	$\Theta_0 = M \cdot V_1$	$M = K_M \dfrac{i_{\phi_0} 2\omega_c + i_{\phi_0 - \pi/2} m e^{-j(\pi/2)}}{jm(1 + jTm)}$	$\lambda = \lambda(m)$ $\lambda(0) = 0$ ϕ_0 is the phase of the carrier for $m = 0$. The fixed winding of the motor should be excited with a voltage of frequency ω_c and a phase $\phi_0 - (\pi/2)$. These conditions have been assumed here.
Induction Tachometer $\cos(\omega_c + \alpha)$ On Reference Winding	$\theta = K \cos(mt + \phi)$	$\Theta = Ke^{j(mt + \phi)}$	$V_0 = T\Theta$	$T = -K_\tau [i_{\alpha - \pi/2} m e^{-j(\pi/2)} - i_\alpha (1/\omega_c) m^2]$	If it is desired that the main output term have a sine carrier, take $\alpha = 0$. For a cosine carrier, take $\alpha = \pi/2$.

The phase of the carrier portion of the motor input voltage is the phase-shift to which the carrier is subjected by the network, namely, $\phi_2(m)$. Hence,

$$\phi_0 = \phi_2(0). \tag{55}$$

But also from Table I, it is seen that

$$\phi_2(m) = \tfrac{1}{2}[\gamma^+(m) + \gamma^-(m)]$$
$$= \tfrac{1}{2}\{\angle G[j(\omega_c + m)] + \angle G[j(\omega_c - m)]\}. \tag{56}$$

Hence,

$$\phi_0 = \phi_2(0) = \angle G(j\omega_c). \tag{57}$$

Therefore, when the stabilizing network has been specified, the phasing of the motor is also specified. That is, ϕ_0 is known.

Referring now to Fig. 2 where a servo with tachometric feedback or an output lead-stabilized servo is shown, the following expression may be written:

$$\Theta_0 = K_A[V_\epsilon - T\Theta_0]\cdot M. \tag{58}$$

Writing

$$V_\epsilon = i_0\epsilon e^{jmt} \tag{59}$$

$$\Theta_0 = \theta_0 e^{jmt}, \tag{60}$$

it follows that

$$\frac{\theta_0}{\epsilon} = K_A \frac{i_0\cdot M}{1 + K_A T\cdot M}. \tag{61}$$

By reference to Table I, this may be expanded. The expansion leads to

$$\frac{\theta_0}{\epsilon} = \frac{2\omega_c K_A K_M}{jm\left[\left(1 + 2\omega_c K_A K_r K_M + \frac{1}{\omega_c} K_A K_r K_M m^2\right) + jTm\right]}. \tag{62}$$

These two examples serve to illustrate the ease with which the transfer functions of carrier-frequency servomechanisms may be derived by use of the transfer operators of Table I. Note that the expressions derived by this scheme are equally valid whether or not the carrier frequency drifts, and therefore the formulas derived for the loop transfer functions may be used to study the off-frequency behavior of the servo.

VIII. Conclusions

The generalization of the concept of transfer function to that of transfer operator permits a complete unification of the techniques of analysis of carrier-frequency servos and dc servos. The transfer operators are combined in the same manner as are transfer functions except that addition and subtraction are replaced respectively by vector addition and subtraction. Multiplication is replaced by the inner product.

The transfer operator of a given component fully describes its behavior in a servo loop, both with regard to its normal behavior, and with regard to its behavior when the carrier frequency shifts. Since, however, the theory is a linear theory, large shifts of the carrier frequency, which result in saturation of the components, cannot be handled.

Finally it should be remarked that the concepts presented here are not restricted to servo theory but are applicable to the analysis of any suppressed-carrier amplitude-modulated transmission system.

Improvement in Gain Stability of the Superheterodyne Mixer Through the Application of Negative Feedback*

GAIL E. BOGGS†

Summary—**The superheterodyne mixer may be stabilized by the employment of the difference-frequency voltage as negative feedback. This results in increased gain stability and, for the case of the mixer couple, in increased gain-bandwidth product. Two mixer circuits using feedback are described. Generalized design curves are shown and a design procedure is outlined. A description of the experimental mixers and a discussion of the experimental results conclude the paper.**

I. Introduction

IN THE USE of field intensity and other frequency-selective measuring equipment, it is highly desirable to have the instrument retain gain calibration over reasonably long periods of time.

One method of mixer stabilization wherein the feedback is applied at signal frequency has been presented by Tucker.[1] Another method is here presented in which the difference frequency is used to stabilize the mixer as an amplifier.

Two types of mixers will be discussed, together with the derivation of their gain equations in order to show the improvement in stability. In addition, a design procedure is given to aid the design engineer.

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† National Bureau of Standards, Central Radio Propagation Laboratory, Washington 25, D. C.

[1] D. G. Tucker, "Frequency changers and amplifiers with constant gain," Proc. I.R.E., vol. 37, pp. 1324–1327; November, 1949.

II. Theory of a Feedback Mixer

In general, it is desirable to use a large voltage of oscillator frequency in order to minimize changes in conversion gain with variations in applied voltage at signal frequency. As a result, the mixer transconductance is driven from maximum on one-half cycle to zero on the other. Obviously the instantaneous transconductance is a function of oscillator voltage. This is shown in Fig. 1.[2]

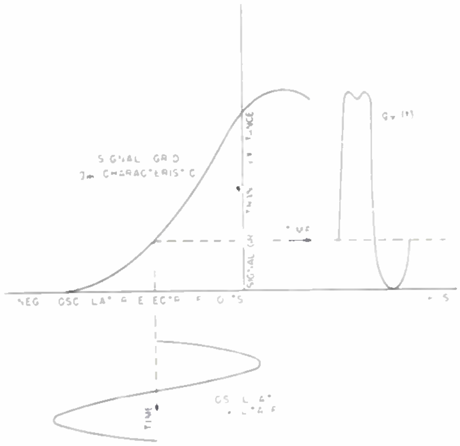

Fig. 1—Transconductance as a function of oscillator voltage.

The instantaneous transconductance can be expressed by the Fourier series

$$g_m = b_0 + a_1 \sin \omega_2 t + a_2 \sin 2\omega_2 t + \cdots + b_1 \cos \omega_2 t + b_2 \cos 2\omega_2 t + \cdots, \quad (1)$$

where ω_2 is the angular frequency of the oscillator voltage.

For the type of switching or modulating function generally encountered, the transconductance as a function of time, $g_m(t)$, may, as an approximation, be represented by a square wave. Now by well-known methods of Fourier analysis,

$$g_m = \frac{\hat{g}_m}{2} + \frac{2\hat{g}_m}{\pi}\left[\sin \omega_2 t + \tfrac{1}{3}\sin 3\omega_2 t + \tfrac{1}{5}\sin 5\omega_2 t + \cdots\right], \quad (2)$$

where $\hat{g}_m$ is the maximum value of transconductance.

While (2) may often be used in mixer analysis, it has been shown experimentally that sufficient accuracy cannot be obtained when using tubes such as the 6SA7 and 6SB7-Y. Therefore, the switching function for the 6SB7-Y was determined experimentally and is shown in Fig. 2. In this case, the transconductance may be represented by the series,

$$g_m = b_0 + a_1 \sin \omega_2 t + a_3 \sin 3\omega_2 t + a_5 \sin 5\omega_2 t + \cdots + b_2 \cos 2\omega_2 t + b_4 \cos 4\omega_2 t + b_6 \cos 6\omega_2 t + \cdots, \quad (3)$$

where the constants are obtained by graphical integration. The constants a_5 and b_6 are neglected since, with graphical integration, they contribute very little to the result. Equation (3) may now be applied to the analysis of a simple feedback mixer.

[2] E. W. Herold, "The operation of frequency converters and mixers for superheterodyne reception," Proc. I.R.E., vol. 30, pp. 84–103; February, 1942.

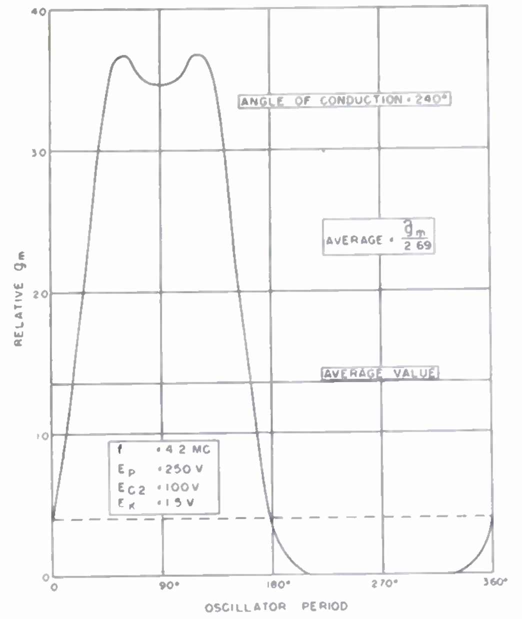

Fig. 2—Switching function 6SB7-Y

Consider the circuit shown in Fig. 3. Let the signal input voltage at angular frequency ω_1 be

$$e_s = E_s \sin \omega_1 t. \quad (4)$$

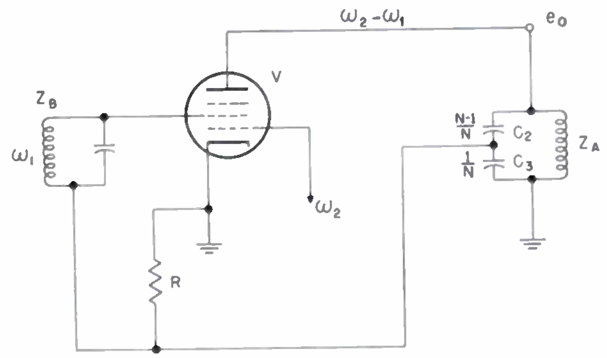

Fig. 3—Schematic diagram of feedback mixer.

Since the output circuit Z_A is tuned to the difference frequency $(\omega_2 - \omega_1)$, only this component need be considered in the ouput voltage, e_0, thus,

$$e_0 = E_0 \cos (\omega_2 - \omega_1)t. \quad (5)$$

The input circuit Z_B is tuned to the frequency of the incoming signal (ω_1), and hence, its impedance at the difference frequency may be neglected. The feedback voltage is, therefore, $(1/N)e_0$, where N is the stepdown ratio of the tuned output circuit. The grid-to-cathode voltage may now be given as

$$e_g = E_s \sin \omega_1 t + \frac{1}{N} E_0 \cos (\omega_2 - \omega_1)t. \quad (6)$$

The output voltage of the mixer is $g_m e_g Z_A$.

Substituting the expressions given in (3) and (6) in the foregoing, expanding, and neglecting all terms which do not contain the difference frequency, yields

$$e_0 = \left[-b_1 \frac{1}{N} E_0 \cos(\omega_2 - \omega_1)t + \frac{a_1 E_s}{2} \cos(\omega_2 - \omega_1)t \right] Z_A.$$

Collecting terms and with the aid of (5),

$$E_0 = \frac{a_1 E_s Z_a}{2\left(1 - \frac{b_0 Z_a}{N}\right)}. \tag{7}$$

Now by definition, the gain with feedback is,

$$A_f = \frac{E_0}{E_s}.$$

Therefore,

$$A_f = \frac{a_1 Z_a}{2\left(1 - \frac{b_0 Z_a}{N}\right)}. \tag{8}$$

For simplification let

$$b_0 = k_0 \frac{\hat{g}_m}{\pi} \quad \text{and} \quad a_1 = k_1 \frac{\hat{g}_m}{\pi},$$

where k_0 and k_1 are constants determined by the switching function of the mixer tube selected.

Substituting

$$A_f = \frac{k_1 \hat{g}_m Z_a}{2\left(\pi - \frac{k_0 \hat{g}_m Z_a}{N}\right)}. \tag{9}$$

This expression for the gain of the mixer with feedback is quite similar to the gain expression for a single-stage feedback amplifier.

In (9) it can be seen that when $(k_0 \hat{g}_m Z_a/N) \gg -\pi$, the gain to a good approximation is no longer a function of the transconductance. Thus it is shown that the use of the difference-frequency voltage for feedback results in an improvement in the mixer gain stability with changing transconductance.

It should be noted that the constant k_1 is directly proportional to the conversion transconductance, while k_0 is a relative measure of the average value of amplifier transconductance. k_0 and k_1 are subject to variation due to changes in the shape of the switching function which may result from a change in oscillator voltage as well as other causes. While this may be considered as a limiting factor for stability improvement, experimental results indicate that these constants tend to change together. With a high degree of feedback, examination of (9) indicates little change in A_f with changes in the value of the constants, provided that the changes of k_0 and k_1 are in the same direction and of like percentage. Unfortunately, the stability of the switching function varies with tube types, and this factor must be considered when employing feedback mixers.

III. A Simple Feedback Mixer

Referring again to Fig. 3, the conversion transconductance g_c is, by definition $a_1/2$. Therefore, the complex gain of the stage as a mixer without feedback is $(k_1 g_m/2\pi) Z_a$. This will be denoted by A. Substituting in (9) yields

$$A_f = \frac{k_1 A}{\left(k_1 - \frac{2k_0 A}{N}\right)}, \tag{10}$$

where $A = A/1+jx$; A is the gain at f_0 and $x = Qu$; $u = (f/f_0) - (f_0/f)$.

Substituting in (10) and simplifying gives,

$$A_f = \frac{A}{\left(1 - \frac{2k_0 A}{Nk_1}\right) + jx}. \tag{11}$$

By definition,

$$B = -\frac{2k_0 A}{Nk_1},$$

where B is the feedback factor; therefore,

$$A_f = \frac{A}{(1+B) + jx}. \tag{12}$$

For design purposes, it is convenient to have an expression for normalized gain in order to determine the selectivity characteristic for varying degrees of feedback. Let the normalized gain be denoted by a.

Now,

$$a = A_f \frac{1+B}{A}.$$

Expanding and squaring reals and imaginaries to obtain the magnitude yields

$$a = \left[\frac{1}{1 + \left(\frac{x}{1+B}\right)^2} \right]^{1/2}. \tag{13}$$

Equation (13) may be plotted in order to allow the design engineer to readily obtain the selectivity characteristic. Note, however, that at the half-power points $(a = 1/\sqrt{2})$, $x/(1+B)$ is equal to one.

Design Procedure:

Given Δf, f_0, A_{f0}, and g_c, where A_{f0} is the center frequency gain with feedback, $f_0 = f_2 - f_1$, and Δf is the bandwidth between half-power points,

(a) Determine values of k_0 and k_1 for the tube under consideration. These constants may be determined experimentally as described.

(b) Choose value of $1+B$ and calculate Q from $Q(\Delta f/f_0) = 1+B$ at half-power points. If Q is impractically large, choose a smaller value of $1+B$ and redesign.

(c) Calculate $C_1 = g_c Q/\omega_0 A_{f0}(1+B)$, where $1/C_1 = = (1/C_2) + 1/C_3)$.

(d) Calculate $A = A_{f0}(1+B)$.

(e) Calculate $N = 2k_0 A/k_1 B$.

(f) Calculate $C_2 = NC_1/N - 1$

$C_3 = NC_1$.

IV. The Mixer Couple

In applying voltage feedback to either an amplifier or mixer, the bandwidth is usually increased. Since a large amount of feedback is generally required to substantially improve the gain stability, it is apparent that a very high Q is necessary for a relatively narrow bandwidth.

If feedback is applied over two stages, a relatively narrow bandwidth with improved flatness will result, using a practically obtainable coil Q. The material following is an adaptation from a paper by G. F. Montgomery,[3] and many of the equations are obtained directly from this work.

The circuit of the mixer couple is shown in Fig. 4. Since a pentagrid mixer is again employed, the switching function given in (3) will be used as a basis for the analysis.

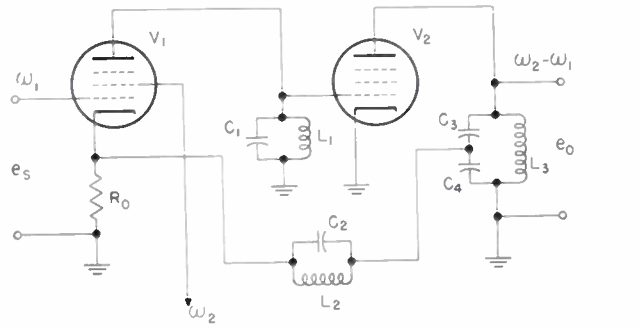

Fig. 4—Schematic diagram of mixer couple.

Let the input voltage at angular frequency ω_1 be

$$e_s = E_s \cos \omega_1 t.$$

Since the output voltage contains only the difference frequency,

$$e_0 = E_0 \sin (\omega_2 - \omega_1)t.$$

Neglecting local feedback at the cathode, the signal grid voltage is

$$e_g = e_s \frac{e_0 R_0}{(R_0 + Z_0)N};$$

hence,

$$e_0 = \frac{g_{m_1} g_{m_2} Z_1 Z_3 e_s}{1 + \dfrac{R_0}{(R_0 + Z_2)N} g_{m_1} g_{m_2} Z_1 Z_3}.$$

Substituting for g_{m1} and expanding trigonometric functions,

$$A_c = \frac{E_0}{E_s} = \frac{k_1 \hat{g}_{m_1} g_{m_2} Z_1 Z_3}{2\left(\pi + \dfrac{R_0}{N(R_0 + Z_2)} k_0 \hat{g}_{m_1} g_{m_2} Z_1 Z_3\right)}. \quad (14)$$

A_c is the complex voltage gain of the mixer couple. Upon inspection of (14), it is seen that with a high degree of feedback, the gain is largely independent of the tube transconductance.

[3] G. F. Montgomery, "Intermediate frequency gain stabilization with inverse feedback," Proc. I.R.E., vol. 38, pp. 662–667; June, 1950.

Since the balance of this derivation is quite lengthy and follows a pattern similar to that already presented,[3] it will be omitted.

Design Procedure:

Given Δf, f_0, A_{c0}, g_c, and g_{m_2}, where $f_0 = f_2 - f_1$ and A_{c0} = voltage gain of mixer couple at f_0,

(a) Determine values of k_0 and k_1 as before.

(b) Choose value of $1+B$. In Fig. 5 find $Q_1(\Delta f/f_0)$, for $a = 1/\sqrt{2}$.

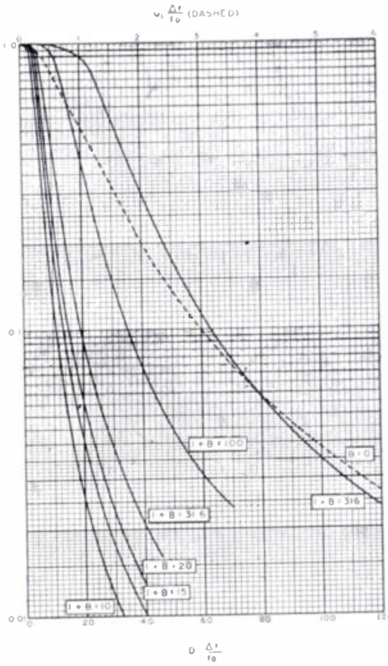

Fig. 5—Normalized gain of mixer couple.

(c) Calculate Q_1. If Q_1 is impractically large, choose a smaller value of $1+B$.

(d) Calculate

$$C_1 = \frac{Q_1}{\omega_0} \sqrt{\frac{g_c g_{m2}}{A_{c0}(1 + B)}}.$$

(e) Calculate $R_1 = Q_1/\omega_0 C_1$.

(f) In Fig. 6 find P and A/N for chosen $1+B$.

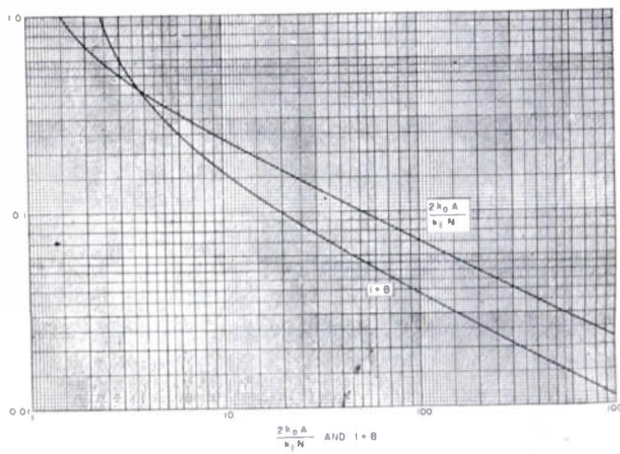

Fig. 6—Design factor chart.

(g) Calculate $A=(1+B)A_{c0}$ and N.
(h) Choose R_0 from tube data.
(i) Calculate

$$R_2 = \left(\frac{2k_0A}{k_1BN} - 1\right)R_0.$$

(j) Verify $N^2R_2 \gg R_1$. If this is not true, choose a smaller value of $1+B$ and redesign.
(k) Calculate $Q_2=(AP/BN)Q_1$.
(l) Calculate $C_2=Q_2/\omega_0R_2$.
(m) Calculate $C_3=NC_1/N-1$
$C_4=NC_1$.

V. Experimental Results

In the case of either mixer, the electrical arrangement represented in the circuit diagram must be duplicated as closely as possible if results are to match the predictions of the design.

For the simple feedback mixer, the experimental verification is not difficult when the switching function is known. A 6SB7-Y tube was selected because of its high conversion transconductance. The design parameters were:

Given $\Delta f=33$ kc, $f_0=450$ kc, $Af_0=3.7$
$g_c=740\ \mu$ mhos
$k_0=1.17$, $k_1=1.75$, as determined by graphical integration of the switching function.

Calculated $1+B=20$, $Q=270$, $C_1=955\ \mu\mu$f, $R_1=100k$, $1/N=0.194$, $C_2=1{,}185\ \mu\mu$f, $A=74$ and $C_3=4{,}930\ \mu\mu$f.

The experimental results agreed well with prediction, although the bandwidth was slightly narrow, measuring 31 kc.

It was noted that the bandwidth varies with oscillator injection voltage as would be expected since this varies A. The oscillator excitation was observed to be quite noncritical insofar as gain with feedback was concerned.

In Fig. 7 is shown the result of varying the plate-supply voltage. The calculated curve is plotted with reference to the zero feedback curve, on the assumption that the only variable is the tube transconductance. To the extent that this assumption may hold true, the results are thought to be in good agreement with prediction.

The following table will give the reader a good idea of what may be expected with a value of $1+B=20$. Examination of this table indicates a considerable improvement in gain stability. It is apparent that a large change in conversion transconductance results in a relatively small change in voltage gain.

TABLE I

g_c	Δg_c	Calculated		Experimental	
		A_{c0}	ΔA_{c0}	A_{c0}	ΔA_{c0}
770 μmhos	−0 db	3.67	0 db	3.70	0 db
545	−3	3.60	−0.16	3.70	0
385	−6	3.51	−0.40	3.65	−0.12
244	−10	3.35	−0.80	3.48	−0.56
77	−20	2.68	−2.74	2.81	−2.36

Before further discussion of the mixer couple experimental results, a few general comments will be made. Due to the increased gain-bandwidth product obtained with the couple, the gain per stage will be larger than for a zero-feedback amplifier. Hence adequate precautions to reduce regeneration, such as shielding and power-lead decoupling, are recommended. The screen grid of the mixer stage should be bypassed directly to the cathode since the formulas for R_0 and C_1 will not be correct if the screen is bypassed to ground.

With the usual values of R_0, the impedance of the feedback coil may be very small. With a low impedance it may very likely be quite difficult to obtain the required

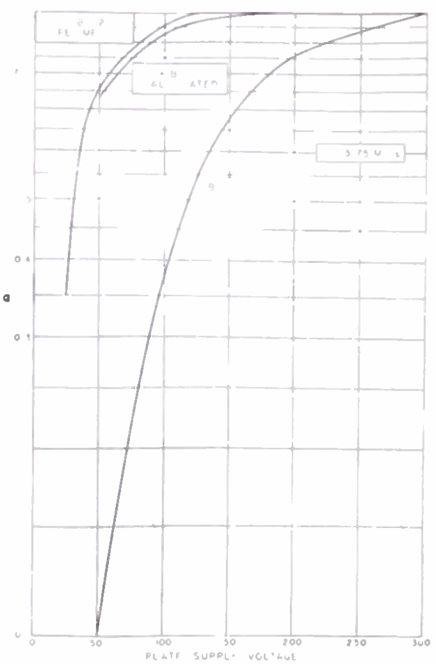

Fig. 7—6SB7-Y gain stability

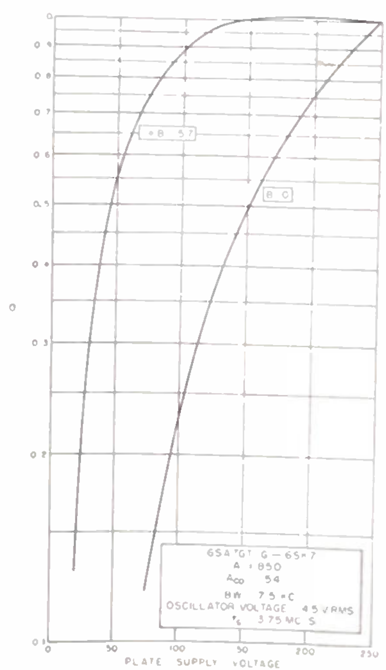

Fig. 8—Mixer-couple gain stability.

Q_2. In that event it is permissible to increase R_0, provided that the correct bias is maintained. In the derivation presented, it was assumed that the internal cathode impedance of the mixer would be much larger than R_0. Hence, if R_0 is increased, it is quite possible that the cathode impedance may be less than R_0; therefore, the feedback and bandwidth will be reduced. This point should be borne in mind when experimental work is being carried on with the mixer couple.

The experimental mixer couples using 6SA7 or 6SB7-Y mixer tubes followed by a 6SK7 amplifier were found to agree with the calculated results within one db. A normalized curve showing gain variations with plate supply voltage for the 6SA7GT/G mixer is given in Fig. 8. Similar curves for the 6SB7-Y are shown in Fig. 9.

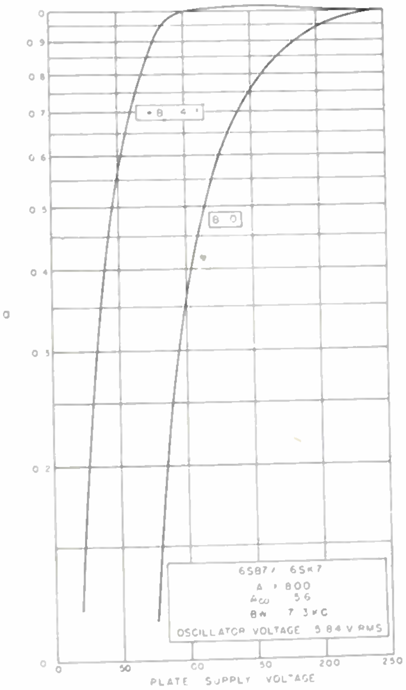

Fig. 9—Mixer-couple gain stability.

Again the oscillator excitation voltage affects A, but is noncritical for A_{c0}. In Fig. 9 it is observed that A_{c0} increases slightly at a reduced plate supply voltage. This effect came about due to slightly deficient oscillator excitation in the experimental setup used.

VI. Conclusion

It has been shown that the difference-frequency voltage may effectively be used as feedback to stabilize the mixer gain. While the oscillator voltage must be carefully adjusted to obtain the correct bandwidth, it is apparent that the gain with feedback, A_{c0}, is relatively independent of this adjustment.

For the simple mixer, high gain is required in order to apply sufficient feedback to effect an improvement in gain stability. For this reason a tube with a high-conversion transconductance should be selected. Also the tuned plate circuit should have high impedance and high Q.

Comparing the mixer couple to a cascade, synchronous, single-tuned mixer-amplifier arrangement, the couple provides improved gain stability, greater flatness, and steeper skirts. Careful attention to the design and construction of the mixer couple is necessary if the results are to closely approximate the calculated values.

The plate circuits are peaked in normal fashion with the feedback line broken. When feedback is applied, the feedback coil is adjusted for a symmetrical response about the center frequency. A peaked response of insufficient bandwidth indicates either that Q_2 is too high or that the cathode impedance of the first tube is too small. If R_0 is fairly large, the latter condition may be due to the internal cathode impedance of the tube. A response with a dip in the center indicates the opposite condition to that just described. In obtaining a flat-top response, the value of R_0 is often a useful final adjustment. When a flat-top response is attained, it will be necessary to adjust the oscillator voltage for the correct A and bandwidth.

If the mixer tube is operated with a fairly large value of R_0, it is desirable to connect the suppressor to the cathode to prevent biasing of the suppressor grid. In the case of the 6SB7-Y and 6SA7 metal tubes, the suppressor is internally connected to the shell. Hence, the above connection leaves the shell "hot" for rf which is undesirable. The metal types can be made to operate satisfactorily but require a larger value of R_0 than would be anticipated. For this reason, it is recommended that tubes be selected which do not have an internal suppressor-shell connection.

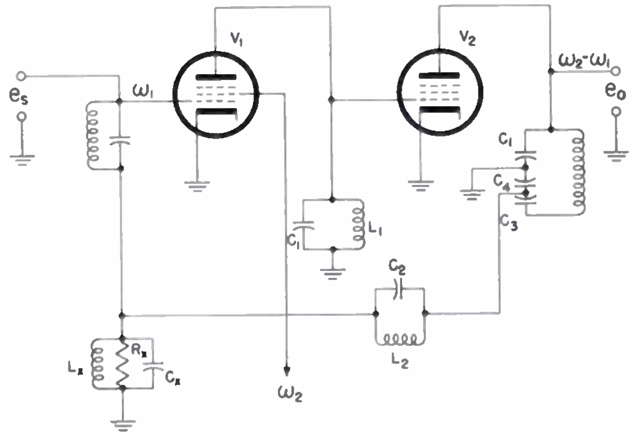

Fig. 10—Schematic diagram of reversed-phase mixer couple.

The above problems, as well as the local feedback at signal and oscillator frequency due to the large cathode impedance, may be eliminated by using the circuit arrangement shown in Fig. 10. In this case the phase of the output voltage is reversed in the output tank and the feedback is applied to the signal grid. The proper terminating impedance for the Beta circuit is provided by tuned circuit consisting of shunt connected L_x, C_x, and R_x. Since this circuit must provide an essentially resistive termination over the IF pass band, $Q_x \ll Q_2$. R_x must be chosen such that it will present a sufficiently low impedance at the lowest signal frequency and at the same time allow practical realization of Z_2. The analysis of this circuit is identical to the preceding case, allowing for the voltage loss at the output tank.

Mutual Coupling of a Slot with a Dipole Antenna*

WALTER J. SURTEES†, MEMBER, IRE

Summary—A parameter is defined from which may be obtained the mutual coupling between a radiating slot, cut in a plane perfectly-conducting sheet, and a dipole fed at its base on the conducting plane. Using a slot cut in a sheet of copper and fed by a waveguide, experimental values of this parameter were obtained for various positions of the dipole relative to the slot. These values are plotted and compared with the theoretical ones, very good agreement being obtained.

I. Introduction

The theory of slots cut in rectangular waveguides has been treated elsewhere.[1-3] Recently, papers dealing with slots cut in a plane, perfectly-conducting sheet have been published.[4-7] Values for the input admittance of the slot agree for these cases, assuming the slot is narrow, its length is almost a half-wavelength, the field in the slot is transverse to the long dimension and varies sinusoidally along the slot, the sheet is perfectly-conducting and infinitely thin, and the face containing the slot is an infinite plane.

A slot in a waveguide may be made to radiate by inserting a probe in the guide near the slot. The probe may also act as an antenna to control the radiation pattern. The effect of the probe may be analyzed by considering an infinite set of images of the array and computing the coupling of all the images to the slot. Since the two nearest images will be about one-half and one wavelength from the slot, the mutual is not great and a good first approximation will be obtained by taking only the coupling between one slot and one dipole.

This paper gives an account of how this coupling may be determined when the dipole is thin, and with the same assumptions for the slot and sheet as indicated above.

II. Development of Theory

Consider an infinite, perfectly-conducting sheet coinciding with the XZ plane of a rectangular co-ordinate system. There is a slot of length $2L_1$ along the Z-axis whose width is $2a$ (see Fig. 1). If $2a$ is small compared with the wavelength and with the length of the slot, the electric field distribution along the slot can be assumed to be sinusoidal. By using Babinet's Principle[8] and using the field equations for the complementary dipole (see Brown),[9] the field components, at a point P, for a thin, center-fed slot of arbitrary length are as follows:

$$H_z = \frac{-jV_1}{2\pi Z_0 \sin kL_1}\left[\frac{\exp(-jkr_1)}{r_1} + \frac{\exp(-jkr_2)}{r_2} - \frac{2\exp(-jkr_0)}{r_0}\cos kL_1\right] \tag{1a}$$

$$H_\rho = \frac{jV_1}{2\pi Z_0 \rho \sin kL_1}\left[\frac{(z-L_1)}{r_1}\exp(-jkr_1) + \frac{(z+L_1)}{r_2}\exp(-jkr_2) - \frac{2z}{r_0}\exp(-jkr_0)\cos kL_1\right] \tag{1b}$$

$$E_\phi = \frac{-jV_1}{2\pi\rho \sin kL_1}\left[\exp(-jkr_1) + \exp(-jkr_2) - 2\exp(-jkr_0)\cos kL_1\right]. \tag{1c}$$

Where

$Z_0 = 120\pi$ ohms, the impedance of free space,

V_1 = the voltage difference between the edges of the slot, at its center,

$k = 2\pi/\lambda = \omega/c$,

ω = applied angular frequency,

$r_1, r_2, r_0, z, \rho, \phi$ are as defined in Fig. 1.

All the components are to be multiplied by the time factor $\exp(j\omega t)$.

Now consider a dipole antenna of length L_2, parallel to the Y-axis, mounted at the point (z, d) on the plane sheet. The component of electric field intensity along it is $E_\phi \cos\phi$, where E_ϕ is given by (1c). This field will induce a current along the length of the dipole.

The configuration may be considered as a four terminal network.[10] Let the slot be fed at terminals 1-1, and the dipole at terminals 2-2. V_1 and V_2 are the applied voltages at these terminals, and cause currents I_1 and

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† University of Toronto, Toronto, Ontario, Canada.

[1] A. F. Stevenson, "Theory of slots in rectangular waveguides," *Jour. Appl. Phys.*, vol. 19, pp. 24–38; January, 1948.

[2] W. H. Watson, "Waveguide Transmission and Antenna Systems," Oxford Univ. Press, London, England; 1947.

[3] A. L. Cullen, "Laterally-displaced slot in rectangular waveguide," *Wireless Eng.*, vol. 26, pp. 3–10; January, 1949.

[4] J. L. Putnam, B. Russell, W. Walkinshaw, "Field distribution near a centre-fed half-wave radiating slot," *Jour. IEE*, vol. 95, pp. 282–289; July, 1948.

[5] J. L. Putnam, "Input impedances of centre-fed slot aerials," *Jour. IEE*, vol. 95, pp. 290–294; July, 1948.

[6] J. W. Crompton, "Impedance characteristics of some slot aerials," *Jour. IEE*, vol. 97, pp. 39–44; January, 1950.

[7] N. A. Begovich, "Slot radiators," Proc. I. R. E., vol. 38, pp. 803–806; July, 1950.

[8] H. G. Booker, "Slot aerials and their relation to complementary wire aerials," *Jour. IEE*, vol. 93, pp. 620–626; 1946.

[9] G. H. Brown, "Directional antennas," Proc. I. R. E., vol. 25, pp. 78–145; January, 1937.

[10] E. C. Jordan, "Electromagnetic Waves and Radiating Systems," Prentice-Hall, Inc., New York, N. Y., pp. 347 ff.; 1950.

I_2 to flow. Then, following the notation of Guillemin,[11] the general mesh equations are:

$$I_1 = g_{11}V_1 + g_{12}I_2 \tag{2a}$$

$$V_2 = g_{21}V_1 + g_{22}I_2. \tag{2b}$$

Now:

$g_{12} = -g_{21}$,

g_{11} = admittance at terminals 1-1 when 2-2 are opened.

g_{22} = impedance at terminals 2-2 when 1-1 are shorted.

If the slot and dipole are not too near, then approximately:

$$\left.\begin{aligned} g_{11} &= Y_{11}, \text{ the self-admittance of the slot alone,} \\ g_{22} &= Z_{22}, \text{ the self-impedance of the dipole alone.} \end{aligned}\right\} \tag{3}$$

When the dipole is not directly excited but the slot is, $V_2=0$; then from (2), Y_i^{-1}, the slot input impedance is:

$$Y_i = \frac{I_1}{V_1} = Y_{11} + \frac{(g_{21})^2}{Z_{22}}. \tag{4}$$

Applying the reciprocity theorem in a manner similar to Jordan,[11] the mutual coupling coefficient g_{21}, referred to the feed terminals is

$$g_{21} = \frac{j}{2\pi \sin kL_1 \sin kL_2} \int_0^{L_2} [\exp(-jkr_1) + \exp(-jkr_2) - 2\cos kL_1 \exp(-jkr_0)] \frac{\cos\phi}{\rho} \sin k(L_2 - y)dy. \tag{5}$$

In the analytic and experimental work, L_1 and L_2 where chosen to be resonant. Thus, (5) becomes:

$$g_{21} = \frac{j}{2\pi} \int_0^{\lambda/4} [\exp(-jkr_1) + \exp(-jkr_2)] \cdot (\cos\phi/\rho) \cos ky\, dy. \tag{6}$$

On carrying out the integration, the following equation for g_{21} is obtained. (See Appendix for details).

$$\begin{aligned} g_{21} = \frac{j}{8\pi} (&\exp(kd)\{\exp(jkz)[Ei(W_{23}) - Ei(W_{11}) \\ &+ Ei(W_{22}) - Ei(W_{32})] \\ &+ \exp(-jkz)[Ei(W_{21}) - Ei(W_{31}) \\ &\quad + Ei(W_{24}) - Ei(W_{12})]\} \\ &- \exp(-kd)\{\exp(jkz)[Ei(W_{53}) - Ei(W_{41}) \\ &+ Ei(W_{52}) - Ei(W_{62})] \\ &+ \exp(-jkz)[Ei(W_{51}) - Ei(W_{61}) \\ &+ Ei(W_{54}) - Ei(W_{42})]\}). \end{aligned} \tag{7a}$$

where:

$$\begin{aligned} W_{11} &= kd + jk(R_1 - 2L + z), \\ W_{12} &= kd + jk(R_2 - 2L - z), \\ W_{21} &= kd + jk(R_1 - z), \\ W_{22} &= kd + jk(R_2 + z), \\ W_{23} &= kd + jk(R_1 + z), \\ W_{24} &= kd + jk(R_2 - z), \\ W_{31} &= kd + jk(R_1 + 2L - z), \\ W_{32} &= kd + jk(R_2 + 2L + z), \\ W_{41} &= -kd + jk(R_1 - 2L + z), \\ W_{42} &= -kd + jk(R_2 - 2L - z), \\ W_{51} &= -kd + jk(R_1 - z), \\ W_{52} &= -kd + jk(R_2 + z), \\ W_{53} &= -kd + jk(R_1 + z), \\ W_{54} &= -kd + jk(R_2 - z), \\ W_{61} &= -kd + jk(R_1 + 2L - z), \\ W_{62} &= -kd + jk(R_2 + 2L + z), \\ R_1 &= (d^2 + 2L^2 - 2Lz + z^2)^{1/2}, \\ R_2 &= (d^2 + 2L^2 + 2Lz + z^2)^{1/2}, \\ j &= (-1)^{1/2},\ k = 2\pi/\lambda,\ L = \lambda/4, \\ \lambda &= \text{wavelength}, \end{aligned} \tag{7b}$$

d and z are distances giving the position of the dipole relative to the slot (see Fig. 1),

$Ei(W) = \int_W^\infty [\exp(-v)/v]dv$, the exponential integral of complex argument.

When the dipole is located on the X-axis, that is, at a point along a line bisecting the length of the slot such that $z=0$, (6) becomes

$$g_{21} = \frac{+j}{\pi} \int_0^{\lambda/4} \{[\exp(-jkr_1) \cos\phi \cos ky]/\rho\} dy. \tag{8}$$

Either integrating this expression, or setting $z=0$ in (7), the following is obtained

$$g_{21} = \frac{-j}{4\pi} \{\exp(kd)[Ei(W_1) - 2Ei(W_2) + Ei(W_3)] - \exp(-kd)[Ei(W_4) - 2Ei(W_5) + Ei(W_6)]\} \tag{9a}$$

where

$$\begin{aligned} W_1 &= kd + jk(R - 2L), \\ W_2 &= kd + jkR, \\ W_3 &= kd + jk(R + 2L), \\ W_4 &= -kd + jk(R - 2L), \\ W_5 &= -kd + jkR, \\ W_6 &= -kd + jk(R + 2L), \\ R &= (d^2 + 2L^2)^{1/2}. \end{aligned} \tag{9b}$$

III. Basis for Experimental Method

Babinet's Principle[8] can be applied to establish a relationship between the self-admittance, Y_{11}, of a slot and the self-impedance, Z_{22}, of its complementary dipole. It is

$$Z_{22} = Y_{11}(Z_0/2)^2 \tag{10}$$

where, Z_0 is the impedance of free space.

[11] E. A. Guillemin, "Communication Networks," vol. II, p. 137, John Wiley and Sons, New York, N. Y.; 1935.

Now, (4) may be rearranged to give

$$(g_{21})^2 = Y_{11}Z_{22}[(Y_i/Y_{11}) - 1]. \tag{11}$$

The self-impedance of a quarter-wavelength, grounded dipole is known to be about 36.6 ohms, resistive. So that Y_{11}, the self-admittance of a half-wavelength slot radiating on one side of a conducting plane, may be obtained from (10) and substituted into (11). The ratio Y_i/Y_{11} may be obtained directly from measurements of the electric field standing wave ratios for the input admittance Y_i of the slot with the dipole present, and the input admittance Y_{11} of the slot with the dipole absent.

IV. Description of Experimental Method and Equipment

The experimental equipment used was the following. A 5,800 to 7,500 mc klystron oscillator, with its associated power supply and modulator, was employed as the signal source. The output was fed into standard $1\frac{1}{2}$ by $\frac{3}{4}$ inch waveguide. A variable attenuator, wavemeter, slotted section with tunable probe, and a standard standing wave meter were also used. The final section of the waveguide was soldered to the center of a four foot square copper sheet. A slot 2.10 centimeters in length and 0.1 centimeter in width, symmetrically located with respect to the walls of the waveguide, was cut in the copper sheet. In order to mount the dipole, small holes were drilled in the sheet at about one-sixteenth wavelength spacings on two lines, corresponding to the X-axis and the line z, as shown in Fig. 1.

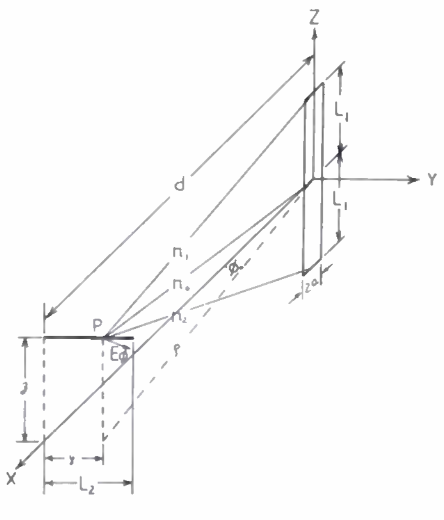

Fig. 1—Co-ordinate system for a slot of length $2L_1$, cut in an infinite-conducting plane coinciding with $Y=0$, and a dipole of length L_2 fed from this plane.

The frequency at which the input impedance of the slot was purely resistive, was found. The source frequency was varied and the standing wave ratio in the waveguide, as well as the shift in the minimum, when the slot was completely short-circuited, were observed. The resonant frequency of the slot was 6,732 mc. At this frequency, the slot length was about 0.47 wavelengths, which agrees favorably with the value given by Putnam[5] and Crompton.[6] The oscillator was set at this frequency and the standing wave ratios and the positions of the minima were then measured for a thin dipole, half the length of the slot, mounted at different positions on the conducting sheet. Measurements were repeated with dipoles of slightly longer length, and dipoles having diameters of one-half and one and one-half times the width of the slot.

V. Experimental Results

From the measured standing wave ratios and the positions of the electric field minima, the ratio Y_i/Y_{11}, for different values of d and z, was found. These values were inserted in (11) and g_{21} was computed. The theoretical values were obtained from (7) and (9). All these results are plotted in Fig. 2 and Fig. 3. It appears that the value of g_{21} determined experimentally is the same as that determined from theoretical considerations, within the limits of experimental error.

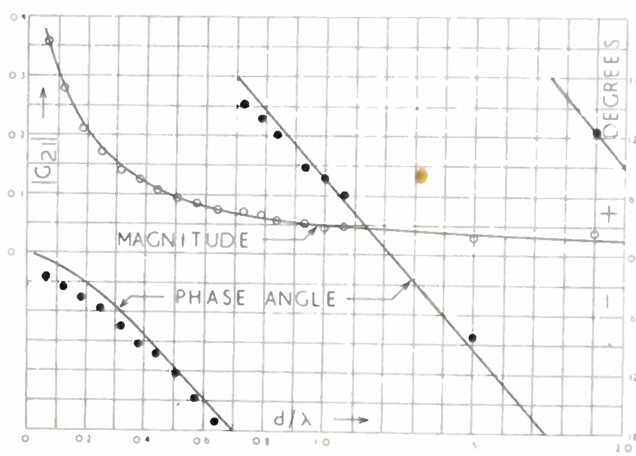

Fig. 2—Magnitude and phase angle of the mutual coupling parameter g_{21}, for $z=0$, as a function of spacing d/λ.
——— Theoretical values,
○, ● experimental values for the magnitude and phase angle, respectively.

