Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

July, 1952

Volume 40 Number 7

Empire State, Inc.

Co-ordinated Broadcasting

In the first installation of its kind in the world, the nearby antennas of 5 individual television transmitters and of 3 frequency-modulation stations, towering 1500 feet above the New York streets, radiate their mutually noninterfering outputs to the metropolitan area.

PROCEEDINGS OF THE I.R.E.

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

The Institute of Radio Engineers

1952
WESTERN ELECTRONIC SHOW AND CONVENTION

Aerial view of Long Beach on the Pacific Ocean

1 p.m. to 9:45 DAILY

MUNICIPAL AUDITORIUM

LONG BEACH, CALIFORNIA

AUGUST 27-28-29, 1952

SPEAKERS AND PAPERS

MR. ARTHUR A. COLLINS, President, Collins Radio Company, "Keynote Address"

DR. I. S. COGGESHALL, General Traffic Manager, Western Union Telegraph Company, "The Transmission of Intelligence in Typescript"

MR. MURRAY G. CROSBY, President, Crosby Laboratories, "Long-Range Communication Trends"

DR. R. M. PAGE, Associate Director of Research for Electronics, Naval Research Laboratory, "Comparative Study of Modulation Methods"

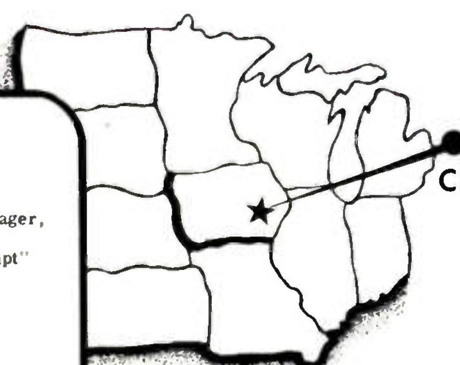
MR. L. MORGAN CRAFT, Vice President, Collins Radio Company "Design Trends in Communication Equipment"

MR. GEORGE Q. HERRICK, Chief, Division of Radio Facilities, Plans and Development, Broadcast Service, U. S. Dept. of State "The Voice of America in the Electronic War"

BANQUET SPEAKER

DR. LLOYD V. BERKNER, President, Associated Univeristies, Inc. "The Evolution of Communications"

MR. AL GRAF, Director of Region 5, Comments on Region 5 Activities



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CEDAR RAPIDS SECTION, IRE

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Table of Contents will be found following page 64A

Meetings with Exhibits

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

△

August 27, 28 & 29, 1952

Western Electronic Show and IRE Regional Convention Municipal Auditorium
Exhibits: Heckert Parker, 215 American Avenue, Long Beach, Calif.

△

September 8-12, 1952

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

△

September 19-20, 1952

Cedar Rapids IRE Technical Conference Roosevelt Hotel, Cedar Rapids, Iowa.
Exhibits: Lauren K. Findley, Collins Radio Co., Cedar Rapids, Iowa.

△

Sept. 29, 30, Oct. 1, 1952

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

△

December 10, 11 & 12, 1952

Joint IRE-AIEE Computers Conference Park Sheraton Hotel
Exhibits: Perry Crawford, 373 Fourth Avenue, New York City.

△

February 5, 6 & 7, 1953

Southwestern IRE Conference Plaza Hotel, San Antonio, Tex.
Accept Exhibits

△

March 23, 24, 25 & 26, 1953

Radio Engineering Show Grand Central Palace, New York City
Exhibits Manager: Wm. C. Copp, 303 W. 42nd St., New York 36, N.Y.

△

April 25, 1953

NEREM—New England Radio Engineering Meeting University of Connecticut, Storrs, Conn.
Accept Exhibits

△

May 11, 12 & 13, 1953

National Conference on Airborne Electronics Hotel Biltmore, Dayton, Ohio.
Exhibits: Paul D. Hauser, 1430 Gascho Drive, Dayton 3.

NEW! SPRAGUE Blue Jacket☆ wire-wound RESISTORS

WITHSTAND SEVERE HUMIDITY!

MEET JAN-R-26A!

Designed to withstand the rigid Characteristic G humidity tests of the most stringent specification of them all—JAN-R-26A—Sprague's new Blue Jacket Wire-Wound Resistors give trouble-free service in military electronic and electrical equipment exposed to extremely damp climates!

These outstanding new members of the Sprague resistor family are now available in tab terminal styles RW29 through RW39 in wattage ratings up to 166 watts.

You'll find the complete Blue Jacket Story with performance specifications in Engineering Bulletin 110, just off the press. Get your copy without delay.

YOU'LL KNOW THESE REMARKABLE RESISTORS BY THEIR VITREOUS ENAMEL BRIGHT BLUE JACKETS

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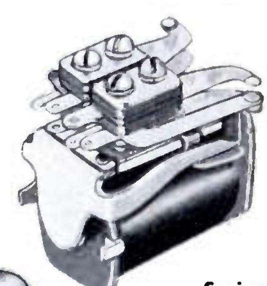

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Series 695 D.C.

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- A.G.C. maintains output constant within ± ¼ db for a ± 3 db variation in the r-f source.
- Wide VSWR ranges of 1:1.3, 1:3, 3:10, 10:30, and 30:100.
- High sensitivity—1.0 microvolts input for full scale deflection.
- Input circuit — Provides for either crystal or bolometer operation.
- Low input noise level of 0.03 microvolts.

Type 275 Amplifier is a high gain linear audio amplifier designed to accurately indicate voltage standing wave ratios. The application of expansion circuit techniques provides a full scale deflection of 1:1.3. This means greater accuracy for low VSWR measurements.
The unit may be operated as either a broadband amplifier over the range of 300 to 3000 c.p.s. or as a narrow band amplifier at 500, 1000, and 1300 c.p.s. The square law meter, calibrated to read directly in db, and the high voltage gain of 140 db make this amplifier particularly suitable for microwave attenuation measurements with a bolometer r-f detector.
Inquiries invited—address Dept. R-3.

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Monitors any four frequencies anywhere between 25 mc and 175 mc, checking both frequency deviation and amount of modulation. Keeps the "beam" on allocation, guarantees more solid coverage, too!
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Dawn or dusk, it doesn't matter. These heroes of the high wires arrive to stop power trouble before it starts. Their "nose for disaster" is in the service truck, in the mobile radio unit which often relies on JK crystals and monitors to keep their assigned radio frequency on the beam!
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