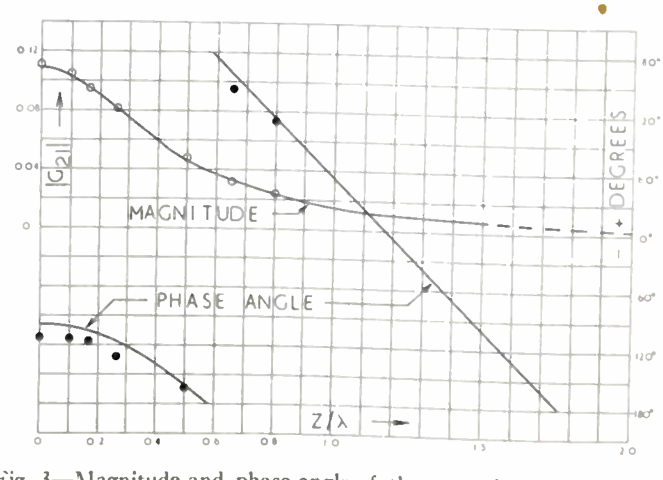

Fig. 3—Magnitude and phase angle of the mutual coupling parameter g_{21}, for $d=0.44\lambda$, as a function of z/λ.
——— Theoretical values,
○, ● experimental values for the magnitude and phase angle, respectively.

For positions of the dipole at large distances from the slot, the absolute value of g_{21} approaches zero. This is to be expected, since the field from the slot falls off with distance from the slot, the induced current along the dipole will decrease with an increase in spacing. Thus the input admittance of the slot is only its self-admittance in the limit as d becomes very large, (see (4)).

Measurements were made with the dipole placed inside the slot opening. With a thick dipole, the slot was short-circuited; when the dipole was thin, the absolute value of g_{21} was approximately 0.48. This was lower than expected, thus showing that the voltage distribution along the slot was disturbed.

The error in the measured values of the imaginary component of g_{21} compared to the value obtained from theory, for close spacings, is analogous to the error in predicting accurately, by the "induced emf method," the reactive component of the self-impedance of an antenna which is assumed infinitely thin.

No noticeable differences in the results were seen when the diameter of the dipole was increased or decreased by one-half. When the length of the dipole was increased slightly, the change to g_{21} was very small, but it indicated that a small reactive component was added to the value of Z_{22}.

VI. Conclusions

Since the theory has been developed for infinitely thin slot and dipole antennas, any application, for example, the possibility of controlling the radiation pattern of slot antennas, will be limited to very thin antennas. For a slot cut in the wall of a waveguide, some error would be made due to the thickness of the walls and due to the face containing the slot being finite and not infinite planes as considered here.

VII. Acknowledgment

The author wishes to express his appreciation for the support extended by the Defence Research Board of Canada, and for facilities given by the Department of Electrical Engineering, University of Toronto, without which this paper would not have been possible. Also, appreciation is extended to Professor George Sinclair of the University of Toronto for his aid and assistance, to John H. Craven for his aid in parts of the experimental work, and to James Wilbur who carried out some of the necessary computations.

Appendix

Evaluation of the Mutual Parameter g_{21}

From (6), the expression for the mutual parameter is:

$$g_{21}=\frac{j}{2\pi}\int_0^{\lambda/4}[\exp(-jkr_1)+\exp(-jkr_2)](\cos\phi/\rho)\cos ky\,dy,$$

where:

$\cos\phi=d/\rho$, $\rho^2=d^2+y^2$,

$r_1{}^2=\rho^2+(L-z)^2$, $r_2{}^2=\rho^2+(L+z)^2$.

$L=$ half length of the slot $=\lambda/4$,

d and z determine spacing of the slot and dipole,

$y=$ distance along the dipole.

Now replace $\cos ky$ by $1/2$ $[\exp(jky)+\exp(-jky)]$, and note that

$$\frac{1}{\rho^2}=\frac{1}{r^2-(L-z)^2}=\frac{1}{2r(r+[L\mp z])}+\frac{1}{2r(r-[L\pm z])},$$

where the upper sign before z is taken when $r=r_1$, and the lower sign is taken when $r=r_2$. Thus g_{21} becomes the sum of eight integrals. A typical integral is the following:

$$J_2=\frac{jd}{4\pi}\int_0^{\lambda/4}\frac{\exp(-jk[r_1-y])dy}{2r_1(r_1-L+z)}.$$

Let $u^2=d^2+(L-z)^2$, then, $r_1{}^2=u^2+y^2$.

Changing the variable to s, by letting $us=r_1-y$, and $u/s=r_1+y$, and substituting into J_2, a new form is obtained. Expanding the denominator in partial fractions and letting

$$w=s-b,\quad\text{and}\quad v=s-c,$$

where,

$$b=\frac{1}{u}(L-z+jd),\quad\text{and}\quad c=\frac{1}{u}(L-z-jd),$$

the following is obtained,

$$J_2=\frac{1}{8\pi}\left\{-\exp(-jkub)\int_{W_{71}}^{W_{11}}[\exp(-w)/w]dw+\exp(-jkuc)\int_{W_{81}}^{W_{41}}[\exp(-v)/v]dv\right\}.$$

The limits W_{11} and W_{41} have the same values as those given in (7b). W_{71} and W_{81} have not been defined as they cancel in the final expression for g_{21}.

Thus, using the above results, and employing similar transformations to the remaining integrals in the expression for g_{21}, the value for this parameter as given by (7a) is obtained.

1952 IRE National Convention Program

WALDORF-ASTORIA HOTEL and GRAND CENTRAL PALACE—MARCH 3–6

Registration

Members and visitors may register at either the Waldorf-Astoria Hotel or Grand Central Palace at the following hours:

	Waldorf-Astoria	*Grand Central Palace*
Mon.	9 A.M.–5 P.M.	11:00 A.M.–9 P.M.
Tue.	9 A.M.–8 P.M.	9:30 A.M.–9 P.M.
Wed.	9 A.M.–6 P.M.	9:30 A.M.–6 P.M.
Thur.	9 A.M.–1 P.M.	9:30 A.M.–9 P.M.

Technical Sessions

Over 200 technical papers will be presented in 43 sessions. A schedule of sessions is listed below; 100-word summaries of pa pers appear in the following pages.

Exhibits

The Radio Engineering Show, featuring 347 engineering exhibits, will occupy four floors of Grand Central Palace. A list of exhibitors and their products starts on page 1A of this issue. Exhibits will be open during the Palace registration hours noted above.

Principal Events

The **Annual Meeting,** to be held at 10:30 A.M. on Monday in the Jade Room of the Waldorf, is for the entire membership. It will feature a novel presentation of 40 years of IRE by Alfred N. Goldsmith and John V. L. Hogan, two of the co-founders, entitled "The IRE: From Acorn to Oak."

A "get-together" **Cocktail Party** will be held on Monday evening from 5:30 to 8:00 in the Grand Ballroom of the Waldorf. Tickets may be purchased at $3.80 each.

The **President's Luncheon,** on Tuesday at 12:45 P.M. in the Starlight Roof of the Waldorf, will honor IRE President Donald B. Sinclair. Special tables will be reserved for Professional Group members. Tickets are available at $5.75 each.

The **Annual Banquet,** to be held in the Grand Ballroom at 6:45 P.M. on Wednesday, will feature a major address by Charles E. Wilson, Director of Defense Mobilization. The 1952 IRE awards will be presented at this time. Tickets are on sale at $12.00 each.

An outstanding **Women's Program** of tours and shows has been arranged for the four days. Women's registration begins at 9:30 A.M. on Monday in the East Foyer of the Waldorf.

SCHEDULE OF TECHNICAL SESSIONS*

	Belmont-Plaza	Waldorf-Astoria			Grand Central Palace	
	Moderne Room	*Grand Ballroom*	*Astor Gallery*	*Jade Room*	*Maroon Room*	*Blue Room*
Mon. P.M. 2:30–5	*Symposium:* Subaudio Instrumentation (1–5)	*Symposium:* Management of Research and Development (6–9)	*Symposium:* Transistor Circuits (10–13)	Information Theory I—Coding Procedures (14–18)	Audio (19–23)	*Symposium:* New Developments in Telemetering (24–28)
Tues. A.M. 10–12:30	Instrumentation I—High-Frequency Instrumentation (29–33)	Television I—General A (34–37)	Circuits I (38–42)	Information Theory II—Noise Statistics and Signal Detection (43–47)	Microwaves I—Waveguides A (48–52)	*Symposium:* Television Broadcasting; Audio and Video Systems (53–57)
Tues. P.M. 2:30–5	Instrumentation II—Electronic Measurements A (58–62)	Television II—Color (63–67)	Circuits II and Information Theory III (68–74)	Medical Electronics (75–80)	Microwaves II—Waveguides B (81–85)	*Symposium:* Television Station Construction and Theater Conversion (86–89)
Tues. Eve. 8–10:30		*Special Symposium:* Present Status of NTSC Color Television Standards (90)				
Wed. A.M. 10–12:30	Instrumentation III—Electronic Measurements B (91–95)	Television III—General B (96–100)	Circuits III (101–106)	Propagation (107–111)	Microwaves III—Filters and Circuits (112–116)	*Symposium:* Digital Computers in Control Systems (117–121)
Wed. P.M. 2:30–5	Antennas I—General (122–126)	*Symposium:* UHF Receivers I (127–132)	Circuits IV (133–138)	Electron Tubes I—Power Output and Gas Tubes (139–143)	Radar and Radio Navigation (144–148)	*Symposium:* Magnetic Core Memory Devices for Digital Computers (149–153)
Thurs. A.M. 10–12:30	Antennas II—Microwave A (154–158)	*Symposium:* UHF Receivers II (159–162)	Feedback Control (163–167)	Electron Tubes II—Small High-Frequency Tubes (168–172)	*Symposium:* The Integration of Electronic Equipment with Airframe Design (173–176)	Digital Computers (177–181)
Thurs. P.M. 2:30–5	Antennas III—Microwave B (182–186)	Radio Communication Systems (187–191)	Circuits V (192–198)	Electron Tubes III—Cathode-Ray Tubes (199–203)	*Symposium:* What's New in Mobile Radio (204–207)	*Symposium:* Reliability of Military Electronic Equipment (208–211)

* Numbers in parenthesis following session titles refer to summaries of technical papers on the following pages.

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.

SYMPOSIUM

Subaudio Instrumentation

(Organized by Professional Group on Instrumentation)

Chairman, ERNST WEBER

(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

This symposium will include five papers by workers in the subaudio field and a subsequent round-table discussion. The papers describe instruments for measurements of physical motion and associated electrical signals. Methods of analysis and synthesis as applied to the subaudio frequency range will also be covered. The papers and discussion will stress both available instrumentation and the need for new methods and equipment.

1. DIRECT SYNTHESIS APPLIED TO SUBAUDIO FREQUENCY SYSTEMS

J. R. MOORE

(North American Aerophysics Laboratory, Downey, Calif.)

2. GENERATING EQUIPMENT FOR SUBAUDIO FREQUENCIES

E. H. GAMBLE

(Curtiss-Wright Corporation, Carlstadt, N. J.)

3. SUBAUDIO FREQUENCIES IN PETROLEUM EXPLORATION

W. M. RUST, JR.

(Humble Oil and Refining Company, Houston, Tex.)

4. OSCILLOGRAPHIC INSTRUMENTATION FOR THE SUBAUDIO FIELD

P. S. CHRISTALDI

(Allen B. DuMont Laboratories, Clifton, N. J.)

5. INSTRUMENTATION FOR HIGH-POWER HYDRAULIC SERVO DEVELOPMENT

HAROLD GOLD

(NACA, Lewis Propulsion Laboratory, Cleveland, Ohio)

SYMPOSIUM

Management of Research and Development

(Organized by Professional Group on Engineering Management)

Chairman, W. L. EVERITT

(University of Illinois, Urbana, Ill.)

A panel composed of representatives with broad backgrounds of experience and accomplishments in research and development activities will discuss their philosophies of the management of research and development activities. Following the presentation of the papers the audience will be invited to participate with questions and comments. Panel members presenting papers are:

6. W. R. G. BAKER (General Electric Company, Syracuse, N. Y.)
7. R. D. BENNETT (U. S. Navy Ordnance Laboratory, White Oak, Silver Springs, Md.)
8. G. N. THAYER (Bell Telephone Laboratories, Inc., Murray Hill, N. J.)
9. R. L. COLE and O. G. TALLMAN (Griffiss Air Force Base, Rome, N. Y.)

SYMPOSIUM

Transistor Circuits

(Organized by Professional Group on Circuit Theory)

Chairman, C. H. PAGE

(National Bureau of Standards, Washington, D. C.)

This symposium will present a survey of transistor circuits particularly designed to illustrate new aspects of circuit analysis resulting from the use of these elements.

10. TRANSISTOR OPERATION: ELEMENTS

(a). EQUIVALENT CIRCUITS

J. A. MORTON

(Bell Telephone Laboratories, Murray Hill, N. J.)

(b). PARAMETER MEASUREMENT

V. P. MATHIS

(General Electric Company, Syracuse, N. Y.)

(c). STABILIZATION OF OPERATING POINTS

R. F. SHEA

(General Electric Company, Syracuse, N. Y.)

11. TRANSISTOR BAND-PASS AMPLIFIERS

R. P. MOORE

(Radio Corporation of America, Camden, N. J.)

12. TRANSISTOR OSCILLATORS

J. S. SCHAFFNER

(General Electric Company, Syracuse, N. Y.)

13. TRANSISTOR PULSE CIRCUITS

J. H. FELKER

(Bell Telephone Laboratories, Whippany, N. J.)

Information Theory I—Coding Procedures

Chairman, M. J. E. GOLAY

(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

14. EFFICIENT CODING

B. M. OLIVER

(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Modern communication theory has shown that the only way to reduce the channel capacity (loosely, the bandwidth) required for electrical messages, is to translate them into a language (signal) in which the successive symbols (sample amplitudes) are more nearly independent of one another, and in which the different symbols (sample amplitudes) are used with nearly equal probability. In this way all the possible waveforms which the channel can handle (with a given peak power) become useful in that they now all represent possible messages.

This paper will present several possible methods of achieving such a translation, or coding, of signals.

15. TELEVISION SIGNAL STATISTICS

E. R. KRETZMER

(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Practical attempts at television bandwidth reduction require a knowledge of the picture signal statistics, such as signal amplitude distribution and autocorrelation. These two functions have been measured for a number of typical television pictures. Amplitude distributions have been determined also for error signals resulting from various types of linear prediction. Autocorrelation has been measured both along the scanning lines and at various angles with these lines.

Typical results of these measurements will be presented. They give us a fair idea of the bandwidth compression theoretically achievable (through reduction of channel capacity required) by means of linear decorrelation.

16. CODING WITH LINEAR SYSTEMS

J. P. Costas
(General Electric Company,
Syracuse, N. Y.)

Message transmission over a noisy channel is considered. Two linear networks are to be used; one at the sending end for treatment of the message before transmission and the second network at the receiving end of the channel for message recovery. For a given allowable average signal power and for a specified channel noise, a design method is developed which yields optimum coding and decoding networks. Special cases of the design equations are shown to yield the Optimum Filter Formula of Wiener and the Optimum Compensator Formula of Lee. The results of a numerical example are given and discussed.

17. PREDICTIVE CODING

Peter Elias
(Harvard University, Cambridge, Mass.)

Predictive coding is a method for coding message terms, in which the transmitter and the receiver store past message terms, and from them estimate the value of the next term. The transmitter codes the difference between the message term and its predicted value, and transmits this coded "error." The receiver decodes the error term and adds it to its prediction, reproducing the correct message term.

This coding procedure is compared with Shannon-Fano ideal coding methods. A criterion for the selection of predictors for predictive coding is given, and the use of Wiener's optimum linear predictor is discussed.

18. EXPERIMENTS WITH LINEAR PREDICTION IN TELEVISION

C. W. Harrison
(Bell Telephone Laboratories, Inc.,
Murray Hill, N. J.)

The correlation present in a signal makes possible the prediction of the future of the signal in terms of the past and present. Linear prediction, although it does *not* make full use of the past, is nevertheless remarkably effective with certain signals, and is also appealing because of its relative simplicity. Here the prediction for the next signal sample is simply the sum of previous signal samples each multiplied by an appropriate weighting factor.

This paper will describe the apparatus used for some experiments on linear prediction of television signals, and will discuss the results obtained to date.

Audio

Chairman, L. L. Beranek
(Massachusetts Institute of Technology,
Cambridge, Mass.)

19. MICROPHONES FOR THE MEASUREMENT OF SOUND PRESSURE LEVELS OF HIGH INTENSITY OVER WIDE FREQUENCY RANGES

J. K. Hilliard
(Altec Lansing Corporation,
Beverly Hills, Calif.)

This paper describes measurements and transmission of sound pressure levels of very high intensities in the subaudible, audible, and super audible ranges. In some cases the Altec standard 21B microphone is used, and for the extremely high intensities a modified type is used. These microphones will withstand high ambient temperatures. They may be used directly in the sound field or with probe tubes. They are used to measure blasts in excess of one atmosphere. Photographs of waveforms on blasts up to an intensity of 202 db, reference 0.0002 dynes per square centimeter, will be shown, at which point the open circuit output is approximately 100 volts. This output corresponds to a voltage ratio of 10^{10} times the minimum intensity which the human ear can detect.

20. AN INSTRUMENT FOR MEASURING THE TIME-DISPLACEMENT ERROR OF RECORDERS

E. N. Dingley, Jr.
(Armed Forces Security Agency,
Washington 25, D. C.)

There is described an instrument designed to provide a graphic record of the time variation of the time-displacement error of sonic recording-reproducing equipment. The utility of this instrument for measuring per cent flutter and flutter index is discussed.

21. A METHOD FOR MEASURING THE CHANGES INTRODUCED IN RECORDED TIME INTERVALS BY A RECORDER-REPRODUCER

J. F. Sweeney
(Armed Forces Security Agency,
Washington 25, D. C.)

Recorders are being increasingly used in instrumentation problems where the recorded information is in the form of a sequence of time intervals. In order to evaluate the suitability of various recording equipments for these applications, a technique has been developed for measuring the changes introduced in recorded time intervals by the recorder. By the use of this technique, a continuous permanent graph is produced of the variations in recorded time intervals introduced by the recording equipment. This graph may also be used to find the instantaneous frequency deviation, flutter repetition rates, and flutter wave shapes.

22. APPLICATION OF ELECTRIC CIRCUIT ANALOGIES TO LOUDSPEAKER PROBLEMS

B. N. Locanthi
(California Institute of Technology,
Pasadena, Calif.)

Electric circuit analogies which describe the three major types of loudspeaker systems currently in use today will be discussed. These are: (1) direct radiator in an infinite baffle; (2) direct radiator in a reflex enclosure, including the effect of the mutual impedance between the port and direct radiator on the radiating side of the enclosure; and (3) horn loudspeaker, treating the horn as a distributed parameter system.

Response curves obtained from the electric circuit analogies will be discussed showing the effects of amplifier source impedance, electromechanical coupling coefficient, and diaphragm suspension compliance on efficiency and bandwidth.

23. A SOUND-SURVEY METER

Arnold Peterson
(General Radio Company,
Cambridge, Mass.)

A simple, compact, low-cost noise meter has been developed as a companion to high performance instruments for use by audio and acoustical engineers. Its performance closely approximates that of standard sound-level meters, but it has simplified operating features. The amplifier and metering circuit are stabilized by negative feedback, and a continuous level control is provided. Its diminutive size, good characteristics, ease of servicing, and simplicity of operation make it ideal for many applications, some of which are economically feasible only with such a low-cost instrument. Applications include sound system measurements, noise surveys, product noise measurement, and experiments in acoustics.

SYMPOSIUM
New Developments in Telemetering

(Organized by Professional Group on Radio Telemetry and Remote Control)

Chairman, C. H. Hoeppner
(Raytheon Manufacturing Company,
Waltham, Mass.)

The Symposium will deal with the problem of recording FM-FM telemetering signals on magnetic tape, and ways and means of compensating for and avoiding the errors introduced by even small values of flutter and drift which give trouble because of the 7.5 per cent deviation used. Other new developments will be discussed.

24. NEW DEVELOPMENTS IN TELEMETERING

C. H. Hoeppner
(Raytheon Manufacturing Company,
Waltham, Mass.)

The problems of storing and reproducing telemetered data are defined. The need for advances in recorders and error compensating systems is discussed and some new developments in this field are noted. Two general methods presently in use for processing telemetering data are outlined and the sources of error in each are discussed. The advantages of sampling and coding at various points in the system are considered.

25. RECENT ADVANCES IN MAGNETIC RECORDING FOR TELEMETERING APPLICATIONS

W. T. Selsted
(Ampex Electric Corporation,
Redwood City, Calif.)

The problem of recording FM-FM telemetering is stated. The development of a four-channel magnetic tape telemetering recorder with a peak-to-peak flutter of less than 0.1 per cent is described. Some other examples of special applications of magnetic tape recorders to recording telemetered data will also be described.

26. FAIRCHILD MODEL 150 TELEMETERING DATA RECORDER

C. F. Kezer
(Fairchild Recorder Equipment Company, Whitestone, N. Y.)

The design considerations of a new rack-mounted four-channel telemetering tape recorder suitable for FM-FM signals are discussed. Problems for which solutions are presented include: flutter of less than 0.1 per cent peak to peak; a speed control system which compensates for tape dimensional changes and other long-term speed variations from a signal recorded on the tape; multi-channel operation; 80 kc frequency response; and large tape reels. Methods of measuring flutter will also be described.

27. RECORDING TELEMETERING DATA

M. V. Kiebert
(Bendix Research Laboratories, Detroit, Mich.)

The telemetering data recording system is analyzed in view of various instrumentation requirements. Four basic reasons for recording telemetering signals are defined: (1) data recording at locations not ordinarily requiring concurrent information presentations, (2) insurance against loss of information by concurrent presentation systems as a result of equipment failure or malfunction, (3) condensed data storage of information on high density media, and (4) concurrent transducer media for data reduction. Calibration curve compensator, and scale factor correction operating system are outlined and described. Typical records are shown and compared with concurrent presentation.

28. TELEMETERING BY PULSE-CODE MODULATION

B. D. Smith
(Melpar Company, Alexandria, Va.)

The possibility of using pulse-code modulation in a radio telemetering system is considered. The basic methods of binary coding are considered and the methods which are most suited to an airborne telemeter are discussed. Quantization errors and methods of nonlinear quantization with the feedback type of coding are described. The advantages and disadvantages of pulse-code modulation over FM or other modulation systems are discussed with respect to a telemetering system in which automatic data reduction is contemplated. Methods of data recording and processing by means of ground equipment are discussed.

Instrumentation I—High-Frequency Instrumentation

Chairman, I. G. Wolff
(Radio Corporation of America, Princeton, N. J.)

29. VHF Q-MEASUREMENT TECHNIQUES

D. M. Hill
(Boonton Radio Corporation, Boonton, N. J.)

A Q-Meter covering the frequency range 20 to 260 mc will be described. An improved voltmeter circuit permits measurement of both low-Q and high-Q components and also enables small changes in Q to be measured more accurately. A new variable air condenser has been designed to reduce errors due to inductance and series resistance in the standard condenser.

Measurement techniques and errors due to circuit residuals and frequency effects will be discussed.

30. A HIGH SENSITIVITY METHOD FOR MEASURING CONDUCTANCE AND CAPACITANCE AT RADIO FREQUENCIES

W. C. Freeman, Jr.
(Boonton Radio Corporation, Boonton, N. J.)

A new and more sensitive instrument for measuring conductance and capacitance at radio frequencies has been designed. Operating on the resonant circuit substitution principle, this Conductance Meter employs a calibrated dc variable resistance as the load of a diode rectifier connected across the resonant circuit. This provides an accurate substitution method for r.f. conductance with a direct-reading calibration which is independent of frequency. The capacitance substitution is provided by a precision variable capacitor designed to have low change of conductance with dial setting.

The sensitivity and stability requirements for the measurement of such high quality dielectrics as Polystyrene and Teflon are analyzed and the methods used to meet these requirements are described.

31. A MEAN-SQUARE VACUUM-TUBE VOLTMETER

L. A. Rosenthal and G. M. Badoyannis
(Rutgers University, New Brunswick, N. J.)

By means of a nonlinear resistor network it is possible to obtain an instantaneous "squarer," or the output current of the network is proportional to the instantaneous square of the input voltage. A dc ammeter in the output circuit indicates the average current which is directly proportional to the mean-square voltage. Upon combining the "squarer" with a preamplifier, a mean-square voltmeter results. There are no zero-setting or balancing controls and the output scale is linear in power and square-law in voltage. The squaring action is accurate to ± 2.5 per cent for a current range of 50 to 1, and the upper frequency limit is approximately 500 kc. As a voltmeter the circuit is useful in the study of complex waveforms, and as a squarer it can find interesting applications such as frequency doublers and simple computers.

The theory of operation, principles of design, performance data, and applications are discussed.

32. A NEW TECHNIQUE FOR THE EVALUATION OF LEAKAGE AND RADIATION FROM SIGNAL GENERATORS

W. A. Stirrat
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Heretofore, the procedure for evaluating tolerable leakage and radiation from signal generators has been of insufficient rigor and has led to a very arbitrary establishment of allowable limits. An investigation recently concluded for the Signal Corps resulted in the establishment of those limits compatible with the conduction of tests on Army receivers.

However, the limits were given in terms of pick-up voltages which are of an extremely low order over certain frequencies of the spectrum and cannot readily be measured with any degree of accuracy.

Techniques have been evolved at SCEL which circumvent the difficulty of such measurement.

33. A WIDE-BAND SWEEP GENERATOR

F. P. Blecher
(Polytechnic Research and Development Company, Inc., Brooklyn, N. Y.)

A wideband sweep generator is described which features a fundamental frequency coverage of 40 to 950 megacycles, one volt output, a minimum of ten megacycles sweep width, and provision for introducing an external marker. The instrument is completely shielded and employs a modified type of waveguide operating beyond cutoff attenuator which provides output voltages between one volt and ten microvolts. Over most of the frequency range, the sweep width is adjustable between zero and twenty megacycles with practically no change in the center frequency. The instrument may also be employed as a C.W. oscillator with low harmonic distortion over the entire frequency range.

Design specifications are discussed and the new techniques responsible for the special characteristics of the instrument are explained.

Television I—General A

Chairman, A. V. Loughren
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

34. GAMMA CORRECTION IN COLOR TELEVISION

S. Applebaum
(General Electric Company, Syracuse, N. Y.)

Gamma correction in color television is desirable to correct the tonal and chromaticity distortion which results from the use of nonlinear display tubes. Correction should be made at the transmitter in the primary colors of the receiver. A nonlinear color receiver causes some noise cross talk from the chromaticity channels to the luminance channel. Despite this, the over-all noise sensitivity is less than that of a linear color receiver. The effect of precorrection upon the nonlinear compatible monochrome picture is to linearize the tonal rendition but attenuate somewhat the display of saturated colors.

35. THE SPECIFICATION AND CORRECTION FOR NONLINEARITY OF CATHODE-RAY TUBES

R. C. Moore
(Philco Corporation, Philadelphia, Pa.)

Correction for nonlinearity of cathode-ray tubes is particularly important in color television. The problems of interpreting data on cathode-ray tube characteristics in order to specify a suitable corrector are considered. It is shown that exponent value of assumed power law is not critical.

The requirements on corrector circuits to satisfy a desired specification are considered. A particular circuit used in a laboratory color-signal source is described and its operating characteristics discussed. It is characterized by good performance and stability.

36. COLORIMETRIC MEASUREMENTS IN COLOR TELEVISION SYSTEM

S. W. Moulton
(Philco Corporation, Philadelphia, Pa.)

A subjective type colorimeter is described that fulfills an increasingly important need for a method to measure quantitatively and simply the output of color television receivers. The spectral similarities of its two fields increases its accuracy and facilitates measurements. The question of the observer's spectral response is eliminated.

37. FRAME SYNCHRONIZATION FOR COLOR TELEVISION

Donald Richman
(Hazeltine Corporation, Little Neck, N. Y.)

The proposed NTSC standards for color television utilize the technique of color phase alternation wherein the phase sequence of color in the color subcarrier is reversed on alternate fields. An arrangement for automatically synchronizing the direction of reversing at the receiver and at the transmitter is desirable to its operation. It will be shown that the present monochrome FCC synchronizing waveform contains the necessary signal information for reliable frame synchronizing. This paper describes practical circuits for frame synchronization, and discusses theoretical considerations relating to the problem of frame synchronization.

Circuits I

Chairman, H. J. Carlin
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

38. NETWORK ALIGNMENT TECHNIQUE

J. G. Linvill
(Massachusetts Institute of Technology, Cambridge, Mass.)

An experimental method by which variable components in a network can be adjusted to bring the frequency or time response of the network into correspondence with the response of a standard network will be described. The technique involves, first, consideration of the difference of the response of the network being adjusted from that desired, and second, consideration of the changes in response due to motions of the elements of a perfect network taken one at a time. The steps are so simple experimentally that not even a slide rule is required for the whole process. Experimental data and oscilloscope pictures will be presented.

39. NETWORK ANALYSIS BY A NEW SEMI-AUTOMATIC COMPUTER

R. L. Bright and G. H. Royer
(Carnegie Institute of Technology, Pittsburgh, Pa.)

A new analog computer which yields directly the complex natural frequencies of any linear system and the locus of these natural frequencies as system parameters are varied is described. General theory details of a unit for linear systems up to the sixth order are discussed. The calculator also provides a very simple method of indicating the locus of the complex natural frequencies as any of the system parameters are varied. Using currents fed from a computer, natural frequency of the system is indicated by a servo-controlled probe whose co-ordinates are the values of the real and imaginary parts of a complex root of the system.

40. NETWORK ANALYSIS BY TWO NEW COMPUTERS

D. Herr
(Hughes Aircraft Company, Culver City, Calif.)

Two new economical analogue computers to be used as *desk-side tools* in the design and analysis of servomechanisms, networks, amplifiers, and analogous systems are described. In a certain sense the two computers are complementary in their use and with the use of a differential analyzer.

The first computer is a modern version of an algebraic-equation solver earlier described. The second computer is a mechanization of the "locus-of-roots" method of analysis and design previously explained by W. R. Evans. Each uses two new computer components: one a high-precision, universal-frequency ac induction resolver, and the other an ac induction potentiometer.

41. NETWORK RESPONSE CHARACTERISTICS USING THE COMPLEX PLANE SCANNER

J. R. Ragazzini and G. Reynolds
(Columbia University, New York, N. Y.)

The complex plane scanner is an electronic analogue computer which electrically scans the complex plane by equivalently running a point along the imaginary axis. By multiplying voltages equivalent to the magnitudes of the distances, from the poles and zeros to the running point, the magnitude of the transmission is obtained. Multiplications and divisions are carried out with logarithmic networks, and the magnitude of the transmission in db is the magnitude of an ac voltage. If a perturbation is added, the phase slope of the transmission characteristic is obtained. The display of the complex plane scanner is by oscilloscope. Thickness of the curve is proportional to the phase slope.

42. RESONANCE CHARACTERISTICS BY CONFORMAL MAPPING

P. M. Honnell and R. E. Horn
(Washington University, St. Louis, Mo.)

The analytical expression $f(\lambda)=a\lambda+b+c/\lambda=\sigma+j\omega$, may be called the "resonance function" since it represents the impedance of the series-connected LRS circuit and the admittance of the parallel-connected CGP circuit, among others. The conformal mapping of the λ plane onto the $f(\lambda)$ plane yields a figure of considerable usefulness in clarifying the meaning of the generalized impedance of the series circuit and admittance of the parallel circuit for complex frequencies. This will be illustrated by typical examples, together with other applications of the figure to more general circuit configurations.

A brief discussion of the mapping of the reciprocal of the resonance function will be included for completeness.

Information Theory II—Noise Statistics and Signal Detection

Chairman, A. G. Clavier
(Federal Telecommunications Laboratories, Inc., Nutley, N. J.)

43. DISCUSSION OF A METHOD OF EXPANDING NOISE AUTOCORRELATION FUNCTION IN A POWER SERIES

F. W. Lehan
(California Institute of Technology, Pasadena, Calif.)

It is shown that under certain conditions the autocorrelation function of a random noise source may be expanded in the following power series:

$$\rho(\tau)=1-\frac{n^2(\pi\tau)^2}{2!}+\frac{(nn')^2(\pi\tau)^4}{4!}-\frac{(nn'n'')^2(\pi\tau)^6}{6!}+\cdots$$

where

n = the average rate of occurrence of zeros in the noise function
n' = the average rate of zeros in the first derivative, etc.

The limitations and usefulness of this expansion are discussed.

44. A PROPOSAL FOR THE DETERMINATION OF COHERENCE IN A SIGNAL FIELD

B. S. Melton
(Headquarters, U. S. Air Force, Washington, D. C.)
and
P. R. Karr
(National Bureau of Standards, Washington, D. C.)

A simple statistical procedure, based on multiple coincidence of field polarities as indicated by several detectors, offers a criterion of the existence of a signal below the general noise level, the effectiveness of the procedure depending upon the signal duration. As an example, with six detectors in a field where there is a sine wave signal whose power is one-twelfth the noise power, there will be 80 per cent more polarity agreements than for the noise alone. A scheme is shown to illustrate how an electrical system can perform the analysis and present the results continuously.

45. THE RESPONSE OF LINEAR SYSTEMS TO RANDOM NOISE

B. Gold
(Hughes Aircraft Company, Culver City, Calif.)
AND
J. P. Ruina
(Brown University, Providence, R. I.)

When random noise is suddenly applied to a linear system with N outputs, a complete description is obtained by means of an N dimensional probability distribution function. This function is in general time dependent and may or may not reach a stationary value.

In this paper formulas are derived, with the aid of circuit theory concepts, with which one can obtain all the (co)variances of the desired distribution as functions of time. In the steady state ($t\rightarrow\infty$) the results obtained become the well-known results for the stationary output of a linear system with input noise.

The method is applied to several problems.

46. CORRELATOR FOR LOW FREQUENCIES

V. J. Guethlen
(Goodyear Aircraft Corporation, Akron, Ohio)

A number of electronic correlators capable of performing the necessary mathematic operations to obtain correlation functions have been reported on in the past. In general, existing electronic correlators are both expensive and not suited to the processing of data containing periodicities of low frequency. This paper describes a correlator that can be readily assembled from commercially available subassemblies. It is capable of computing the auto- or cross-correlation functions of data having low frequency components. The storage of data and provisions for the introduction of a variable delay required in computation is instrumented by a magnetic tape recorder. This work was sponsored by an Air Force contract to study radar terrain-return signals.

47. OPTIMUM TECHNIQUES FOR DETECTING PULSE SIGNALS IN NOISE

D. L. Drukey
(Hughes Aircraft Company, Culver City, Calif.)

The problem treated is that of processing N samples of signal plus noise or noise alone so as to obtain the maximum probability of correctly identifying a pulse-type signal as present, consistent with a given probability of calling noise a signal. The signal and noise are treated statistically, and the effect of signal amplitude fluctuations is taken into account. In addition to specifying optimum processing methods, the false alarm probabilities and probabilities of detecting a signal have been calculated for these optimum systems and for several different signal characteristics.

Microwaves I—Waveguides A

Chairman, G. C. Southworth
(Bell Telephone Laboratories, Inc., Holmdel, N. J.)

48. MICROWAVE WIRING

D. D. Grieg and H. Engelmann
(Federal Telecommunication Laboratories, Nutley, N. J.)

A novel approach to microwave transmission results from replacing the familiar waveguide or coaxial line with a signal wire or strip supported above a ground plane.

This structure is approximately equivalent to a parallel-wire system, but by virtue of an "image" of the wire in the ground plane, adequate symmetry is achieved without dimensional criticalness. The spread of the field about the conductor is small and the losses approximate those that characterize a coaxial structure.

Printed-circuit techniques are particularly applicable to these microwave transmission systems and make possible the construction of compact, rugged, and relatively inexpensive microwave components.

49. SIMPLIFIED THEORY OF TEM PROPAGATION ALONG CONDUCTOR-GROUND-PLANE TRANSMISSION SYSTEMS

F. Assadourian and E. Rimai
(Federal Telecommunication Laboratories, Nutley, N. J.

Characteristic impedance, power distribution, and transmission losses for transverse electromagnetic (TEM) propagation in infinite transmission systems comprising a conductor above ground were treated by an electrostatic approach using conformal mapping in the image plane.

The cases of wire and zero-thickness strips of varying height-to-width ratios above infinite ground are discussed. An assumed small thickness permits calculation of strip losses without greatly disturbing the fields except at the strip edges. The effect of finite ground width is also treated. Finally, the effect of a slot in the ground plane is examined for both wire and strip above ground.

50. MICROWAVE COMPONENTS FOR CONDUCTOR-GROUND-PLANE TRANSMISSION SYSTEMS

J. A. Kostriza
(Federal Telecommunication Laboratories, Nutley, N. J.)

A new type of microwave transmission line has been developed that is adaptable to fabrication of wide-band microwave components of reasonable cost. It employs an open waveguide system consisting of two parallel conductors, one acting as a line and the other as a ground conductor.

Detailed descriptions are given of experimental techniques and results in the construction of components for operation in the 5,000-megacycle region. Among these components are transducers to coaxial lines, loads, and pads; crystal mounts; "rat races"; directional couplers; and shunt T-junctions. Data are included for wire-above-ground, "printed," and "sandwich" lines.

51. METHOD FOR OPEN WAVEGUIDE STANDING-WAVE MEASUREMENTS

S. W. Attwood and G. Goubau
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

In order to measure standing-wave ratios at microwave frequencies on open waveguides, a new indirect method had to be employed because there were no satisfactory standing-wave detectors available for these types of guides. The method employs a thin dielectric disk which intercepts the field and is moved along the guide. The disk causes a change of input impedance which depends on the location of the disk. The ratio of maximum to minimum impedance change is proportional to the square of the standing-wave ratio. Two methods for determining this ratio will be discussed.

52. NEW GUIDED WAVE TECHNIQUES FOR THE MILLIMETER WAVELENGTH RANGE

A. G. Fox
(Bell Telephone Laboratories, Inc., Holmdel, N. J.)

The results of preliminary experiments with new guided wave techniques at 48 kmc make it appear that such will be very useful in the millimeter wavelength range. These techniques provide the only practical means yet known for obtaining flexible transmission links in this range. They afford an easy way of obtaining directional couplers having practically any desired value of coupling loss.

SYMPOSIUM

Television Broadcasting; Audio and Video Systems

(Organized by Professional Group on Broadcast Transmission Systems)

Moderator, W. B. Lodge
(Columbia Broadcasting System, New York, N. Y.)

53. FIXED AND MOBILE TV LIGHTING

Emero Fiorentino
(WJZ-TV, New York, N. Y.)

Television lighting of today has emerged from the laboratory stage where it was merely a physical device used to activate the electro-sensitive systems. It is now an excellent example of the marriage between the artistic and highly technical elements of a production, a marriage so necessary in the television industry.

Because of the rapid advancement that television lighting has been going through, many of its old principles are still adhered to religiously by many people. These people undoubtedly neither have had the time nor the money nor, most important of all, information concerning new developments which has not been made available and meaningful to them. By the same token, those who are employing better lighting techniques have had their hands full with that job and consequently there has been little time for elaborate, large-scale discussion of the lighting problems and techniques of today.

Through this paper, there will be provided information concerning these lighting techniques; the adaptation of motion picture lighting and the many facets peculiar to lighting in TV.

54. THE TRANSIENT RESPONSE OF TV TRANSMITTER-RECEIVER SYSTEMS

J. RUSTON
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

The results of an experimental investigation of the transient response of TV transmitters are presented and substantiated as far as practicable by partial theoretical analyses. In a vestigial sideband system, the transmitter response must be considered in relation to the over-all response of the transmitter and receiver. The effect of the transmitter characteristics on the system response is considered for various types of receiver characteristics. Methods for correcting possible deficiencies in the transmitter are suggested. The possibility of including correction in the transmitter for common deficiencies in the receiver is discussed, and doubt is expressed regarding its desirability. A method is suggested whereby a change of transmitter and receiver standards would largely eliminate the transient distortion inherent with the present standards.

55. MEASUREMENT OF TV FIELD INTENSITIES BY HELICOPTER

J. G. PRESTON
(American Broadcasting Company, New York, N. Y.)

Recent use of a helicopter to measure the horizontal field intensity pattern radiated by a typical vhf TV broadcast antenna installation is reported. Field intensity recordings of unusual quality were obtained through measurement on a circular flight course at an appropriate radius from the antenna. The measurements disclosed pattern noncircularity in excess of the tolerance permitted by the antenna specifications. Through the use of appropriate assumptions, it is found possible to verify mathematically the measured pattern and to determine antenna phasing changes necessary to bring the pattern within the original specifications. Subsequent helicopter measurements of the rephased antenna confirmed performance within specifications.

56. HIGH POWER UHF KLYSTRON FOR TV SERVICE

J. J. WOERNER
(Eitel McCullough, San Bruno, Calif.)

Within the past few years it has become recognized that the cascade klystron amplifier will satisfy the specific requirements of television transmission in the ultrahigh frequency range.

The subject paper will discuss the factors involved in adapting this type of tube to the more general requirements of commercial television broadcast service. A developmental 5-kilowatt klystron designed particularly for this use will be described in detail. Features of this tube are externally tuned cavities to extend the operating range of frequency, and magnetic focussing of a type to allow greater access to the tuning and loading arrangements.

Slides of photographs and drawings of the tube and circuit details will be shown.

57. AN ULTRA-HIGH-FREQUENCY TELEVISION TRANSMITTER

E. G. MCCALL AND T. P. TISSOT
(Radio Corporation of America, Camden, N. J.)

This paper describes a TV transmitter including a 1-kw visual transmitter and a 0.5-kw aural transmitter for operation in the uhf TV band, channels 14 to 83.

A description is given of the power output stage of both visual and aural transmitter which utilizes a single high-gain air-cooled tetrode in a coaxial cavity circuit. The method of video modulation is discussed along with the phase shift FM direct crystal-controlled modulator used in the aural transmitter.

A discussion of the frequency stability requirements at uhf is given to summarize the conditions necessary for satisfactory operation of inter-carrier-sound type receivers.

Instrumentation II—Electronic Measurements A

Chairman, E. P. FELCH
(Bell Telephone Laboratories, Inc., New York, N. Y.)

58. MEASUREMENT OF IMPEDANCE AND ADMITTANCE

B. SALZBERG AND J. W. MARINI
(Naval Research Laboratory, Washington 25, D. C.)

A method is described for direct, rapid, and simultaneous measurement of the real and imaginary components of self or transfer impedance or admittance. The method employs a mixer whose signal and oscillator voltages are derived from the voltage across and current through the unknown, respectively. When the unknown is supplied by a sinusoidal constant current source, the mixer dc output is linearly related to the resistance. If the oscillator voltage phase is shifted 90° the mixer dc output is linearly related to the reactance. Apparatus using this method and operating from 50 kc to 5 mc has been built. With appropriate modifications of detail the method appears to be applicable even to microwaves.

59. ACCURATE RF MICROVOLTS

M. C. SELBY
(National Bureau of Standards, Washington, D. C.)

The questionable accuracy of rf microvolts has been of great concern to the radio fields for many years. There is an urgent need for a simple, yet reliable, source of microvolts for measurements in generaland for radio receiver sensitivity determinations in particular. Extremely simple devices, which seem to satisfy that need most adequately, were recently developed. These devices provide constant voltage sources of accurate microvolts for a range of 1 to 10^5 and wider at all frequencies to 300 mc and higher. They are adaptable for balanced and unbalanced sources. Their electrical constants are simply determined by using known dc voltages and currents. Basic principles, design features and applications are discussed.

60. AUTOMATIC SWITCHING APPLIED TO INTERELECTRODE CAPACITANCE MEASUREMENTS

R. E. GRAHAM
(Sylvania Electric Products, Inc., Kew Gardens, N. Y.)

A new equipment is described which greatly reduces the time and effort required to measure electron-tube interelectrode capacitances. The switching required for a series of measurements is performed by shielded relays which are actuated by punched cards and connected so that the resultant stray capacitance across the measurement terminals is held to a negligible amount. Design principles, description of circuits, and physical layout are given.

61. MEASUREMENTS OF MILLIMETER RADIATION WITH THE PNEUMATIC HEAT DETECTOR

HANS THEISSING, H. J. MERRILL, AND J. M. MCCUE
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Theoretical considerations indicate that a very thin layer of certain metals absorbs electromagnetic radiation throughout a wide range of the spectrum independently of wavelength. The absorption of such layers used as the absorbing area of a Golay pneumatic heat detector is evaluated to indicate the "grayness" of the detector. The calibrated detector is used in an optical type spectroscope to evaluate the average power emitted at the fundamental and higher harmonics of a 1.25-cm magnetron source as a function of the frequency. The signal equivalent to noise on the 12 mm diameter detector is of the order of 10^{-8} watts.

62. AUTOMATIC SMITH-CHART IMPEDANCE PLOTTER

K. S. PACKARD
(Airborne Instruments Laboratory, Inc., Mineola, N. Y.)

This system plots the impedance of a component by measuring its reflection coefficient as a termination on a 50-ohm line. A directional coupler is used to sample the direct and reflected waves on this line. By means of an electronic circuit $K \cos \theta$ and $K \sin \theta$, where K is the amplitude and θ the phase angle of the reflection coefficient, are obtained and applied to the deflection plates of a cathode-ray tube. Thus, K is plotted on a polar diagram and a Smith Chart overlay gives the impedance of the component. Two continuously sweeping oscillators, one for the 100 to 200 mc band, or one for the 200 to 400 mc band, are used so impedance is plotted over these frequency ranges.

Television II—Color

Chairman, E. W. ENGSTROM
(Radio Corporation of America, Princeton, N. J.)

63. REQUISITE COLOR BANDWIDTH FOR SIMULTANEOUS COLOR-TELEVISION SYSTEMS

Knox McIlwain
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

It has been known for many years that the acuity of the human eye for color differences is less than for changes in brightness. It has been shown that this fact can be used to reduce the bandwidth required for simultaneous color television systems. The experiments reported here relate to psychophysical measurements made by both skilled and lay observers to determine just how far this reduction can be carried without objectionable deterioration of the reproduction. It is shown that under the conditions tested approximately one megacycle is sufficient for most color reproduction, provided four megacycles is available for brightness detail.

64. COLORIMETRIC ELECTRONICS

F. J. Bingley
(Philco Corporation, Philadelphia, Pa.)

The paper considers the relations between electrical circuits and colorimetric quantities in a color-television system. These relations are described with particular reference to a compatible color-television system of the type employing a modulating signal made up of a main monochrome component plus a color carrier. The various types of monochrome components and their colorimetric makeup are discussed. The characteristics of the color carrier and its representation by various vector systems is discussed in detail. The interrelations between these vector systems and the colorimetric quantities which they represent are demonstrated. Graphical methods are used to illustrate the analytical processes used.

65. THE GENERATION OF COMPATIBLE COLOR SIGNALS FOR RESEARCH AND FIELD TESTING

Joseph Fisher
(Philco Corporation, Philadelphia, Pa.)

This paper will describe the equipment currently being used in the research laboratories of the Philco Corporation for the generation of compatible color signals. The centralized equipment consists of a number of units such as a crystal controlled timing unit, color flying-spot scanner, burst pedestal generator and combiner. Distribution amplifiers are incorporated to feed the composite color signal to various laboratories, and microwave equipment is provided to relay the color signal to television station WPTZ for experimental broadcasts. The routing of the signal will be illustrated with block diagrams and some of the circuits used will be discussed.

66. A UNIVERSAL SCANNER FOR COLOR TELEVISION

G. R. Tingley, R. D. Thompson, and J. H. Haines
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

This paper describes a flying-spot scanner for 35 mm transparencies providing three simultaneous gamma-corrected signals for any proposed color-television system. The optical system employs a grey-faced cathode-ray tube operating at 30 kv, high definition scanning yoke, focus modulation, $f/1.5$ objective lens, high efficiency dichroic mirror beamsplit, and three phototubes. Pictures produced have high detail contrast and signal-to-noise ratio. Horizontal resolution approximates 600 lines for compatible color and 500 lines for FCC color. Rates corresponding to either of these or to intermediate values may be switch selected. A field-sequential composite color signal at the selected rate is also produced by a keyer.

67. VESTIGIAL SIDEBAND TRANSMISSION OF THE COLOR SUBCARRIER IN NTSC COLOR TELEVISION

W. F. Bailey
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

In NTSC color television, the picture is encoded as a luminance signal and two additional signals representative of the color.

A color subcarrier is resolved into two synchronous components differing in phase by 90°, and each component is amplitude modulated by a color signal. For compatibility, the color subcarrier is near the high edge of the video band. Vestigial sideband transmission results in undesired color cross talk because of the quadrature distortion produced.

Reversal of the sequence of the color subcarrier components (color phase alternation) at field rate reverses the quadrature distortion on adjacent lines in the picture, resulting in essentially complete visual cancellation of the distortion.

Circuits II and Information Theory III

Chairman, W. R. Bennett
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

68. NETWORKS FOR DETERMINATION OF POWER SPECTRA MOMENTS

S. H. Chang and W. H. Lob
(Northeastern University, Boston, Mass.)

In speech analysis studies, it is often desirable to compute the first few moments of short-time power spectra. The advantages and limits of the moment method will be explored first. Then three possible schemes of automatic computation by analogue means will be discussed. One of the schemes, employing a square-law device and filters of various rising characteristics, i.e., 10, 20, 30, etc., db/decade has been tested and will be described in more detail. Possible applications for other time series, and for the computation of the moments of distributions other than power spectra, will also be mentioned.

69. NONLINEAR FILTER DESIGN ON MAXIMUM LIKELIHOOD BASIS

T. G. Slattery
(Melpar, Inc., Alexandria, Va.)

This paper presents the theory of designing nonlinear filters using the statistical technique of the "maximum likelihood estimate." The limitations of linear filters in separating signal from noise are reviewed and the advantages of nonlinear techniques are stated. The fact that filtering can be accomplished by estimating constants is stressed, and the advantages of this method are stated. The theory of the maximum likelihood estimate is then developed and its relationship to the Wiener type of best linear filter and the Singleton type of best nonlinear filter is described.

The example of the method is that of separating a sine wave of unknown amplitude and frequency from Gaussian noise.

70. OPTIMUM LINEAR SHAPING AND FILTERING NETWORKS

R. S. Berkowitz
(University of Pennsylvania, Philadelphia, Pa.)

Given a linear communication system consisting of a transmitter, channel, and receiver, we consider the problem of jointly optimizing the transmitter and receiver transfer characteristics. Signal power spectrum, noise, input power spectrum, power output of transmitter, and distortion of signal at transmitter are considered as being specified. The general criterion of optimization used is that the equivocation due to the noise be at a minimum. Results obtained consist of expressions for optimum transmitter and receiver characteristics in terms of the specified quantities when equivocation is monotonically related to either of two simple measures of the noise component of the receiver output.

71. A GENERALIZED THEORY OF FILTERING AND MULTIPLEXING

L. A. Zadeh
(Columbia University, New York, N. Y.)
and
K. S. Miller
(New York University, New York, N. Y.)

A theory of filtering and multiplexing based on the class properties of signals is formulated. An ideal filter is defined as one which can extract a signal belonging to a specified class from the sum of two or more signals belonging to some other specified classes. A study is made of the conditions under which two or more multiplexed signals can be separated by means of linear (variable) and nonlinear filters. These conditions are formulated in geometrical terms via the function space representation of signals and, also, in analytical terms by using the concept of λ domain.

72. FILTER TRANSFER FUNCTION SYNTHESIS

G. L. Matthaei
(University of California, Berkeley, Calif.)

Techniques based on the electrostatic potential analogy provide a means for synthesizing filter transfer functions to meet a variety of restrictions which may be imposed by circuit considerations. To illustrate, an example is presented of the synthesis of a filter function to be realized in an LC network, and then synthesis technique for obtaining an analogous function which meets the requirements for realization in an RC network is shown. A comparison is readily made between "efficiencies" of LC and RC realizable filter functions.

73. FILTERS OF MAXIMUM BANDWIDTH-IMPEDANCE RATIO PRODUCT

T. J. O'Donnell
(Gulf Research and Development Company, Pittsburgh, Pa.)
AND
E. M. Williams
(Carnegie Institute of Technology, Pittsburgh, Pa.)

This paper is concerned with simplified methods for solving the problem of wide-band impedance transforming network synthesis in which the product of the bandwidth and impedance ratio is maximized. Conditions are developed for maximum impedance bandwidth transformations in cascades of similar transformable band-pass filter sections for two cases of special interest. One case, when applied to a standard "three-element" section is shown to lead to a network which is the lumped circuit analog of the experimental transmission line. Some experimental illustrations are given, in particular, a four-to-one ratio impedance transformer fleet from 54 to 88 mc is described.

74. A BAND-PASS FILTER USING SIMULATED TRANSMISSION-LINE ELEMENTS

A. D. Frost and C. R. Mingins
(Tufts College, Medford, Mass.)

The use of mutually coupled coaxial transmission lines as a band-pass filter operating at frequencies of 1,000 mc and above has been described by Karakash and Mode. By utilizing various forms of synthetic transmission-line structures, the principles involved have been successfully applied to the design of filters operating in the frequency range from 2 to 8 mc. A number of these are shown together with their transmission characteristics. An important feature of this approach to the filter problem is the electrical and mechanical simplicity of the units involved.

Medical Electronics

Chairman, Britton Chance
(University of Pennsylvania, Philadelphia, Pa.)

75. NEW ELECTRONIC TECHNIQUES FOR SPECTROPHOTOMETRY

C. C. Yang
(University of Pennsylvania, Philadelphia, Pa.)

In the course of designing a simple and compact recording spectrophotometer in the visible and ultraviolet region, new circuits have been developed for direct reading of the optical density of a solution with respect to a solvent sensitivity, accurately, and rapidly. A time sharing scheme using optical switching and a feedback circuit for ratio measurement are used to obtain the per cent absorption of the sample solution. Two analogue computing circuits are described to convert the absorption into optical density electronically. Particular emphasis is placed on the measurement of small optical density changes, and a reproductibility in optical density of about 2×10^{-4} is obtained.

76. APPLICATION OF MICROWAVES IN PHYSICAL MEDICINE

J. F. Herrick
(Mayo Foundation, Rochester, Minn.)

A study of the effects of microwaves on certain living tissues of experimental animals prior to the acceptance of microwaves for medical diathermy will be presented.

The measurement of dielectric constant and dissipation factor of various freshly excised tissues were necessary for understanding the experimentally observed temperature distribution produced in these tissues by microwave diathermy. Results of measurements at three frequencies, i.e., 1,000, 3,000 and 8,600 megacycles will be given.

The design of a "transformer" for increasing the transfer of microwave power into tissues will be described as an example of the utility of dielectric data.

77. DESIGN PROBLEMS IN THE ABSOLUTE OXIMETER

R. H. Taplin
(Canadian Marconi Company, Montreal, Canada)

The percentage measurement in vivo of O_2 in the hemoglobin of humans is carried out photometrically but precise results have been difficult to obtain. The work described here results from an investigation into the causes of error mainly from an instrument design point of view. It is confined to the electronic type of oximeter of Goldie and later developed by Paul which gives the greatest flexibility. Large photometric errors are overcome by the new earpiece described, and the required design parameters of stability in earpieces, electronic amplifiers, discriminators, and light modulators are discussed. It seems possible to achieve a long term instrumental drift better than ±3 per cent and a short term of ±1 per cent for several hours.

78. TELEVISION MICROSCOPY IN THE ULTRAVIOLET

V. K. Zworykin, L. E. Flory, and R. E. Shrader
(Radio Corporation of America, Princeton, N. J.)

The differential absorption of tissues and micro-organisms in the ultraviolet is an important aid in medical research and diagnosis. The television microscope, provided with reflective optics and an ultraviolet sensitive vidicon, translates ultraviolet intensities directly into brightness differences and circumvents the time lag inherent in photographic methods. With the aid of a pulsed ultraviolet source, a suitable wavelength selecting system and a color receiver partial images in three different primary colors, corresponding to three selected ultraviolet wavelengths, may be presented to the eye in rapid succession. This permits an immediate recognition of localized differential absorptions by color differences. Problems arising in the construction of the above device are discussed.

79. RECORDING MULTI-AXIAL PROJECTION OF VECTORCARDIOGRAMS: THE AXOSTAT

B. P. McKay, W. E. Romans, D. A. Brody, and R. C. Little
(University of Tennessee, Memphis, Tenn.)

Present electrocardiographic instrumentation necessitates recording from many different electrodes to determine transition points which delineate the indices of ventricular gradient. With the "Axostat" these transition points are located by recording from only four electrodes. A patient is connected through appropriate switches to a wye (Wilson central terminal) and inverted delta resistance network. A sliding contact on the delta network allows voltages to be taken through 360 degrees with respect to the center of the wye. The relation of these voltages to the normal electrocardiographic potentials is a calculable function for the angular position of the contact. This has shown a need for calibrated preamplification in order that recordings may be taken at normal electrocardiograph amplitudes. Circuitry is described for the above networks and preamplifier. For clinical application simplified resistive networks are shown which do not require preamplification and can be used directly with many standard electrocardiographs.

80. CONTINUOUS INTEGRATING COUNTING-RATE SYSTEM FOR RADIOACTIVITY

Mones Berman and Salvatore Vacirca
(Sloan-Kettering Institute, New York, N. Y.)

An instrument is described that continuously computes the counting rate of a source by integrating the total number of counts and giving an accuracy at any time characteristic to the number of counts collected. The count integration is obtained by converting the counts into a voltage proportional to its number. This voltage is continuously balanced against another voltage which varies linearly with time. The latter voltage is obtained by driving a potentiometer with a synchronous motor. For balance between the two voltages to be maintained, the total voltage across the potentiometer must be proportional to the integrated counting rate and is, hence, a measure of it. Adjustment of this voltage for balance conditions is made either manually or automatically.

Microwaves II—Waveguides B

Chairman, A. G. Kandoian
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

81. NONUNIFORM TRANSMISSION LINES

J. G. Gurley
(Hughes Aircraft Company, Culver City, Calif.)

Transmission lines with tapered mechanical dimensions or electrical properties are analyzed in terms of interactions between modes. A rigorous solution of the electromagnetic field problem is obtained by considering all modes; when only one mode has appreciable amplitude then the results are consistent with quasi-static transmission line concepts.

A by-product of this investigation is an expression for characteristic impedance which is valid for waveguides as well as for two-conductor lines. Successful measurements were made on a rectangular waveguide with a vertical dielectric web of tapered thickness.

82. THE OPTIMUM PISTON POSITION FOR COAXIAL-TO-WAVEGUIDE TRANSDUCERS

W. W. Mumford

(Bell Telephone Laboratories, Inc., Holmdel, N. J.)

A coaxial line can be matched to a waveguide by means of a probe antenna located ahead of a short circuiting plunger. An impedance match can be achieved by varying any two of the following three dimensions; (a) the probe length; (b) the piston position; (c) the off-center position of the probe.

This paper points out that there is, theoretically, an optimum piston position for greatest bandwidth, and presents some evidence corroborating this theory. Bandwidths of ±10 per cent to the 1 db SWR have been realized by fixing the piston position at its optimum and varying (a) and (c) above to obtain a match.

83. BROAD-BAND RIDGED AND FLATGUIDE COMPONENTS 10-40 KMC

Samuel Hopfer

(Polytechnic Research and Development Company, Inc., Brooklyn, N. Y.)

Two basic transmission systems have been selected for the construction of broadband components in the range from 10–40 kmc/sec, namely, the ridged guide and the flat guide. The former is a single mode system and useful for broadband receiver work. The components to be discussed are: Tuner, Crystal Mount, Attenuator, and Tapers. The flat guide system useful for transmission work is capable of supporting higher modes. The type of discontinuities employed in the construction of components do not generate higher modes. The flat guide components to be discussed are: Tuner, Crystal Mount. The equivalent circuit of a third mode generator such as the inductive window is given.

84. STEP-TWIST WAVEGUIDE COMPONENTS

Henry Schwiebert and H. A. Wheeler

(Wheeler Laboratories, Great Neck, N. Y.)

The step twist is a sequence of short sections of rectangular waveguide twisted about their common axis, having graduated dimensions and angles determined by techniques newly developed for this purpose. The fixed twist supplants the much longer twisted waveguide with advantages in wideband matching in shorter space by greater freedom of dimensioning and greater reproducibility. A 90° fixed twist, handling the rated 40 per cent frequency bandwidth of the waveguide within 0.3 db SWR, is made in a length only 3/2 the guide width. The step twist has been adapted to rotary designs in lengths proportional to the required refinement of wideband matching.

85. WAVEGUIDE MATCHING TECHNIQUE

W. C. Jakes, Jr.

(Bell Telephone Laboratories, Inc., Holmdel, N. J.)

This paper presents the results of a theoretical and experimental study of a new waveguide matching technique. The main advantage of the method described is that the technique may be applied at any distance away from the discontinuity causing the original mismatch and a broad-band match may still be obtained.

Design curves are included which give the required parameters for the technique and the power loss for a given initial mismatch and desired VSWR reduction. Experimental confirmation of the theory is also presented.

SYMPOSIUM

TV Station Construction and Theater Conversion

(Organized by Professional Group on Broadcast Transmission Systems)

Moderator, R. F. Guy

(National Broadcasting Company, New York, N. Y.)

86. THE NEW WOR-TV STUDIO AND TRANSMITTER BUILDING AT 60TH STREET AND COLUMBUS AVE., NEW YORK CITY

J. R. Poppele

(WOR-TV, New York, N. Y.)

87. NEW BUILDING AND TECHNICAL FACILITIES AT WCAU-TV, PHILADELPHIA

J. G. Leitch

(WCAU, Philadelphia, Pa.)

88. THE WFAA-TV PLANT, DALLAS, TEXAS

C. L. Dodd

(Dallas News-The Dallas Journal, Dallas, Tex.)

The foregoing trio of papers, 86, 87, and 88 will offer a detailed report on the physical, mechanical, and electrical problems involved in the preparation, construction, installation, and operation of new TV broadcasting facilities.

89. THEATER-TV CONVERSIONS

(a). NBC PROGRAM

A. A. Walsh

(National Broadcasting Company, New York, N. Y.)

(b). CBS PROGRAM

A. B. Chamberlain

(Columbia Broadcasting System, New York, N. Y.)

(c). ABC PROGRAM

J. M. Middlebrooks

(American Broadcasting Company, New York, N. Y.)

In (a), (b), and (c) above, there will be offered a comprehensive survey of all of the interesting aspects relating to the physical, mechanical, electrical, audio, and video equipment requirements in typical converted legitimate theatres or motion picture theatres in New York City.

Topics to be covered will include lease or purchase negotiations, typical conversion schedules, major modifications usually necessary, scenery and prop handling, typical floor plans (before and after conversion), and layout arrangements (stage and audience areas, control room, equipment room, studio-mc interconnections, etc.)

SYMPOSIUM

Present Status of NTSC Color-Television Standards

Chairman, A. G. Jensen

(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

90. A panel of leading color-television engineers will discuss the most recent work of NTSC panels towards the preparation of standards for the NTSC color-television system. The discussion will include any available results of the field tests currently being conducted under such standards.

Instrumentation III—Electronic Measurements B

Chairman, R. R. Batcher

(Radio Television Manufacturers Association, New York, N. Y.)

91. A RASTER SWEEP OSCILLOGRAPH FOR PRECISION TIME MEASUREMENTS

H. B. Steinhauser

(Allen B. DuMont Laboratories, Inc., Clifton, N. J.)

The equipment described provides a means for analyzing the time relationships between electrical pulses. A raster type presentation is used in order to combine the advantage of a fast sweep for accurate time determination with a long time base for extended range of measurements. Incoming video signals produce vertical deflection, while the raster sweep is triggered from an external pulse. The display is photographed and the oscillogram enlarged to obtain reading accuracy commensurate with measurement accuracy. The complete unit is mounted in a dust-tight cabinet with pressure ventilation. Slide mounts allow each individual chassis to be pulled out, tilted upward 90°, and locked in position for servicing.

92. PRECISION AUTOMATIC TIME MEASUREMENT EQUIPMENT

D. W. Burbeck and W. E. Frady

(Hughes Aircraft Company, Culver City, Calif.)

Equipment using a new method for automatically measuring time differences to an average accuracy of 0.008 microsecond over the range of 25 to 600 microseconds is described. The principle of operation involves two steps—first, the number of 2.0 megacycle pulses occurring in the interval to be measured are counted, then the interpolation between the 2 mc pulses is carried out by a pulse counting vernier technique which gives an over-all accuracy of 0.008 microsecond for the measurement. The vernier interpolation provides high precision without exceeding a 2 mc counting rate.

The final result of the measurement appears as a number in a binary counter. The range of the system may be readily extended by increasing the frequency stability of the 2 mc frequency source.

93. A ROTATING-BEAM CEILOMETER SYSTEM

R. H. GUENTHER AND L. W. FOSKETT
(U. S. Weather Bureau, Washington 25, D. C.)

This ceilometer system indicates cloud heights every 24 seconds. It works by triangulation and comprises a detector, projector, and indicator. The detector is a lead-sulphide cell with parabolic reflector. The projector includes a modulated light source with another reflector. The detector "sees the zenith" and the projector, by suitable rotation, scans the detector field of view. The indicator is a 5-inch cathode-ray tube with a peripheral height scale. The indicator spot moves 360 degrees while the projector scans the detector field of view. Detected signals deflect the spot radially toward the scale, thus indicating cloud height. Measurements up to 2,000 feet are made consistently, both day and night.

94. A POLAR-COORDINATE CATHODE-RAY OSCILLOGRAPH FOR USE WITH THE ROTATING-BEAM CEILOMETER

M. T. NADIR AND M. B. KLINE
(Allen B. DuMont Laboratories, Inc., Clifton, N. J.)

A polar-coordinate cathode-ray oscillograph designed specifically for cloud-height measurements in the U. S. Weather Bureau's new rotating-beam ceilometer system is described. An automatically synchronized circular baseline is produced on a Type 5CP7-A cathode-ray tube, with a selected 36° sector scanned by the searchlight spread out over a 360° circle. The signals picked up by the photocell detector of the ceilometer system produce radial deflection simultaneously with automatically produced altitude markers. Using a 500 foot baseline, a range of altitude from zero to beyond 14,000 feet may be indicated. Accuracies of a few feet in altitude are attained on the 400 foot range.

A number of the novel circuits used in the oscillograph are described. The unique mechanical and optical system utilizes unitized design to facilitate changes of baseline for different installations and maintenance in the field.

Standard oscillograph-record cameras may be used for making photographic records.

95. AN ELECTRONIC FRINGE INTERPOLATOR FOR AN OPTICAL INTERFEROMETER

R. D. HUNTOON
(National Bureau of Standards, Corona, Calif.)

In the interferometer method, precision length measurements require careful interpolation between fringes. This electronic interpolator generates an FM wave by vibrating a plate with an amplitude of about $\lambda/4$. The exact amplitude is determined by analyzing the wave, and the result is displayed on a cro indicator.

Interpolation to about 1/100 fringe has been accomplished.

Television III—General B

Chairman, I. J. KAAR
(General Electric Company, Syracuse, N. Y.)

96. THE PROBLEM OF INTERLACE IN TELEVISION RECEIVERS

J. DE LEON
(Allen B. DuMont Laboratories, Inc., East Paterson, N. J.)

With the continuing trend toward larger picture tubes in television receivers, more attention is necessarily directed toward improvement of interlace.

Various types of noninterlace are examined, with special attention to the form evidenced as stable pairing. This form is generally attributed to interfering voltage coupled to the vertical oscillator from horizontal sweep circuits. It is shown that similar misinformation may be obtained from the incoming sync signal.

Means of reduction of interfering pickup are discussed, together with certain methods of reducing susceptibility of vertical oscillator circuits to interference. Tools and techniques applicable to experimental investigations are described.

97. A METHOD OF EVALUATING THE PERFORMANCE OF A TELEVISION PICTURE TUBE AND ITS ASSOCIATED COMPONENTS

JULIUS GREEN
(Philco Corporation, Philadelphia, Pa.)

The paper describes a method for evaluating the distribution of light intensity across a scanning line of a picture tube, the equipment used, and the results obtained. The method employed involves illuminating only a single line of a standard television pattern, and imaging it through a fine optical slit onto a photocell.

There is produced on a cathode-ray oscilloscope a trace the ordinates of which are proportional to the distribution of luminous intensity on the picture tube.

An adaptation of this method to measure the relation between contrast and resolution is described.

98. CHARACTERISTICS AND PERFORMANCE OF TELEVISION CLAMPING CIRCUITS

A. J. BARACKET
(Federal Telecommunication Laboratories, Nutley, N. J.)

Clamping circuits are used to insert direct-current information into studio camera chains, picture monitors, and receivers. They permit the use of coupling circuits having relatively poor low-frequency response, which tend to suppress the effects of hum and microphonics. Direct-current error, recovery time, synchronization loading, and noise bursts are considered under extreme noise and signal variations. Immunity to noise and recovery time are functions of the clamp coupling capacitance and the duration of the clamping pulses. It is shown that compromises may often be employed to satisfy the requirements of both immunity to noise and good recovery time.

99. COLOR-TELEVISION SYNCHRONIZING-GENERATOR CIRCUITS

I. KRAUSE, A. J. BARACKET, AND H. DELL
(Federal Telecommunication Laboratories, Nutley, N. J.)

Binary frequency dividers with feedback loops and accurate gating and keying circuits are employed in generating timing signals and in shaping pulses for field-sequential color television. Generating and positioning of the color pulse during the red field are described. The 48-cycle frame pulses are locked to the 60-cycle power-line frequency by separate circuits. Crystal control is also provided at ten times the equalizing pulse rate. 583-kilocycle marker signals, 0.6-microsecond wide, are generated at a 2.16-kilocycle burst rate and may be mixed with blanking pulses for checking sweep linearity of color monitors and receivers.

100. PRINTED UNIT ASSEMBLIES FOR TELEVISION

W. H. HANNAHS AND NORMAN STEIN
(Sylvania Electric Products, Inc., Bayside, N. Y.)

A one-stage module is described for TV receiver construction. Etching and silk-screening techniques are combined to produce a completely "printed" unit assembled by solder-dipping, which interconnects with other stages without wire. Design data and performance are given for a 25 mc IF. This type of construction is usable for most of the stages in present receivers and is expected to save critical materials. Resistors are fabricated to commercial tolerances.

Circuits III

Chairman, H. A. WHEELER
(Wheeler Laboratories, Great Neck, N. Y.)

101. THE EFFECTIVE BANDWIDTH OF VIDEO AMPLIFIERS

F. J. TISCHER
(Stockholm, Sweden)

Investigation of some relations existing between transient response and transfer function on a steady-state basis of video amplifiers shows that it is possible to derive directly from the complex transfer function an "effective" bandwidth as a figure of merit for the relative utility of the amplifier for transient application. The effective bandwidth is independent of the shape of the amplitude characteristic and takes account of the phase characteristic. Important relations are investigated.

102. TRANSIENT RESPONSE OF CATHODE PEAKED VIDEO AMPLIFIERS

J. H. Mulligan, Jr.
(New York University, New York, N. Y.)
AND
L. Mautner
(Hughes Research and Development Laboratories, Culver City, Calif.)

Consideration is given to the effect produced on the step function transient response of a video amplifier when cathode peaking is added. Conclusions are drawn concerning the effects produced on the transient response of a video amplifier of general form when this type of peaking is used. The results are obtained primarily by utilizing the theory of the dependence of step function response upon pole and zero locations in the complex frequency plane.

To illustrate the analysis presented, examples will be given showing the effect of cathode peaking on the transient response of several common video amplifier configurations.

103. VARIABLE BANDWIDTH-AMPLIFIER DESIGN FOR HIGH RATE OF CUTOFF AND LARGE BANDWIDTH VARIATIONS

M. Dishal
(Federal Telecommunications Laboratories, Nutley, B. J.)

Variable-bandwidth intermediate-frequency amplifiers are often required. For large bandwidth variation and high rate of cutoff in the narrow-band condition, the usual overcoupled double-tuned circuit with variable coupling combined with a single-tuned circuit ($n=3$) will not do.

Exact design data will be given for variable-bandwidth intermediate-frequency amplifiers using cascades of 1 to 4 different double-tuned circuits ($n=1$ to 8). The design methods are based on the known desired location of the poles of the required response shape, and the information will be presented in a graphical form that is quite simple to use.

104. COUPLING CIRCUITS HAVING FLAT-AMPLITUDE CHARACTERISTICS

A. B. Macnee
(University of Michigan, Ann Arbor, Mich.)

Adjustment of peaking circuits for flat amplitude response is considered, such adjustment being desirable when wide bandwidths without good transient response are required. Modified versions of the series-peaking and series-shunt-peaking circuits are analyzed. The circuit parameters necessary to give an optimum flatness of the amplitude response curve are presented for all ratios of input to output capacity. Gain-bandwidth factors of from $\sqrt{2}$ to 2.7 are achievable with circuits adjusted in this manner. Modification of the circuits for good transient response will also be discussed.

105. OSCILLATOR SYSTEMS CONTROLLED BY PHASE DETECTOR REACTANCE TUBE

J. C. Tellier and G. W. Preston
(Philco Corporation, Philadelphia, Pa.)

Conditions under which a phase-detector reactance-tube controlled oscillator will synchronize with a reference signal will be analyzed and a basic equation derived. If certain relations hold among the parameters of the equation, synchronization will occur for a given value of initial frequency difference between oscillator and reference signals, for any value of initial phase difference. These relations will be presented graphically. An approximate analytic expression for this curve is used to obtain an equation which describes the lock-in performance in terms of the cutoff frequency of the RC network and the permissible static phase error after synchronization.

106. ESSENTIAL INSERTION LOSS

D. R. Crosby
(Radio Corporation of America, Camden, N. J.)

The idea of essential insertion loss has both theoretical and practical advantages. Unlike the usual insertion loss, it is the property of a two-port net and so is independent of the terminations of the net.

In certain important cases at microwave frequencies, it is more easily measured with conventional test equipment than insertion loss. The theory yields useful theorems which are not true for insertion loss, and is descriptive of important filter properties concerned with microwave duplex filters.

Propagation

Chairman. A. E. Cullum, Jr.
(Consulting Radio Engineer, Dallas, Tex.)

107. THE POLARIZATION OF VERTICALLY INCIDENT LONG RADIO WAVES

J. M. Kelso, H. J. Nearhoof, R. J. Nertney, and A. H. Waynick
(Pennsylvania State College, State College, Pa.)

The polarization of long electromagnetic waves in the ionosphere is considered. The analytical expressions relating the $N-\nu$ distributions of the ionosphere to the characteristic polarization are developed. The polarization of the electromagnetic wave leaving the ionosphere is treated in two ways. In Part I the wave is treated as a single magneto-ionic component. It is shown that under certain conditions the wave does behave as a single component which attains a "limiting polarization" at some N value $N\neq 0$. It is shown, however, that this analysis "breaks down" under certain conditions and, in fact, wave polarizations occur which are not characteristic polarizations associated with any $N-\nu$ value.

It is shown in Part II that a model consisting of electrons D and E-regions is capable of giving polarization results which are in good agreement with experiment.

108. RADIO TRANSMISSION BEYOND THE HORIZON IN THE 40–4,000 MC BAND

Kenneth Bullington
(Bell Telephone Laboratories, Inc., New York, N. Y.)

Reliable signals have been received at distances of several hundreds of miles at frequencies of 500 and 3,700 mc. The median signal levels are 50–90 db below the free space field but are hundreds of db (in one case 700 db) stronger than the value predicted by the classical theory based on a smooth spherical earth with a standard atmosphere. Antenna gains and beamwidths are maintained to a first approximation and no long delayed echoes have been found. The experimental results are compared with other available data.

109. TROPOSPHERIC PROPAGATION DATA ON FREQUENCIES BETWEEN 92 AND 1,047 MC AT DISTANCES FAR BEYOND THE HORIZON

G. R. Chambers, J. H. Chisholm, J. W. Herbstreit, and K. A. Norton
(National Bureau of Standards, Washington, D. C.)

Preliminary measurements have been made at distances between 50 and 400 miles of the signal power received from conventional 3 kw vhf transmitters and from a specially designed 4 kw continuous-wave crystal-controlled 1,047 mc klystron transmitter located on Cheyenne Mountain in Colorado. The distance ranges at which the measurements could be made were increased by confining the radiated energy to a narrow frequency band. Contrary to expectation from either standard atmosphere or duct theory, the measured attenuation rate (db per mile) was not constant at large distances beyond the horizon but increased markedly and continuously out to the maximum distance covered.

There was some evidence that the received signals were propagated at least in part via some scattering mechanism, but this scattering was not sufficient at a point far beyond the horizon to appreciably reduce the gain (24 db) or change the pattern expected with plane waves for a 10-foot paraboloidal antenna at 1,047 mc.

110. STATISTICAL FLUCTUATIONS OF RADIO FIELD STRENGTH FAR BEYOND THE HORIZON

S. O. Rice
(Bell Telephone Laboratories, Inc., New York, N. Y.)

When a sinusoidal radio wave of extremely high frequency is sent out by a transmitter, the wave received far beyond the horizon is often observed to fluctuate. Here some of the statistical properties of this fluctuation are derived on the Booker-Gordon assumption; namely that the received wave is the sum of many little waves produced when the transmitter beam strikes "scatterers" distributed in the troposphere. Expressions are obtained for the periods of the fluctuations in time, in space, and in frequency. These expressions extend closely related results obtained by Booker, Ratcliffe, and others.

111. SOME CONSIDERATIONS IN THE USE OF HIGHLY DIRECTIONAL ANTENNAS ON SOURCES OF COMPARABLE ANGULAR SIZE TO THE BEAMWIDTH

D. O. McCoy
(Collins Radio Company, Cedar Rapids, Iowa)

Particularly in the field of radio astronomy, it is necessary to receive energy from sources that subtend angles comparable to that of the beamwidth. Under this condition, both the gain and the effective beamwidth of the antenna are altered. The discussion begins with a review of some of the fundamentals involved, revealing the nature of the variations encountered with non-point sources. The results of a theoretical analysis of the problem are presented. The discussion ends with a bit of philosophy comparing radio antennas with optical systems.

Microwaves III—Filters and Circuits

Chairman, D. D. King
(The Johns Hopkins University, Baltimore, Md.)

112. FURTHER TRANSMISSION ANALYSIS OF HYBRID RINGS

H. T. Budenbom
(Bell Telephone Laboratories, Inc., Whippany, N. J.)

A paper presented at the 1948 IRE National Convention gave an analysis of the hybrid ring as a re-entrant transmission line. The present paper is in many respects a summary of advances in the study of four-arm series hybrid rings since that paper. Work by L. J. Cutrona and H. Kahn is mentioned; the remainder of the work is from Bell Laboratories sources. The second section of the paper gives: (a) procedure for including the effect of conductor resistance; (b) discussion of peak conjugacy and impedance characteristics, the latter including discussion of optimum waveguide annulus width; (c) effective equivalent "T" section length; and (d) phase effects of impedance in conjugate leg. The paper concludes with discussion of a multiplying property of the four-arm series hybrid ring.

113. RESONANT CAVITY BAND-PASS FILTERS—PRACTICAL ADJUSTMENT TO PREDICTED PERFORMANCE

D. DeWitt, M. Klein, and T. J. Potts, Jr.
(Radio Receptor Company, Inc., Brooklyn, N. Y.)

Using the procedures given here, band-pass filters, using three or four coupled resonant circuits at microwave frequencies, are simple to design and adjust. Design curves and a systematic procedure for adjusting each coupling and tuning parameter to realize the theoretical characteristics are presented. Calculations are based on lumped circuit representations of resonant cavities. The adjustment procedure involves the location of standing-wave minima on a slotted section preceding the filter with the filter termination uncoupled. Tables of instructions are presented based on the behavior of the input reactance of the equivalent lumped network at various stages of adjustment. Cavities are assumed lossless, isochronous narrow-band approximations are made, and lumped coupling is assumed.

114. SYNTHESIS OF NARROW-BAND DIRECT-COUPLED FILTERS

H. J. Riblet
(Microwave Development Laboratories, Inc., Waltham, Mass.)

A general synthesis procedure for the design of narrow-band direct-coupled filters is based on an approximate first-order equivalence between direct and quarter-wave coupled filters. Thus a quarter-wave coupled filter, whose bandwidth is a few per cent wider than required, serves as a prototype. The approximations underlying the general synthesis procedure for quarter-wave coupled filters, given by Lawson and Fano and applied by Mumford to the maximally flat case, are re-examined and justified. The transmission characteristics of a three- and a six-cavity filter, each of total Q of about 50, are computed exactly, with excellent agreement with the design performance.

115. ON HIGH-*K* DIELECTRIC CAVITIES

H. M. Schlicke
(Allen-Bradley Company, Milwaukee, Wis.)

This paper is concerned with resonance phenomena in cylinders in high-K dielectrics. These "cavities" are extremely small in terms of the wavelength. All four combinations of radial and axial impedance being zero (metallized dielectric) or infinite (interface dielectric/air) are investigated. In contradistinction to conventional cavity theory, dealing only with metallic boundaries, quasi-degenerated TE modes are realizable for nonmetalized faces of the cylinder. The proper mode for high-K disk condensers is also derived. The "cavities" are easily tunable by magnetic rods. A dielectric spiral, evolved from the quasi-degenerated TE_{010}^{**} mode, and encircling a magnetic rod constitutes a novel type of antenna.

116. A DUAL-CHANNEL COLINEAR ROTARY JOINT

E. O. Hartig
(Goodyear Aircraft Corporation, Akron, Ohio)

This paper describes a dual-channel colinear rotary joint designed for use at X band. The input and output transitions are identical and consist of two rectangular waveguides coupling into orthogonal TE_{11} modes in circular waveguide. Between the two transitions is a 180 degree differential phase shifter which rotates at half the speed of the output junction.

The matrix theory of this rotary joint has been derived. A theoretical study has also been made to determine the effect of misalignment of the elements, variations in the differential phase shift and mismatch of the load.

A rotary joint of this type has been fabricated, tested, and shown to have greater than 35 db decoupling between channels and a very low VSWR over a 6% frequency band.

SYMPOSIUM
Digital Computers in Control Systems

(Organized by Professional Group on Radio Telemetry and Remote Control)

Chairman, J. W. Forrester
(Massachusetts Institute of Technology, Cambridge, Mass.)

The Symposium will deal with the application of digital computers and digital techniques to the solution of a large scale control problem in which information is available at remote points and control functions are to be performed at remote places. The four operations of coding, communicating, computing, and display will be covered.

117. DIGITAL COMPUTERS IN CONTROL SYSTEMS

J. W. Forrester
(Massachusetts Institute of Technology, Cambridge, Mass.)

Control and data collection systems are rapidly becoming larger in physical extent and more complex in the tasks which are assigned. For long distance transmission of multi-channel precision data, pulse-code digital modulation systems have been employed. With new advances in analog-to-digital translation devices and in high-speed digital computers, the all-digital integrated system is fast becoming a reality. The new systems need for their successful execution a keen awareness of the "systems engineering" task. Proper displays and the coupling between persons and the system become increasingly important.

118. CODERS

R. P. Mork
(Raytheon Manufacturing Company, Waltham, Mass.)

A description is given of several practical digital coding methods for use in control systems, and their advantages are assessed. Coding methods suitable for use with both electrical and mechanical input data are included. The paper discusses the problem of coding at high speeds, as required when multiplexing several data channels for transmission over a single link, with a presentation of some unusual techniques for increasing maximum coding speeds and reducing the average quantizing error.

119. DATA TRANSMISSION LINKS

P. Pontecorvo
(Raytheon Manufacturing Company, Waltham, Mass.)

Some of the general problems encountered in the transmission of information in digital form over microwave links are discussed. The solution of some of these problems in one particular application are described in more detail.

120. THE DIGITAL COMPUTER AS A CONTROL ELEMENT

C. R. Wieser
(Massachusetts Institute of Technology, Cambridge, Mass.)

When a digital computer is used to control a large-scale physical system, the computer's capability for carrying out logical as well as arithmetic operations makes it a key element in planning the entire system. Before the question of "how to compute," comes the question of "what to compute," and this must be decided, exactly and in detail, before either the system behavior or the computer requirements can be precisely defined. The desired functional behavior of the system must be broken down into a pattern of simple logical steps, which can be translated into instructions for the computer to follow.

121. DISPLAY ELEMENTS

B. S. Benson
(Benson-Lehner Corporation, West Los Angeles, Calif.)

The flow of information between digital devices and human beings is discussed with reference to communication and control. A survey of available equipment and future trend is also discussed.

Antennas I—General

Chairman, L. C. Van Atta
(Hughes Aircraft Company, Culver City, Calif.)

122. OPTIMUM PATTERNS FOR ARRAYS OF NONISOTROPIC SOURCES

George Sinclair
(University of Toronto, Toronto, Canada)
and
F. V. Cairns
(National Research Council, Ottawa, Canada)

The problem of synthesizing the optimum pattern for an array of nonisotropic sources will be discussed. The mathematical conditions to be satisfied by the polynomial representing the pattern space factor will be given. It is found that trial-and-error methods yield satisfactory approximations to the optimum polynomial for arrays having up to six or eight elements. Arrays containing large numbers of elements present a more difficult problem, and various approximate methods for finding the optimum polynomial are discussed. The current distributions for a number of optimum arrays will be given.

123. A GEOMETRICAL METHOD OF ANALYZING THE EFFECTS OF SITE REFLECTIONS ON DIRECTION-FINDING SYSTEMS

G. A. Deschamps
(Federal Telecommunication Laboratories, Nutley, N. J.)

Analysis of the response of even a simple loop direction finder to multiple rays of arbitrary polarizations and directions is difficult. Poincare's spherical representation of ellipses permits its expression in geometrical form and general conclusions may be drawn sometimes. The effect of continuous phase change between two rays is thus interpreted.

More important is the relation between statistical optics with partially polarized light and direction finding through randomly phased site reflections, for which the notion of "partially directed" waves was introduced.

Stokes parameters from statistical optics permit analyzing such problems as finding time averages of response of one antenna.

124. THE RADIATED FIELDS OF PULSE-EXCITED DIPOLE ANTENNAS

C. S. Roys
(University of Massachusetts, Amherst, Mass.)

Since there is little material available concerning the general performance of antennas with pulse excitation, the writer has developed two methods of analysis.

In the first method the input current was first expressed in Fourier integral form. This, together with the results available for antennas with cw excitation, resulted in a corresponding integral formulation for the radiated field.

The second method consisted of finding the incident and reflected current surges along the antenna. The resultant current moment was obtained next by integration. This gave an instantaneous form of the so-called "Schelkunoff radiation vector." Formulas for the oscillogram of the radiated field could then be determined by application of Maxwell's equations.

125. AN EXPERIMENTAL INVESTIGATION OF THE CORNER REFLECTOR ANTENNA

E. F. Harris
(The Antenna Research Laboratory, Inc., Columbus, Ohio)

More than 1,000 measured radiation patterns for the corner configuration have been taken for corner angles from 10° to 270° and for dipole-to-corner spacings of 0.1 to 3 wavelengths. Both *H*-plane and *E*-plane patterns are shown for each configuration employing semi-infinite sheet sides.

Certain specific corners have been investigated for the effects on pattern of finite side lengths and side height, and integration patterns run to compute absolute gain of the units. Effects of spines and spacing of grid construction are investigated.

126. AN OMNIDIRECTIONAL SLOT ANTENNA ARRAY

A. J. Hoehn and S. I. Cohn
(Armour Research Foundation, Chicago, Ill.)

A longitudinally-polarized, omnidirectional antenna array is described in which the elements are formed from a coaxial line by slots running around the periphery of the outer conductor. The complete unit consists of a four-element, colinear, broadside array designed to operate at 2,500 megacycles per second. It is approximately 14 inches long, has a diameter of ⅞ inch including radome, and is strong, lightweight, and adaptable to low-cost manufacture. The array is capable of handling an average power of at least ten watts and has a power gain greater than four.

SYMPOSIUM
UHF Receivers I

(Organized by Professional Group on Broadcast and Television Receivers)

Chairman, D. E. Harnett
(General Electric Company. Syracuse, N. Y.)

127. UHF HYBRID RING MIXERS

W. V. Tyminski and A. E. Hylas
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

Application of hybrid rings to mixers reduces local oscillator radiation and noise, suppresses radio-frequency interference, decreases cross-modulation products, and provides a degree of high-pass filter action. In the 470 to 890 mc band the performance of hybrid ring mixers followed by low-noise IF amplifiers, compares favorably with RF amplifiers and conventional mixers.

A practical low cost uhf hybrid ring mixer has been developed which overcomes size and bandwidth limitations. Data on uhf hybrid ring performance will be presented, and comparisons will be made with other uhf receiving methods.

128. UHF TUNERS

M. F. Melvin
(P. R. Mallory & Co., Inc., Indianapolis, Ind.)

The problems encountered in uhf tuning and conversion can best be met through the use of a continuous tuner. The techniques of application of circuitry to a continuous tuner of the variable-inductance type, yield simpler solutions than are obtainable with other methods of tuning. This continuous tuning technique lends itself to uhf converter application as well as vhf-uhf "front end" design.

129. THE DESIGN AND PERFORMANCE OF A COMPACT UHF TUNER

H. F. Rieth
(Kingston Products Corporation, Kokomo, Ind.)

A uhf converter or tuner continuously covering the frequency range of 470–890 mc, using curved tuned lines as a basic tuning device, is described. The mechanical design lends itself to the small basic package with simplicity of production and alignment. Mechanical layout and microwave measurement technique will be briefly described. The design is primarily intended for use as a uhf converter or as a tuner in a television chassis. Performance data and curves will be submitted.

130. A UHF-VHF TURRET TUNER FOR TELEVISION RECEIVERS

Albert Cotsworth, Max Beier, John Bell and James White
(Zenith Radio Corporation, Chicago, Ill.)

A turret-type tuner capable of both uhf and vhf reception has been described. A new tuner of this type is described which combines small size, ease of strip replacement, good performance, low radiation, and low cost.

131. AN 82-CHANNEL TURRET TUNER

A. M. Scandurra
(Kollsman Instrument Corporation, Elmhurst, N. Y.)

Experiments have been conducted to determine the feasibility of discrete channel selections in the uhf range similar to that presently used by most vhf tuners. The paper presents the engineering problems and describes some of the unsuccessful attempts as well as the reasons for choosing the final circuitry and mechanical arrangement. A discrete channel uhf tuner covering all television channels will be discussed in detail. The paper covers the circuit problems, vacuum tube problems, characteristics of the uhf-vhf tuner (noise figure, image rejection, etc.) and the fundamental aspects of probability of successful reception under field conditions. Due to the many possible solutions, the development of the 82-channel turret tuner will discuss the technical, economical, and field requirements. The characteristics of a turret tuner as compared to a continuously tunable structure will be analyzed. Photographs of experimental tuners utilizing these principles and measurements of performance will be presented.

132. RF PERFORMANCE OF A NEW UHF TRIODE

H. W. A. Chalberg
(General Electric Company, Owensboro, Ky.)

This paper is being presented in conjunction with the paper "A UHF Amplifier Tube for Television Tuners" which will be presented in the technical session on Small High-Frequency Tubes.

The details of techniques of measurements and the dynamic results obtained with the uhf triode will be discussed. Performance characteristics in the vhf and uhf television bands will be covered, and a comparison of gain and noise figures will be made with tubes now available for vhf amplifier service.

Circuits IV

Chairman, L. A. Zadeh
(Columbia University, New York, N. Y.)

133. DISPERSION IN TRANSMISSION SYSTEMS

M. J. Di Toro
(Federal Telecommunications Laboratories, Nutley, N. J.)

The novel quantum methods of Gabor show intuitively satisfying formulations for delay and dispersion of the impulse transient response of transmission systems whose amplitude and phase steady-state transfer characteristics are neither flat nor linear. Dispersion or pulse lengthening in the impulse transient response results from the weighted phase distortion and from ripples or mean-square fluctuations in amplitude response. In all transmission systems, dispersion gives rise to intersymbol interference with consequent limitation in the amount of transmissible information.

134. NETWORK SYNTHESIS FOR SPECIFIED TRANSIENT RESPONSE

W. H. Kautz
(Stanford Research Institute, Stanford, Calif.)

Three methods for the synthesis of finite, lumped-parameter networks for the case in which the desired behavior is prescribed in transient terms rather than the more usual frequency (gain and/or phase) characteristics will be presented. Refinements on the already-known "time-domain" approach, certain time-frequency-domain relationships which permit a "frequency-domain" approximation while retaining control over the network transient response, and a new method employing both time- and frequency-domain approximations, form the basis for the procedures.

135. TRANSFORMS FOR LINEAR TIME-VARYING NETWORK FUNCTIONS

J. A. Aseltine and D. L. Trautman
(University of California, Los Angeles, Calif.)

Methods for finding integral transformations appropriate for linear time-varying systems will be discussed. These transformations reduce the differential equations describing the variable system to algebraic ones. Many features of the Laplace transform method for fixed systems have counterparts. In particular, these transforms lead to a "system function" through which analysis and synthesis methods can be formulated. Theorems and tables of transform-pairs are developed for systems with two kinds of parameter variation. This procedure facilitates, for example, synthesis of networks having a transient response of the form t^n if L and C vary as t, and $(\sin bt)/t$ if R varies as $1/t$.

136. PARALLEL-TUNED CIRCUIT PERIODICALLY SWITCHED TO A DC SOURCE

L. J. Giacoletto
(Radio Corporation of America, Princeton, N. J.)

A parallel-tuned circuit periodically connected to a source of linear direct-current energy (e.g., by means of an electron tube or switch) is fundamental to a large group of energy conversion circuits. A complete analysis embraces many voltage and current variations including sinusoidal, saw-tooth, and more complex wave shapes depending upon circuit parameters and switch period. The analytic results can be used to analyze sinusoidal and nonsinusoidal oscillators, class-C amplifiers, and many pulsed circuits.

137. A HIGHLY ACCURATE VARIABLE TIME DELAY SYSTEM

Y. P. Yu
(Allen B. DuMont Laboratories, Clifton, N. J.)

A variable time-delay system having time jitter less than 0.0001 microsecond, equivalent to 2.5 parts per million of its maximum time delay, will be described. Bilateral elements and distributed amplifier to compensate losses are used. The signal pulse travels through each delay element many times to increase total time delay. An experimental unit having total time delay up to 100 microseconds in steps of one microsecond has been satisfactory. Usual causes of time jitter such as variations of cutoff characteristics, noise, hum, and fluctuations of supply voltages, cannot affect the accuracy of the system.

138. RC TIME DELAY CIRCUIT OF VERY HIGH TIME CONSTANT

R. G. Roush
(The Johns Hopkins University, Baltimore, Md.)

Performance of the cathode-follower circuit in timing applications in which its input impedance is used in conjunction with a capacitor to give a high effective time constant without prohibitively large component values will be analyzed. The simple cathode-follower circuit may be incorporated into any resistance-capacitance timing circuit using the differentiator configuration common to the multivibrator. Such a circuit will permit the monostable multivibrator to be used for the generation of adjustable delay periods of many minutes with good accuracy for most applications.

Electron Tubes I—Power and Gas Tubes

Chairman, Dayton Ulrey
(Radio Corporation of America, Harrison, N. J.)

139. METHOD FOR PREDICTION OF MAGNETRON CHARACTERISTICS RELATING FREQUENCY AND OPERATING ANODE VOLTAGE TO POWER OUTPUT

H. W. Welch, Jr.
(University of Michigan, Ann Arbor, Mich.)

An approximate method has been developed for determining the shape and density of the spokes of electronic space charge when large rf potentials exist in the magnetron. With the estimation of space-charge configuration which this method makes possible, induced current theory and resonant circuit analysis can be applied to calculate frequency-power and operating anode potential-power characteristics. Relatively simple equations for these characteristics are presented. Characteristics for typical values of the variables of magnetron design are given. The theory has been applied to experimental data with reasonably good argument.

140. A NEW PULSE KLYSTRON AMPLIFIER FOR THE 960–1,215 MC REGION OF AIR NAVIGATION AIDS

C. Veronda
(Sperry Gyroscope Company, Great Neck, N. Y.)

A pulse klystron amplifier, capable of delivering over 20 kilowatts of peak power with a gain of over 200 times, is described. The design concepts and the salient points of the development are discussed.

This three-resonator cascade amplifier was designed to be used as the final stage of the ground transponder in the Civil Aeronautics Authority's Distance Measuring Equipment. The klystron is "space-charge" focused, and is designed for use with a duty cycle which is variable from zero up to 1%. Because of the unusual beam requirements in this tube, the bunching voltage required is comparatively high. Therefore, the measurements on this tube are of special interest.

141. UHF POWER TUBES

P. T. Smith
(Radio Corporation of America, Princeton, N. J.)

The relative performance of structurally equivalent triodes and tetrodes have been determined with wide-band circuits at uhf. Most of the data was obtained with 5 kw peak power, silver-soldered ceramic envelope insulated triodes with grounded grid circuits. The performance of the triodes shows good agreement with large signal theory. At 900 mc the transit time loading of the grid, with normal tetrode operation, exceeds that of the grounded grid triode when the cathode-to-grid spacing is 0.015 inch and the grid-to-screen is 0.035 inch. With 5 kw output the performance of the triode at 900 mc is comparable with that at low frequencies, merely requiring large electron emission densities.

142. HIGH-FREQUENCY PERFORMANCE OF ELECTRON MULTIPLIERS

R. R. Law, D. A. Jenny and F. H. Norman
(Radio Corporation of America, Princeton, N. J.)

Design features and performance data of a developmental power electron multiplier intended to give several kilowatts output at frequencies up to 900 megacycles will be described. The poor performance of this tube at higher frequencies led to a critical study of other developmental multi-stage electrostatic electron multipliers. These tests will also be described. To explain the results it is suggested that secondary electron emission exhibits a simple exponential decay characteristic with a time constant of about 3×10^{-10} second. The engineering aspects of such a fundamental limitation to the high-frequency performance of practical electron multipliers will be discussed.

143. FACTORS AFFECTING THE LIFE OF HYDROGEN THYRATRONS

M. R. Zinn
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

The life of a hydrogen thyratron in a line-type modulator circuit is a function of the various circuit parameters in the grid and plate circuits. With the grid circuit conditions maintained constant, it is found that the average life of the tubes under various conditions of the plate circuit can be correlated with three dissipation factors: the power dissipated in the plate during the pulse, the power dissipated in the plate during the inverse voltage, and the power dissipated in the cathode due to the passage of pulse current. These factors have been successfully correlated with the results of life tests. The results of the analysis indicate the circuit parameters to be controlled by the application engineer and the tube characteristics of interest to the tube manufacturer.

Radar and Radio Navigation

Chairman, H. Busignies
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

144. DESIGN OF SMALL RADAR LINE-TYPE MODULATORS WITH AC CHARGING CIRCUITS

J. F. Clayton and S. J. Krulikoski, Jr.
(Bendix Aviation Corporation, Detroit, Mich.)

The use of a dc driven inductor-alternator and hydrogen thyratron switch permits the construction of compact ac charging pulse modulators with repetition frequencies up to several thousand pulses per second. The high output impedance of the inductor-alternator is used as the charging inductance with a low leakage reactance step-up transformer. The hydrogen thyratron is automatically phased and triggered from a non-linear coil in the charging circuit, no tubes being required. The dc drive on the inductor-alternator may result in some variation in the frequency of the ac voltage. Curves are presented showing the charging voltage step-up ratio as a function of circuit Q and applied frequency for the cases of one and two cycle charging. Over-voltage protection by means of circuit tuning is discussed briefly.

145. A HIGH QUALITY PICTURE DISPLAY UNIT

R. T. Petruzzelli
(Allen B. DuMont Laboratories, Inc., Clifton, N. J.)

The units described in this paper are high quality television monitors developed primarily for use in an electronic display system, providing bright television pictures of airport surveillance radar and mapping information. The equipment provides a simultaneous display of the above information on a bright display console at a brightness level sufficient for operational use in an airport traffic control tower.

The unit incorporates highly linear sweeps with good over-all focus and a highlight brightness of 60 foot-lamberts. The video amplifier has a bandwidth of 8 megacycles, with very good low-frequency response, provided by the use of keyed diode clamp circuits.

146. ANALYSIS OF AN AUTOMATIC RADAR RANGE-TRACKING SYSTEM

E. F. Grant
(W. L. Maxson Corporation, New York, N. Y.)

An analysis is made of a conventional automatic radar range-tracking system to determine the relationship between indicated range jitter due to noise and the system parameters and signal-to-noise ratio. In particular, it is shown for an intermediate-frequency amplifier with a gaussian frequency characteristic (approached by a synchronous-tuned or linear phase-shift amplifier) that there is a bandwidth which will minimize the jitter due to noise for a fixed signal power.

The method used is that of tracing the power spectrum of white thermal noise through the IF amplifier, detector, integrators, boxcar generator and servo amplifier (closing the loop), and finally integrating the power spectrum of indicated range for a fixed target to find the mean-square error in indicated position.

147. THE WIND-FINDING RADAR SYSTEM

A. D. Emurian
(Signal Corps Engineering Laboratories, Belmar, N. J.)

The Wind-Finding Radar System comprises the 3 cm Automatic Tracking Radar Set AN/CPS-10(XE-1) with its automatic wind computer and the windborne balloon-target train.

This paper outlines the evolution of the system into its present form; describes the basic features which fit it uniquely to the task of charting upper air winds with meaningful accuracy, quickly and cheaply; lists its essential technical and operational characteristics and relates them to the governing objectives which dictated its development.

Mention is made of present development to increase the system's wind-altitude capability and accuracy, to simplify it, and to balance the system package.

148. POWER REQUIREMENTS FOR LONG-RANGE NARROW-BAND NAVIGATION SYSTEMS IN THE LOW-FREQUENCY BANDS

N. Marchand, A. Jacobs, D. Cawood
(Sylvania Electric Products, Inc., Bayside, N. Y.)

Calculations of radiated power requirements for various distances are made for the frequency range 0.1 to 2.0 mc. Generalized curves are given for various signal-to-noise ratios, using nighttime noise. Correction factors for the different latitude zones are included. The curves are drawn for one cps bandwidth, but since the power will be directly proportional to the bandwidth, they are transferable to any usable system. Ground constants are taken from $\sigma=2\times10^{-14}$ to $\sigma=5\times10^{-11}$ emu, with ϵ varying from 10 to 80. The equations used for these curves are derived from the surface-wave equations given by K. A. Norton. Ground-to-ground transmissions were used. Curves for height-gain factors beyond line-of-sight are given to allow transformation of the charts for use at various heights beyond line-of-sight.

SYMPOSIUM

Magnetic Core Memory Devices for Digital Computers

(Organized by Professional Group on Electronic Computers)

Chairman, C. V. L. Smith
(Office of Naval Research, Washington, D. C.)

149. AN ANALYSIS OF MAGNETIC DELAY LINE OPERATION

E. A. Sands

(New York, N. Y.)

Since the first publication of the Harvard Computation Laboratory Progress Reports, there has been considerable interest in the use of magnetic components in digital computers and business machine systems. Particular interest has centered around the use of magnetic delay lines as a word-storage medium in a large scale, low access time internal memory. The difficulty in the design of magnetic delay lines (and magnetic components under current pulse conditions, in general) has been in representing the input impedance of a magnetic core by a simple analytical expression. In this paper, a simple equivalent input impedance of a magnetic core under current pulse conditions is derived, and this equivalent impedance is used in the analysis of the operation of a magnetic delay line.

150. DESIGN OF A HIGH-SPEED SHIFT REGISTER USING MAGNETIC BINARIES

Max Fishman

(Transducer Corporation, Boston, Mass.)

The availability of highly oriented nickel-iron alloys has resulted in a greatly increased interest in the application of magnetic materials in static memory devices. However, for high-speed applications, the use of highly oriented magnetic alloys has created many problems. At the expense of slightly more complicated circuitry, it is often possible to use non-square-loop material in the memory device and thus effect a large increase in speed of operation. An example of such circuitry is described for an application in a high-speed shift register which can be designed around magnetic ferrite cores or ultra-thin magnetic tape-wound cores.

151. MAGNETIC MATRIX SWITCHES

K. H. Olsen

(Massachusetts Institute of Technology, Cambridge, Mass.)

A matrix switch using ferromagnetic cores as nonlinear elements is a straightforward solution to many pulse switching problems that are difficult or impossible with a crystal matrix switch. Besides having arbitrary input and output impedances, it is often more efficient and useful than its crystal counterpart because power is supplied only to the selected output. The magnetic core is inexpensive, rugged, and apparently reliable for an indefinitely long life.

A pair of these switches is being developed for use with a multi-dimensional magnetic-ferrite memory where they will significantly decrease the required number of vacuum tubes and crystal diodes.

152. STATIC MAGNETIC MATRIX MEMORY AND SWITCHING CIRCUITS

J. A. Rajchman

(Radio Corporation of America, Princeton, N. J.)

Information bits are stored in terms of the direction of magnetization of a multitude of saturated cores connected in a matrix array. Access to any core, for registry or interrogation, is by simultaneous excitation of its defining matrix lines. Bi-valued signals, identifying the information bit and corresponding core, select these lines by activating magnetic switches, likewise composed of saturable cores. This memory is characterized by an access time of several microseconds, indefinitely long storage requiring no holding power, and the possibility of large storage capacity at low cost. Results with experimental models of 256 cores will be discussed.

153. THE FERRO-RESONANT FLIP-FLOP

Carl Isborn

(Computer Research Corporation, Hawthorne, Calif.)

The bi-stable nature of the ferro-resonant circuit has been utilized to produce flip-flops which count at rates of 100,000 pulses per second. These flip-flops are composed of reactive elements and, therefore, consume little power. At the same time they are amplifiers and, therefore, are capable of delivering considerable sustained power into a load; thus, enabling them to drive indicator lights, diode or resistor matrices, operate relays, etc.

Two inputs and two outputs are provided in these flip-flops. Thus, either single-input carry-type binary counters, or double-input parallel gated counters may be constructed.

Antennas II—Microwave A

Chairman, A. G. Fox

(Bell Telephone Laboratories, Inc., Holmdel, N. J.)

154. GAIN OF ELECTROMAGNETIC HORNS

E. H. Braun

(Naval Research Laboratory, Washington, D. C.)

Recent experimental evidence indicates that the measured gain of pyramidal electromagnetic horns may be considerably in error if the measurements are carried out at short distances, and the aperture to aperture separation between horns is used in the gain formula: $G=(4\pi R/\lambda)\sqrt{P_R/P_T}$.

Further experimental verification of this effect has been obtained, and a theory has been developed which is in good quantitative agreement with present experimental data and demonstrates the physical reasons why the previous "far field" criterion of $2D^2/\lambda$ is invalid.

Pending further experimental confirmation of the theory, curves will be calculated from which the error in gain measured at any distance may be obtained and applied as a correction.

155. A RAPID-SCAN CIRCULARLY SYMMETRICAL PILLBOX ANTENNA

Walter Rotman

(Air Force Cambridge Research Center, Cambridge, Mass.)

An X-band antenna capable of high speed scanning through 360° by rotation of a small central hub, with a beam narrow in azimuth and shaped in elevation, is described. Construction is of the waveguide-fed, double-layer pillbox type, but with a circular radiating aperture instead of the customary linear one. Spherical aberration of the antenna is markedly reduced by a circularly symmetric dielectric lens and/or multiple waveguide feeds. The elevation pattern is adjustable to some extent by surface-wave techniques making use of metal-backed dielectric or corrugations. Continuous scan as well as simultaneous scan in several directions is possible. The antenna may be either flush-mounted or enclosed within a radome. Constructional details and applications are discussed.

156. METHOD FOR SIDE-LOBE REDUCTION

C. J. Sletten

(Air Force Cambridge Research Center, Cambridge, Mass.)

A method is shown for compounding primary feeds in the focal region of microwave lenses and reflectors in order to modify the secondary pattern of high-gain systems. The basic idea is to examine the field structure produced in the focal region of a lens or reflector when it is illuminated by a plane wave, and to tailor the primary feed to this field structure. The main object has been to reduce side-lobe levels while preserving as much as possible of the gain and beam width.

Experimental models used to produce side lobe lower than 20 db down on a spherical reflector are described. The method of beam shaping described here can be used to advantage in a wide variety of antenna applications.

157. TOLERANCES ON PARABOLOIDAL REFLECTORS

John Ruze

(Massachusetts Institute of Technology and Air Force Cambridge Research Center, Cambridge, Mass.)

The effect of mechanical deviations from a parabolic surface on the antenna gain and on the side-lobe level are analyzed. It is assumed that these mechanical errors are randomly distributed over the reflector surface with a known or measurable mean-square error. The far field radiation pattern is expressed as the transform of the auto-correlation function of the aperture distribution. Analysis indicates that on a statistical average the undistorted pattern is modified by the addition of spurious radiation whose angular distribution depends on the error correlation interval, and whose magnitude is proportional to the mean-square error and to the square of the correlated interval. Graphs will indicate the expected side-lobe level and the resulting reduction in gain.

158. DESIGN OF DIELECTRIC WALLS FOR OPTIMUM TRANSMISSION

R. M. Redheffer

(University of California, Los Angeles, Calif.)

and

B. Golvin

(Hughes Aircraft Company, Culver City, Calif.)

Optimum design of a dielectric wall is discussed in terms of minimizing the number of parameters whose values must be specified arbitrarily. The case of the symmetrical three-layer sandwich wall is discussed rather completely. Such factors as transmission and reflection, dipolarization, and incidence-angle bandwidth are considered. We obtain expressions for optimum values of linear dimensions and dielectric constants for the cases where (a) there is only a single polarization, and (b) the dipolarizing effect of the dielectric wall must be considered. Tolerances are discussed in terms of transmission (or reflection) and phase contours; analytical and graphical means are given by which tolerances may be estimated. A set of design curves is given for sandwich walls, covering the range of dielectric constants thought to be physically realizable.

SYMPOSIUM UHF Receivers II

(Organized by Professional Group on Broadcast and Television Receivers)

Chairman, D. D. Israel
(Emerson Radio and Phonograph Corporation, New York, N. Y.

159. PRACTICAL TV ANTENNAS FOR UHF RECEPTION

E. O. Johnson
(RCA Victor Division, Camden, N. J.)
AND
J. D. Callaghan
(RCA Service Company, Inc., Camden, N. J.)

Requirements for the reception of television signals on the ultra-high-frequency, or uhf, band (470–890 mc) are much the same in many respects as on the existing very-high-frequency, or vhf band (54–216 mc). For the more difficult fringe areas, or locations where reflections are severe, special types of antennas are needed, just as they are in vhf. Of the wide variety of special uhf antennas designed and tested by RCA engineers and RCA Service Company technicians during field tests in Washington and Stratford, near Bridgeport, Conn., from 1948 to the present, several types have proved so outstanding in their simplicity, economy, and performance, that it is felt they will find additional widespread use where maximum performance and reliability are the customer's primary considerations.

160. AMPLIFIERS FOR UHF DISTRIBUTION SYSTEMS

T. Murakami
(RCA Victor Division, Camden, N. J.)

A theoretical and experimental study has been made of the use of single channel grid-separation amplifiers in ultra-high-frequency television distribution systems. The theoretical gain and noise figure curves for several currently available tubes are shown with the corresponding experimental curves. The over-all noise figure of an amplifier followed by a receiver or another amplifier is calculated for various gains and noise figures for the first amplifier. Curves of over-all gain and noise figure of a distribution system consisting of a single tube feeding a multiple number of tubes is shown, assuming equal distribution of power to each of the driven tubes. An experimental distribution amplifier using lumped circuit constants and coupled circuits is described.

161. COMPARISON OF PRESENT-DAY UHF AND VHF TELEVISION RECEIVERS

R. A. Varone
(Admiral Corporation, Chicago, Ill.)

A comparison of present-day television receivers will be made from the point of view of their noise figure. These figures will be interpreted in more commonly used laboratory terminology, and resulting differences in reception will be discussed. Present limitations of uhf receivers and the probable need of new electronic tools will be indicated.

162. ROUND-TABLE DISCUSSION: RELATIVE ASPECTS OF THE VARIOUS METHODS OF UHF TUNING

Moderator, Lewis Winner
(*Television Engineering*, New York, N. Y.)
Introductory Remarks, W. B. Whalley
(Sylvania Electric Products Inc., Bayside, N. Y.)

Feedback Control

Chairman, G. S. Brown
(Massachusetts Institute of Technology, Cambridge, Mass.)

163. STABILITY THEOREMS FOR FEEDBACK SYSTEMS

J. F. Koenig
(National Bureau of Standards, Washington, D. C.)

Eight stability theorems which give new aspects of design of linear feedback systems will be presented. Five enter a new design method called the "root-trajectory" method which gives information not obtainable from any other method of feedback system synthesis. The root-trajectory method gives the trajectories of the roots of any nth order characteristic equation as two or three interesting parameters are varied simultaneously in any arbitrary manner.

164. STABILIZATION OF NONLINEAR FEEDBACK CONTROL SYSTEMS

R. L. Cosgriff
(Ohio State University Research Foundation, Columbus, Ohio)

Often the desired static relationship between output and input of a system is a nonlinear function. Corresponding function generators can be made quite accurate by feedback; however, such systems cannot be stabilized in the same manner as is common for linear systems. A method of stabilization using perturbation techniques has been developed which effectively linearizes these nonlinear systems. This method is an extension to the conventional methods used in the analysis of linear systems. Other nonlinear systems, which cannot be stabilized by linear methods, can frequently be stabilized by the method developed.

165. RATE-LIMITED CONTROL SYSTEM NOISE

I. H. Van Horn and R. G. Wilson
(Goodyear Aircraft Corporation, Akron, Ohio)

The effects of noise in rate-limited stabilization systems have been studied using analytical, topological, and simulation techniques. Simulation has been found to yield results agreeing very satisfactorily with those obtained by other means. Nonlinear techniques used in the simulation will be presented. This study shows the importance of considering the effects of noise in the design of rate-limited closed-loop systems.

166. EXPERIMENTAL STUDIES ON SERVOMECHANISMS

A. V. Cohee
(Navy Department, Indianapolis, Ind.)

Laboratory equipment for the study of instrument servomechanism response and measurements on servos having common ailments will be discussed. Nonlinear effects, of little importance in high-power servos, can cause serious errors in some cases. Equipment to be described was designed for accurate evaluation of high-speed, high-accuracy instrument servos under dynamic conditions. Examples of nonlinear effects include errors at creep velocities, errors due to nonlinear error signals, and double-valued frequency response.

167. AFC SYSTEM ANALYSIS BY ELECTROMECHANICAL ANALOGUE

D. Leed
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

In measurement set applications and in communication systems, frequent use is made of automatic frequency-control devices. The operation of commonly employed circuits for automatic frequency control will be analyzed from the viewpoint of the (electrical frequency: angular velocity) mechanical analogue. Concepts of lock-in range, the effect of detector time constant and of strains imposed on the frequency control loop will be discussed.

Electron Tubes II—Small High-Frequency Tubes

Chairman, E. F. Carter
(Sylvania Electric Products Inc., New York, N. Y.)

168. A HIGH-GAIN KLYSTRON AMPLIFIER FOR RELAY SYSTEMS

G. Bernstein
(Sperry Gyroscope Company, Great Neck, N. Y.)

A 3-cavity klystron amplifier (SAC-41) operating at 4,000 mc has been developed to meet the needs of microwave relay work. The tube will provide 10 watts of output power at an anode potential of 700 volts and has a power gain of 30 db. It was developed to overcome the fading problem in microwave relay link transmission. It is tunable over the common-carrier band of 3,700–4,200 mc and features an ion-focussed high-current beam which makes possible a low operating voltage. Performance is centered so that there is less than a 1 db drop in power output at the ends of the tuning range. The tube is adaptable to high-gain wide-band operation as either a power amplifier or a high-level mixer.

169. FM DISTORTION IN REFLEX KLYSTRONS

Theodore Moreno and R. L. Jepsen
(Varian Associates, San Carlos, Calif.)

A theoretical and experimental study of the distortion properties of reflex klystrons will be reported. This study is particularly concerned with FM application of these tubes. These distortion properties are of particular importance when reflex klystrons are used in FM relay or transmitter service. A novel experimental method has been devised to measure the absolute value of FM distortion. These measurements agree very well with the theoretical calculations. Design techniques for reflex klystrons of minimum distortion will be discussed.

170. THE MEASUREMENT OF CATHODE INTERFACE IMPEDANCE

H. B. Frost
(Massachusetts Institute of Technology, Cambridge, Mass.)

Cathode interface impedance may be encountered in vacuum tubes after several thousand hours of life when alloys containing considerable silicon are used for cathode base material. This impedance may be measured by a new method which gives greatly improved accuracy and which uses a transconductance bridge in conjunction with a resistance-inductance network whose impedance complements the equivalent resistance-capacitance network of the interface impedance. At least two time constants (0.1 to 0.5 microsecond and 0.5 to 3.0 microseconds, respectively) are associated simultaneously with the interface impedance. An associated potential barrier between 1 and 2 electron volts has been measured.

171. UHF AMPLIFIER TUBE FOR TELEVISION TUNERS

C. E. Horton and Hsiung Hsu
(General Electric Company, Owensboro, Ky.)

A nine-pin miniature grounded-grid amplifier tube has been developed to operate as the radio-frequency amplifier in uhf television tuners. Satisfactory gain, bandwidth, and noise characteristics and good isolation between input and output circuits are obtained throughout the 470–890 megacycle band.

This paper treats the special problems encountered in this development and discusses the methods employed in arriving at a useful solution. The techniques of the measurements involved in the development of this tube are discussed in detail.

172. MICROWAVE CONVERSION AND DETECTION EMPLOYING ELECTRON TUBES

A. B. Bronwell, J. May, and I. C. Nitz
(Northwestern University, Evanston, Ill.)

Experiments on vacuum tubes as detectors and converters of modulated microwave signals have shown that these have low noise levels and greater sensitivity than crystal detectors. The principle of operation is different from that of the conventional operation of vacuum tubes. In conventional tubes, the electron transit time is a small fraction of the period of the ac wave. The tubes to be described here have large transit times, often exceeding several times the period of the ac wave. Also, the microwave signal is radiated into the interelectrode space, instead of being applied to the tube terminals.

Solutions of the resulting electron dynamic equation will be presented in graphical form for several typical cases.

SYMPOSIUM

The Integration of Electronic Equipment with Airframe Design

(Organized by Professional Group on Airborne Electronics)

Chairman, L. B. Hallman, Jr.
(Wright Air Development Center, Dayton, Ohio)

173. THE INTEGRATION OF ELECTRONIC EQUIPMENT WITH AIRFRAME DESIGN

A. F. Coombs and C. W. Dix
(General Electric Company, Syracuse, N. Y.)

Progress in co-ordinating the design of airborne radar equipment with the airframe is discussed, emphasizing such basic design characteristics as accessibility, form factor, heat dissipation, altitude, shock, vibration, rf noise, and primary power supply. A review of the scope of present-day installation problems follows, giving specific examples of co-ordinated design. Suggestions for improving current design practices include an early exchange of information among all agencies concerned, close liaison as the design unfolds, and throughout, an appreciative understanding of each other's problems.

174. THE UNIQUE AIRPLANE ENVIRONMENT EFFECT ON ELECTRONIC EQUIPMENT

D. T. Geiser
(Boeing Airplane Company, Wichita, Kan.)

Effects of altitude, temperature, and humidity are reviewed as individual problems, and the results are compared to typical high-performance airplane environment. Transitions of environment are shown as important as static environment and require special design care. Some design precautions are discussed. Brief environment transition testing is shown useful as a design tool.

175. ELECTRONIC COMPONENTS FOR AIRBORNE REQUIREMENTS

F. E. Wenger
(Wright Air Development Center Dayton, Ohio)

This paper describes the environmental space factor, and reliability requirements of airborne electronic component parts. The parameters for these requirements are described and the need for them substantiated. Approaches being used to solve some of the problems are described, as well as the general trends along which component development should be directed. The purpose of this paper is to present the importance of the availability of adequate component parts for airborne electronic equipments, and the philosophy upon which their future development should be based.

176. HEAT DISSIPATION FROM AIRBORNE ELECTRONIC EQUIPMENT

Louis Possner
(Hughes Aircraft Company, Culver City, Calif.)

This paper will present a brief discussion of the over-all problem, the reasons for the present interest, and review of research in this field. Design criteria for the dissipation of heat from electronic equipment will be presented, weight and power requirements of cooling systems will be discussed, and various methods will be compared. The advantages and disadvantages of high operating temperatures will be given, and need for higher operating efficiencies will be shown.

Digital Computers

Chairman, J. G. Brainerd
(University of Pennsylvania, Philadelphia, Pa.)

177. THE CADAC

W. E. Dobbins
(Computer Research Corporation, Hawthorne, Calif.)

Recently a small electronic digital computer, known as CADAC, or CRC-102, has been constructed for the Air Force. The CADAC is a universal type computer, uses a three-address code, and operates in a binary fixed-point number system. The memory is a magnetic drum, with space for 1,024 words. Physically, the machine occupies a space of 2½ feet by 4½ feet, and is 5 feet high. Its weight is 500 pounds, mostly in the power supply. The small size and reduced number of components (180 tubes and 2,500 diodes) are the unique features of this computer.

178. ANALYSIS OF CONTROL SYSTEMS INVOLVING DIGITAL COMPUTERS

W. K. Linvill
(Massachusetts Institute of Technology, Cambridge, Mass.)

A digital computer operates on sampled signals. To analyze the operation of a computer in a control system which operates largely on continuous signals, one should describe sampling, desampling, and computer operation in familiar control-system terms.

Sampling is analogous to impulse modulation. The whole mixed system viewed in the frequency domain is analogous to a system having one part operating on direct signals and another part operating on suppressed-carrier amplitude-modulated signals. Desampling is like ripple filtering in demodulation, and linear computer operation can be characterized by transfer functions. Frequency analysis allows evaluation of the interaction between the computer and the rest of the system and intelligent adaptation of the computer into the system.

179. FREQUENCY ANALYSIS OF DIGITAL COMPUTERS USED IN CONTROL SYSTEMS

J. M. Salzer
(Hughes Aircraft Company, Culver City, Calif.)

This paper discusses the analysis and synthesis of linear real-time digital-computer programs in the frequency domain. Such programs correspond to linear difference equations, and can be characterized in the frequency domain by a transfer function, which is rational in e^{-sT} (where e is the Naperian base of logarithm, s the complex frequency variable, and T the constant time interval of sampling). This contrasts with linear analog filters, whose transfer functions are rational in s.

Conventional techniques of frequency analysis are adaptable to digital filters: the amplitude, phase and locus of the program are defined, and stability can be studied in the complex plane. Synthesis of programs becomes as systematic as that of networks, and the method finds use in the design of computers, analog-digital systems, as well as numerical processes.

180. A VERY RAPID ACCESS MEMORY USING DIODES AND CAPACITORS

A. W. Holt
(National Bureau of Standards, Washington, D. C.)

An electrostatic memory for computers is described which utilizes the principle of regeneration to store binary information upon discrete capacitors, access being through two diodes. It seems possible to have fractional microsecond access for reading any digit in the matrix. Power efficiency is superior to other forms of electrostatic memory, and is mainly limited by present characteristics of germanium diodes. Emphasis is placed on the fact that only two-terminal devices are used in the memory proper, thus allowing promising design flexibility and minimum maintenance.

181. THE CHARACTRON

J. T. McNaney
(Consolidated Vultee Aircraft Corporation, San Diego, Calif.)

The Charactron is a special-purpose cathode-ray tube incorporating a design which is unique among tubes of this type. A matrix containing character-shaped openings is located between the electron gun and the fluorescent screen. A stream of electrons directed through the matrix openings results in a shaped beam that provides a presentation of characters on the screen of the tube where they can be read or photographed.

Among the more general applications of the Charactron are: (1) data conversion and tabulation of analog or digital information, (2) computer readout, (3) high-speed printing, (4) high-speed communications, and (5) monitoring and message display equipments.

Antennas III—Microwave B

Chairman, P. H. Smith
(Bell Telephone Laboratories, Inc., Whippany, N. J.)

182. A MICROWAVE LUNEBERG LENS

G. D. M. Peeler, D. H. Archer, K. S. Kelleher
(Naval Research Laboratory, Washington, D. C.)

A two-dimensional microwave model of the Luneberg lens has been designed employing the TE_{10} mode. It consists of two 36-inch-diameter, almost-parallel, conducting plates; the spacing between plates is filled with polystyrene and varies with the radius r to give the desired index of refraction $n=\sqrt{2-r^2}$. Due to symmetry about the axis, this lens has radiation patterns with constant gain and good side-lobe level as a feed horn scans over the circumference. Experimental patterns in the two principal planes show good agreement with computed patterns.

183. RADIATION FROM METAL-LOADED WAVEGUIDES TERMINATED IN A GROUND PLANE

R. E. Webster and M. H. Cohen
(Ohio State University Research Foundation, Columbus, Ohio)

Radiation from small apertures in a ground plane is considered. Measurements have been made on apertures excited by metal-loaded guides suitable for radiating circular polarization. Dielectric loading and combination metal-dielectric loading were also considered as schemes for reducing the cutoff frequency of the exciting waveguides. Parameters affecting bandwidth and aperture reflections are discussed, and experimental techniques for obtaining the effective aperture impedance are described. A method of calculating the aperture impedance from the guide dimensions for certain loading configurations is also presented.

184. MUTUAL COUPLING BETWEEN SLOT RADIATORS

M. J. Ehrlich, C. W. Curtis, and R. Fawcett
(Hughes Aircraft Company, Culver City, Calif.)

In the design of slot arrays with critical radiation patterns, mutual coupling between radiators is an important quantity. Application of Babinet's principle to P. S. Carter's relationships of the self and mutual impedance of parallel dipoles, and normalization of the data with respect to the feed waveguide, furnishes theoretical results. The self and mutual admittances of the two slots located on an infinite ground plane are measured as a function of slot separation and orientation. The theoretical and experimental values are in excellent agreement within the experimental error.

In addition, the coupling between two longitudinal shunt slots, displaced axially on the broad face of a rectangular waveguide, has been measured. The coupling is found to be a negligible magnitude as compared to variations due to manufacturing tolerances.

185. OFF-AXIS CHARACTERISTICS OF PARABOLOIDS AND SPHERES

K. S. Kelleher
(Naval Research Laboratory, Washington, D. C.)

Information is presented on the radiation patterns of paraboloids and spheres fed by a point source. A series of paraboloidal reflectors of various focal lengths, each 30 inches in aperture diameter, were evaluated at a wavelength of 3.2 cm. For each reflector an investigation was made of the patterns at various positions of feed horns in front of the reflector. Data was obtained on the gain, beamwidth, and side-lobe level of the radiation patterns as a function of aperture illumination and f/D ratio. Other quantities evaluated included beam shift as a function of feed displacement and f/D ratio. A similar type of information was obtained from a series of spherical cap reflectors of various radii.

186. A BROAD-BAND AXIALLY SYMMETRIC VERTEX FEED

F. L. Hennessey
(Naval Research Laboratory, Washington, D. C.)

A vertex feed, designed to illuminate a parboloidal reflector antenna at microwave frequencies, is discussed. Certain advantages over vertex feeds presently in use are pointed out. A small splash plate of special geometry, placed at the end of a circular waveguide extending through the vertex of the reflector, directs the energy back onto the reflector and provides a match to space of VSWR<1.5 over at least a twenty per cent frequency band. The complete axial symmetry of the feed permits the use of either linear or circular polarization and provides mechanical advantages in narrow-angle rapid-scanning systems.

Radio Communication Systems

Chairman, W. M. Goodall
(Bell Telephone Laboratories, Inc., Deal, N. J.)

187. A RADIO RELAY SYSTEM EMPLOYING A 4,000-MC THREE-CAVITY KLYSTRON AMPLIFIER

J. J. Lenehan
(Western Union Telegraph Company, New York, N. Y.)

This paper discusses the application of this tube as an amplifier in a relay system already in operation when the amplifier became available. The reasons for using the tube, the design necessary to incorporate it into existing circuitry, and its performance characteristics are described. The practical problems of tube alignment, life, and maintenance as encountered in system operation are discussed.

188. AN FM MICROWAVE RADIO RELAY

R. E. Lacy and C. E. Sharp
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

The design features of an 8,000–8,500 mc radio relay are reviewed. The innovations described are the result of research and engineering accomplished for the design of a military radio relay system.

A mechanically and electronically tuned cw communications magnetron is included which provides a carrier power in excess of 50 watts, capable of being frequency modulated. A unique frequency stabilization circuit maintains the carrier center frequency, improves the linearity of the modulation, and greatly reduces the carrier-noise frequency variations by virtue of the inverse feedback introduced.

A novel duplexing antenna system, comprised of a waveguide hybrid tee, a waveguide mast structure, and off-center fed parabolic reflector antenna assembly, and a waveguide cavity preselector for the receiver are described.

189. NONSYNCHRONOUS PULSE MULTIPLEX SYSTEM WITH RANDOM SAMPLING

J. R. Pierce and A. L. Hopper
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

This system uses the same frequency and approximately the same repetition rate for each transmitter. Pulse groups carrying samples from a given transmitter are "tagged" for identification at the receiver. Interferences between transmitters are reduced by sampling at random times. When pulses overlap, the receiver is disabled and the sample is lost. Noise due to lost samples is minimized by holding one sample until another is received.

Such a system allows an unlimited number of channel assignments in a given frequency band although only a limited number can be used simultaneously.

An experimental two-channel system was built and tested.

190. EXALTED-CARRIER AND SINGLE-SIDEBAND DIVERSITY RECEIVERS

M. G. Crosby
(Crosby Laboratories, Mineola, N. Y.)

Recent developments are described which have resulted in two simplified receiving systems for long range communication. The first is a contact double-sideband exalted-carrier detector unit which may be connected to an ordinary communications-type receiver. The second is a single-sideband adapter unit useable in the same manner. The advantages of exalted-carrier detection and the problems involved in the change from double-sideband to single-sideband techniques are discussed. The requirements of a diversity combining system are outlined and a new type of exalted-carrier or single-sideband diversity combiner is described which provides optimum performance.

191. COUNTER CIRCUIT FOR A BROAD-BAND MULTIPLEX RECEIVER

A. R. Vallarino, H. A. Snow, and C. Greenwald
(Federal Telecommunication Laboratories, Nutley, N. J.)

Counter receivers, operating at one megacycle with a modulating frequency exceeding 150 kilocycles, were developed for frequency-division multiplex subcarrier systems. Harmonic distortion of less than 0.2 per cent is not critically dependent on the values of the passive (resistors and capacitors, only) and active elements. There are no tuning adjustments.

Each limiter stage employs a double-cutoff cathode-coupled triode to generate square waves having excellent symmetry between positive and negative halves and rise times shorter than 0.04 microsecond.

The counter-discriminator is a cathode-follower variation of previously used counters, and permits relatively small output tubes to be employed.

Circuits V

Chairman, H. L. Krauss
(Yale University, New Haven, Conn.)

192. ANALYSIS OF MEASUREMENTS ON MAGNETIC FERRITES

C. D. Owens
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Ferrites are unique nonmetallic materials with new combinations of magnetic, electric, and dielectric properties. For this reason, methods commonly used for measuring and expressing core loss coefficients in magnetic cores are not very satisfactory. Resonance type effects associated with frequency and dimensions are also present in ferrites. Measurements of permeability, magnetic Q, and dielectric properties versus frequency, temperature, and dimensions are discussed. The need for standardization of measurements and magnetic data for the convenient use of design engineers is brought out. The product μQ of the magnetic material is shown to be conveniently related to the quality factors of inductance coils and transformers. The μQ product measured on a closed magnetic core remains essentially constant when discrete air gaps are inserted for lowering permeability, and therefore is a useful and practical parameter for evaluating ferrite core materials to be used in assembled cores containing gaps.

193. MAGNETIC AMPLIFIER PERFORMANCE ANALYSIS

D. Lebell and B. Bussell
(University of California, Los Angeles, Calif.)

The analysis of magnetic amplifier circuits is facilitated and extended by application of the differential analyzer computer. Effects of magnetic hysteresis, rectifier leakage, and reactor nonlinearities have been studied for the parallel connected self-feedback amplifier.

As output data, the computer indicates the average value of load current and plots waveforms of current and flux. This data shows the shift in transfer characteristics due to hysteresis, the decrease in gain (slope or transfer curve) due to rectifier leakage, and the "rounding" of the transfer curve due to curvature of the magnetic characteristic. Results point to corrections which can be applied to improve the simplified analysis or design by compensating for these effects.

194. BARIUM TITANATE PROPERTIES

A. I. Dranetz
(Gulton Manufacturing Company, Metuchen, N. J.)

Useful relative dielectric constants up to 6,000 can be realized by barium titanate ceramics, and the materials may be made to have a nonlinear dielectric constant, now being investigated for circuits such as modulators and dielectric amplifiers. The nonlinear materials exhibit a remanent polarization and can be used as the basis of various new memory devices.

The titanates also have piezoelectric characteristics. Instruments such as high-frequency accelerometers, ultrasonic transducers, microphones, phonograph pickups, hydrophones, underwater projectors and displacement gauges are using these materials, and components such as acoustic delay lines are in development.

195. A FERROELECTRIC AMPLIFIER

H. Urkowitz
(Philco Corporation, Philadelphia, Pa.)

"Ferroelectric" refers to nonlinear dielectrics characterized by a charge-versus-voltage relationship exhibiting hysteresis and dielectric saturation. A capacitor using barium-strontium titanate ceramic was used in a single-tuned circuit to which a high-frequency current was applied. A low-frequency signal was applied to the capacitor. The resulting amplitude-modulated high-frequency voltage was detected. Analysis shows conditions for maximum amplification. Power gains of about 60 were obtained.

With only the high-frequency voltage applied, the output of the amplifier was fed back to the input, and sustained oscillations of low frequency were obtained. Frequency of oscillation could be varied continuously over a wide range.

196. GERMANIUM DIODE TRANSIENT RESPONSE

J. H. Wright
(National Bureau of Standards, Washington, D. C.)

Whisker-contact germanium diodes are valuable as fast switching devices because of their high forward conductance and low capacities. Since inherent limitations on switching speeds are not well known, experiments were made to study recovery times after 0.03 microsecond switching transients from: (a) forward to back, (b) back to forward, (c) neutral to back, and (d) neutral to forward conduction. Quite serious effects occur for (a) and (b), which are the important cases for computer and other switching applications. Conductances vary by factors of 3.0 to 10.0 times normal, with 0.5 to 1.0 microsecond for 90 per cent recovery. The mechanism will be discussed and transient tests recommended.

197. GERMANIUM DIODE TESTING PROGRAM

D. J. Crawford and H. F. Heath
(International Business Machines Corporation, Poughkeepsie, N. Y.)

This paper will present a review of an extensive germanium diode testing program. A discussion of the various type-approval and production tests is given along with the means for effecting them. A result of this program is the formalizing of criteria for a better computer diode.

198. AN ANALYSIS OF CRYSTAL DIODES IN THE MILLIVOLT REGION

W. B. Whalley, N. P. Salz, and C. Masucci
(Sylvania Electric Products Inc., Bayside, N. Y.)

Crystal diodes, when operated as detectors for signal inputs of the order of a few millivolts, present new problems regarding their exact behavior. Methods of rapidly testing the crystal performance in this region have been devised and the equipment required for the measurements has been designed and constructed.

A theoretical analysis will be presented of a type of crystal application which requires the detection of very small differences in large magnitude voltages. Data will be given for a number of types of crystal diodes tested for use in this application.

Electron Tubes III—Cathode-Ray Tubes

Chairman, R. R. Law

(Radio Corporation of America, Princeton, N. J.)

199. THE ANATOMY OF CONTRAST RANGE IN CATHODE-RAY TUBES

J. H. Haines

(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

AND

R. E. Mueller

(formerly Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

In metallized cathode-ray tubes the contrast range is degraded principally by two effects; normal reflections and halation. The contrast is controlled by three variables: faceplate transmission, faceplate reflection, and optical contact of the phosphor to the faceplate.

Detail contrast is defined as the brightness ratio between a large illuminated field and a small unexcited area in this field. Measurements on two flying-spot scanner tubes, identical except for faceplate transmission, showed only 10 to 1 detail contrast for a clear faceplate, but 100 to 1 for a 66 per cent transmission faceplate.

It is shown the detail contrast is half that measured at edge of a uniform raster. Simple measurement techniques are outlined.

200. THE SELFOCUS TUBE

A. Y. Bentley, K. A. Hoagland, and H. W. Grossbohlin

(Allen B. DuMont Laboratories, Inc., Clifton, N. J.)

By making use of an electron lens of improved gometry, it has been possible to mass-produce television cathode-ray tubes in which focus is maintained without the necessity of external magnetic fields or focus voltage sources. The advantages of the new lens structure as compared to conventional electrostatic focus lenses are discussed. The performance characteristics of a typical Selfocus picture tube are described and compared to the widely used magnetic-focus prototype. It is shown that equivalent, and in some cases, superior performance is attained with the Selfocus tube.

201. A NEW HIGH-SPEED CATHODE-RAY TUBE

H. J. Peake and R. W. Rochelle

(Naval Research Laboratory, Washington, D. C.)

A cathode-ray tube, DuMont Type K-1056, has been developed to fulfill the need for displaying transient waveforms which require spot velocities up to thousands of inches per microsecond. The development of the tube is discussed and the characteristics of the production model are given. A "figure of merit" is introduced which allows evaluation of the performance of a cathode-ray tube in a transient display application.

202. THE DEFLECTRON—A NEW SYSTEM FOR ELECTROSTATIC DEFLECTION

Kurt Schlesinger

(Motorola, Inc., Chicago, Ill.)

The conventional method of electrostatic deflection is limited to angles of less than 30° and to beams of low intensity, because of its inherent pattern and focus distortion and small aperture.

This paper presents a new approach to the problem of wide-angle electrostatic deflection. Deflector electrodes of a composite type have been developed, which yield biaxial deflection simultaneously, i.e., with a common center of deflection. Two forms of these "deflectron" units are presented; one circular and one rectangular. The circular deflectron lends itself to radar and oscilloscope applications, while the rectangular deflectron was developed for picture tubes. "Deflectrons" are made of glass, using photographic methods. The lecture will describe the tubes, their associated circuits, and include a demonstration of a deflectron in operation.

203. FIELD PLOTTING AS A TOOL IN DEFLECTION YOKE DESIGN

E. Sieminski

(Sylvania Electric Products, Inc., Bayside, N. Y.)

Deflection yoke design is predominantly empirical in practice. Mathematical analysis loses its utility because assumptions must be made in considering the many factors which affect a useful result. This paper describes a study of magnetic deflection field distributions, emphasizing their value in promoting a more precise design of deflection yokes. Field plots of windings in various configurations are compared for analysis. Measurement procedures are described and equipment, built for this investigation, is illustrated and treated in detail. The technique is particularly of value in that it helps to bridge the gap between the complexity of a theoretical approach and the inadequacy of an empirical solution.

SYMPOSIUM

What's New in Mobile Radio

(Organized by Professional Group on Vehicular Communications)

Chairman, Austin Bailey

(American Telephone and Telegraph Company, New York, N. Y.)

204. MOBILE RADIO PROBLEMS RESULTING FROM NEW TECHNIQUES

E. L. White

(Federal Communications Commission, Washington, D. C.)

The development of equipment giving satisfactory performance on frequencies above 30 mc initiated a phenomenal growth in the mobile service. This growth has been further accelerated by the development of equipment for operation above 100 mc. As a result, in many services and in many frequency bands the numbers of stations has increased to an extent that congestion is serious. Relief must be found. The development of mobile equipment for the bands above 400 mc offers the opportunity for the transfer of a considerable portion of the communications load from the lower frequencies. Developments of new circuits and components permitting closer spacing between channels will result in more economic use of the frequency bands available for mobile use. Better circuit discipline and more efficient operating procedures will permit the transmission of a greater volume of intelligence during available circuit time. The degree to which mobile radio can serve our economy depends to a great extent upon success in achieving maximum frequency utilization, and is a challenge to the designer, the manufacturer, and the user.

205. APPLICATION OF VOICE-FREQUENCY TONE SIGNALLING TO MOBILE RADIO SYSTEMS

C. L. Roualt

(General Electric Company, Syracuse, N. Y.)

Certain theoretical concepts relative to feedback-type selective amplifiers have been reduced to practice, resulting in two basic elements useful in a signalling system. These are a tone generator and a selective amplifier. The tone generator possesses excellent stability and the selective amplifier permits ready attainment of Q's of 200 in the voice frequency band 300–3,000 cycles per second. These basic elements have been employed in a number of signaling systems for signalling from a base station to individual or groups of mobile units and from the mobile units to the base station. The essentially electronic nature of these systems contributes to exceptional speed of operation and excellent reliability.

206. DISPATCHER'S WAYSIDE-TO-TRAIN RADIO CONTROL SYSTEM

S. D. Burton

(Bendix Radio Division, Baltimore, Md.)

Control of the movement of trains over a railroad division and how direct radio communication between a dispatcher and train crews can greatly expedite train operation with safety will be considered. The original requirements set forth by the Northern Pacific Railroad for a complete dual-channel automatic train communication system and how these requirements were resolved engineering-wise to a series of unitized equipments which can be grouped to fulfill the system requirements of any railroad will be outlined. The paper will close with a step-by-step description of the operation of a typical system.

207. NEW DEVELOPMENTS IN ARMY MOBILE COMMUNICATION EQUIPMENT

J. H. Durrer

(Coles Signal Laboratory, Ft. Monmouth, N. J.)

As a result of an intensive post war development program by the Signal Corps Engineering Laboratories in conjunction with private industry, large scale production of an entirely new series of forward combat area mobile and portable communication equipments has begun. The features of these new equipments, including ease of channel changing, crystal saving, flexibility, interchangeability, remote control, retransmission facilities, and selectivity and stability characteristics will be discussed.

SYMPOSIUM

Reliability of Military Electronic Equipment

(Organized by Professional Group on Quality Control)

Chairman, J. R. STEEN

(Sylvania Electric Products Inc., New York, N. Y.)

208. DISCUSSION OF THE COMPLEXITY AND UNRELIABILITY OF MILITARY EQUIPMENT AND THE NEED FOR SIMPLIFICATION AND INCREASED LIFE

A. S. BROWN

(Stanford Research Institute, Stanford, Calif.)

The paper is a discussion and criticism of the complexity, unreliability, and short life of military electronic equipment. It is a proposal to add to the usual military characteristics, requirements for maximum of simplicity, longevity, reliability, and usability as a goal for engineers. Examples of World War II equipment, its maintenance and operation during that time, are given and its status during the subsequent years. Some suggestions are offered for maximum improvement in the future using transistors, ruggedized tubes, and other devices. Recommendations are made for streamlining research and development in the Armed Forces and urging of earlier co-operative decisions.

209. MAINTENANCE MINIMIZATION IN LARGE ELECTRONIC SYSTEMS

W. D. MCGUIGAN

(Stanford Research Institute, Stanford, Calif.)

Some of the effects of unitized packaging on reliability and economy in electronic systems are discussed. Examples are given both of circuit designs and panel layouts which tend to simplify the maintenance and recognition of performance. The factors governing the division of a system into packages and the design of a fault-location system are outlined. The paper concludes with an outline of some of the system factors requiring further investigation in the quest for reliability.

210. THE RELIABILITY PROBLEMS IN MISSILE DEVELOPMENT

A. C. PACKARD AND R. WELLER

(U. S. Naval Air Missile Test Center, Point Mugu, Calif.)

The efforts to develop a satisfactory guided missile have forced us to operate in an environment which taxes the ingenuity of man and about which relatively little is known. Operating a new power plant in a newly designed vehicle at heretofore unattained speeds with a complex guidance and control system never before attempted has placed tremendous demands on equipment designers and technique developers.

211. APPLICATION ENGINEERING FOR IMPROVED ELECTRONIC RELIABILITY IN GUIDED MISSILES

W. T. SUMERLIN

(Philco Corporation, Philadelphia, Pa.)

The means for transplanting standardized, accepted electronic equipment and systems from their customary environments into guided missiles merits special consideration in the realm of application engineering. Of most general interest, perhaps, is a review of such, tailored to fit the viewpoint of an electronic manufacturer who faces his first production of some typical guided missile. Indocrinating this manufacturer's organization is considered, and items such as circuit design, component selection, fabrication techniques, testing facilities, and supervision are discussed in the light of high reliability in the special environment applicable. Examples are given.

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WOMEN'S ACTIVITIES COMMITTEE

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Mrs. P. B. Harkins, *Vice-Chairman and Treasurer*

Mrs. S. L. Bailey, Mrs. Roger McSweeny, Mrs. I. S. Coggeshall, Mrs. Ogden Prestholdt, Mrs. R. V. Gould, Mrs. Carl Scholz, Mrs. D. B. Sinclair

Contributors to Proceedings of the I.R.E.

Fred Assadourian was born on April 13, 1915, in Panderma, Turkey. New York University conferred the B.S. degree on him in 1935, the M.S. in 1936, and the Ph.D. degree in mathematics in 1940.

Fred Assadourian

From 1937 to 1942, Dr. Assadourian instructed in mathematics at New York University and, from 1942 to 1944, he was an associate professor of mathematics at Texas Technological College.

Engaged as a research engineer at the Westinghouse Research Laboratories from 1944 to 1946, Dr. Assadourian worked on pulse transformers. Since 1946, he has been a development engineer at Federal Telecommunication Laboratories, where he is doing theoretical work in electronics and communication.

Dr. Assadourian is a member of the American Mathematical Society and of Phi Beta Kappa.

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William S. Bachman was born on October 29, 1908, at Williamsport, Pennsylvania. He is a graduate of Tower Hill School in Wilmington, Delaware, and received a degree in electrical engineering from Cornell University in 1932.

W. S. Bachman

In 1934, Mr. Bachman joined the Radio Receiver Engineering Department of the General Electric Company where he worked on loudspeakers, FM and AM radio receivers, and phonograph combinations. In 1946 he joined Columbia Records, Inc., where he is currently Director of Engineering and Development.

Mr. Bachman received the Charles A. Coffin award for work on feedback amplifiers, and is also the designer of the G.E. Variable Reluctance Phonograph Reproducer. He contributed to the development of the Long Playing Microgroove record, and was responsible for getting it into commercial production.

Mr. Bachman is a member of Eta Kappa Nu and of Tau Beta Phi.

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Gail E. Boggs was born in Chicago, Illinois on April 22, 1921. He attended the Illinois Institute of Technology in 1942, and was graduated from the George Washington University with the degree of B.E.E. in 1948. At present, he is enrolled in the Graduate School of the University of Maryland.

From 1941 to 1942, Mr. Boggs was employed by the Belmont Radio Corporation of Chicago. He entered the Armed Forces in 1943, and was assigned to the Communications Laboratory of the Office of Strategic Services in 1944, serving as design eingineer on military communications equipment until the end of 1945.

Gail E. Boggs

Since 1948, he has been employed by the National Bureau of Standards in the Experimental Ionospheric Research Section of CRPL in Washington, D. C. During this period, his work has involved the investigation of gain stability problems encountered in field intensity recording.

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Edwin L. Chinnock was born in Brooklyn, New York on August 25, 1916. After attending Stevens Institute of Technology, he was employed by The Electrical Industries Manufacturing Co. In 1939 Mr. Chinnock joined the radio research department of the Bell Telephone Laboratories at Holmdel, N. J. He was called to active duty with the U. S. Navy in 1940, and was assigned to a classified group which received a special unit citation for their work during the war.

E. L. Chinnock

After being discharged as a Chief Radioman in 1945, Mr. Chinnock returned to the radio research department of the Bell Telephone Laboratories at Holmdel, N. J., where he has done research on microwave relay systems, the close-spaced triode, and the measurement of noise figures.

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Frank G. Cole (S'48–A'51) was born in Newton, Mass. on April 30, 1927. He received the B.S. and M.S. degrees in electrical engineering from the Massachusetts Institute of Technology, the latter in 1950. He attended the cooperative course in conjunction with the Philco Corporation.

Frank G. Cole

Mr. Cole was a Radio Technician in the Navy from 1945 to 1946. He has worked for the General Electric Receiver Department from 1950 to the present on TV sweep systems engineering, and primarily on the developmemt of the wattmeter. His present assignment in the Government Section is with Airborne radar.

Mr. Cole is a member of Eta Kappa Nu.

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For a photograph and biography of J. M. Diamond, see page 438 of the April, 1951, issue of the Proceedings of the I.R.E.

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J. J. Ebers (S'46–A'48) was born in Grand Rapids, Mich., on November 25, 1921. He received the B.S. degree from Antioch College in 1946, his education having been interrupted by three years' service in the U. S. Army. He obtained the M.S. degree in electrical engineering from Ohio State University in 1947, and the Ph.D. in 1950. Since 1947 he has been an instructor in the electrical engineering department of this university, as well as a research associate for The Ohio State University Research Foundation. Recently he has received an appointment as assistant professor.

J. J. Ebers

Dr. Ebers is a member of Eta Kappa Nu, Sigma Xi, and the American Physical Society.

❖

Donald E. Garrett (S'48–A'49) was born on April 24, 1922, in McLoud, Okla. He attended the University of Kansas from 1939 to 1941, and received the B.S. degree in electrical engineering from the University of Washington in 1948 after two years, spent there. In 1950, he received the S.M. degree in electrical engineering from the Massachusetts Institute of Technology.

D. E. Garrett

From 1941 to 1944, Mr. Garrett worked for the Boeing Airplane Company in Seattle, Wash., on control, armament, and miscellaneous electrical problems on the B-17 Flying Fortress, the C-97 Strato Freighter, the B-50, and the Stratocruiser. After two years (1944 to 1946) as an electronic technician in the Navy, he rejoined the Boeing Company, and from 1948 until 1950, he worked on guided missiles for the Research Laboratory of Electronics at M.I.T.

Since 1950, Mr. Garrett has worked on TV sweep circuit, high voltage, and meas-

Contributors to Proceedings of the I.R.E.

urement problems for the General Electric Receiver Department. His present assignment is color television.

Mr. Garrett is a member of Sigma Xi and RESA.

Raymond F. Guy (A'25–M'31–F'39) was born in Hartford, Conn., on July 4, 1899. In 1916 he entered radio professionally with the Marconi Wireless Telegraph Company, and during World War I he served overseas with the Signal Corps of the United States Army. Upon being discharged, he entered Pratt Institute, from which he graduated with an electrical engineering degree in 1921.


RAYMOND F. GUY

In the same year Mr. Guy was engaged as a broadcast engineer for WJZ. In 1924 he became a member of the engineering staff of the RCA Research Laboratories, where he supervised engineering, development, and construction of standard and short-wave broadcasting apparatus, stations, and systems, and participated in RCA's earliest television development.

In 1929 Mr. Guy transferred to the National Broadcasting Company to direct its frequency allocations engineering, and the planning, design, and construction of all NBC transmitting facilities. He is now Manager of Radio and Allocations Engineering for NBC.

Mr. Guy has been active in Institute affairs for over twenty years: he was a Director starting in 1943, Treasurer in 1947, and President in 1950, and is currently Senior Past President and Membership Relations Co-ordinator. He is a charter member of the Radio Pioneers, a life member of the Veterans' Wireless Operator's Association, a Fellow of the Radio Club of America, a member of the Radio Executives Club, and an associate of the Association of Federal Communications Engineers.

Arthur A. Hauser, Jr. (M'48) was born in Dayton, Ohio, on January 28, 1920. He received his B.S. degree from the Massachusetts Institute of Technology in 1942, and his M.S. degree from New York University in 1950. From 1941 to 1942, Mr. Hauser was an assistant in the Physics Department at the Massachusetts Institute of Technology. He joined the Sperry Gyroscope Company in 1942 and remained with them until 1946, during which period he was successively an assistant project engineer and project engineer. In 1946 he became an instructor in the Department of Mathematics at the Rensselaer Polytechnic Institute. In 1947 Mr. Hauser returned to Sperry, where he served successively as project engineer, senior project engineer, and research engineer.


A. A. HAUSER, JR.

At present, Mr. Hauser is engineering section head for Electronics at the Sperry Gyroscope Company.

Robert S. Hoff (A'46–M'50) was born in Delavan, Ill., on March 12, 1920. He received the B.S. degree in electrical engineering from the Texas A. and M. College in 1941 and the M.S. degree in engineering from the University of Florida in 1950.


ROBERT S. HOFF

From 1941 to 1946 Mr. Hoff served as an officer with the Signal Corps, in which capacity he was engaged in staff work pertaining to radar installation and maintenance and to radio intelligence training, organization, and equipment development guidance. In his last year of service as Organization and Training Officer, in the grade of Major, of an agency of the Army Communications Service, Mr. Hoff was awarded the Legion of Merit.

After leaving the Army, Mr. Hoff joined the Engineering and Industrial Experiment Station, University of Florida, as leader of an Air Force-sponsored low-frequency atmospheric noise and wave-propagation project. He is a member of Commission IV, Terrestrial Noise, of the American Section of URSI, and has presented papers at technical meetings of that organization.

Mr. Hoff joined the Ordnance Development Division of the National Bureau of Standards in 1950, and is presently an engineer in charge of a development project.

C. W. Horton was born on September 23, 1915, at Clerryvale, Kan. He received the B.A. degree with honors in physics in 1935, and the M.A. degree in 1936, both from The Rice Institute, in Houston, Tex. In 1945 he was awarded the Ph.D. degree from the University of Texas.


C. W. HORTON

Dr. Horton was a research associate at the Underwater Sound Laboratory, Harvard University, from 1943 to 1945. For this work he received a Development Award from the U. S. Navy Bureau of Ordnance, and a Certificate of Appreciation from the Office of Scientific Research and Development.

Since 1945 he has been a research physicist at the Defence Research Laboratory and an associate professor of physics at the University of Texas, Austin, Tex. He is a member of the American Physical Society, the American Geophysical Union, and the Society of Exploration Geophysicists.

Craig C. Johnson was born in San Marcos, Texas, on February 27, 1924. He entered the University of Texas in 1941 and graduated in 1948 with a B.S. degree in mechanical engineering, having spent three years, 1943 through 1945, as an Air Force pilot.


CRAIG C. JOHNSON

After graduation, Mr. Johnson joined the Defense Research Laboratory at the University of Texas as a research engineer engaged in guided missile work for the United States Navy. Since June, 1951 he has been employed as a research engineer at the North American Aerophysics Laboratory in Downey, Calif.

While at college Mr. Johnson was a student member of the American Society of Mechanical Engineers. He is a member of Phi Eta Sigma, Pi Tau Sigma, and Tau Beta Pi. In 1947 he was awarded a Westinghouse Achievement Scholarship.

Raymond C. Johnson (A'46–M'49) was born in Galveston, Texas on September 29, 1922, and was educated at Texas A. and M. College, where he received the B.S. degree in electrical engineering in 1946, and at the University of Florida, where he took the M.S. degree in electrical engineering in 1949. He completed the Army training course at the Long Lines Inside Plant and the Army electronics training course at Harvard and the Massachusetts Institute of Technology.


R. C. JOHNSON

As an assistant research professor at the University of Florida, Mr. Johnson is currently devoting his time to classified research in electronics for the Federal government. Prior to 1946, when he joined the staff of the College of Engineering, he had two years of Army radar experience with the Signal Corps Engineering Laboratories,

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where he worked on new radar equipment. This included one year in the European and Pacific war theaters.

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

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Hugh LeCaine was born in Port Arthur, Ontario, Canada, on May 27, 1914. He received the B.Sc. degree in engineering physics from Queen's University, Kingston, Ontario, in 1938, and the M.Sc. degree in physics from the same University in 1939. The following year was spent in research work at Queen's University on a National Research Council Studentship. In 1940 he joined the staff of the National Research Council of Canada, where he now holds the position of assistant research officer

Hugh LeCaine

From 1948 to date, Mr. LeCaine has been on leave-of-absence, and has been pursuing graduate studies at the University of Birmingham, England, on a National Research Council Fellowship.

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Chester M. McKinney (S'43–A'45–M'49) was born on January 29, 1920, in Cooper, Texas. He received the B.S. degree in physics from the East Texas State Teachers College in 1941, followed by the M.A. in 1947 and the Ph.D. in 1950 in phydisc from the University of Texas. Mr. McKinney served as a radar officer in the Air Force from 1942 to 1946. He was a research physicist at Defense Research Laboratory of The University of Texas from 1946 to 1950, and an assistant professor of physics at Texas Technological College from September, 1951, to the present.

C. M. McKinney

Mr. McKinney was recalled to active duty in the Air Force, in June, 1951.

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Horst A. Poehler (S'41–A'44) was born in Gera, Germany on October 9, 1917. He received the B.E.E. degree from the Polytechnic Institute of Brooklyn in 1939, pursued one year of graduate study at the Moore School of the University of Pennsylvania, 1939–1940, received the M.A. degree in Pure Science from Columbia University in 1942, and received the Ph.D. degree from Columbia in 1948.

Horst A. Poehler

From 1942 to 1944 he was associated with the Westinghouse Electric Corporation, Bloomfield, New Jersey, where he worked on vacuum tube problems. From 1944 through 1946, he held the position of project engineer at the International Electronics Laboratories, New York, and was concerned with instrumentation and control problems in industry. In 1946 he was awarded a graduate fellowship and returned to Columbia University to resume graduate studies.

Since 1948 Dr. Poehler has been a staff member with the General Precision Laboratory, Inc. at Pleasantville, New York. His work here is in the field of electro-mechanical measuring and computing circuits.

Dr. Poehler is a member of the American Institute of Electrical Engineers, Eta Kappa Nu, Pi Mu Epsilon, and Tau Beta Pi.

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Henry J. Riblet (A'45) was born on July 21, 1913 at Calgary, Canada. He received the B.S. degree in 1935 and the Ph.D. degree in 1939 from Yale University.

Henry J. Riblet

From 1939 to 1942, Dr. Riblet taught mathematics, first as instructor at Adelphi College, and later as assistant professor at Hofstra College. From 1942 to 1945 he was at the Radiation Laboratory, where he was in charge of one of the three design sections of the antenna group. From 1945 to 1949 he was in charge of the antenna and rf groups at the Submarine Signal Company. He is now employed by the Microwave Development Laboratories.

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Walter J. Surtees (S'42–A'45–M'48) was born in Ottawa, Ontario, Canada, on January 1, 1922. He received the B.Sc. degree in Electrical Engineering in 1943 from Queen's University, followed by the M.A.Sc. degree in electrical engineering in 1947, from the University of Toronto.

From 1943 to 1946 Mr. Surtees served as signals officer with the Royal Canadian Air Force and from 1947 to 1949 he was an instructor in the Department of Electrical Engineering of the University of Toronto.

Walter J. Surtees

At present Mr. Surtees is a research assistant and graduate student at the University of Toronto. The research for his paper herein printed was conducted in the laboratories of the Department of Electrical Engineering at the University, under contract, in part, between it and the Defense Research Board.

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Richard Theile was born in Halle, Germany, on March 23, 1913. He attended the Realgymnasium and University in Marburg and the Technische Hochschule in Berlin, and received the degree of Ph.D. in 1938 at the University of Marburg. He joined the Telefunken Company in 1936. Since that time he has been engaged in research and the development of television pick-up devices, camera tubes, and photo-multipliers, as well as other projects important to this field. From 1946 Dr. Theile was a lecturer in electronics at the University of Marburg, and he is at present engaged in television research at the laboratories of Pye Limited, in Cambridge, England.

Richard Theile

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Frederick H. Townsend (SM–'49) was born in London, England on September 26, 1911. He received his technical education at the Northampton Polytechnic Institute, London. In 1930 he joined the Valve Department of Standard Telephones and Cables Ltd., and in 1931 went to A. C. Cossor Ltd., as an assistant in their Research Department, where he remained until 1938. From 1938 to 1946 he was second-in-charge of the Vacuum Laboratory of Pye Ltd., in Cambridge, England. In 1946 he became Chief Vacuum Engineer and Manager of Cathodeon Ltd., the vacuum tube subsidiary of the Pye organization.

F. H. Townsend

Mr. Townsend is also an Associate Member of the British organization, the Institution of Electrical Engineers, and is currently Chairman of the Cambridge Radio Group of the Institution.

Institute News and Radio Notes

Technical Committee Notes

The **Standards Committee** will hold a meeting during the 1952 IRE National Convention on Tuesday, March 4, at 8:00 a.m.

The Executive Committee, at its meeting on December 4, approved the change of names of the **Committee on Modulation Systems** to that of the **Committee on Information Theory and Modulation Systems** and the **Committee on Electron Tubes and Solid-State Devices** to that of the **Committee on Electron Devices.** At this meeting the appointment of A. G. Jensen as Chairman of the new United States Committee (of CCIR), to handle questions involved in the work assigned to Study Group XIV (Vocabulary) of the **Committee on International Radio Consultative**, was confirmed.

The **Standards Committee** convened on December 13, under the Chairmanship of A. G. Jensen. J. W. Horton's paper, "Fundamental Considerations Regarding the Use of Relative Magnitudes," was approved by the Committee and will appear in the PROCEEDINGS. E. A. Laport's idea for an Electronics Encyclopedia was endorsed by the Committee. The proposed "Standards on Electron Tubes: Methods of Testing Gas-Filled Radiation Counter Tubes," together with the "Gas-Filled Counter-Tube Definitions," which were approved at this meeting by the Standards Committee on November 8, will be sent back to the sponsoring Committee (**Electron Devices**) for their approval. A new ruling was adopted by the Standards Committee that, after being approved, all definitions and methods of testing would be sent back to the sponsoring Committee for its approval. If the Committee does not agree with the changes made by the Standards Committee, they may so advise the Committee. Revision of the Standards Committee Manual was also discussed.

A meeting of the **Committee on Antennas and Waveguides** was held on November 13, under the Chairmanship of A. G. Fox. A major part of the meeting was devoted to a discussion of "Q" definitions, both "of a medium" and "of a waveguide." Notes on the "Q" concept, as written by Mr. Fox, were used as starting points.

The **Facsimile Committee** convened on November 2, under the Chairmanship of R. J. Wise. Mr. Wise proposed the establishment of a Subcommittee to develop and present a suitable test pattern, and asked for ideas and suggestions which might be helpful to this group. It was suggested that means for measuring definitions be given particular emphasis. The subject was discussed at some length, and a number of suggestions were made as to methods of supplementing the test pattern or to alternatives which might provide a better indication of definitions.

The **Committee on Navigation Aids** convened on November 19, under the Chairmanship of P. C. Sandretto. General Sandretto reported, to the Committee Members present, on the action taken at the November 8 meeting of the Standards Committee, relative to the controversial term "Fruit Pulse." He reported that the Definition which the Committee on Navigation Aids has proposed had been accepted by the Standards Committee. The Committee then turned its attention to a review of navigation terms.

The **Committee on Measurements and Instrumentation** convened on November 16, under the Chairmanship of F. J. Gaffney. Three topics for standards work by the **Subcommittee on Video-Frequency Measurements** were proposed: (a) measurement of impedance at video frequencies, (b) video measurements employing transient techniques, and (c) measurements of time for video signals. It was requested that these proposals be reviewed and comments sent to the Chairman of the Subcommittee. Attention was called to the meeting on Electrical Insulation, which was held at the National Bureau of Standards on October 29–31, and sponsored by the National Research Council. A brief summary will be contained in the Annual Review. The Chairman reported that most of the Annual Review reports were now in, and the remaining ones have been promised shortly. A new method of measurement of the dielectric properties of gases, developed by the Bureau of Standards was reported on by J. L. Dalke.

The **Committee on Wave Propagation** convened on December 11, under the Chairmanship of H. G. Booker. The proposed, "Standards on Wave Propagation: Standards of Measuring," drawn up by Subcommittee 24.1, was considered, and approved. The document, "Tropospheric Propagation: A Selected Guide to the Literature," was also considered, and approved. Discussion was given to a document entitled, "Definitions of Terms Relating to Propagation in the Troposphere," which was adopted unanimously at the Plenary Session of the CCIR on July 3, 1951. The feeling of the Committee was that, while there was no violent reaction to any of the definitions concerned, nevertheless the document could not be considered as an improvement upon the IRE Tropospheric Definitions, published in the November, 1950, issue of the PROCEEDINGS. The Committee felt that the proper course was to consider the CCIR definitions in connection with the next revision of the IRE definitions and that no special action should be taken at the present time. It was reported that the Radio Astronomy Definitions, which were approved at the last meeting of the Committee, will be published in the near future.

At the December 7 meeting of the United States National Committee of the International Electrotechnical Commission, W. R. G. Baker, representing the RTMA, and F. B. Llewellyn, representing the IRE, were re-elected to their respective posts.

At the meeting of the Committee on Electrical Standards, of the American Standards Association, held on December 7, C. R. Harte and W. R. G. Baker were re-appointed Chairman and Vice-Chairman, respectively. Dr. Baker is the new Vice-Chairman of the Communications and Electronics Division of the Committee on Electrical Standards, and Harry Brown has relieved Sidney Withington as Vice-Chairman of the Power Division of the ESC. W. R. G. Baker and L. G. Cumming were appointed as members of the Executive Committee. Other members of the Executive Committee will be C. R. Harte, P. H. Chase, M. Brandon, J. J. Pilliod, R. C. Sogge, H. Brown, and Colonel Ice.

British Radio Show Slated

The ninth annual private exhibition of British components, valves, and test gear for the radio, electronic, and telecommunication industries is scheduled for April 7–9, 1952, at the Grosvenor House, London, Eng.

An exhibition is being promoted to show the advances in design and development of British instruments. A warm welcome is extended to those interested in this country.

Calendar of COMING EVENTS

1952 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6

IRE Connecticut Valley Section, Electronics Industry Day, Storrs, Conn., April 5

Annual British Radio Component Show, Grosvenor House, London, England, April 7–9

Radio and Television Show, Manchester, England, April 23–May 3

IRE Cincinnati Section, Spring Technical Conference, Cincinnati Engineering Societies Building, Cincinnati, Ohio, April 19

URSI-IRE Spring Meeting, National Bureau of Standards, Washington, D. C., April 21–24

IEE Television Convention, London, England, April 28–May 3

IRE-AIEE-RTMA Symposium on Progress in Quality Electronic Components, Washington, D. C., May 5–7

IRE New England Radio Engineering Meeting, Copley Plaza Hotel, Boston, Mass., May 10

IRE National Conference on Airborne Electronics, Hotel Biltmore, Dayton, Ohio, May 12–14

4th Southwestern IRE Conference and Radio Engineering Show, Rice Hotel, Houston, Tex., May 16–17

Radio Parts and Electronic Equipment Show, Conrad Hilton Hotel, Chicago, Ill., May 19–22

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29

National Electronics Conference, Chicago, Ill., September 29–October 1

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 27–29

Institute News and Radio Notes

Professional Group Notes

An invitation has been issued to all IRE Professional Groups to attend the 1952 Centennial of Engineering, to be held in Chicago, on September 3–13, 1952. All Groups interested in participating in this event should advise the IRE Headquarters as soon as possible.

The IRE Professional Groups Administrative Committees who have registered with Headquarters to hold meetings at the 1952 **IRE National Convention** are scheduled as follows: **Airborne Electronics,** Thursday, March 6, Moderne Room, Belmont Plaza Hotel; **Antennas and Propagation,** Wednesday, March 5, Jade Room, Waldorf Astoria Hotel; **Broadcast Transmission Systems,** Wednesday, March 5, Ballroom, Waldorf Astoria Hotel; **Circuit Theory,** Monday, March 3, Astor Gallery, Waldorf Astoria Hotel; **Electronic Computers,** Wednesday, March 5, Blue Room, Grand Central Palace; **Engineering Management,** Monday, March 3, Grand Ballroom, Waldorf Astoria Hotel; **Information Theory,** Monday, March 3, Jade Room, Waldorf Astoria Hotel; **Vehicular Communications,** Thursday, March 6, Maroon Room, Grand Central Palace. According to a policy set up by the IRE Headquarters, these meetings will be held on the mornings of the Convention days, to be adjourned no later than 9:30 a.m.

The **IRE Professional Group on Airborne** Electronics will hold a luncheon during the 1952 IRE National Convention, on March 5, in the Baroque Room of the Belmont Plaza Hotel, at 12 noon.

The **IRE Committee on Professional** Groups will hold a meeting during the 1952 IRE National Convention on the morning of Tuesday, March 4, in the Grand Ballroom of the Waldorf Astoria Hotel.

W. H. Doherty, Chairman of the **IRE Technical Program Committee** of the 1952 IRE National Convention, has announced that all IRE Professional Groups except **Nuclear Science** will be represented at the Convention by technical sessions or symposia.

The dates of the IRE National Conference on Airborne Electronics, to be held in Dayton and sponsored by the **IRE Professional Group on Airborne Electronics,** have been changed to May 12–14, 1952, in order to avoid conflict with the Symposium on Components to be held in Washington, D. C.

Dr. George Sinclair, Chairman of the **IRE Professional Group on Antennas and Propagation,** has announced the election of A. H. Waynick as Vice Chairman of the Group for the current year. The material for the first issue of this Groups publication "Transactions" will be available for distribution shortly.

The **IRE Professional Group on Audio,** under the Chairmanship of B. B. Bauer, has announced that beginning with the January, 1952, issue of the Audio Newsletter, the last page of the publication will be devoted to institutional advertisements by various companies in the industry. The price for a listing in 6 issues will be $25. The Group has recently made available for distribution to the membership, "Transactions of the Professional Group on Audio," comprising four papers as follows: "Horn Loaded Loudspeakers," by D. J. Plach and P. B. Williams; "Wear of Phonograph Needles," by B. B. Bauer; "A Selective Automatic Phonograph Mechanism," by J. C. Kiefer and A. G. Bodoh; and, "An Electronic Music Box," by E. L. Kent.

Under the Chairmanship of D. D. Israel, the **IRE Professional Group on Broadcast and Television Receivers,** announced the distribution to the entire membership of an assessment notice levied at a cost of $2.00 per member, to defray expenses incurred through publication of technical papers of interest to its members. Members were requested to return payment forms with check to the IRE Headquarters by January 15, 1952.

The petition for the formation of the **IRE Professional Group on Electronic Computers** has been approved by the IRE Executive Committee. Acting-Chairman M. M. Astrahan has announced the appointment of the Administrative Committee Members of the Group as follows: one-year terms—W. F. Gunning, H. H. Sarkissian, E. G. Andrews, and J. H. Pomerene; two-year terms—G. W. Downs, H. D. Huskey, J. H. Howard, and M. M. Astrahan; three-year terms—W. D. Caldwell, T. A. Rogers, J. R. Weiner, and S. N. Alexander.

IRE Headquarters has announced the approval of the **IRE Professional Group on Electron Devices** Constitution, by the IRE Executive Committee, on December 4, 1951. The Acting Chairman of the Group is G. D. O'Neill of Sylvania Electric Products Company, Bayside, N. Y.

IEE Convention Plans Advance

Further plans have been announced for the convention on "The British Contribution to Television," sponsored by the Institution of Electrical Engineers. The convention, to which all IRE members have been invited, is scheduled for April 28 to May 3, 1952, London, England.

The number of technical papers on which discussion will take place will be between 60 and 80, many of which have already been submitted. Technical sessions covering all aspects of television from the program production to the viewer will be supported by interesting demonstrations of television equipment, including large-screen projection television and, possibly, an early Baird 30-line equipment which is at present being reassembled.

During the convention, visits of inspection will include the BBC television studios and the latest television transmitters, the post office research station, and commercial organizations manufacturing television equipment of the London-Birmingham co-axial cable link and others.

Information concerning procurement of registration forms and additional program plans may be found on page 1467, of the November, 1951, issue of the Proceedings.

Notice! Members

The IRE Professional Group on Electronic Computers announces that Members who have not previously indicated interest can obtain application forms from the Membership Chairman, J. R. Weiner, Eckert-Mauchly Computer Corporation, 3747 Ridge Avenue, Philadelphia 32, Pa. The Group will sponsor a Symposium on Magnetic Core Memory Devices at the 1952 IRE National Convention. Information about the Group will be available at the Professional Group desk during the Convention.

URSI/IRE Spring Meeting Scheduled

A meeting of the USA National Committee of the International Scientific Radio Union (URSI), and the IRE Professional Group on Antennas and Propagation will be held at the National Bureau of Standards, Washington, D. C., on April 21–24, 1952.

Sessions will be held concerning the following topics: radio measurement methods and standards, tropospheric radio propagation, ionospheric radio propagation, terrestrial radio noise, radio astronomy, antennas and waveguides, radio waves and circuits (including general theory), and electronics (tubes and semiconductors).

A preliminary program and advance registration forms will be available after March 10, 1952. These and further information concerning the meetings may be obtained from A. H. Waynick, Secretary, U.S.A. National Committee of URSI, Pennsylvania State College, State College, Pa.

Over 900 Attendance at IRE-AIEE Conference

More than 900 engineers, scientists, and mathematicians attended the joint IRE-AIEE Computer Conference held in Philadelphia, December 10–12, 1951. The meeting afforded the first opportunity for manufacturers and users of large-scale digital computing equipment to exchange information on results obtained from completed machines. Ten computers were described in some detail including two machines located in Great Britain, with papers and discussions presented by the various representatives and authorities of the machines.

The papers and discussions which were contributed to the oral presentations will be collected and bound in a Proceedings which should be available soon. These written papers are more complete than the oral presentations. These Proceedings will be sold for $3.50 a copy and may be obtained from the AIEE or IRE Headquarters.

Information on the program schedule and machines discussed at the conference can be found on page 1466, in the November, 1951, issue of the Proceedings.

IRE People

Patrick E. Sullivan (M'48) has been appointed assistant manager of the Buffalo Tube Works of the General Electric Company. Mr. Sullivan has been works engineer at the Buffalo GE plant since 1947.

P. E. SULLIVAN

A native of Detroit, Mr. Sullivan graduated from the University of Detroit in 1942, with a B.S. degree in electrical engineering. He joined the General Electric Company in 1942 as a trainee in the test engineering program at Bridgeport, and then transferred to the Buffalo plant as a quality engineer. He was appointed assistant works engineer in 1946, and became a works engineer in 1947.

He is a member of the American Society for Quality Control.

❖

Raymond Collins (A'40) who was the assistant general manager of radio stations WFAA and WFAA-TV, Dallas, Texas, died recently at his home after an illness of several months.

Mr. Collins was born in 1907, and studied electrical engineering at the Southern Methodist University. He joined WFAA in 1928, becoming the technical supervisor in 1935. During World War II, he participated in radar development and research at the Radio Research Laboratory at Harvard University. In 1944, he was one of 16 engineers chosen to perfect courtermeasure radar equipment at the request of General Dwight D. Eisenhower. This work consisted of jamming the German radar network along the French coast before the invasion by the allied forces.

After the war, Mr. Collins returned to WFAA where he was in charge of technical operations until his death. Mr. Collins was a charter member of the Dallas IRE Section.

Carl E. Scholz (M'26–SM'43) has been elected vice president and chief engineer of the American Cable and Radio Corporation. Prior to this appointment Mr. Scholz has served in the same capacity for three operating subsidiaries of AC&R; the Mackay Radio and Telegraph Company, the Commercial Cable Company, and All America Cables and Radio, Incorporated.

Mr. Scholz has been associated with The International Telephone and Telegraph Corporation, an affiliate of AC&R, since 1917, when he joined the Federal Telegraph Company as an engineer. With the exception of a few years in South America, Mr. Scholz has spent most of his career at IT&T headquarters in New York, working on engineering and design problems for the various subsidiary companies. Mr. Scholz is also a member of the American Institute of Electrical Engineers.

❖

Ross Gunn (A'35–SM'47–F'49) has received the 1951 Air Safety Award from the Flight Safety Foundation, Incorporated. The award was made to Dr. Gunn in recognition of his successful research in the causes of precipitation static in aircraft radio communications and for the practical solution developed under his direction which has contributed to increased safety in the air.

ROSS GUNN

Dr. Gunn, who is the Director of Physical Research, United States Weather Bureau, Washington, D. C., has specialized in the invention and development of new electrical instruments and electronic devices, including early radio control apparatus for airborne missiles. In 1945, he was cited by the Secretary of the Navy for Distinguished Civilian Service, in connection with the development of the atomic bomb.

Dr. Gunn was born in Cleveland, Ohio, in 1897. He received the degrees of B.S.E.E., and M.S., in physics from the University of Michigan in 1920 and 1921, respectively, and the Ph.D. degree from Yale in 1926. He was a wireless operator on the Great Lakes during 1915–1917, and was a special instructor in radio courses at the University of Michigan in 1918. From 1920 to 1922, he was an instructor in physics at the University of Michigan, and then became a radio research engineer in the U. S. Air Service, in 1922–1923. As an instructor in physics at Yale University in 1923 to 1927, he was put in charge of the high-frequency laboratory during 1926–1927.

In 1927 he joined the Naval Research Laboratory and became their chief physicist where he worked as the superintendent of the mechanics and electricity division, superintendent of the aircraft electrical research division, and technical director of the Army-Navy Precipitation-Static Project.

Dr. Gunn was presented the IRE Fellow Award in 1949, for his, "long service and many technical contributions in the radio and electronics fields." He is a fellow of the American Physical Society, and a member of the National Academy of Sciences.

Ralph R. Shields (M'50), formerly engineer for Sylvania test equipment merchandising, has been appointed merchandising supervisor for the television picture tube division of Sylvania Electric Products, Incorporated.

RALPH R. SHIELDS

A native of Pennsylvania, Mr. Shields received his engineering degree in 1938. During World War II, he served as administrative engineer of the Special Studies Branch, Signal Corps Engineering Laboratories, where he supervised research and development projects which eliminated radio interference from army vehicles.

Mr. Shields has authored many engineering and business paper articles on the technical and economic aspects of television servicing and servicing instruments.

❖

Allen D. Cardwell (A'14–VA'39) inventor and founder of an electrical instrument manufacturing concern, died recently at the Nassau Hospital Mineola, L. I., N. Y. He was 63 years of age.

Mr. Cardwell, who retired in 1945, received early recognition as an inventor and designer in the printing telegraph field and for inventing a device to speed stock market tickers. A pioneer in the guided-missile field, he devised, in 1924, a system for such missiles that met the acceptance test of the British Government.

A holder of many patents in electronics, he was credited with having invented the first low-loss condenser to be manufactured in quantity. He was co-inventor with Ralph Batcher of an automatic calibration system used during World War II, for the calibration of high-frequency meters. He received a citation from the government for his work.

In retirement, Mr. Cardwell had been working on pilot models of physical therapy equipment to be used in the rehabilitation of disabled veterans. Mr. Cardwell was a native of Rochester, N. Y.

William Fingerle, Jr., (S'35–A'38–SM'47) has recently joined the Budelman Radio Corporation. In his new position he will be engaged in carrying out development contracts in the FM communications and multichannel relay fields.

Mr. Fingerle was born on April 9, 1914, in New York, N. Y. He studied electrical engineering at the Massachusetts Institute of Technology where he received the B.S. degree in 1936. He then joined the Duro-Test Lamp Company as assistant chief engineer. In 1937 he became associated with the Link Radio Corporation where he served in a capacity as head of the department of development, design, and manufacture of high power FM, emergency, and television services, and as chief engineer.

Mr. Fingerle was also active in the development and application of military radio equipment during World War II.

❖

William E. Osborne (A'41) has been elected President and General Manager of Resdel Engineering Company of Los Angeles. He was previously Director of Electronics at the Hycon Manufacturing Company of Pasadena, California, for several years, where he organized the Electronics Division.

W. E. Osborne

Mr. Osborne was engaged in electronic and radar work after receiving the E.E. degree from Queen's College, Melbourne University, Australia, in 1925. Now a United States citizen, he served with both the British and Australian Forces during the war, as head of the Radiophysics Branch of the Australian Army, and as radar liaison officer to the United States and British Governments.

In 1945–1947, he was associated with Gilfillan Brothers, Incorporated, of Los Angeles, as Principal Radar Design Engineer. Mr. Osborne holds a number of patents in the television, radar (GCA), infrared, and nucleonics fields.

❖

J. P. Coughlin (S'43–A'46) has been appointed as manager of aircraft and electronic transformer sales of the General Electric Company. Mr. Coughlin has been with General Electric since 1941, when he was graduated from Pratt Institute with a degree in electrical engineering. Following service on the test course and as a design engineer for distribution transformers, he transferred to the specialty transformer division. He later became the assistant manager of this division, a position he held until his recent appointment.

Books

Les Tubes Electroniques a Commande par Modulation de Vitesse by R. Warnecke and P. Guenard

Published (1951) by Gauthier-Villars, 55 Quai des Grands Augustins, Paris, France. 792 pages, +22 pages +2-page errata. 476 figures.

R. Warnecke is the technical director of the electronics department, and P. Guenard is chief of the laboratory, Center of Technical Research, Campagnie General de Telegraphie Sans Fil, Paris, France.

The scope of this book is adequately described in the authors' own foreword. They state that they have attempted to review all theoretical and technical literature on velocity modulation tubes that was available to them. As they both belong to the Laboratories of the Compagnie Generale de TSF, they have naturally based their work primarily on what has been accomplished there in this field, but they have also carefully studied a great number of other published or still unpublished material, with a particular mention of W. W. Hansen's and E. Feenberg's unpublished notes which have been graciously communicated to them. The bibliography included in the book contains 385 items carefully chosen to cover all important aspects of the problem.

The authors are careful to indicate that no complete deductive theory of any of the tubes will be found in their book. In fact, they claim that no such theory can be made, as restrictive assumptions have to be introduced to deal with practical cases and various facets of the problems involved in order to arrive at results applicable to actual structures. The approximations utilized are, however, very clearly mentioned everywhere and the book undoubtedly constitutes a basic exposition of all principles underlying the design of this type of uhf and vhf tubes which remains of great importance in microwave techniques, and to the development of which the authors have themselves contributed considerably. It will be recalled that Dr. Warnecke was given an IRE Fellow Award in 1950 for his engineering and research contributions to vacuum-tube theory and design in France. Dr. Guenard is also a well known French expert in the field and both have filed a number of patent applications and published extensively.

The book is divided into seven parts. The first deals with the fundamentals of electronic behavior in velocity modulated tubes treating such questions as bunching and debunching of electrons, and influence of transit time in the interaction processes between fields and electrons. The second covers the essential facts about cavity resonators and the influence of shape and gaps on their properties. In the third part are gathered the basic theoretical explanations of the various types, such as two and three cavity amplifiers, frequency multipliers, reflex klystrons, and velocity modulated oscillators. The fourth part describes a number of structures and gives very useful data on the important components, as well as details on the less conventional devices such as coaxial line oscillators and various others due to Heil, Ludi, Hahn and Metcalf, and Coeterier. The fifth part comes back on theoretical considerations, incorporating refinements and additional information with a view to approach correct design more closely. Here are found interesting chapters on how beams are focused, the way electrons behave in nonuniform fields, the influence of secondaries, space charge, electronic hysteresis, and relativity theory. Noise level and scaling problems are also discussed, together with the authors' method of designing a tube for specified characteristics together with specific examples. The sixth part is devoted to problems which arise in connection with circuitry associated with the tubes, load, matching, bandwidth, and power supplies. The final part is a review of what has been accomplished so far and what the authors suggest could be accomplished in the future to improve output, obtain higher frequencies, broader bandwidths, or reduce the noise level.

This book should indeed be of very great interest and value to all concerned with the development of velocity modulation tubes, and this has played, since the pioneering work of the Varian brothers, and continues to play, an important part in microwave techniques. It is clearly written, illustrated, indexed, and very well presented by one of the best French publishers of technical books. It is to be hoped that a translation will be made of this remarkable work into English, to make it available to a wider circle of students, engineers, and scientists specializing in this field or interested in related subjects.

A. G. Clavier
Federal Telecommunication Laboratories
Nutley, N. J.

Books

TV and Electronics As a Career by Ira Kamen and Richard H. Dorf

Published (1951) by John F. Rider Publisher, 480 Canal Street, New York 13, N. Y. 304 pages+6-page index+15-page appendix+x pages. 136 illustrations. 5¾×8¾. $4.95.

Ira Kamen is the Director of Electrons, Brach Manufacturing Corporation, Newark, N. J. Richard H. Dorf is an Audio and Television Consultant, New York, N. Y.

The scope of this new book includes detailed information of what must be known and learned in order to qualify for employment in almost every department of radio and electronics. In the main, the class of readers to which the book will appeal are high school and junior college students desiring authoritative information about what to study and how to go about securing a position in the wide field of radio and electronics. It is a readable work despite a considerable amount of repetition unavoidable in a book with five contributors.

The main divisions of the text include: selecting a career in TV and electronics; television broadcasting; AM and FM broadcasting and communications; radio and television manufacturing; electronic engineering; television servicing; and electronics in the armed services. The excellent illustrations in each of the chapters are helpful. The chapter on TV broadcasting, by J. R. Poppele, the experienced chief engineer of WOR-TV, might perhaps have appeared to better advantage as one of the later chapters, rather than as the second chapter of the book.

The need for guide books of this kind becomes evident with the authors' statement that the number of engineering graduates decreased by approximately fifty per cent between the years 1946 and 1950, and that a further decline is anticipated in the next two or three years. The situation would seem to be serious when it is declared that "the need for service personnel will outstrip any possible supply." One cause cited for the dearth of competent applicants for positions is "the high personal requirements." It may be that radio and TV can profit from the experience in training men of older arts such as telegraphy, telephony, and railroading. Writing about assistant chief engineers in broadcasting, Mr. Poppele states, "An engineering degree is not a necessity; indeed, few assistant chiefs hold them."

In obtaining the requisite knowledge to qualify for employment in important broadcast stations and studios, emphasis is placed upon the advice to get a start at a small station, anywhere, and to learn the techniques and routines on a small scale. As much time as possible should be devoted to studying radio and television textbooks and to reading current trade journals in the field.

This book is of value to those who seek information about the requirements of employment in radio and TV, but it also contains a considerable amount of solid information for men already established in any of the various departments.

DONALD MCNICOL
Communication Engineer
25 Beaver Street
New York, N. Y.

Radio and Television Receiver Circuitry and Operation by Alfred A. Ghirardi and J. Richard Johnson

Published (1951) by Rinehart Books, Inc., 232 Madison Avenue, New York 16, N. Y. 621 pages+19-page index+27-page glossary+xvi pages. 411 figures. 6×9. $6.00.

Alfred A. Ghirardi is a radio and electronic engineering consultant and a technical writer and editor. J. Richard Johnson is a technical editor at Rinehart Books, Inc., New York, N. Y.

This is a new and up-to-date book, written primarily for servicemen, but covering such a wide range of subjects that it would be a valuable reference book for a design engineer, since many engineers allow themselves to become so specialized in one phase or another of electronic development that they lose sight of what is occurring in related fields.

The subjects covered are as follows: Amplitude modulation and AM signals, frequency modulation and FM signals, rf amplifiers and trf receivers, am superheterodyne receivers, AM detector and avc systems, FM receivers, pushbutton tuning and afc systems, af amplifiers, loudspeakers, radio receiver power supply systems, tv principles and the tv receiver, receiving antenna systems, home recorders, phonopickups and record players, automatic record changers, and the mechanical construction of receivers.

The authors have done an excellent job of describing and explaining this subject matter while using an absolute minimum of mathematics.

There are rather obvious errors in two or three of the schematic diagrams (such things as blocking capacitors and grounds omitted) which would not ordinarily deserve comment if the book were not intended primarily for readers who might not realize that the grounds and blocking condensers should be there. Also, the section on oscillator tuning and tracking is somewhat misleading in that it gives the impression that all receivers use only two point tracking, while, as a matter of fact, practically all receivers with equal capacity-tuning gang sections are designed to track at three points, particularly in the am broadcast tuning range.

Aside from these minor points, the book should be of valuable assistance to servicemen, technicians, and those who wish to gain some knowledge of present-day radio and television without delving too deeply into the mathematics of the principles involved.

ALDEN PACKARD
American Radio and Television, Inc.
P.O. Box 328
North Little Rock, Arkansas

Fundamentals of Acoustics by Lawrence E. Kinsler and Austin R. Frey

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Avenue, New York 16, N. Y. 499 pages+5-page appendix+3-page glossary+5-page index+vii pages. 165 figures. 5¾×8¾. $6.00.

Lawrence E. Kinsler and Austin R. Frey are professors of physics, United States Naval Postgraduate School, Annapolis, Md.

This book is intended as a text for graduate students in physics or electrical engineering rather than as reference material only. Consequently, there is much attention given to the tutorial aspects of the presentation. The physical situations are given much more attention than are accorded purely analytical considerations; the numerous illustrations and problems are particularly useful.

The scope of the text covers in similar order the topics that Morse has in his more mathematical "Vibrations and Sound." However, Kinsler and Frey have placed greater emphasis on the mechanism of the physical picture, thus making it easier reading for the student whose background is primarily engineering. Although this point of view sometimes leads to minor lapses from rigor, the presentation, on the whole, is sufficiently complete and clear in its purpose. There are few repetitions, and no glaring inconsistencies.

One of the valuable features is a discussion of loudspeakers, microphones, and horns. It is much more thorough than Morse, and more basic than that found in Olson's "Elements of Acoustical Engineering." Nevertheless, the text fails to integrate its treatment with the realities of the design, construction, testing, performance, and use of these devices. Audio engineers will still have to rely on the journal literature for much of the information on these instruments.

A useful chapter on absorption in tubes, porous solids, and fluids is included in this book. Both impedance and propagation concepts are utilized. The section on psychoacoustics is conventional, with emphasis on the difference of loudness, loudness level, and intensity. Architectural acoustics is presented more or less as a summary of useful results from the work of Morse and Bolt. The reviewer would have preferred more attention given to the treatment of underwater acoustics since it appears too restricted in security and space considerations. However, the following section on ultrasonics does give much that is of additional interest in underwater sound. Tables of constants, functions, and symbols conclude the book.

Few errors or controversial statements are found in the book, and the terms used follow the recommendations of the ASA; however, the use of "magnetomôtive impedance" for the motional mechanical impedance seems misleading. An interchanged parenthesis and exponent of 2 on page 42 is an obvious error; the effect of standing waves in speaker cabinets causes less change in the radiation than is implied by the statements on page 288; and, finally, the term dbm could well have been introduced on page 352.

On the whole, the text offers a somewhat clearer and more detailed exposition than similar ones. The reviewer believes that engineers could use this for reference as well as for teaching.

VINCENT SALMON
Stanford Research Institute
Stanford, Calif.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with That Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger(†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

ACOUSTICS AND AUDIO FREQUENCIES

534.2:551.556 1

Wind Noise and the Transmission of Sound in the Open Air—F. Spandöck (*Z. angew. Phys.*, vol. 3, pp. 228–231; June, 1951.)

534.231 2

The Radiation of Pulses by Plane Piston Diaphragms in a Rigid Wall—F. A. Fischer. (*Acustica*, vol. 1, no. 1, pp. 35–39; 1951. In German.) When pulses are radiated by a piston diaphragm, direction-dependent distortion occurs because the propagation of the various spectral components varies in accordance with the frequency dependence of the radiation pattern. Calculation shows that a radiated unit pulse can be considered as composed of individual pulses from elementary areas of the diaphragm, and gives a representation of the shape of the diaphragm. For a pulse of arbitrary shape, the resultant is a product depending on both diaphragm shape and pulse shape.

534.231+621.396.67]:778.3 3

A Photographic Method for Displaying Sound-Wave and Microwave Space Patterns—W. E. Kock and F. K. Harvey. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 564–587; July, 1951.) Description of the method, with many illustrations of actual sound-wave and microwave fields. A similar probe method, for recording on Teledeltos paper, has been described by Iams (3590 of 1947).

534.232:534.321.9 4

Construction of High-Power Ultrasonic Transmitters—E. Skudrzyk. (*Elektrotech. u. Maschinenb.*, vol. 68, pp. 173–178 and 202–212; April 1 and 15, 1951.) Discussion of both quartz and magnetostriction oscillators. Suitable types of holder for high power quartz oscillators are described, and an oscillator transmitting waves through a duralumin membrane, for therapy applications, is illustrated. Simple equivalent-circuit theory of quartz oscillators is outlined and more rigorous theory is considered. Equivalent-circuit theory is also given for magnetostriction oscillators and a numerical example is calculated. Methods of measuring the output of ultrasonic oscillators are indicated.

534.232:534.321.9 5

An Efficient Low-Power Ultrasonic Generator—F. Pirker. (*Radio Tech* (Vienna), vol. 27, pp. 175–180; April, 1951.) Description of the construction and circuit arrangement of a generator which provides an output of 50 w at an electroacoustic efficiency of 60 per cent. Full details are given of the preparation and mounting of the quartz crystal, which is energized from a Hartley circuit using a 40-w pentode.

534.321.9 6

Ultrasonics in Air, and Applications—F. Canac and V. Gavreau. (*Acustica*, vol. 1, no. 1, pp. 2–16, 1951. In French.) Electrodynamic and magnetostrictive generators and receivers are described, and investigations of the reflection and diffusion of ultrasonic waves by cylinders and polycylindrical surfaces are reported. Practical acoustical applications discussed include the study of hall acoustics by means of models.

534.321.9:061.3 7

Ultrasonics in Fluids—E. G. Richardson. (*Nature* (London), vol. 168, pp. 106–107; July 21, 1951.) Brief report of an international conference held in Brussels in June, 1951. About thirty-five papers dealing with physical measurements and theories in the field of ultrasonics were presented.

534.321.9:534.373]:538.221 8

The Influence of Magnetization of Ultrasonic Attenuation in a Single Crystal of Nickel or Iron-Silicon—Levy and Truell. (*See* 169.)

534.44 9

An 8000-c/s Sound Spectrograph—O. Gruenz. (*Bell Lab. Rec.*, vol. 29, pp. 256–261; June, 1951.) Spectral density in 45-cps and 300-cps bandwidths at frequencies up to 8 kc is permanently recorded on charts. Energy distribution with frequency is presented in two ways: (a) as discrete horizontal markings corresponding to different frequencies, for any selected 5-ms period of integration, with a maximum of 35 db relative amplitude; (b) as a continuous record of intensity, indicated by relative darkness of trace, in a frequency/time rectangular-co-ordinate framework.

534.75 10

Nonlinear Characteristics of the Ear—G. Haar. (*Funk u. Ton*, vol. 5, pp. 248–257, May, 1951.) Comparative tests of the response of the ear to two pure notes of different frequencies f_1 and f_2 received simultaneously, show that difference tones formed in the ear are physically real vibrations which occur even when they are inaudible. The predominant difference frequency is $2f_1 - f_2$.

534.833.1 11

A Tentative Method for the Measurement of Indirect Sound Transmission in Buildings—E. Meyer, P. H. Parkin, H. Oberst, and H. J. Purkis. (*Acustica*, vol. 1, no. 1, pp. 17–28; 1951. In English.)

534.845.2 12

Acoustic Behavior of a Porous Material—A. Bressi and G. G. Sacerdote. (*Alta Frequenza*, vol. 20, pp. 28–33, February, 1951.) Measurements are reported of the acoustic absorption of "betamianto." The results differ from those for other porous materials in that an absorption maximum is exhibited at low frequencies even for a thin layer.

534.846 13

Acoustics and Sound Exclusion—W. A. Allen and P. H. Parkin. (*Arch. Rev.* (London), vol. 109, pp. 377–384; June, 1951.) Survey of the planning, structure, and finishing materials of the Royal Festival Hall, London, with consideration of both the internal acoustics of the auditorium and the exclusion of noise from the outside.

534.846.6 14

Investigation of Sound Diffusion in Rooms by Means of a Model—T. Somerville and F. L. Ward. (*Acustica*, vol. 1, no. 1, pp. 40–48; 1951. In English.) An experimental comparison was made of the effects produced by rectangular-, semicylindrical- and triangular-section diffusing elements applied to the room walls. The investigation covered both steady-state and pulse operation. All three types of diffuser caused reduction of irregularity in both types of characteristic, the effect produced by the rectangular-section diffusers being greatest.

621.395.61/.62 15

Standards of Transducers: Definitions of Terms, 1951—(PROC. I.R.E., vol. 39, pp. 897–899; August, 1951.) Reprints of this Standard, 51 IRE 20 S2, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy.

621.395.61/.62:621.395.92 16

Acoustic Transducers for Hearing Aids—W. Güttner. (*Fernmeldetech. Z.*, vol. 4, pp. 227–234; May, 1951.) A review of piezoelectric transducers. For the production of crystal microphones, the most suitable materials are Rochelle-salt and ammonium-phosphate crystals, and polarized $BaTiO_3$ ceramic. The properties of such microphones are analyzed and their parameters evaluated. For crystal earpieces using Rochelle salt, all possible constructions are described. From their mechanical equivalent representations, the frequency characteristics of the sound pressure are calculated and compared with measurement data.

621.395.61:621.385.82.029.3 17
Increasing the Efficiency of the High-Power Thermionic Cell by Superposition of a Strong Field obtained from a High Voltage of High Frequency—Klein. (*See* 280.)

621.395.616 18
Electrical Input Resistance of Capacitor Microphone—U. Kirschner. (*Arch. elekt. Übertragung*, vol. 5, pp. 273–278; June, 1951.) Calculation of the effective resistance is based on the work of Braun (1087 of 1945) and Weymann (470 of 1944). The results are applied to determine the parameters of a capacitor microphone to be used as a sound-pressure receiver with a circular characteristic.

621.395.623.7 19
Progress in the Development of Electrodynamic Loudspeakers—G. Buchmann and K. Küpfmüller. (*Fernmeldetech. Z.*, vol. 4, pp. 253–261; June, 1951.) The frequency response curve, the sensitivity, and the nonlinear distortion together give a good idea of the quality of a loudspeaker. Suitable methods for determining these quantities are described. The optimum response curve is found to be one which falls off at both the lower and the higher frequencies. With suitable design, cheap loudspeakers can be produced with a good response curve over a wide band of frequencies.

621.395.623.7 20
Acoustic Boosting of the Lower Audio Frequencies by means of Bass Resonators with Phase Inversion—H. Gemperle. (*Radio Tech.* (Vienna), vol. 27, pp. 245–249; June, 1951.) The method makes use of a resonator forming part of the loudspeaker housing, and coupled acoustically to the back of the vibrating membrane. Optimum dimensions of the resonator are calculable for any particular loudspeaker. Practical tests showed an increase in loudness of frequencies in the range 30 to 100 cps of about 8 db when the resonator was designed for a frequency of 60 cps.

621.395.623.7 21
Acoustic Coupling of the Diaphragms in Duo-Cone Loudspeakers—G. B. Madella. (*Alta Frequenza*, vol. 19, pp. 267–276; October–December, 1950.) Consideration of an idealized case indicates that radiated power as a function of diaphragm velocity is appreciably affected by the degree of coupling.

621.395.625+621.396.62:061.4 22
The Radio Exhibition at the Vienna Spring Fair—(*See* 250.)

621.395.625.3 23
Recording Demagnetization in Magnetic Tape Recording—O. W. Muckenhirn. (PROC. I.R.E., vol. 39, pp. 891–897; August, 1951.) "An analysis of the magnetic tape recording process employing supersonic excitation is presented by considering the effect of the spatial distribution of the magnetic field around the recording head air gap on the magnetic history of an unmagnetized element of tape as it tracks across the recording head. This leads to an effect which is termed "recording demagnetization," and serves to explain certain performance characteristics. An experimental technique developed for the measurement of this recording demagnetization is described, as is the method of measuring the air-gap field distribution. Finally, the correlation of the measurements of the recording demagnetization with normal recording performance characteristics is reported."

534.852:621.395.625.3 24
Magnetic Recording—M. Alixant. (*Radio Tech. Dig.* (France), vol. 5, no. 3, pp. 147–161; 1951.) A bibliography of articles and patents complementary to that included in the book "Magnetic Recording" by Schuh and Mikhnewitch published in 1950.

621.395.625.3 25
The R.C.A. 10794 High-Fidelity Magnetic Recording Head—M. Rettinger. (*Radio Tech. Dig.* (France), vol. 5, no. 3, pp. 131–141; 1951.) French version of 807 of 1951.

621.396.645.029.4:621.018.78†:534.861 26
Audio-Frequency Amplifiers with Adjustable Non-linearity (Distortion Circuits)—G. Hoffmann. (*Funk. u. Ton*, vol. 5, pp. 169–176; April, 1951.) Two amplifier circuits having nonlinear characteristics largely independent of frequency, and designed for investigation of the audibility of distortion tones produced in the transmission of music, are described.

ANTENNAS AND TRANSMISSION LINES

621.315.2 27
Interaxial Spacing and Dielectric Constant of Pairs in Multipaired Cables—J. T. Maupin. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 652–667; July, 1951.) Experiments show that so far as the interaxial spacing and dielectric constant are concerned, actual cable conditions can be represented by two parameters of an ideal cable. These two parameters can thus be readily obtained for a commercial cable by calculation from the measured inductance, mutual capacitance, and capacitance to ground of the pair.

621.315.2:621.396.97 28
Modern Broadcasting Cables to C.C.I.F. Recommendations—E. A. Pavel. (*Fernmeldetech. Z.*, vol. 4, pp. 150–157; April, 1951.) The essential features of the former and the recent C.C.I.F. recommendations are compared and tabulated, with particular reference to the widening of the transmission band required in connection with the introduction of usw broadcasting.

621.392:621.396.677 29
Approximation Methods in Radial Transmission Line Theory with Application to Horns—A. E. Laemmel, N. Marcuvitz, and A. A. Oliner. (PROC. I.R.E., vol. 39, pp. 959–965; August, 1951.) "The radial transmission line theory provides a rigorous description of the propagation of electromagnetic energy in certain cylindrical regions. The complexity of direct calculation may be avoided by a simpler, although approximate, description based on the asymptotic identity of radial and uniform transmission line formulas at large radii. Approximate expressions are developed for input admittance, frequency sensitivity, and higher mode interaction effects on radial lines. Estimates of the order of accuracy are included, and applications are made to sectoral horns."

621.392.09 30
Propagation of an Electromagnetic Wave Guided by a Metal Surface with Dielectric Coating—J. F. Colin. (*Onde élect.*, vol. 31, pp. 245–256; May, 1951.) Propagation along a cylindrical conductor is considered, assuming a perfect conductor and a loss-free dielectric. The conditions are deduced for a single possible mode of propagation, the TM_{01} mode. If the frequency, or the dielectric constant, or the thickness of the dielectric are increased beyond certain limits, conditions are obtained in which other modes are possible. The effective cross section of the conductor in transmitting power is much greater than its physical dimensions, owing to the distribution of energy in the surrounding fields. Expressions are deduced for the effect of losses and for the maximum power transmitted, and curves are drawn to assist in the solution of practical cases. The conditions for propagation of TE waves are briefly discussed.

621.392.2.09+621.396.11 31
Characteristic Impedance, Power, Voltage and Current in Transmission along Lines, in Waveguides and in Free Space—O. Zinke. (*Funk u. Ton*, vol. 5, pp. 225–238; May, 1951.) Neglecting losses and assuming purely progressive waves of sinusoidal form, the following relations are derived: (a) the ratio of the mean field strengths E_m/H_m at any point in the field cross section perpendicular to the direction of propagation is independent of the dimensions of the system, and is termed the elementary characteristic impedance z_e; (b) the characteristic impedance Z_e of the system is equal to $z_e h_m/b_m$, where h_m is the effective height, and b_m the effective breadth of the field; (c) the power transfer $N = E_m \cdot H_m \cdot h_m \cdot b_m$, i.e., the product of mean power per cm^2 and the effective cross section S of the field. For rectangular waveguides, S agrees exactly with the actual cross section. For coaxial cables and narrow strip conductors the agreement is only approximate.

621.392.26† 32
Experimental Investigation of the Reflections produced in a Waveguide by any Dielectric—L. R. Noriega. (*Radio franç.*, no. 6, pp. 1–9; June, 1951.) Reprint. See 3038 of 1949.

621.392.26† 33
A Study of Single-Surface Corrugated Guides—W. Rotman. (PROC. I.R.E., vol. 39, pp. 952–959; August, 1951.) Two types of structure were investigated experimentally as transmission lines and as radiators: (a) a flat grooved plate fed by a waveguide; (b) a circular corrugated cylinder fed by a coaxial line. Modification of type (b) results in a spirally grooved rod with similar properties. Measured field parameters agree with those predicted from theory. For properly designed structures, the energy is essentially bound to the corrugated surface. Little radiation occurs, and the attenuation of the traveling wave is due mainly to losses in the metal. The effect of filling the corrugations with solid dielectric is analyzed.

621.392.26†:621.39.09 34
Refraction of Evanescent Waves—N. Carrara. (*Alta Frequenza*, vol. 19, pp. 164–174; August, 1950.) An account of experiments in which evanescent centimeter em waves are incident on a plane separating two dielectrics, the reflected wave being also evanescent, while the refracted wave is of ordinary type. The results indicate that Snell's laws and Fresnel's formulas are valid even when the angles of incidence and reflection are complex.

621.392.26†:621.39.09 35
Propagation of U.H.F. Electromagnetic Waves in Conducting Waveguides of Rectangular Cross-Section: Part 1—H. Frühauf. (*Elektrotechnik* (Berlin), vol. 5, pp. 263–269; June, 1951.) Under certain conditions, energy is propagated in tubular conductors as plane waves. Vector analysis is applied to derive simple formulas, applicable by mathematical technicians to practical cases, for the field components of E and H waves, the limiting frequency, the wavelength in the rectangular waveguide, and the phase velocity.

621.396.67 36
Progressive-Wave Aerials—A. F. Huerta. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 23–37; June, 1951.) Theory is given relating to a progressive-wave antenna terminated by its characteristic impedance. The gain is calcu-

lated, and the effect of finite ground conductivity is discussed. The results are applied to explain the properties of rhombic antennas and of the Marconi series-phase antenna.

621.396.67 37
A Short Proof that an Isotropic Antenna is Impossible—H. F. Mathis. (PROC. I.R.E., vol. 39, p. 970; August, 1951.)

621.396.67 38
Tuning and Matching of Aerials—A. F. Huertas. (*Rev. Telecomun.* (Madrid) vol. 6, pp. 38–49; March, 1951.) Various methods of tuning and matching are illustrated and analyzed. Particular cases considered include the excitation of a directive group of antennas and applications of "balun" devices.

621.396.67:621.316.761.2 39
Increase of the Bandwidth of Aerials by means of Compensation Circuits—R. Goublin. (*Onde élect.*, vol. 31, pp. 233–244; May, 1951.) The problem is reduced to that of determining an antenna and associated network such that, in the complex diagram of the system, its input impedance falls within a circle defined by the magnitude of the maximum permissible SWR in the feeder line. Using this construction, the general cases are discussed of the matching of any antenna to a network, and of the increase of antenna bandwidth by compensation circuits, or by sections of transmission lines. Self-compensated arrays are described, and compensation methods are demonstrated by practical examples.

621.396.67:621.397.5 40
Directional Operation Planned for KRON-TV—R. A. Isberg. (*Tele-Tech.*, vol. 10, pp. 26–27, 76; July, 1951.) An eight-section channel-four R.C.A. TF-DA "Super-Gain" antenna is mounted on a 200-foot tower on San Bruno Mountain, four miles inland from the San Francisco coast. Radiation is at present omnidirectional, but may be modified to give 100 kw radiated power inland and 3kw radiated power towards the ocean. Feeder and antenna assembly are described in outline.

621.396.67+534.231]:778.3 41
A Photographic Method for Displaying Sound-Wave and Microwave Space Patterns—Kock and Harvey. (*See* 3.)

621.396.67.029.51 42
A Long-Wave Aerial—L. C. Garcia. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 82–85; March, 1951.) Description of an antenna forming part of the equipment of the new Spanish Transradio station at Parets del Vallés, Barcelona. The antenna is of the inverted-L type, the horizontal portion having a mean height of 86 m and being supported from steel lattice masts. The operating frequency of 79.15 kc will give a working radius of 1,750 km with fields$>$30 μv/m, sufficient for reception on an undulator.

621.396.67.029.62 43
Transmitting Aerials for U.S.W. Broadcasting—W. Berndt. (*Telefunken Ztg.*, vol. 24, pp. 6–21; March, 1951.) The characteristics desirable in antennas for broadcasting in the frequency band 87.7 to 100 mc are enumerated, and the properties of simple and beamed radiator elements are described, the conditions being determined for selection of the optimum distance between the elements. Methods of feeding multi-element antennas are described. Practical installations of U-antennas and double-slot antennas are illustrated and discussed.

621.396.67.029.63 44
Sidefire Helix U.H.F.-TV Transmitting Antenna—L. O. Krause. (*Electronics*, vol. 24, pp. 107–109; August, 1951.) A traveling-wave helical antenna giving a power gain of 20 at 500 mc, with a bandwidth of 20 mc, consists of four vertically stacked bays, each including a right- and a left-hand helix placed end to end and fed at their junction. Beam tilt can be produced by mechanically rotating one portion of the antenna relative to the other.

621.396.67.029.64 45
U.H.F. Dipoles—P. E. Vincelet. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 15, pp. 43–49; April, 1951.) In the construction of linear arrays, dipoles, including slotted dipoles, are particularly useful. A simplified method of calculating the characteristics of dipole arrays is outlined, and admittance characteristics of dipoles used under various conditions are given.

621.396.671 46
[E.m.] Waves in a Flat Horn—B. L. Rozhdestvenski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 77, pp. 221–224; March 11, 1951. In Russian.) A mathematical discussion is presented of the excitation of a flat horn by waves, either plane or cylindrical, arriving from a waveguide coupled to the horn. A conformal transformation is used which enables the problem to be reduced to a differential equation with partial derivatives. With certain simplifying assumptions, a finite solution of this equation can be obtained. The solution is sufficiently accurate for practical purposes, but its accuracy can be increased by using the method of successive approximations. The method is applicable also to circular and coaxial horns.

621.396.676 47
Radio-Frequency Current Distributions on Aircraft Structures—J. V. N. Granger and T. Morita. (PROC. I.R.E., vol. 39, pp. 932–938; August, 1951.) Models having scale factors of 43:1 to 72:1 were used, and the magnetic field surrounding the excited model was explored by means of a small loop. The results are shown diagrammatically for several types of aircraft excited by antennas of different types.

621.396.676 48
Experimental Study of Slot Aerials—J. Desauty and R. Rocherolles. (*Radio franç.*, no. 6, pp. 14–17; June, 1951.) Signals from a slot antenna located under an aircraft, were received by a similar antenna on the ground. Polar diagrams are given. The results are in conformity with theory.

621.396.677 49
Lenses for Microwaves—F. L. Garrido. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 65–75; March, 1951.) A concise review of the properties of various types of lens, including those formed of homogeneous dielectric, those with plane-parallel conductors, lenses of artificial dielectric, and others in which the index of refraction varies with distance from the axis.

621.396.677 50
Path-Length Microwave Lens—D. G. Kiely. (*Wireless Eng.*, vol. 28, pp. 248–250; August, 1951.) Experimental investigation of the *E*-plane radiation patterns of a cylindrical path-length lens shows an asymmetric main beam. This cannot be caused solely by an asymmetric distribution of field amplitude across the aperture, but must be due also to the presence of asymmetric phase curvature, contrary to the experience of Kock (3058 of 1949). The path-length lens, although having many *H*-plane applications, is fundamentally unsuited for phase-correction in the *E*-plane.

621.396.679.4 51
New Method of Top Feed for Antifading Broadcasting Masts—H. Graziadei. (*Fernmeldetech. Z.*, vol. 4, pp. 159–167; April, 1951.) Methods adopted to reduce fading effects in the service area of a broadcasting antenna are discussed, and the advantages of feeding an antenna at a point near the top are pointed out. Practical methods for top feeding a vertical antenna are described, and the results of measurements of the vertical distribution of radiation for antennas fed at the foot and top, respectively, are shown in graphs.

621.392.26† 52
Felder und Wellen in Hohlleitern [Book Review]—H. H. Meinke, Publishers: R. Oldenburg, München, Germany, 1949, 148 pp., 15 DM. (*Frequenz*, vol. 5, p. 53; February, 1951.) A detailed exposition in relatively simple form of the propagation of waves in waveguides. Useful for practising engineers as well as for students.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.3:621.396.621.54 53
New Diagrams for 'Ganging' Calculation—H. Kerbel. (*Funk u. Ton*, vol. 5, pp. 287–296; June, 1951.) Diagrams facilitating the determination of the parameters of ganged capacitors in superheterodyne receivers are given. From the data of the variable capacitors, completely independent parameters can be derived. The values of the parallel (trimmer) and series capacitors are obtained by multiplication of the corresponding parameters by the capacitance variation ΔC of the variable capacitor, the inductance by division by ΔC. The magnitude and distribution of the distributed capacitance are taken into account. A method of measuring the error curve is also described.

621.3.015.7†:621.387.4† 54
A Mechanical Kick-Sorter (Pulse Size Analyzer)—S. G. F. Frank, O. R. Frisch, and G. G. Scarrott. (*Phil. Mag.*, vol. 42, pp. 603–611; June, 1951.) Each pulse causes a small steel ball to be propelled along an inclined board; the ball describes a parabolic path, which is determined by the pulse amplitude, and lands in one of thirty parallel grooves. A histogram of the pulse size distribution is thus built up. The mechanical construction and associated electronic circuits are described. The latter eliminate pulses which are either too large or too small to be recorded, or which follow too closely upon the previous pulse.

621.3.015.7†:621.3.018.78† 55
Pulse Distortion—S. H. Moss. (*Proc. IEE*, vol. 98, pp. 398–400; September, 1951.) Summary of IEE Monograph No. 5. For a linear low-pass process, certain invariants which completely define the pulse shape exist for any input pulse. These are always increased by the same invariants imposed by the system to yield the invariants of the output waveform. The pulse invariants correspond to parameters used to describe statistical distributions. The theory in generalized form applies also to wave packets.

621.3.016.35 56
Application of Bode's Relation to the Calculation of a Nyquist Diagram—A. Herrent and G. Novgorodsky. (*HF* (Brussels), no. 9, pp. 229–236, 255; 1951.) In applying Nyquist's stability criterion to a closed system, it is necessary to know the frequency/phase and frequency/amplitude characteristics of the network over a very wide frequency range. The practical difficulties involved, especially at low frequencies, are eliminated by the use of Bode's relation which relates the phase and amplitude characteristics. A quick and easily applied graphical method is described for evaluating the integral in Bode's formula, and hence the Nyquist diagram may be drawn. The method has been verified experimentally in the case of a low-frequency amplifier with resistive feedback.

621.314.2:621.392.6 57
Contributions to the Theory of the Differential Transformer—G. Schmitt. (*Frequenz*, vol. 5, pp. 39–44; February, 1951.) Equations

and equivalent circuits for the differential transformer are developed from the theory of six-pole networks. The stop-band attenuation is calculated and its frequency variation investigated. Design conditions are discussed for making the stop-band attenuation as great as possible.

621.315.592†:537.312.6 58
Experiments on Thermistors of Italian Manufacture—P. Lombardi and M. Vallauri. (*Alta Frequenza*, vol. 20, pp. 51–67; April, 1951.) Basic theory of electronic conduction in semiconductors is outlined, and variation of resistance with temperature is considered. Measurements on specimen thermistors are reported, and the values of their time constants are deduced. Applications both familiar and unfamiliar are discussed.

621.316.726.078.3 59
U.H.F. Discriminator and its Application to Frequency Stabilization [of klystron]—G. Pircher. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 15, pp. 37–42; April, 1951.) A modification of Pound's stabilization circuit (1690 of 1947) is described which avoids the use of magic-T junctions. See also 2758 of 1951.

621.316.761.2:621.396.67 60
Increase of the Bandwidth of Aerials by means of Compensation Circuits—Goublin. (*See* 39.)

621.318.4 61
Mean Winding Lengths—L. Hubl. (*Elektrotechnik*, (Berlin), vol. 5, pp. 175–179; April, 1951.) Methods and formulas are given for determining the mean length of choke and transformer windings on laminated cores. Tables and diagrams enable the required values to be found directly for many practical cases without much calculation, the values in the tables being used for interpolation.

621.318.572 62
Transition of an Eccles-Jordan Circuit—J. R. Tillman. (*Radio Tech. Dig.* (France), vol. 5, nos. 3 and 4, pp. 143–146 and 203–216; 1951.) French version of 2119 of 1951.

621.318.572 63
Electronic Switch—S. F. Pinasco. (*Rev. Telegr.* (Buenos Aires), vol. 39, pp. 333–335; 353; June, 1951.) Description, with detailed circuit diagram, of a switching circuit consisting of a multivibrator, which produces the voltages necessary for alternate blocking and unblocking of the signal, and an amplifier giving the required amplitude.

621.318.572:621.385.5 64
High-Speed Sampling Techniques—B. R. Shepard. (*Electronics*, vol. 24, pp. 112–115; August, 1951.) Condensed version. For full paper see *Proc. Nat. Electronics Conference* (Chicago), vol. 6, pp. 255–265; 1950. An arrangement is described which is suitable for transmitting observations from a number of measuring elements in rapid cyclic succession, using only one channel. Signals of low intensity are considered. A rotating-radial-beam-tube commutator of the type described by Skellett (3167 of 1944) is connected to the individual circuits in turn. The noise generated by this scanning operation is discussed; its magnitude determines the amount of preamplification required. Scanning rates of 10^5 elements /s are obtainable.

621.319.4 65
Working Characteristics and Performance of Ceramic Capacitors—A. Danzin. (*Ann. Radioélect.*, vol. 6, pp. 156–179; April, 1951.) The performance of capacitors of a great variety of types formed from the ceramic dielectrics previously described (132 of 1951) is discussed in detail. Manufacturing processes are described and tables are given analyzing the general characteristics: stability, dielectric loss, endurance under varying climatic conditions, temperature coefficient, and overload characteristics.

621.392 66
The Effect of Alteration of Network Parameters in a Particular Type of Reactance Network on the Resonance Frequency of the Network—T. O'Callaghan. (*Frequenz*, vol. 5, pp. 44–47; February, 1951.) The networks considered are of low-pass shunt-capacitance ladder type, with n unequal capacitances and $n+1$ unequal inductances. Two cases are distinguished: (a) where available component magnitudes do not coincide exactly with design requirements for a desired performance of the network; (b) where it is desired to eliminate an undesired resonance occurring in an existing network. The method developed is based on considering half-sections of the network as separate resonant systems.

621.392.001.4 67
Determination of Amplitude and Phase Characteristics of Linear Networks by means of Square Waves—Müller. (*See* 209.)

621.392.5:512.831 68
Transformation of the Quadripole-Network Matrix to the Diagonal Form—H. Schulz. (*Arch. elekt. Übertragung*, vol. 5, pp. 257–266; June, 1951.) Description of a method which systematizes all types of linear quadripole, with application to a low-loss transformer with fixed coupling.

621.392.52 69
The Four-Circuit Band-Pass Filter and Bandwidth Switching by Back Coupling—G. Hentschel. (*Funk. u. Ton*, vol. 5, pp. 281–286; June, 1951.) The formulas for calculating the resonance curves of four-circuit filters with arbitrary values of coupling and damping are derived. By introducing auxiliary coupling from the fourth to the first circuit, bandwidth switching is effected very simply. The filter then approximates to a two-circuit filter with enhanced coupling.

621.392.52 70
Transfer Properties of Single- and Coupled-Circuit Stages With and Without Feedback—J. W. Muehlner. (PROC. I.R.E., vol. 39, pp. 939–945; August, 1951.) "Universal curves of the reciprocal of the complex response function are studied for systems with one or two tuned circuits. The normalized parabola for two circuits is presented with a net of loci for the origin of co-ordinates within it. The ratio Q_1/Q_2, the amount of fixed detuning, the amount of coupling, and amplitude and phase of any additional feedback independently displace the origin with respect to its position in case of two identical, uncoupled circuits without feedback. Complex cases may be easily analyzed. Different methods incorporating regeneration or feedback loops are discussed which allow practical bandwidth control within wide limits."

621.392.52:621.395.44 71
Group-Combining Filters for Twelve and Twenty-Four Circuit Carrier Systems—D. W. Robson. (*GEC Telecommun.*, vol. 3, no. 1, pp. 45–51; 1948.) Complementary high-pass and low-pass crystal filter units are described for combining and separating the two sets of twelve channels in the 12-to 60-kc and 60-to 108-kc bands, respectively, in a twenty-four-channel carrier system. An attenuation of >40 db is achieved for the two stop bands of, respectively, 60.3 to 108 and 12 to 59.7 kc, with a pass-band loss level to within ± 0.5 db.

621.392.53 72
Waveform Systems and 'Time Equalizers'—D. C. Espley. (*Wireless Eng.*, vol. 28, pp. 251–258; August, 1951.) A new approach to the solution of circuit problems where the available transmission information is given in terms of distorted waveforms. Some classical transforms relating frequency and time characteristics can be replaced by simpler algebra which enables the physical interpretation to be kept in view. The amplitude and phase characteristics of a system with unknown internal structure are evaluated simply from the nonredundant networks which are arranged to simulate the actual system waveform distortions. The insertion characteristics of the waveform-correcting networks, or "time equalizers," follow directly. The correction of some systems may require the use of both "time equalizers" and the better known "steady-state" equalizers.

621.392.6.062 73
N-Terminal Switching Circuits—E. N. Gilbert. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 668–688; July, 1951.) Shannon's method of synthesis of two-terminal switching networks (2456 of 1949) is generalized to apply to N-terminal networks which use selector switches with any number of positions.

621.396.611.1 74
Constant Oscillatory Circuits for High Frequencies—K. Schreck. (*Fernmeldetech. Z.*, vol. 4, pp. 145–149; April, 1951.) The design and properties of various fixed and variable circuits are described, including ceramic components, which give a frequency constancy of the same order as that obtained with quartz crystals. The methods of compensation specially developed for new types of circuit are described.

621.396.611.1 75
Internal Resonance in Circuits containing Nonlinear Resistance—R. E. Fontana. (PROC. I.R.E., vol. 39, pp. 945–951; August, 1951.) The simultaneous presence of two oscillations locked in synchronism is investigated for circuits containing a nonlinear element. Such a circuit is said to be "internally resonant," the ratio of the two frequencies being that of two integers. The solution of the appropriate equations and the conditions for stable equilibrium are discussed.

621.396.611.1 76
Oscillation Excitation in Circuit with Periodically Varying Capacitance—W. Taeger. (*Funk u. Ton*, vol. 5, pp. 300–309; June, 1951.) Starting from a particular form of the modulation function of the capacitance variation which is rich in harmonics (sawtooth function) and which has the advantage of affording an exact solution of the differential equation describing the phenomena, the conditions are investigated under which the natural damping of the circuit can be eliminated, and the currents and voltages follow a simple sine law.

621.396.611.1 77
Method of Determining the Forced Oscillations produced by Periodic E.M.F. of Arbitrary Waveform—O. M. Bogatyrew. (*Elektrotechnik* (Berlin), vol. 5, pp. 194–197; May, 1951.) A formula for the induced current is derived which involves an integral with limits convenient for calculation. The formula is developed by a simple application of the Duhamel integral.

621.396.611.3 78
Equations for the Natural Frequencies of Coupled Circuits from the Quadripole Viewpoint—H. Rukop and H. Kaiser. (*Telefunken Ztg.*, vol. 24, pp. 64–74; June, 1951.) Various simple and more complex circuits are considered, and formulas for the coupling factors are derived and tabulated together with relations between the circuit parameters. Applications of the theory to circuits involving the various types of coupling are described.

621.396.611.4 79
Separation of Degenerate Oscillation Modes in Perturbed Rectangular Cavities—F. Bosinelli. (*Alta Frequenza*, vol. 19, pp. 244–259; October–December, 1950.) A rectangular resonant cavity is considered, the walls of

which have a slight arbitrary deformation. The perturbed resonance frequencies and, in particular, the frequency displacements of two degenerate modes, are determined theoretically. The principles followed in the case of twofold degeneracy can be generalized for N degenerate modes. The frequency shifts are calculated for the two degenerate modes TE_{emn} and TM_{emn} in a rectangular cavity which has been deformed into a trapezoidal cavity by tilting one of its walls through a small angle θ. This type of perturbation separates the two modes: to a first approximation, the relative frequency shifts are proportional to θ. Similar results are obtained for the two modes TE_{102} and TE_{201} when they are degenerate.

621.396.615:621.316.726 80
Difference Oscillators and their Constancy—W. Herzog. (*Arch. elekt. Übertragung*, vol. 5, pp. 279–283; June, 1951.) Arrangements are described which produce two frequencies simultaneously and, after rectification, emit the difference frequency. Bridge types of oscillator with similar or dissimilar arms are considered. In the latter case, the use of a crystal as one branch of the circuit results in high constancy of the difference frequency. A simple oscillator is also described which uses a 130-kc crystal provided with three electrodes; the resulting difference frequency is 1.3 kc.

621.396.615.018.424† 81
Seven-League Oscillator—F. B. Anderson. (PROC. I.R.E., vol. 39, pp. 881–890; August, 1951.) A bridge-type RC oscillator is described which is continuously variable over a very wide frequency range in one sweep of a two-gang linear-potentiometer control. The output is available in four phases, and the frequency is an approximately logarithmic function of the potentiometer setting. The design of the bridge is discussed in terms of circle diagrams of transmission through RC networks. Tentative frequency limits are 0.01 cps and 10 mc. Accuracy of setting and stability of frequency are discussed.

621.396.615.029.422/.426 82
A Balanced RC Oscillator—D. A. Bell. (*Electronic Eng.*, vol. 23, pp. 274–275; July, 1951.) A Wien-bridge feedback circuit with a balanced maintaining amplifier is used to produce frequencies between 0.1 and 20 cps having a very small even-harmonic content.

621.396.615.14 83
Limiting Range for the Generation of Oscillations with Grid-Controlled Valves—L. Ratheiser. (*Radio Tech.* (Vienna), vol. 27, pp. 168–170; April, 1951.) Continuation of paper abstracted in 1819 of 1948. Circuit arrangements and construction of various types of usw oscillator are discussed.

621.396.615.17 84
Frequency Multiplier with Selective Network in the Grid Circuit—A. Taraboletti. (*Alta Frequenza*, vol. 19, pp. 221–230; October–December, 1950.) Relative amplitudes of the various harmonics appearing in the anode circuit are calculated for the case of an aperiodic and of a selective grid circuit. Experimental results confirm the superiority of the selective arrangement.

621.396.619.23 85
Oscillator Circuits as Frequency Modulators—Mansfeld. (*See* 297.)

621.396.645 86
Mathematical Treatment of Control Curves for Amplitude-Controlled Amplifiers—J. Hacks. (*Telefunken Ztg.*, vol. 24, pp. 51–54; March, 1951.) The dependence of the output voltage on the input voltage is investigated (a) for linear dependence on the slope of the tube characteristic (hexode with grid-3 control), (b) for exponential dependence (pentode, or hexode with grid-1 control). Both cases admit of accurate calculation. Experimental and theoretical curves are compared and the design of a practical IF amplifier with forward and backward control is shown.

621.396.645 87
Cathode-Follower Input Impedance—J. E. Flood. (*Wireless Eng.*, vol. 28, pp. 231–239; August, 1951.) The input conductance G of a cathode follower with a capacitive load decreases with increasing frequency, passes through zero, and becomes negative. The circuit with the grid lead returned to a point on the cathode resistance is analyzed; an expression is obtained for the frequency at which G is zero. Methods of raising this frequency are discussed; the most satisfactory involves a series grid resistance. Experimental work with a triode-connected EF 91 is described. Discrepancies between predicted and observed behavior are attributed to Miller effect, resulting from the unavoidable anode load presented by the anode decoupling capacitor.

621.396.645:621.392.52 88
New Method of Calculating High-Frequency Filters with Tchebycheff Type of Amplification—H. Edelmann. (*Arch. elekt. Übertragung*, vol. 5, pp. 284–292; June, 1951.) With the aid of Tchebycheff polynomials, the amount of amplification, and hence the complex amplification, is determined by choice of the null points of the reciprocal of the complex amplification. Circuit parameters can then be chosen so that the null points agree with requirements. The method is applicable to amplifiers with any number of stages and circuits. Typical examples considered are (a) single-circuit amplifier stage, (b) two-circuit amplifier stage with coupled filters.

621.396.645.018.424† 89
Design and Construction of Wide-Band I.F. Amplifiers—L. Gérardin. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 15, pp. 19–36; April, 1951.) The general design of IF amplifiers for use in radar receivers is discussed. Large bandwidth may be achieved in amplifiers of normal type by the use of certain artifices, but the method has limitations. A distributed amplifier is described which has a pass band of 10 to 150 mc at ±3 db, with an over-all gain of 72 db. The calculation of the over-all noise factor of a radar receiver is described in an appendix.

621.396.645.018.78† 90
Nonlinear Distortion in Push-Pull Class-B Amplifiers with Choke Output—F. Böttcher. (*Telefunken Ztg.*, vol. 24, pp. 39–48; March, 1951.) Detailed mathematical analysis of the choke output circuit, with numerical calculations for a 40-kw tube. Approximate formulas are shown to give results which are sufficiently accurate for most purposes.

621.396.645.029.4:621.3.018.78†:534.861 91
Audio-Frequency Amplifiers with Adjustable Nonlinearity (Distortion Circuits)—Hoffmann. (*See* 26.)

621.396.645.2.018.422† 92
Calculation of High-Frequency Amplifier Circuits with Relatively Narrow Pass Band (Radio Filter Circuits)—W. Cauer. (*Arch. Elektrotech.*, vol. 40, no. 2, pp. 88–199; 1951.) A general method of calculation is developed whereby any symmetrical frequency characteristic of an amplifier containing only single-circuit coupling networks can be realized in respect of phase and amplitude by an amplifier circuit having a prescribed number of stages with multiple-circuit coupling. Detailed treatment is given for two- and three-circuit reactive networks. Application of the method to purely resistive networks is considered.

621.396.645.35 93
D.C. Amplifier with Reduced Zero-Offset—W. McAdam, R. E. Tarpley, and A. J. Williams, Jr. (*Electronics*, vol. 24, pp. 128–132; August, 1951.) Essentially the same paper as published in *Proc. Nat. Electronics Conference* (Chicago), vol. 6, pp. 277–285; 1950. An amplifier for current measurement, using a calibrated 1-m Ω woven-wire resistor, has a zero offset (i.e., input required to bring output to zero) less than 1.6 times the peak-to-peak value of thermal-fluctuation voltages, corresponding to 10^{-12} a on current ranges and 1 μv on $2=10^{10}-\Omega/v$ voltage ranges. Over-all dc feedback is used.

621.396.645.37 94
Parallel Voltage Feedback—K. H. R. Weber. (*Elektrotechnik*, (Berlin), vol. 5, pp. 180–182; April, 1951.) Series and parallel voltage feedback are discussed with the aid of diagrams, and are shown to be essentially similar in operation. Parallel voltage feedback involves two damping quantities, which can be determined from formulas given. Amplification with parallel feedback is not directly measurable, but must be calculated from a simple formula involving the two damping quantities and a certain voltage ratio. A particular case is noted in which the output voltage is zero.

621.396.645.371 95
Distortion and Power Output of Negative-Feedback Amplifier Output Valves in Class-A Operation—R. Siegert. (*Telefunken Ztg.*, vol. 24, pp. 100–117; June, 1951.) A theoretical and experimental investigation, illustrated by many curves, of the distortion remaining after application of negative feedback, first for the case of a phase-constant external resistance, and then for a frequency-dependent external resistance such as a loudspeaker.

621.396.645.371:621.3.015.3 96
Transient Phenomena in Wide-Band Feedback Amplifiers—O. B. Lur'e. (*Elektrotechnik*, (Berlin), vol. 5, pp. 171–174; April, 1951.) German version of 78 of 1950.

GENERAL PHYSICS

537.291 97
On the Motion of Gaseous Ions in a Strong Electric Field: Part 1—G. H. Wannier. (*Phys. Rev.*, vol. 83, pp. 281–289; July 15, 1951.) The Boltzmann form of the kinetic theory of gases is used to determine the motion of positive ions in a gas under the influence of a static, uniform electric field. The ion density is assumed to be vanishingly low, and the field to be of a strength such that the energy it imparts to the ions is not negligible in comparison with their thermal energy. The velocity averages are studied, rather than the details of velocity distribution.

537.52 98
Cold Emission of Electrons in Spark Gaps—F. L. Jones and E. T. de la Perrelle. (*Nature* (London), vol. 168, pp. 160–161; July 28, 1951.) The results of measurements of the electron emission from a Ni cathode, oxidized by previous sparking in air, are interpreted as supporting the view that the electrons have their source in the oxide surface layer, and are extracted by a field process.

537.52 99
Physics of Electrical Discharge—F. L. Jones. (*Nature* (London), vol. 168, pp. 140–142; July 28, 1951.) Report of a symposium on "Some Aspects of Discharge Physics" held on March 29 to 31, 1951, at Swansea.

537.52 100
The Distribution of Electron Energies in a Discharge constricted by its Self-Magnetic Field—J. E. Allen. (*Proc. Phys. Soc.*, vol. 64, pp. 587–589; June 1, 1951.) The distribution of electron velocities in a high-current arc in which the radial electric field can be neglected, is shown to differ from the Maxwellian, owing to the action of the self-magnetic field. In particular, the proportion of high-velocity electrons is reduced.

537.521 101

Microsecond Transient Currents in the Pulsed Townsend Discharge—J. A. Hornbeck. (*Phys. Rev.*, vol. 83, pp. 374–379; July 15, 1951.) The variation of photoelectric emission current with time was observed using an experimental gas-filled tube and a pulsed light source. The method is useful for studying fundamental parameters and processes in the noble gases.

537.525 102

Mechanism of Secondary Ionization in Low-Pressure Breakdown in Hydrogen—F. L. Jones and D. E. Davies. (*Proc. Phys. Soc.*, vol. 64, pp. 519–527; June 1, 1951.) The breakdown mechanism was investigated by examining the dependence of the shape of cathode emission curves on the deposition of electro-positive or electro-negative atoms on cathodes of Ni, Al, Ag, Cu, and Mo. "Results are analyzed in terms of the varying effective work function of the surface due to the deposition of atoms. It is concluded that the high photoelectric emission from cathodes of low effective work function was due to low-energy photons generated in the prebreakdown currents, but that for clean metals, the secondary emission was due to impact of positive ions. Photo-emission due to any high-energy photons generated in the ionization currents was negligible in comparison."

537.525.5 103

The Potential Field in and around a Gas Discharge, and its Influence on the Discharge Mechanism—W. Finkelnburg and S. M. Segal. (*Phys. Rev.*, vol. 83, pp. 582–585; August 1, 1951.) Measurements on the cathode and anode fall regions in carbon arcs are described; the thickness of each region is <0.01 cm. The distorting effect of the nonuniform space-charge distribution on the potential field is discussed. Potential probe measurements in low- and high-current carbon arcs are in good agreement with the theoretical analysis, and prove the transitional region between the distorted potential field inside and the undistorted potential field outside the discharge to be a fairly thin one.

537.533:538.56 104

Radio Wave Generation by Multistream Charge Interaction—J. Feinstein and H. K. Sen. (*Phys. Rev.*, vol. 83, pp. 405–412; July 15, 1951.) Theories of the excitation of plasma oscillations in a two-electron-beam system are reviewed. Analytical and graphical methods are used to determine the ranges of parameters within which the growth of waves is possible, taking into account the effect of thermal motion. The growth of waves is possible, within a narrow frequency range, even when the beam injection velocity is much smaller than the mean thermal velocity. Relevance to solar phenomena is discussed. A new theory is advanced of the mechanism by which the longitudinal oscillations are converted into transverse electromagnetic radiations.

537.533:538.56 105

The Relativistic Theory of Electro-Magneto-Ionic Waves—V. A. Bailey. (*Phys. Rev.*, vol. 83, pp. 439–454; July 15, 1951.) "The relativistic equations governing an ionized gas pervaded by static electric and magnetic fields, and the corresponding equations for small perturbations, are derived. The equations for plane perturbations are then obtained, and several important cases are developed in detail. Frequency bands in which growing wave modes occur are also determined. By means of certain rules of transformation, the theory is also used to study the plane waves which can occur in interpenetrating double streams of electrons. The results obtained are formally similar to those obtained for waves in an ionized gas. In the absence of static magnetic fields, and with the effects of collisions neglected, it is found that, in either an ionized gas or in interpenetrating double streams of electrons, certain waves propagated obliquely to the drift motions may both grow and possess Poynting fluxes. These fluxes are such that certain initial disturbances can lead to the escape of amplified electromagnetic energy from an ionized medium. The exchange of momentum and energy, between the streams of electrons and ions and the growing waves, is discussed by means of the momentum-energy tensors of the charged particles and of the electromagnetic field. It is concluded that both theory and observation lend support to the hypothesis suggested previously that a notable part of cosmic noise and strong solar noise originates as electro-magneto-ionic waves in magnetized ionized regions."

537.533:621.385.029.63/.64 106

Waves in Electron Streams and Circuits—J. R. Pierce. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 626–651; July, 1951.) A review of some of the assumptions made, and general problems involved, in analysis of the properties of electron streams coupled to circuits. The reasons for using a wave approach are explained. "The propagation constant of the wave is obtained in terms of the properties of the electron stream and the impedance of the circuit. Some general properties of waves are described. The importance of fitting boundary conditions in the solution of an actual problem is discussed, and examples, including that of "backward-gaining" waves, are considered."

537.533.8 107

A Theory of Secondary Electron Emission from Metals—E. M. Baroody. (*Phys. Rev.*, vol. 78, pp. 780–787; June 15, 1950.) A theory is formulated which is based on the Sommerfeld free-electron model, momentum transfer between electrons and lattice being accounted for by introducing a finite mean free path for elastic scattering. An inverse relation between emission and work function is deduced which is in qualitative agreement with experiment. The theory also accounts for the velocity distribution of the secondary electrons, and is consistent with their observed angular distribution.

537.533.8 108

A Comparison of Theories of Secondary Emission—J. J. Brophy. (*Phys. Rev.*, vol. 82, pp. 757–758; June 1, 1951.) Comparison of results deduced from the quantum-mechanical theory of Wooldridge (147 of 1940) and the free-electron theory of Baroody (107 above).

537.533.8 109

Secondary Emission of Electrons from Liquid Metal Surfaces—J. J. Brophy. (*Phys. Rev.*, vol. 83, pp. 534–536; August 1, 1951.) Curves showing the secondary-emission ratio of Bi, Ga, Pb, and Hg as a function of primary energy were obtained experimentally for the metal in both the solid and liquid states. The characteristics for the two states are similar to one another and to the curves for other pure metals. The observed maximum values of secondary-emission ratio are compared with values predicted from theory.

537.533.8 110

The Application of Wooldridge's Theory of Secondary Emission—E. M. Baroody. (*Phys. Rev.*, vol. 83, pp. 857–858; August 15, 1951.) Conclusions are reached in general agreement with those of Brophy (108 above). For Wooldridge's theory, see 147 of 1940.

538.123 111

The Analytical Expressions for the Vector Potential of a Rotationally Symmetrical Magnetic Field—A. Moussa and J. Lafoucrière. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 139–141; July 9, 1951.

538.311:621.318.423:513.647.1 112

The Excitation of a Helical Conductor—S. Kh. Kogan. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 489–492; September 21, 1950. In Russian.) It is assumed that a voltage is applied to an infinitely small element of the helix at $L=0$. The distribution of voltage along the helix is then given by equation (2), and the boundary condition, taking into account the finite conductivity of the helix, is given by equation (3). An integral equation (3b) determining the distribution of the current is derived, and its solution is reduced to the evaluation of an integral (top of p. 491.) For practical purposes it may be assumed that only one wave is propagated along the conductor.

538.566 113

Properties of Inhomogeneous Plane Electromagnetic Waves—G. Bonfiglioli. (*Alta Frequenza* vol. 19, pp. 259–266; October–December, 1950.) Discussion of the propagation of plane-polarized, inhomogeneous waves, i.e., with electric/magnetic field vectors not constant along the wavefront, in an isotropic unbounded space. For any frequency there is an infinite and continuous series of propagation modes. This result is related to the theory of propagation in waveguides and to the theory of total reflection.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7 114

World-Wide Observations of Solar Activity—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 93–95; July, 1951.) A general account of the National Bureau of Standards' system for collecting information about the various types of solar activity.

523.72 115

Measurements of Solar Extreme Ultraviolet and X-Rays from Rockets by means of a $CaSO_4$-Mn Phosphor—R. Tousey, K. Wetanabe, and J. D. Purcell. (*Phys. Rev.*, vol. 83, pp. 792–797; August 15, 1951.) Samples of a $CaSO_4$-Mn phosphor, which was insensitive to wavelengths above 1,340Å were exposed to sunlight in V2 rockets. Response in the bands 0–8Å, 1,050 to 1,340Å, and 1,230–1,340Å was measured using Be, LiF, and CaF_2 filters. X-rays were observed on one flight which reached 128 km, during which a sudden ionospheric disturbance occurred. Energy at wavelengths between 1,050 and 1,340Å was observed at heights down to 80 to 90 km on all four flights. Values of the total radiation intensity in two ranges of the solar spectrum are calculated for a height between 82 and 127 km. The radiation between 795 and 1,050Å reaching this region has an intensity well above that produced by a 6,000°K black-body sun.

523.72:621.396.822 116

Solar Radio-Frequency Emission from Localized Regions at Very High Temperatures—J. H. Piddington and H. C. Minnett. (*Aust. Jour. Sci. Res., Ser.*, vol. 4, pp. 131–157; June, 1951.) The slowly varying component, S, of the solar rf radiation is correlated with sunspot data and degree of circular polarization. It is suggested that the S component is due to thermal emission from localized regions at temperatures of about 10^7°K, often near sunspots. The radiation from a model hot region is examined in detail; the derived emission spectrum and polarization characteristics agree reasonably with observations. Electrons and protons would probably be emitted thermionically from the hot regions, would travel to the earth with average velocities of a few hundred kilometers per second, and could form the slow corpuscular radiation deduced from terrestrial magnetic data.

523.72:621.396.822 117

Solar Noise—M. Nicolet. (*Scientia*, vol. 85, pp. 37–41 and 71–77; March–April, 1950. In French.) A survey paper. Observations on rf waves emitted by the sun are described and discussed in relation to known solar and terrestrial phenomena. The noise from the quiet sun originates in the outer regions; centimeter waves originate in the chromosphere, and meter waves in the corona. When the sun is disturbed, as indicated by the appearance of sunspots, the noise level rises. With the variation of solar activity, bursts are superposed on the steady noise level. These are sometimes extremely intense, being then associated with unusual phenomena such as SW fade-outs, or eruptions of cosmic rays consisting of charged particles accelerated by the electric field induced by the varying magnetic field of the sunspots. Temperatures of the order of 10^{6}°K for the quiet sun, and 10^{14}°K for the most intense bursts are indicated.

523.746 "1950" 118
Final Relative Sunspot-Numbers for 1950—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 56, pp. 439–441; September, 1951.)

523.746"1951.0/.06" 119
Provisional Sunspot Numbers for April–June 1951—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 56, p. 445; September, 1951. *Z. Met.*, vol. 5, p. 288; September, 1951.)

523.854:621.396.822 120
Observations of the Source of Radio-Frequency Radiation in the Constellation of Cygnus—B. Y. Mills and A. B. Thomas. (*Aust. Jour. Sci. Res., Ser.*, vol. 4, pp. 158–171; June, 1951.) The position and some of the properties of this source were determined. Observed fluctuations in intensity were found to correlate with activity in the ionosphere *F* region. There is no evidence that the emission from the source varies.

523.854:621.396.822 121
A Preliminary Survey of the Radio Stars in the Northern Hemisphere—M. Ryle, F. G. Smith, and B. Elsmore. (*Mon. Not R. Astr. Soc.*, vol. 110, no. 6, pp. 508–523, 1950.) The positions and intensities of fifty discrete sources of radio waves were measured, using an interferometer of high resolving power operating on a wavelength of 3.7 m. The results indicate that most of these radio stars are situated within the galaxy, and are distributed with an average density comparable with that of visual stars. They were not identifiable with visual stellar bodies.

538.12 122
On the Stability of Magneto-Hydrostatic Fields—S. Lundquist. (*Phys. Rev.*, vol. 83, pp. 307–311; July 15, 1951.) The stability of static magnetic fields in an electrically conducting liquid is investigated mathematically. Particular attention is given to the case of twisted cylindrical magnetic fields. Instabilities may be caused by the twisting when it introduces an increase of magnetic energy of the same order of magnitude as the energy of the original field. The discussion is relevant to theories of the earth's magnetic field.

538.711 123
Aeromagnetic Survey of Vertical Intensity over the Sound with Apparatus of the BMZ Type—A. Lundbak. (*Tellus*, vol. 3, pp. 69–74; May, 1951.) The magnetometric zero balance was suspended in gimbals in an aircraft flying over the sea area between Denmark and Sweden. Comparison with measurements taken on the ground near the coasts shows that, with certain precautions, accuracies to within ±20 gammas can be obtained.

550.38 "1951.01/.03" 124
International Data on Magnetic Disturbances, First Quarter, 1951—J. Bartels and J. Veldkamp. (*Jour. Geophys. Res.*, vol. 56, pp. 442–444; September, 1951.)

550.38 "1951.04/.06" 125
Cheltenham Maryland Three-Hour-Range Indices *K* for April to June, 1951—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 56, p. 445; September, 1951.)

550.384 126
Variations of the Magnetic Field and Rotation of the Earth—N. Stoyko. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 80–82; July 2, 1951.) From a comparison of recorded data, it is concluded that the variation of the geomagnetic field may be explained by variation of the earth's rotation, and that any rotating body should possess a magnetic field of which at least a part is due to the rotation round its axis.

550.385 127
Principal Magnetic Storms [April–June, 1951]—*Jour. Geophys. Res.*, vol. 56, pp. 446–448; September, 1951.)

551.510.535 128
The Modes of Formation of the Ionospheric Layers—J. H. Piddington. (*Jour. Geophys. Res.*, vol. 56, pp. 409–429; September, 1951.) Ionospheric measurements made during eclipses are analyzed and, together with recently acquired solar data, are used in a reexamination of the theories of the ionosphere. It is shown that: (a) The rates of disappearance of electrons in all the principal layers may be much higher than hitherto believed. If so, then the number of ionizing quanta needed is also greatly increased. (b) There is probably a nonsolar source of electron production in region *E* (and perhaps in region *D*). It may be the electric currents flowing in the upper atmosphere which are also manifested by fluctuations in the earth's magnetic field. (c) Solar ionizing radiation does not appear to originate (principally) either throughout the corona or near the photosphere, but in regions of rapid temperature transition in the chromosphere and lower corona. Excess emission occurs, as has been shown previously, from regions of disturbance, often near sunspots. (d) The emission spectrum of these regions may be strong in the quasi-Lyman spectrum of He II and in X-rays of wavelengths ranging down to about 3Å. During flares, X-rays of wavelength less than 1Å may be emitted. (e) The F_1 and F_2 ionospheric layers may be accounted for if minor modifications of existing theories are made. The *E* and *D* layers may be formed by quanta of a few hundred and a few thousand electron volts, respectively, originating in the chromospheric transition region.

551.510.535 129
A Self-Consistent Treatment of the Oxygen Dissociation Region in the Upper Atmosphere—H. E. Moses and Ta-You Wu. (*Phys. Rev.*, vol. 83, pp. 109–121; July 1, 1951.) A theory is suggested according to which the temperature and the concentrations of oxygen atoms and molecules at a height of 100 to 140 km can be expressed as functions of the height from a knowledge of the temperature and its gradient at a certain height. The theory is illustrated by an actual calculation. The result obtained differs considerably from those of previous researches.

551.510.535 130
Statistical Nature of the Ionosphere—Ya. L. Al'pert and A. A. Aynberg. (*Zh. Eksp. Teor. Fiz.*, vol. 21, pp. 389–400; March, 1951.) The electromagnetic field of a single signal reflected from the ionosphere is regarded as the resultant of the main reflection and a large number of "elementary" signals scattered by the ionosphere irregularities. The variations of the latter signals are of a random nature, and they characterize the degree of irregularity of the reflecting layer. Formulas are derived for determining the displacement velocity of the irregularities and the ratio of the energy of the main reflection to that of the scattered reflections. Results of experiments are given showing that reflection from the F_2 layer is nearly always of this statistical character.

551.510.535 131
Comments concerning the Paper 'Fine Structure of the Lower Ionosphere' by R. A. Helliwell, A. J. Mallinckrodt, and F. W. Kruse, Jr.—R. J. Nertney. (*Jour. Geophys. Res.*, vol. 56, pp. 449–451: September, 1951.) From the results of an extensive series of group-height, absorption, polarization, and phase-height measurements on a frequency of 150 kc, a model of the lower ionosphere has been developed which appears to satisfy the theoretical and experimental results on a diurnal and seasonal basis at this frequency. Another interpretation is suggested for the two apparent "reflection" heights noted by Helliwell et al. (1898 of 1951).

551.510.535 132
The D-Layer of the Ionosphere—A. P. Mitra. (*Jour. Geophys. Res.*, vol. 56, pp. 373–402; September, 1951.) A résumé is given of the present state of knowledge of the *D* region, both theoretical and experimental work being summarized. New theoretical studies of the formation and structure of the *D* region are made, based on contemporary knowledge. It is assumed that the *D* region is produced by ionization of O_2 at the first ionization potential. The densities of electrons and ions are calculated for various heights, taking account of the variation of atmospheric temperature and recombination coefficient with height. The distribution does not follow a Chapman law. The electron density increases continuously with height, although the ion density has a maximum value at nearly the same height as the rate of ion production, as in the Chapman distribution. The variation of height of the reflecting layer with solar zenith angle for long and very long waves is explained, and satisfactory agreement is obtained between experimental measurements of reflection coefficient at low frequencies and values derived from the theory.

551.510.535 133
A Note on Certain Characteristics of the Normal *E*-Layer—C. H. Grace. (*Jour. Geophys. Res.*, vol. 56, pp. 452–454; September, 1951.) A method due to Kelso was used in converting *E*-region equivalent heights, measured at Pennsylvania State College, to true heights. The true height of maximum ionization, h_m, was found not to follow Chapman theory except for an hour after sunrise and an hour before sunset. During most of the day, h_m was relatively constant at about 115 to 120 km. For a period centered on noon, the effective recombination coefficient varies with height, in the range 108 to 120 km, in a manner suggestive of pressure dependence.

551.510.535:523.75 134
Ionospheric Effects of Solar Flares—M. A. Ellison. (*Mon. Not. R. Astr. Soc.*, vol. 110, p. 626; 1950.) Summary of paper in *Publications of the Royal Observatory, Edinburgh*, vol. 1, no. 4, 1950. Five types of sudden ionospheric disturbances, noted during 1949 from continuous radio and magnetic records, were studied in relation to flare outbursts as seen simultaneously by *Hα* light in the spectrohelioscope. The effects on the *D* layer are confined to the illuminated hemisphere, and are independent of the location of the flare on the sun's disk, showing the operative radiation to be ultraviolet light. The flares may accelerate cosmic-ray particles, and these may contribute to the ionization towards the end of the disturbances.

551.510.535:621.396.11 135
Lateral Deflection of the Ray on Reflection at an Inhomogeneous Ionosphere Layer—Rawer. (*See* 244.)

551.510.535:621.396.11.029.51 136
Effects of Ionosphere Disturbances on Low Frequency Propagation—Watts and Brown. (*See* 246.)

551.510.535:621.396.11.029.51 137

A Method for Obtaining the Wave Solutions of Ionospherically Reflected Long Waves, Including all Variables and their Height Variation—J. J. Gibbons and R. J. Nertney. (*Jour. Geophys. Res.*, vol. 56, pp. 355–371; September, 1951.) A method of solution is given for the one-dimensional wave equation $\pi''/\pi = -\kappa_0^2 e^2(x)$ which arises in the application of wave theory to the study of the ionosphere. The method is applied to the problem of reflection from the *E* layer at 150 kc, Chapman distributions with various maximum electron densities being assumed. A brief comparison with experimental observations indicates that the calculated absorption is too low. This supports the theory of a *D* region below the *E* layer. The phase heights are fairly consistent with measured group heights, the sense of the difference agreeing with ionospheric dispersion theory.

LOCATION AND AIDS TO NAVIGATION

621.396.93 138

Radio Direction Finding from the Morphological Viewpoint—W. Stanner. (*Elektron Wiss. Tech.*, vol. 5, pp. 135–176; May, 1951.) A comprehensive review of the subject dealing with (a) arrangements for directional reception of medium, long, and ultra-short waves, (b) arrangements for directional transmission, (c) the consol system, (d) various methods of distance measurements, (e) radar equipment, (f) anti-radar methods, (g) geophysical and astrophysical problems of radio df.

621.396.93 139

'Telegon' Direction Finder—W. Runge, M. Strohhacker, and A. Troost. (*Telefunken Ztg.*, vol. 24, pp. 75–81; June, 1951.) Description of Telefunken equipment for ships, with a wavelength range of 85 to 1,530 m. Circular crossed coils 110 cm in diameter, together with a vertical rod 2.6 m long and coincident with the common diameter of the coils, constitute the antenna system. The special features of the Type-E374N receiver are described, and a simplified circuit diagram is given. The total frequency range is covered in four ranges with adequate overlap. For an account of the goniometer used, see 140 below (Troost and Jankovsky).

621.396.93 140

Recent Goniometer Developments—A. Troost and R. Jankovsky. (*Telefunken Ztg.*, vol. 24, pp. 81–85; June, 1951.) Improvements in goniometer design resulting from the use of iron-cored rotors, iron rings, and shields, are reviewed, and a description is given of the goniometer used in the Telegon receiver Type-E374N (139 above). This is of the iron-shielded type with a high-μ rotor core which gives the requisite inductance of the windings with a relatively small number of turns. The residual bearing error is $<0.15°$. Methods of test are outlined.

621.396.93.029.62 141

Ultra-Short-Wave Direction-Finding Installation PV-1B (Direction-Finding Installation at Kloten Airport)—W. Schoeberlein. (*Bull. schweiz. elektrotech. Ver.*, vol. 42, pp. 226–232; April 7, 1951. In German.) Outline of the principles of operation and features of the single-receiver automatic system in operation since 1949 [see 1078 of 1949 (Cleaver)]. Sources of error in the Adcock fixed-antenna system are briefly discussed. Remote indication introduces an error of not more than 1 per cent.

621.396.93.088 142

Polarization Errors of Radio Direction Finders, a Proposed Classification—P. G. Hansel. (PROC. I.R.E., vol. 39, p. 970; August, 1951.) "Primary instrumental polarization errors" are defined to be those which arise when the response of the antenna system to an "unwanted" primary component of the received wave is large. "Secondary polarization errors" are those which arise when the antenna does not respond to the primary "unwanted" component, but takes a bearing on reradiated waves polarized in the "wanted" direction.

621.396.93.089.6 143

The Calibration of Aircraft Direction Finders with Particular Reference to Site Selection—J. H. Moon. (*Marconi Rev.*, vol. 15, pp. 101–112; 3rd Quarter, 1951.) Experiments are described which show that radio waves are refracted as they pass over reinforced-concrete runways. Errors were observed on a loop direction finder at frequencies between 200 and 700 kc, with waves incident horizontally at an angle of 50° to the line of the runway. The maximum error was about 6° at the edge of the runway and 1.5° at a distance of 100 feet from it.

621.396.933 144

A Straight-Line-Flight Indicator for the Pilot of a Radar-Equipped Aircraft—R. C. Richardson. (*Aust. Jour. Appl. Sci.*, vol. 2, pp. 223–234; June, 1951.) An attachment to shoran airborne equipment to assist the pilot in flying along any predetermined straight-line path. Ranges from the ground beacons are transmitted mechanically from the shoran oscillograph to the aircraft equipment. There they are used to set the position of a probe moving parallel to a glass plate coated with a thin layer of metal through which lines, which represent the desired flight paths, are engraved. Divergence of the flight from one of these lines is indicated to the pilot by a center-zero meter. In trials the pilot was able to control the aircraft so that its average departure from a straight-line path was about ±0.02 mile.

621.396.933 145

Modern Radar Landing Systems—M. Sollima. (*HF* (Brussels), no. 10, pp. 270–286; 1951.) The principles and practice of G.C.A. are described and exemplified by the C.F.T.H. installation at Melsbroeck-lez-Bruxelles airport. This uses a precision radar equipment working on a frequency of 9,375 mc with a peak power of 35 kw. The antennas give a horizontal beam 3.5° wide and 0.6° deep for determination of angular elevation, and a vertical beam of width 0.6° and depth 2° for determination of bearing, the energy in secondary lobes being less than 5 per cent of that in the principal lobe. The construction of the variable-width waveguides to secure fine scanning is described, together with ancillary apparatus.

621.396.933 146

Blind-Landing System Type A. B.—A. Moisson. (*Rev. Tech. Comp. franç. Thomson-Houston*, no. 15, pp. 5–18; April, 1951.) A detailed description of the equipment. See also 145 above (Sollima).

621.396.932/.933.24 147

Radio Beacon 'Sol'—M. Borondo. (*Rev. Telecommun.* (Madrid), vol. 6, pp. 23–25; December, 1950.) Short description and theory of the consol system. See also 2912 of 1946 (Clegg) and 2252 of 1948 (Jessell).

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215:546.431-31 148

Optical Absorption and Photoconductivity in Barium Oxide—W. W. Tyler and R. L. Sproull. (*Phys. Rev.*, vol. 83, pp. 548–555; August 1, 1951.) "Measurements of the optical absorption and photoconductivity in BaO single crystals are reported. The threshold photon energy for both processes is 3.8 ev, and the absorption constant at higher energies is at least 10^5 cm^{-1}. A second increase in the absorption constant begins at about 4.8 ev, and is not accompanied by photoconductivity. The observed dependence of photoconductivity on temperature and electric field strength indicates space charge effects in limiting current flow. The onset of absorption at 3.8 ev is thought to be due to the production of excitons, with the accompanying photoconductivity due to thermal dissociation of excitons or exciton ionization of impurity centers."

535.215:546.431-31 149

Photoconductivity and Photoelectric Emission of Barium Oxide—H. B. DeVore and J. W. Dewdney. (*Phys. Rev.*, vol. 83, pp. 805–811; August 15, 1951.) Measurements were made on sprayed coatings of BaO on Ni. The variation of the characteristics with temperature and during activation is shown in diagrams. The results are discussed with reference to energy levels.

535.215+535.37]:546.472.21 150

A Comparative Study of Photoconductivity and Luminescence—R. H. Bube. (*Phys. Rev.*, vol. 83, pp. 393–396; July 15, 1951.) Measurements are reported of the luminescence and photoconductivity of ZnS as functions of (a) operating temperature during excitation, (b) time, and (c) temperature during thermostimulation. The results are discussed. They indicate different mechanisms for the two processes.

535.215+535.37]:546.482.21 151

Electron Mobility and Luminescence Efficiency in Cadmium Sulfide—L. Gildart and A. W. Ewald. (*Phys. Rev.*, vol. 83, pp. 359–363; July 15, 1951.)

535.37 152

Electron Penetration and Scattering in Phosphors—L. R. Koller and E. D. Alden. (*Phys. Rev.*, vol. 83, pp. 684–685; August 1, 1951.) ZnS films ranging in thickness from 0.1 to 0.45μ were deposited on glass and were irradiated at a constant current density of 0.8μ A/cm^2. The luminescence brightness was measured as a function of beam voltage, using a photomultiplier with cathode ray tube display. The results show that nearly 90 per cent of the beam energy is lost in a distance of half the range of the electrons.

535.37 153

Properties of a $CaSO_4$-Mn Phosphor under Vacuum Ultraviolet Excitation—K. Watanabe. (*Phys. Rev.*, vol. 83, pp. 785–791; August 15, 1951.) An investigation of the long-period phosphorescence excited by wavelengths below 1,350Å.

535.37:537.39:546.281.26 154

Injected Light Emission of Silicon Carbide Crystals—K. Lehovec, C. A. Accardo, and E. Jamgochian. (*Phys. Rev.*, vol. 83, pp. 603–607; August 1, 1951.) The yellow light emitted by certain SiC crystals was investigated as a function of temperature and of current through the crystal. The light intensity varies approximately linearly with current, and experiments under pulsed conditions indicate good hf response. The emission spectrum changes considerably with temperature, but only slightly with current density. The mechanism of the light emission is discussed and attributed to the recombination of carriers injected through $p-n$ boundaries. This process is contrasted with emission from phosphors under bombardment.

537.226 155

The Origin of Ferroelectricity—G. A. Smolenski and N. V. Kozhevnikova. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 519–522; February 1, 1951. In Russian.) The deciding factor for the appearance of ferroelectricity is a certain mutual disposition of oxygen octahedra in the crystal. A table of all known substances which exhibit this effect, and of substances in which this effect can be expected at certain temperatures, is included.

537.226 156

Domain Boundary Motion in Ferroelectric Crystals and the Dielectric Constant at High Frequency—C. Kittel. (*Phys. Rev.*, vol. 83, p. 458; July 15, 1951.)

537.226.1:546.431-31 157

The Dielectric Constant of Barium Oxide—R. S. Bever and R. L. Sproull. (*Phys. Rev.*, vol. 83, pp. 801–805; August 15, 1951.) Measurements were made on BaO crystals for the frequency range 60 cps to 60 mc and temperature range −25° to +60°C. The value found for the dielectric constant in these ranges was about 34, rising at the lowest frequencies.

537.311.33 158

Volume Rectification in Zincite—S. R. Khastgir, S. K. R. Tolpadi, and S. C. Mitra. (*Nature* (London), vol. 168, pp. 162–163; July 28, 1951.) Experiments are reported which indicate that the rectification in zincite depends on the volume and not on the area in contact with the electrodes. A theoretical explanation based on the crystal structure is given.

537.311.33 159

Semiconductors—R. L. Ortueta and E. Y. Garrido. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 12–22; December, 1950.) A concise account of the properties of semiconductors, with theory based on energy-level bands. A table and diagrams give the physical constants and rectification characteristics of Ge.

537.311.33 160

The Electrical Conductivity of Simple p-Type Semiconductors—A. Lempicki. (*Proc. Phys. Soc.*, vol. 64, pp. 589–590; June 1, 1951.) An analysis is outlined, which establishes a complete analogy between electron and hole conduction. In both cases conductivity passes through a maximum with increasing temperature.

537.311.33 161

A Study of Thermoelectric Effects at the Surfaces of Transistor Materials—J. W. Granville and C. A. Hogarth. (*Proc. Phys. Soc.*, vol. 64, pp. 488–494; June 1, 1951.) The surfaces of various specimens of Ge and PbS have been explored with a whisker contact, and the character of the conduction mechanism in relatively small regions has been examined by the polarity of (a) the photo-voltaic effect, (b) the rectification, and (c) the thermoelectric effect. Methods (a) and (b) give results characteristic of the bulk material as determined by Hall effect measurements, whereas method (c) only gives results consistent with (a) and (b) for uncontaminated cleaved surfaces or for surfaces which have been etched after polishing. For polished surfaces, anomalous effects are obtained and point to the existence of a layer on the surface of character different from that of the bulk semiconductor.

537.311.33:538.221 162

On the Dispersion of Resistivity and Dielectric Constant of Some Semiconductors at Audiofrequencies—C. G. Koops. (*Phys. Rev.*, vol. 83, pp. 121–124; July 1, 1951.) Semiconducting Ni/Zn ferrites were investigated. A simple model, comprising highly conductive grains separated by layers of lower conductivity, is proposed to explain the dispersion of ac resistivity and apparent dielectric constant. Dispersion formulas are given. The theory is supported by experimental results.

537.311.33:546.289 163

The Effect of Fast Neutron Bombardment on the Electrical Properties of Germanium—J. W. Cleland, J. H. Crawford, Jr., K. Lark-Horovitz, J. C. Pigg, and F. W. Young, Jr. (*Phys. Rev.*, vol. 83, pp. 312–319; July 15, 1951.) The effect of lattice disorders introduced by bombardment is discussed from the theoretical standpoint, and measurements are reported. For *n*-type Ge, the conductivity under bombardment first decreases and then increases; on passage through the minimum, the material is converted to *p*-type. The conductivity of *p*-type material increases under bombardment, tending towards a saturation value. Figures are quoted for the rate per neutron of carrier removal and introduction in *n*- and *p*-type Ge, respectively. Related temperature effects are described. A footnote mentions that for both *n*- and *p*-type Si, bombardment produces only a decrease of conductivity.

537.311.33:546.289 164

Observations of Zener Current in Germanium *p-n* Junctions—K. B. McAfee, E. J. Ryder, W. Shockley, and M. Sparks. (*Phys. Rev.*, vol. 83, pp. 650–651; August 1, 1951.) Junctions in a Ge single crystal were examined using As as the donor impurity and Ga as the acceptor. The reverse I/V characteristic was measured over a 5-decade range of current, and the critical voltage gradient across the junction was studied by observing the variation of capacitance of the junction with reverse voltage. While the slope of the characteristic in the high-current part of the range agrees well with theory, the critical voltage-gradient predicted is about three times the measured value.

537.312.6:621.315.592† 165

Investigation of Semiconductors at High Temperatures, Refractory Thermistors—N'Guyen Thien-Chi and J. Suchet. (*Ann. Radioélect.*, vol. 6, pp. 99–105; April, 1951.) The materials investigated included the normal refractory oxides, MgO, SiO_2, ZrO_2, etc., and mixtures of these with such metals as Ni, Co, Fe, etc., and the semiconducting oxides of Zn, Ti, Ta, and V. The work has resulted in the commercial production of a wide range of stable, robust thermistors for use at 250 to 300°C, possessing high temperature coefficients (5.5 per cent/°C), and of refractory thermistors operating up to 1,100°C, giving much better control than is obtainable with thermocouples or the usual resistive probes.

537.312.8:546.289 166

The Magnetoresistance Effect in Oriented Single Crystals of Germanium—G. L. Pearson and H. Suhl. (*Phys. Rev.*, vol. 83, pp. 768–776; August 15, 1951.) An extensive experimental study of the effect as a function of the orientation of magnetic field and electric current relative to the crystal axes. "The measurements are internally consistent with existing phenomenological theory based on cubic crystal symmetry, in which terms involving the magnetic field to higher than the second order are neglected. It is shown that such deviations as do occur arise from higher terms in the field, since an extension of the phenomenological theory to the fourth order predicts their symmetry. Relations are established between the experimentally observed phenomenological constants and those constants appearing in existing magnetoresistance electronic theories."

538.221 167

On Ferromagnetic States—J. Giltay. (*Appl. Sci. Res.*, vol. B2, no. 3, pp. 199–216; 1951.) Reprint. See 126 of 1951.

538.221 168

Structure and Properties of Ferrites—F. G. Brockman. (*Elec. Eng.*, vol. 70, pp. 489–494; June, 1951.) 1951 AIEE Winter General Meeting paper. Advances in the understanding of ferromagnetic ferrites subsequent to those reported in 660 of 1950 are reviewed. Particular attention is paid to Néel's theory.

538.221:[534.321.9:534.373 169

The Influence of Magnetization on Ultrasonic Attenuation in a Single Crystal of Nickel or Iron-Silicon—S. Levy and R. Truell. (*Phys. Rev.*, vol. 83, pp. 668–669; August 1, 1951.) Report of measurements made using ultrasonic beams of frequency 5 to 50 mc, with constant magnetic fields, of intensities ranging from zero to saturating field, applied both perpendicular to and parallel to the beam direction. At saturation the beam attenuation is low and constant over the frequency range.

538.632:621.315.592† 170

Resistivity and Hall Constant of Semiconductors—C. N. Klahr. (*Phys. Rev.*, vol. 83, p. 460; July 15, 1951.) Correction to paper noted in 2206 of 1951.

539.23:546.26:537.311.32 171

The Conductivity of Thin Carbon Films—A. Blanc-Lapierre, M. Perrot, and N. Nifontoff. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 141–143; July 9, 1951.) Nonlinear films of resistance 1 to 50 megΩ have been found by Perrot and Lavergne to have different resistance properties over different ranges of values of applied voltage. An experimental investigation of carbon films of resistance <0.5 megΩ is reported here which extends the range of the earlier findings. Voltages of 1 to 40 v were used. Results are shown as families of log R/log I curves. Formulas for the various ranges are developed, and values derived for various parameters are tabulated.

539.23:546.59 172

Influence of the Support on the Crystallization of Very Thin Gold Films—A. Colombani and G. Ranc. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 46–48; July 2, 1951.) Continuation of experimental investigation noted in 2739 of 1951. Support materials studied include KCl, KBr, $NaNO_3$, NaCl, rhodoid, plexiglas, and glass.

546.431.824 173

The Lorentz Correction in Hexagonal Barium Titanate—J. R. Tessman. (*Phys. Rev.*, vol. 83, pp. 677–678; August 1, 1951.) An explanation based on the structure is advanced for the experimentally observed absence of ferroelectricity in the hexagonal modification of $BaTiO_3$. See also 2188 of 1950 (Slater).

548.52:546.431-31 174

Growth and Manipulation of Barium Oxide Crystals—R. L. Sproull, W. C. Dash, W. W. Tyler, and A. R. Moore. (*Rev. Sci. Instr.*, vol. 22, pp. 410–414; June, 1951.)

549.514.51:621.396.611.21.002.2 175

Manufacture of Quartz Piezoelectric Crystals—G. Flórez. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 3–22; June, 1951.) A detailed account of all processes from the initial examination of the raw material and determination of the optic axis up to the deposition of electrode material on the faces of crystal plates and their subsequent mounting.

620.197 176

Some Principles and Practices in Tropicalisation—G. A. Durrance, W. J. D. Harrison, and R. T. Lovelock. (*GEC Telecommun.*, vol. 4, pp. 23–32; 1949.) An account of the adverse effects of various climatic conditions on telecommunications apparatus, and of the means of overcoming them by impregnation, metal coating, the use of sealed components, etc. The importance of suitable mechanical design is stressed.

621.315.61 177

Dielectric Properties of Insulators—K. W. Wagner. (*Alta Frequenza*, vol. 20, pp. 3–27 and 68–79; February and April, 1951.) Theories regarding the relation of the dielectric properties to the structure of the material are reviewed. The influence of temperature and humidity is examined. The effects of drying and impregnation on paper are investigated.

666.1.037 178

A Method of Sealing Sapphire to Glass—H. Rawson. (*Jour. Sci. Instr.*, vol. 28, pp. 208–209; July, 1951.) A modification of the method given in 179 below is described, using hf induction heating.

661.1.037:621.383 179

A Method of Sealing Sapphire to Glass and its Application to Infra-red Photocells—R. P. Chasmar, J. L. Craston, G. Isaacs, and A. S. Young. (*Jour. Sci. Instr.*, vol. 28, pp. 206–207; July, 1951.) The requirements of window mate rials for PbTe cells are discussed; artificially grown sapphire is both suitable and readily available. In the sealing method described, the glass is heated to a high temperature so that it wets the sapphire.

533.5 180

La Technique du Vide [Book Review]—M. Leblanc. Publishers: Armand Colin, 188 pp. (*Bull. Soc. franç. élect.*, vol. 1, p. 252; May, 1951.) ". . . this little treatise will be useful to all those . . . who are interested in vacuum technology."

621.315.616:058 2 181

Modern Plastics Encyclopedia and Engineer's Handbook, 1951 [Book Notice]—Publishers: Plastics Catalogue Corporation, New York, N. Y., 15th edn, 636 pp. A record of developments in plastic materials, machinery, engineering, techniques and applications, with a special section, "Plastics in Defense," giving information on applications of plastics by the U.S. Services and related government bureaus, and on the location of the relevant government offices. The directory section lists all U.S. manufacturers of plastic materials, machinery, and equipment.

MATHEMATICS

51.537.315.6 182

Mathematical Tools for the Solution of Problems of the Distribution of Potential—J. J. R. Moral. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 38–41; June, 1951.) In many cases such problems are most conveniently dealt with in cylindrical or spherical co-ordinates. Laplace's equation is transformed to spherical co-ordinates, and a solution is obtained in terms of Legendre functions.

512.831:621.392.5 183

Transformation of the Quadripole-Network Matrix to the Diagonal Form—Schulz. (*See* 68.)

681.142 184

The New Universal Digital Computing Machine at the University of Manchester—T. Kilburn. (*Nature* (London), vol. 168, pp. 95–96; July 21, 1951.) Detailed description of a machine recently put into service, incorporating improvements over the experimental model described in 133 of 1950.

MEASUREMENTS AND TEST GEAR

537.54:621.396.822 185

Gaseous Discharge Super-High-Frequency Noise Sources—H. Johnson and K. R. De Remer. (Proc. I.R.E., vol. 39, pp. 908–914; August, 1951.) The positive column of a discharge in argon provides a standard noise source at microwave frequencies, the noise temperature being about 15.5 db above 290°K. The discharge tube is inserted diagonally across a waveguide, a good termination being obtained without using tuned elements. Performance data are given for the frequency band 2.6 to 26 kmc. The effects of varying the discharge parameters are discussed.

621.3.011.2(083.74):621.315.212 186

Coaxial Impedance Standards—R. A. Kempf. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 689–705; July, 1951.) "The calibrations of bridge networks used in developmental tests on coaxial cable are obtained by comparison of the networks with calculable standards of impedance consisting of a group of short-length precision copper coaxial lines. The standards are calculable by reason of the availability of precise formulas relating the distributed primary constants to the measurable physical constants and dimensions of the coaxials. This paper outlines the constructional problems and design features of a group of such standards of impedance which provide a range of values over a broad band of frequencies."

621.317.3:621.314.6 187

Measurement of the Temperature Dependence of the Voltage Drop and of the Backward Current in Dry Rectifiers. Measurement Methods and Apparatus—K. Maier. (*Arch. tech. Messen*, no. 185, pp. T71–T72; June, 1951.) Measurements of the static characteristic are carried out with direct voltage, but most other measurements with pulsed steady direct voltage. A method for the simultaneous measurement of the dynamic forward and backward currents is described.

621.317.3:[621.396.615.029.3+534.232 188

The Generation of Audio-Frequency Voltages for Measurement Purposes—O. Schmid. (*Funk u. Ton*, vol. 5, pp. 133–142; March, 1951.) General description of the principles of different types of tone source, including electromechanical devices and, particularly, self-excited oscillators. Basic circuits are shown for the RC-, glow-discharge-tube- and heterodyne-type generators. Their features and advantages are discussed. For high precision, tuning-fork or quartz generators are required.

621.317.3:621.396.822 189

The Measurement of Fluctuation Noise by Diode and Anode-Bend Voltmeters—R. E. Burgess. (*Proc. Phys. Soc.*, vol. 64, pp. 508–518; June 1, 1951.) "The response of a diode voltmeter to a cw sine-wave signal, and to noise, is analyzed for three typical diode characteristics: discontinuous linear, discontinuous parabolic, and exponential. In the first type, the indication is proportional to the rms value of the input voltage, but the constant of proportionality is different for noise and for cw, and depends upon the ratio of the load resistance to diode resistance. For small input signals, the curvature of the diode characteristic is important, and the voltmeter tends to a square-law behavior, with the result that cw, noise or a mixture of cw, and noise, can all be measured in terms of a single calibration. The input conductance of a diode voltmeter is, in general, greater for noise than for cw. It can be very much greater in practical conditions, and this represents a possible source of error in the measurement of noise. The response of the anode-bend voltmeter to cw and to noise is considered for a discontinuous parabolic and for an exponential characteristic. The departure of the cw calibration from a square law, and the difference between cw and noise calibrations, are evaluated in terms of the characteristics."

621.317.318.087:621.395.625.3 190

Magnetic-Tape Recording of Electrostatic Field Changes—N. D. Clarence and D. J. Malan. (*Proc. Phys. Soc.*, vol. 64, p. 529; June 1, 1951.) Brief description of a method of photographic recording which reduces film consumption.

621.317.335.3.087.4† 191

Automatic Measurement Computation and Recording of Dielectric Constant and Loss Factor against Temperature—E. B. Baker. (*Rev. Sci. Instr.*, vol. 22, pp. 376–383; June, 1951.) The terms ϵ' and ϵ'' are recorded as a function of temperature from −23°C to +150°C, at given frequencies between 50 cps and 600 kc, for solids in the form of thin disks. A modified Schering bridge is used. Circuit details are given of the associated equipment.

621.317.337.029.64† 192

Detection of Small Variations in the Quality Factors of a Cavity Resonator—R. Malvano and M. Panetti. (*Alta Frequenza*, vol 19, pp. 231–243; October–December, 1950.) Description and analysis of a dynamic method in which an FM signal is applied to the cavity input. A crystal detector circuit in the output is used to measure the relative variations in output voltage. These are proportional to $\Delta Q/Q$. Errors introduced by the generator and detector are negligible. An example is given of the application of the method in investigation of the absorption spectra of paramagnetic substances.

621.317.35 193

Harmonic Analysis of a Distorted Oscillation after Amplitude Limiting—M. Kolscher. (*Arch. elekt. Übertragung*, vol. 5, pp. 293–299; June, 1951.) In FM oscillations, distortion can be reduced by amplitude limiting. For the case of a sinusoidal interfering oscillation, this effect has hitherto only been investigated theoretically under particular assumptions concerning the frequency ratio and amplitude ratio of the main and interfering oscillations, and the shape of the limiter characteristic. The dependence of the results on the frequency ratio is here made clear, and a new method of analysis is developed which is applicable for all amplitude ratios and characteristics. Numerical calculations for a typical case show what limiter characteristics give the best interference elimination, and what characteristics can be used without the interference becoming appreciably greater.

621.383.27†:621.317.42 194

Investigation of the Conditions for Use of Kubetski's Mosaic Multiplier as an Indicator of the Magnetic Field—Krasnogorskaya. (*See* 298.)

621.317.42:621.385.833 195

Inductance Method for studying the Field Pattern on the Axis of a High-Power Magnetic Electron Lens—C. Fert and P. Gautier. (*Compt. Rend. Acad. Sci.* (Paris), vol 233, pp. 148–150; July 9, 1951.)

621.317.43 196

Measurements on Coils with Iron Cores at Frequencies between 1 c/s and 1 mc/s—H. Wilde. (*Arch. tech. Messen*, no. 183, pp. T39–T40; April, 1951.) Three factors of which account should be taken are hysteresis, magnetic after-effect, and eddy currents. Effective means described for determining these are, respectively: permeability measurement at very low frequencies using a type of mutual-inductance bridge to eliminate the effect of copper losses; extrapolation to zero field strength of the inductance characteristic measured by a Maxwell-Wien bridge; a similar procedure carried out at high frequency using a capacitance-tuned resonance bridge, or one with a folded-slide-wire balancing element.

621.317.44 197

Principles of the Design of Pulse-Transformer Laminations: Pulse Hysteresis Meter—P. Bouvier and B. Daugny. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 15, pp. 51–67; April, 1951.) A hysteresis meter is described for use in investigating the properties of magnetic cores under operating conditions such as obtain in the pulse transformers used in radar equipment. It consists of a modulating unit of peak power 1 MW producing a 0.5-μs pulse across the terminals of a winding on the core tested, and a measurement unit enabling the shape of the voltage pulse, the magnetic field, the magnetic induction, and the hysteresis loop to be obtained.

621.317.7 198

Sensitivity of Electrical Measurement Apparatus and its Enhancement—F. Moeller. (*Arch. tech. Messen*, no. 185, p. T68; June, 1951.)

621.317.7:537.312.6:621.315.592† 199

The Use of Thermistors for Temperature Compensation in Precision Measurement Apparatus—N'Guyen Thien-Chi and J. Suchet. (*Ann. Radioélect.*, vol. 6, pp. 106–108; April, 1951.) The increase of resistance with temperature of the copper coils of electrical measuring apparatus may be compensated by the use in series or in parallel of thermistors possessing negative temperature coefficients.

621.317.7:621.396.619.2 200

A New Modulation Meter—H. Müller. (*Telefunken Ztg.*, vol. 24, pp. 55–60; March, 1951.) A meter suitable for monitoring the modulation of broadcasting transmitters and magnetic-tape, disk, or sound-film equipment is described. The amplitude range is −45 db to +5 db, with an approximately uniform logarithmic scale, and the frequency range is 40 to 15,000 cps. A separate indicating instrument is provided in addition to that in the apparatus. Calibration at 0, 1, and 100 per cent modulation is effected by stabilized voltages derived from the supply unit.

621.317.7.085.34:621.383 201
Some Points in the Design of Optical Levers and Amplifiers—R. V. Jones. (*Proc. Phys. Soc.*, vol. 64, pp. 469–482; June 1, 1951.) Sources of instability in optical levers using photoelectric amplifiers are discussed, and the design of a high-precision instrument is described.

621.317.7.088.6 202
Compensation of A.C. Instruments for Variations in Frequency—J. H. Miller. (*Elec. Eng.*, vol. 70, pp. 494–497; June, 1951.) 1951 AIEE Winter General Meeting paper. A general solution is presented for compensation over the frequency range up to 5 kc. The series resistance of the meter is shunted by a capacitor whose value is rigorously calculated.

621.317.727.088 203
Leakage, Stray Capacitance and Inductance in Potentiometer Comparison Measurements with Direct and Alternating Currents—J. H. Gosselin. (*Rev. gén. élect.*, vol. 60, pp. 244–253; June, 1951.) The effects of insulation faults in dc potentiometers are examined. These arise notably in the auxiliary circuit connections, and can be reduced by suitably locating the galvanometer and switch. In ac potentiometers, stray capacitance gives rise to leakage currents in quadrature with intercircuit voltages, and inductance effects due to ambient magnetic fields are also significant. Compensation methods are described for reducing relative errors resulting from these effects to the order of 10^{-5}.

621.317.733:621.317.382 204
A Self-Balancing Microwave Power Measuring Bridge—L. A. Rosenthal and J. L. Potter. (PROC. I.R.E., vol. 39, pp. 927–931; August, 1951.) 1950 URSI-IRE Meeting paper, "The basic techniques of microwave power measurement are discussed, together with principles of self-balancing bridges for that application. A method of measuring the power change by means of a nonlinear "slave" bridge is presented. Temperature compensation problems as applied to the meter under consideration are discussed. The complete power meter is considered along with operating procedure "

621.317.733.011.21† 205
Impedance Measurement Bridge for Determination of Aerial Impedances in the Medium-Wave Band—K. Fischer. (*Elektrotech. u. Maschinenb.*, vol. 68, pp. 249–252; May 15, 1951.) The instrument is designed particularly for measurement of impedance in which the reactive component is large, the resistive component being balanced by capacitive tuning. Particular attention is paid to screening and compensation of stray capacitances. Tuning reactors are contained in a separate screened unit. The bridge is fed from a tone-modulated hf source enabling an aural-null indication of balance to be used.

621.317.733.011.3/.4 206
An Easily Operated Precision Measurement Bridge for Self-Inductors and Capacitors with Losses—E. Sorg. (*Funk u. Ton*, vol. 5, pp. 202–209; April, 1951.)

621.317.733.025:621.385.832 207
The C. R. Tube as Indicator for A. C. Bridges—R. Oetker. (*Frequenz*, vol. 5, pp. 33–38; February, 1951.) Conditions are investigated for the use of cathode ray tubes as indicators of complex variables in familiar bridge circuits. Particular attention is given to sources of error.

621.317.74:621.397.5 208
Transmission Measuring System—O. D. Engstrom. (*Bell Lab. Rec.*, vol. 29, pp. 264–268; June, 1951.) The apparatus measures phase distortion in the transmission of video signals along transmission lines. The calibrating signal consists of two sine waves with a constant frequency difference of 200 kc, the frequency range extending upwards to 3 mc. The phase shift of the resulting 200-kc envelope for any frequency pair is compared oscillographically with the phase of the envelope resulting from a standard frequency pair of 892 kc and 1,092 kc.

621.392.001.4 209
Determination of Amplitude and Phase Characteristics of Linear Networks by means of Square Waves—J. Müller. (*Fernmeldetech. Z.*, vol. 4, pp. 211–220; May, 1951.) Detailed theory of the method is given. The construction of a square-wave generator for the range 50 cps to 6 mc is described, and also the method adopted for its calibration. The great advantage of the method is that the amplitude and phase characteristics of a network can be determined over a relatively wide band of network frequencies by means of a single square-wave test frequency. Auxiliary equipment comprises a wide-band cro with timebase circuit for a frequency of 1 mc and deflection proportional to time, and a wide-band amplifier. Examples of the application of the method include (a) anode circuit of an amplifier, (b) two-stage low-pass filter, (c) lf transformer, (d) magnetophone, (e) band-pass filters. See also 210 below (Meyer-Eppler).

621.392.001.4 210
Measurement of the Frequency Characteristics of Linear Systems by Single or Repeated Switching Processes—W. Meyer-Eppler. (*Fernmeldetech. Z.*, vol. 4, pp. 174–182; April, 1951.) Nonstationary methods of measurement are discussed, and methods using single pulses, periodic series of pulses, or square waves are described, particular reference being made to low-pass, high-pass, and band-pass filters, and resonators.

621.396.621.001.4 211
Systematic Analysis of the Properties of Radio Receivers—Fromy. (*See* 252.)

621.397.6.001.4 212
Traveling-Wave Amplifier Measurements—F. E. Radcliffe. (*Electronics*, vol. 24, pp. 110–111; August, 1951.) Circuits and procedures for transmission and impedance measurements on wide-band amplifiers are described. Frequency sweeps of 500 mc at 4 kmc with a repetition rate of 60 cps are used, and measurements are accurate to within 2 per cent.

621.317:621.314.6 213
Gleichrichtermesstechnik (Rectifier Measurement Technique) [Book Review]—H. F. Grave. Publishers: Geest and Portig, Leipzig, Germany, 1950, 227 pp., 27 DM. (*Elektron Wiss. Tech.*, vol. 5, p. iv; April, 1951.) One of a series of monographs which deals with the general theory of different types of rectifier and their application in measurement circuits.

OTHER APPLICATIONS TO RADIO AND ELECTRONICS

621.316.578.1 214
A Millisecond Timer—J. M. Sturtevant. (*Rev. Sci. Instr.*, vol. 22, pp. 359–362; June, 1951.) The circuits described produce a time delay of up to 10 seconds, accurate to within 1 ms, and can be used as a stop-clock reading to the same accuracy with manual or electrical control. A test pulse generator for adjustment purposes is also described.

621.365.54/.55† 215
Fundamentals of High-Frequency Heating—G. Lang. (*Bull. schweiz. elektrotech. Ver.*, vol. 42, pp. 289–303; May 5, 1951. In German.) After a short explanation of terminology and an historical review, the thermal aspect of the subject is discussed. The principle of induction heating is explained, and the importance of the depth of penetration is stressed. Dielectric heating is next considered. Different types of generator are reviewed, and their advantages and disadvantages for particular applications are noted.

621.365.54† 216
Additions to the Generalized Theory of Power Transfer by Induction—M. van Lancker. (*Bull. Soc. franç. élect.*, vol. 1, pp. 241–251; May, 1951.) Further development of the theory published in 1950 (3112 of 1950), leading to a complete solution of the problem.

621.38.001.8 217
A Judgment Box—W. H. Alexander. (*Electronic Eng.*, vol. 23, pp. 256–257; July, 1951.) Each factor influencing a decision which must be made is assigned a weight. By summation of voltages corresponding to the ten weighted factors, their weighted mean is shown on a meter.

621.38.001.8 218
Electronic Protection for War Plants—R. Y. Atlee. (*Electronics*, vol. 24, pp. 96–101; August, 1951.) Photoelectric, capacitance, and acoustic alarm systems are discussed. Illustrations of recent installations, including boundary-protection systems, are given.

621.383.001.8 219
Industrial Tristimulus Color Matcher—G. P. Bentley. (*Electronics*, vol. 24, pp. 102–103; August, 1951.) Construction and circuit details are given of a sensitive flicker-photometer instrument incorporating a photomultiplier.

621.384.6† 220
The Yale Linear Electron Accelerator—H. L. Schultz and W. G. Wadey. (*Rev. Sci. Instr.*, vol. 22, pp. 383–388; June, 1951.) Design, construction, and performance of the accelerator are described. Independent cavity resonators driven by 500-kw triode power amplifiers are used, and large currents of 10-MeV electrons are produced.

621.384.6† 221
Operation of a Six-MeV Linear Electron Accelerator—G. E. Becker and D. A. Caswell. (*Rev. Sci. Instr.*, vol. 22, pp. 402–405; June, 1951.) A traveling-wave type accelerator is described which uses a single magnetron as power source and operates at a wavelength of 10.5 cm, with a peak power output of 0.9 MW.

621.384.611† 222
On the Radio-Frequency Requirements of High-Energy Electron Synchrotrons—T. R. Kaiser. (*Proc. Phys. Soc.*, vol. 64, pp. 502–507; June 1, 1951.) Taking into account the finite rate of rise, the required rf voltage for efficient transition from betatron to synchrotron operation is determined for two machines under construction.

621.384.612.2† 223
The Synchrocyclotron at Amsterdam: Part 3—The Electromagnet—F. A. Heyn. (*Philips Tech. Rev.*, vol. 12, pp. 349–364; June, 1951.) Design, construction, and performance of the magnet are described. A method has been found for stabilizing the particle orbits. As a result, 28-MeV deuterons can be produced instead of the planned 25-MeV deuterons. Parts 1 & 2: 2790 of 1951.

621.385.38.001.8 224
Measurement of Ignition Delay in Internal-Combustion Engines—J. Kwasieborski. (*HF* (Brussels), no. 10, pp. 287–292; 1951.) Two thyratrons, fired by impulses derived from the phenomena investigated, control the

functioning of a pentode the anode current of which is measured.

621.385.38.001.8 225

A Simple Photostimulator—R. G. Bickford and W. T. Moffet. (*Electroencephalography clin. Neurophys.*, vol. 3, pp. 251–252; May, 1951.) A pulse generator using a thyratron triggers a flash tube, giving 0.5 to 50 flashes per second. Constructional component and details are given.

621.385.833 226

German Society for Electron Microscopy: Third Annual Conference—(*Nature* (London), vol. 168, pp. 70–71; July 14, 1951.) Brief report of conference held on May 18 to 20, 1951 in Hamburg, at which about eighty papers dealing with the electronoptics and applications of electron microscopes were presented.

621.385.833 227

On the Dioptrics of Electrostatic Electron Lenses—G. Wendt. (*Z. angew. Phys.*, vol. 3, pp. 219–225; June, 1951.) By inversion of the function expressing the potential distribution on the axis of the lens, characteristic relations are derived for the electron paths in basic types of accelerating lens and in the three-element independent lens. Substitution of axis potential for axial co-ordinate in the differential equation for the paraxial rays gives an equation of the Heun type witn polynomial coefficients, the solutions of which are Riemann functions of second order with four branch points. For a similar method see *Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 734–735; February 20, 1950. (Régenstreif.)

621.385.833 228

New Formulae for the Third-Order Aberrations of Magnetic Lenses—R. Sturrock. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 146–147; July 9, 1951.)

621.385.833:621.317.42 229

Inductance Method for studying the Field Pattern on the Axis of a High-Power Magnetic Electron Lens—Fert and Gautier. (*See* 195.)

621.387.4† 230

A Localizing Geiger Counter—S. G. F. Frank. (*Phil. Mag.*, vol. 42, pp. 612–615; June, 1951.) A method is described for locating ionizing particles, based on the fact that the discharge in a Geiger counter spreads along the wire at a constant speed.

621.387.462† 231

Behavior of Space Charge in Diamond Crystal Counters under Illumination: Part 1—A. G. Chynoweth. (*Phys. Rev.*, vol. 83, pp. 254–263; July 15, 1951.)

621.387.462† 232

Behavior of Space Charge Free Diamond Crystal Counters under Beta-Ray Bombardment: Part 2—A. G. Chynoweth. (*Phys. Rev.*, vol. 83, pp. 264–268; July 15, 1951.)

621.39.001.11:6 233

Cybernetics—J. Loeb. (*HF* (Brussels), no. 10, pp. 257–269, 286; 1951.) The development of the subject as a theory of telecommunication signals is outlined, and its application to calculating machines and to servomechanisms is described, together with wider applications in physiology, psychology, and the political economy.

681.142 234

Step Multiplier in Guided Missile Computer—E. A. Goldberg. (*Electronics*, vol. 24, pp. 120–124; August, 1951.) A 4,000-tube analogue computer for simulating the characteristics of a missile-target system includes a high-precision multiplier comprising reversible binary counter and relay-operated conductance networks for maintaining orthogonality between the reference-axis systems of earth and missile as the computation proceeds.

681.142 235

Some Aspects of Electrical Computing—J. Bell. (*Electronic Eng.*, vol. 23, pp. 213–216 and 264–269; June and July, 1951.) An account of the application of various devices in the design of equipment for conversion from polar to cartesian co-ordinates, or vice versa, differentiation, integration, multiplication, division, and computation of variable functions relating to ballistics. These devices include (a) the ipot (inductive potentiometer), a toroidal coil very uniformly wound on a large ring core and provided with sliding contacts, and (b) the magslip resolver, a special type of servomechanism.

PROPAGATION OF WAVES

538.566 236

'Internal' Reflection in a Stratified Medium with Application to the Troposphere—G. Eckart and T. Kahan. (*Arch. Elektrotech.*, vol. 40, no. 2, pp. 133–140; 1951. German version of 718 of 1951.

536.566 237

Properties of Inhomogeneous Plane Electromagnetic Waves—Bonfiglioli. (*See* 113.)

538.566:535.42 238

Diffraction of Electromagnetic Waves near the Earth's Surface in an Optically Inhomogeneous Medium—R. Schachenmeier. (*Arch. elekt. Übertragung*, vol. 5, pp. 267–272; June, 1951.) Making use of the laws of geometrical optics and non-Euclidean geometry, the problem is transformed to render it amenable to calculation by methods already developed. For practical calculations, the actual curvature of the earth's surface is reduced by that of the ray path when the ordinary laws of diffraction can be used.

621.396.11+621.392.2.09 239

Characteristic Impedance, Power, Voltage and Current in Transmission along Lines, in Waveguides and in Free Space—Zinke. (*See* 31.)

621.396.11 240

A V.H.F. Field-Strength Survey on 90 Mc/s—H. L. Kirke, R. A. Rowden, and G. I. Ross. *Proc. IEE*, vol. 98, pp. 343–359; discussion, pp. 378–382; September, 1951.) The effects of the height of the site and of the transmitting antenna, and the profile of the transmission path, are discussed in relation to field-strength measurements for a transmitter at Wrotham, Kent. In builtup areas the variations are less for horizontal than for vertical polarization. In hilly country the minimum field strength occurs on the near-side slope of a valley, and is lower for horizontal than for vertical polarization. Measurements along radials show that, in general, signals of these two polarizations are propagated equally well. By extrapolating from these data, a field-strength contour map of S.E. England has been prepared for a transmitter power of 25 kw and high-gain antenna on a 500-foot mast.

621.396.11 241

The Propagation of Metre Radio Waves beyond the Normal Horizon: Part 1—Some Theoretical Considerations, with Particular Reference to Propagation over Land—J. A. Saxton. (*Proc. IEE*, vol. 98, pp. 360–369; discussion, pp. 378–382; September, 1951.) The relative importance of abnormally high refraction near the surface of the earth, and of reflection from high-level inversion layers, is investigated and illustrated by examples. Of the two mechanisms, the latter is the more likely to give abnormally high field strength at ranges of a few hundred kilometers, especially for low terminal heights. Consideration is also given to the effects of scattering by turbulent eddies in the atmosphere. This is of less importance than other mechanisms of propagation up to distances of 250 km.

621.396.11 242

The Propagation of Metre Radio Waves beyond the Normal Horizon: Part 2—Experimental Investigations at Frequencies of 90 and 45 Mc/s—J. A. Saxton, G. W. Luscombe, and G. H. Bazzard. (*Proc. IEE*, vol. 98, pp. 370–378; discussion, pp. 378–382; September, 1951.) The statistical distribution of quasi-peak field strength as a function of time was determined for propagation over two paths of respective lengths 110 and 270 km at 90 mc, and over one of length 160 km at 45 mc, for a period of two years. The observed field strengths often considerably exceeded the values corresponding to standard atmospheric refraction. Some degree of correlation was found with meteorological data obtained from routine radiosonde ascents. Use of the results in planning vhf broadcasting services is outlined. Part 1: 241 above.

621.396.11:551.510.535 243

Vertical Propagation of Electromagnetic Waves in the Ionosphere—M. N. Saha, B. K. Banerjea, and U. C. Guha. (*Proc. Nat. Inst. Sci.* (India), vol. 17, pp. 205–226; May–June, 1951.) A discussion of the equations for vertical propagation in the ionosphere is given in standardized notation. The electric vector components E_x and E_y are coupled by polarization terms ρ_1 and ρ_2 which are functions of geomagnetic latitude and height. The propagation vectors associated, respectively, with the ordinary and extraordinary waves are given by $(E_x+i\rho E_y)/\sqrt{+\rho^2}$ for two different values of ρ, and are governed by two refractive indices q_0 and q_e and a coupling term ϕ. The five quantities needed to define wave propagation completely are ρ_1, ρ_2, $\dot{\phi}$, q_0, and q_e. The first three of these are discussed here; the last two have been dealt with by Booker (see e.g., 355 of 1935) and others. For F-layer propagation, $\dot{\phi}$ can be neglected everywhere except very near the geomagnetic poles. E-layer propagation is more difficult to calculate.

621.396.11:551.510.535 244

Lateral Deflection of the Ray on Reflection at an Inhomogeneous Ionosphere Layer—K. Rawer. (*Z. angew. Phys.*, vol. 3, pp. 226–227; June, 1951.) General and numerical calculation of azimuth variation caused by lateral ray displacement, based on particular ionosphere models. The deflection is in the direction of stronger ionization, and angular displacement of nearly 10° occurs, corresponding to an increase of 10 per cent in the muf. Good agreement with observations is noted.

621.396.11.012.3 245

Field Power Conversion—R. E. Perry (*Electronics*, vol. 24, p. 134; August, 1951.) A chart is shown for converting propagation data from field-strength to power-density values.

621.396.11.029.51:551.510.535 246

Effects of Ionosphere Disturbances on Low Frequency Propagation—J. M. Watts and J. N. Brown. (*Jour. Geophys. Res.*, vol. 56, pp. 403–408; September, 1951.) Some photographic records of ionospheric reflections obtained with a transmitter giving a peak pulse power of 900 kw on frequencies of 50, 100, and 160 kc, are reproduced and discussed. The disturbed appearance of the E-region night-time echoes is found to be a sensitive indication of storminess. In the daytime the effect of storms is to cause an increase in absorption, with a corresponding weakening or fade-out of E echoes. A graph is given showing correlation between magnetic activity and night-time disturbance of the E layer.

621.396.11.029.51:551.510.535 247

A Method for Obtaining the Wave Solutions of Ionospherically Reflected Long Waves, including all Variables and their Height Variation—Gibbons and Nertney. (*See* 137.)

RECEPTION

519.272.15:621.39.001.11 248

The Electronic Correlator—H. Doizelet. (*Rad. Tech. Dig.* (France), vol. 5, no. 4, pp. 187–200; 1951.) The problem of improving amplifier sensitivity in spite of irreducible background noise is considered in relation to statistical information theory. The autocorrelation function for noise voltage vanishes for large values of the time interval, while the autocorrelation function for a periodic voltage is itself periodic. This affords a method for separating the two voltages. The operation of the correlator constructed at the Massachusetts Institute of Technology is described. See also 730 of 1951 (Lee, Cheatham, and Wiesner).

621.396.61/.62:621.396.67 249

Transmission and Reception of Circularly Polarized Microwaves with a Common Aerial—Ruppel. (*See* 281.)

621.396.62+621.395.625]:061.4 250

The Radio Exhibition at the Vienna Spring Fair—(*Radio Tech.* (Vienna), vol. 27, pp. 195–198; April, 1951.) Review of portable receivers and magnetophone recording equipment exhibited.

621.396.621 251

Short-Wave Receivers—H. Flicker and J. Hacks. (*Telefunken Ztg.*, vol. 24, pp. 27–38; March, 1951.) An account of present-day technique used in the construction of commercial SW receivers. The properties that can reasonably be expected in a receiver are considered, and also how far such requirements are mutually exclusive or concordant. Technical data for twelve modern receivers of German, French, British, and American manufactures are tabulated. The propagation of short waves is discussed in relation to receiver characteristics.

621.396.621.001.4 252

Systematic Analysis of the Properties of Radio Receivers—É. Fromy. (*Onde élect.*, vol. 31, pp. 210–222 and 282–291; May and June, 1951.) A technique of receiver performance evaluation is described which overcomes the inadequacy of the usual specifications of sensitivity. For the receiver hf amplifier, a cw source is used to derive a family of curves, relating antenna input voltage to detector output, for gradually increasing values of amplification. This is linked with a corresponding family of curves for set noise as a function of input voltage. The af amplifier, with gain control at maximum, and the controls in earlier stages adjusted so that set noise at the output terminals is inappreciable, is treated similarly, its characteristics being obtained by varying the voltage applied to the receiver input terminals. The interpretation of the performance characteristics so obtained is discussed, and examples are given.

621.396.621.54:621.3.012.3 253

New Diagrams for 'Ganging' Calculation—Kerbel. (*See* 53.)

621.396.622+621.396.619.2 254

Modulators, Frequency Changers and Detectors using Rectifiers with Frequency-Dependent Characteristics—Tucker. (*See* 282.)

621.396.622 255

The Relative Advantages of Coherent and Incoherent Detectors: A Study of their Output Noise Spectra under Various Conditions—R. A. Smith. (*Proc. IEE*, vol. 98, pp. 401–406; September, 1951.) Summary of IEE Monograph No. 6. Comparison is made of output spectra obtained with incoherent square-law and linear detectors on the one hand, and coherent homodyne and commutator detectors on the other hand, for an input comprising sinusoidal signal plus noise. Input noise from several common types of filter is considered. Curves give the signal-to-noise ratio at the output when no further bandwidth limitation is applied, and when the detector is followed by a narrow-band filter. Considerable advantage can be obtained with a coherent detector when the input signal-to-noise ratio is somewhat less than unity.

621.396.8.029.45 256

Overseas Reception on Medium Waves—K. M. Schwarz. (*Radio Tech.* (Vienna), vol. 27, p. 224; May, 1951.) Near-East stations often heard in Europe are listed. Reception of medium-wave programs from South America is subject to long-period fading (up to several minutes), but is considerably more regular than for SW stations.

621.396.823 257

Radio Influence Tests in Field and Laboratory—500-kV Test Project—G. D. Lippert, W. E. Pakala, S. C. Bartlett, and C. D. Fahrnkopf. (*Elec. Eng.*, vol. 70, pp. 481–486; June, 1951.) A 1951 AIEE Winter General Meeting paper. Report of an investigation of the effect on the unwanted rf radiation from 500-kv lines of weathering, precipitation and variation of voltage, frequency, and conductor diameter.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 258

A History of the Theory of Information—E. C. Cherry. (*Proc. IEE*, vol. 98, pp. 383–393; September, 1951.) Fundamental points in human communication systems which have been summarized by precise mathematical theory are mentioned. The need for economy, which appeared as telegraphy and telephony developed, led to systems of signal compression and early theories of communication, later extended to express "information" quantitatively. Definite accomplishments in the mechanization of processes analogous to thought, as embodied in calculating machines and servomechanisms, are outlined. Evidence from the whole field of scientific observation supports the view that "information plus entropy is an important invariant of a physical system."

621.395.44 259

German Carrier-Frequency System V60—W. Zerbel. (*Fernmeldetech. Z.*, vol. 4, pp. 193–201 and 268–274; May and June, 1951.) The carrier-frequency arrangements which will be used by the German Post Office on its new long-distance cable network are described, with particular reference to the electrical aspects (pre-grouping or pre-modulation system for sixty channels) and the practical construction.

621.395.665.1 260

Instantaneous Compandors—C. O. Mallinckrodt. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 706–720; July, 1951.) Discussion of the theory of the "instantaneous" type of compandor, and evaluation of the noise advantage when instantaneous companding is applied to telephone channels.

621.396.5:621.396.931 261

Experimental Radio-Telephone Service for Train Passengers—N. Monk. (Proc. I.R.E., vol. 39, pp. 873–881; August, 1951.) A series of stationary transmitters and receivers working in the 35-to 45-mc band, and situated along the 440-mile railway from New York to Buffalo, provide links between trains and the normal telephone system. Train attendants were formerly required, but experiments with coin-boxes have been successful. In this case the frequency band 152 to 162 mc is used.

621.396.61/.621]:621.396.931 262

New Equipment for V.H.F. Radio-Telephone Systems—(*GEC Telecommun.*, vol. 3, no. 1, pp. 5–14; 1948.) AM and FM transmitter-receivers for mobile and central stations are described. All units are crystal controlled and will operate at any frequency in the bands allocated between 30 and 184 mc. Miniature tubes and components are used, enabling the complete equipment for one station to be contained in a box 8 inches×18 inches×8 inches. Public-address facilities are incorporated.

621.396.61/.621]:621.396.931 263

High-Frequency Radio Transportable Transmitter and Receiver—(*GEC Telecommun.*, vol. 4, no. 1, pp. 12–22; 1949.) A description of apparatus suitable for cw or mcw telegraphy and AM telephony; the weight of which is 60 lbs. Power supply may be at 12 or 24 v dc, or from 50-cps mains. The transmitter, covering the frequency range 2 to 9.1 mc, comprises a crystal oscillator and modulated amplifier stages, and radiates 25 w on cw telegraphy. The modulation amplifier may be used alternatively for public-address purposes. The superheterodyne receiver covers the band 2 to 20 mc.

621.396.619.14 264

Product Phase Modulation and Demodulation—D. B. Harris. (Proc. I.R.E., vol. 39, p. 907; August, 1951.) Correction to paper noted in 3159 of 1950.

621.396.65 265

New Directional [radio] Links—J. Kornfeld. (*Radio Tech.* (Vienna), vol. 27, pp. 164–167; April, 1951.) Review of American and European wide-band relay networks for long-distance communications and television services.

621.396.65 266

Application of Microwave Channels—R. C. Cheek. (*Elec. Eng.*, vol. 70, pp. 500–503; June, 1951.) A 1951 AIEE Winter General Meeting paper. Relevant factors in the design of microwave communication systems are considered, including the selection of frequency band, the influence of topography on the choice of terminal sites, the calculation of the free-space loss between antennas, which determines the antenna size, and the losses in cables or waveguides between the equipment and antennas.

621.396.65:621.396.619.13 267

Asymmetry of the Frequency Swing, particularly in the case of Wide-Band F.M. Directional Links—P. Barkow. (*Fernmeldetech. Z.*, vol. 4, pp. 168–173; April, 1951.) The cause of the asymmetry and of the consequent distortion is analyzed, and methods of correcting and of compensating the errors of symmetry are described.

621.396.65.029.62:621.396.5 268

Conditions for the Development in France of Two-Way Radiotelephony on Very Short Waves—E. P. Courtillot. (*Onde élect.*, vol. 31, pp. 223–232; May, 1951.) The use of the frequency band 25 to 174 mc for two-way communication between fixed and mobile stations is discussed generally in the light of current practice in the United States and in England. It is essential to arrange development according to a fixed plan, taking full account of economic factors.

621.396.97:621.396.66 269

The Automatic Monitoring of Broadcast Programmes—H. B. Rantzen, F. A. Peachey, and C. Gunn-Russell. (*Proc. IEE*, vol. 98, pp. 329–340; discussion, pp. 340–342; September, 1951.) See 1491 of 1951.

SUBSIDIARY APPARATUS

621.314.1.082.72 270

Electrostatic Direct-Current Transformer of 300 Kilovolts—J. M. Malpica. (*Rev. Sci. Instr.*, vol. 22, pp. 364–369; June, 1951.) A continuous transfer of electric charges is effected by rotating a dielectric between two pairs of metallic brushes, the primary pair being connected to an hv source. Several disks on one shaft are used to form a stepup transformer. The elementary theory and experimental results are given.

621.314.632.1 271

Resistance Stratification in Copper-Oxide Rectifiers: Part 1—Investigations on Massive

Plates with Nonblocking Contacts—F. Rose. (*Ann. Phys.* (Lpz.), vol. 9, pp. 97–123; June 30, 1951.) Measurements of the apparent resistance of massive oxidized copper disks with graphite or silver contacts were carried out at frequencies up to 100 kc and at various temperatures in the range 195° to 303°K. The results obtained are presented in a table, and diagrams, and are discussed with particular reference to Lehovec's method of estimating the conductivity distribution through the oxide layer.

621.314.632.1 272
Resistance Stratification in Copper-Oxide Rectifiers: Part 2—Investigations on Rectifier Disks—F. Rose. (*Ann. Phys.* (Lpz.), vol. 9, pp. 124–140; June 30, 1951.) Measurements were made for the same frequency and temperature ranges as for massive disks (271 above) on rectifier disks of two sorts of copper subjected to various annealing conditions, again using graphite or silver electrodes. The results are shown in diagrams, and Lehovec's method of analysis is applied.

621.316.722.078.3 273
Stabilization of Direct Voltages—H. Günther. (*Funk u. Ton*, vol. 5, pp. 124–132; March, 1951.) Discussion of the maximum stability attainable using an amplifier tube as a series resistance. Suitable conductance characteristics for the controlling pentode are shown. Its anode load resistance should be as large as possible. An additional potentiometer-type control circuit at the input reduces input voltage fluctuations and reduces the effective internal resistance of the control circuit.

621.319.3 274
New Electrostatic Generators—N. J. Félici. (*Onde élect.*, vol. 31, pp. 205–209; May, 1951.) The general principles underlying the functioning of rotary es generators are outlined. The necessity for a fluid dielectric with high breakdown voltage has led to the use of compressed gas, usually air. Pressures of thirty atmospheres enable the power output of a machine to be made two hundred times greater than is possible with normal atmospheric pressure. Figures are quoted for several existing machines, giving outputs of up to 100μ A at 500 kv. Machines are being developed for outputs as high as 30 mA at 500 kv or more. An over-all efficiency of 75 per cent is readily obtained. See also 999 of 1951. (Hémandinquer)

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 275
The Evaluation of Picture Quality with Special Reference to Television Systems: Part 1—L. C. Jesty and N. R. Phelp. (*Marconi Rev.*, vol. 15, pp. 113–136; 3rd Quarter, 1951.) "A new method of assessing the performance of a picture reproducing system is described. The effect of the simultaneous variation of the four parameters, brightness, contrast, resolution, and viewing distance, has been explored. Measurements have been made under conditions of best picture reproduction, with the system maintained at this level of adaptation. Various photographic and television systems have been examined. This work has involved a similar investigation of the behavior of the 'average observer'."

621.397.5:535.62 276
A Color Television System for Industry—H. R. Smith, A. L. Olson, and R. F. Cotellessa. (*Elec. Eng.*, vol. 70, p. 517; June, 1951.) Summary of 1951 AIEE Winter General Meeting paper. Description of a wired system using an Orthicon camera tube and suitable for industrial, commercial, medical, and military applications. A field-sequential system with rotating color disks is used.

621.397.5:535.62](083.74) 277
Plans for Compatible Color Television—D. G. F. (*Electronics*, vol. 24, pp. 90–93; August, 1951.) System standards recommended by the U. S. National Television System Committee are outlined. A composite signal is suggested, including a monochrome component and a color component on a subcarrier. Field tests are to be made in order to formulate definite numerical proposals for a full set of compatible color standards.

621.397.5:535.623 278
Dot-Interlaced Scanning and its Recent Development in American Colour Television—E. Schwartz. (*Fernmeldetech. Z.*, vol. 4, pp. 243–250; June, 1951.) The four-raster system of point scanning is explained, and new pulse technique for this method of scanning is described. Its application to color television is examined in detail, and suitable tubes developed for the purpose in the United States are briefly described.

621.397.5:621.317.74 279
Transmission Measuring System—Engstrom. (*See* 208.)

TRANSMISSION

621.316.726.078.3 280
U.H.F. Discriminator and its Application to Frequency Stabilization [of klystron]—Pircher. (*See* 59.)

621.396.61/.62:621.396.67 281
Transmission and Reception of Circularly Polarized Microwaves with a Common Aerial—W. Ruppel. (*Fernmeldetech. Z.*, vol. 4, pp. 251–253; June, 1951.) By using both directions of rotation of circularly polarized waves, an antenna system can be used for both the transmission and reception of waves of the same frequency. The bridge arrangement for decoupling the energy paths is explained, and a practical example is described which embodies two tank circuits. Arrangements of compact linear elements avoid the use of a gas-filled blocking tube. The principle is, in practice, only applicable to microwaves.

621.396.619.2+621.396.622 282
Modulators, Frequency Changers and Detectors using Rectifiers with Frequency-Dependent Characteristics—D. G. Tucker. (*Proc. IEE*, vol. 98, pp. 394–398; September, 1951.) An approximate method of allowing for frequency dependence of the resistance characteristic of the rectifier can be applied easily to the analysis of circuits where the terminating impedances are finite only at a finite number of modulation-product frequencies, and zero at all other frequencies. Results for a particular set of operating conditions are illustrated by numerical examples based on copper-oxide modulators working at frequencies up to 6 mc. The conversion insertion-loss can be stabilized against temperature variation by a suitable choice of terminating resistance. The working and results are the same for shunt, series, and ring modulators for conditions of low input frequency with high output frequency, and vice versa.

621.396.619.23 283
Oscillator Circuits as Frequency Modulators—W. Mansfeld. (*Funk u. Ton*, vol. 5, pp. 309–316, 351–360, and 411–421; June to August, 1951.) Three different arrangements are investigated theoretically and experimentally, and it is shown that proportional frequency variations can be obtained by modulation of the feedback phase angle. Using a single-tube oscillator circuit, a frequency swing of over 6 per cent was obtained, with linearity over a wide range, but a disadvantage of this modulator circuit is the dependence of the excited frequency on the operating voltages and tube characteristics. A push-pull circuit with push-pull modulation gave much better results, the linearity, frequency swing obtainable, small amplitude variation with modulation voltage, and frequency constancy with variations of operating voltage making such an arrangement useful as a frequency modulator. Another push-pull circuit, incorporating a phase-control arrangement, also gave good results.

TUBES AND THERMIONICS

621.383.27†:621.317.42 284
Investigation of the Conditions for Use of Kubetski's Mosaic Multiplier as an Indicator of the Magnetic Field—N. V. Krasnogorskaya. (*Bull. Acad. Sci.* URSS, *sér. géogr. géophys.*, vol. 15, pp. 43–50; January—February, 1951. In Russian.) A multiplier with a CuS-Cs photocathode and six stages of electron multiplication is described. Each stage consists of a secondary-emission CuS-Cs electrode and a nonemitting Ag electrode. The alternation of dark emitting and bright nonemitting layers resembles a mosaic. The electrons leaving the photocathode are directed by means of an electric and a magnetic field. An additional magnetic field shifts the beam from sensitive to nonsensitive electrodes, thus altering the value of the anode current. Experiments with these multipliers are described. The necessary conditions of stability of the power supplies are derived for a given accuracy of measurement of the intensity of the magnetic field, and limits of possible measurements are established.

621.385.029.63/.64:537.533 285
Waves in Electron Streams and Circuits—Pierce. (*See* 106.)

621.385.3:546.289 286
Electronic Forming in *n*-Germanium Transistors using Phosphorus-Alloy Contacts—J. P. Stelmak. (*Phys. Rev.*, vol. 83, p. 165; July 1, 1951.) Results of measurements emphasize the important role played by the P content of the collector contact point in improving transistor gain when forming is done by pulsed currents.

621.385.5:621.318.572 287
High-Speed Sampling Techniques—Shepard. (*See* 64.)

621.385.82.029.3:621.395.61 288
Increasing the Efficiency of the High-Power Thermionic Cell by Superposition of a Strong Field obtained from a High Voltage of High Frequency—S. Klein. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 143–145; July 9, 1951.) Modifications to the gas-filled thermionic cell described in 593 of 1947 are proposed. An hf voltage applied to a Pt point produces a discharge which heats the ion emitter. The device can be used both as an acoustic and ultrasonic transducer, and as a source of ultraviolet radiation.

621.396.615.141.2:621.385.029.63/.64 289
The Magnetron as a Traveling-Wave Valve—O. Döhler. (*Funk u. Ton*, vol. 5, pp. 146–155 and 257–262; March and May, 1951.) Treatment of the theory, and description of the construction and operation of different types of magnetron. Complementary to paper noted in 1017 of 1951. See also 2064 of 1950 (Warnecke *et al.*).

621.38 290
Electronics [Book Review]—P. Parker. Publishers: Arnold, London, England, 1050 pp., 50s. (*Jour. Sci. Instr.*, vol. 28, pp. 223; July, 1951.) "In this admirably balanced work there is no subject which is over-elaborated, or in which the reader is not brought up to a level just short of research standard."

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621.39.001.5 291
Telecommunications Research: Fundamental Investigations Undertaken by the D.S.I.R.—(*Wireless World*, vol. 57, pp. 431–432; October, 1951.)

621.396:061.4 292
Radio Exhibition Review—(*Wireless World*, vol. 57, pp. 384–395; October, 1951.) Trends in the design of radio, television, and associated apparatus shown at the 18th National Radio Exhibition, held in London during September, 1951, are illustrated and discussed.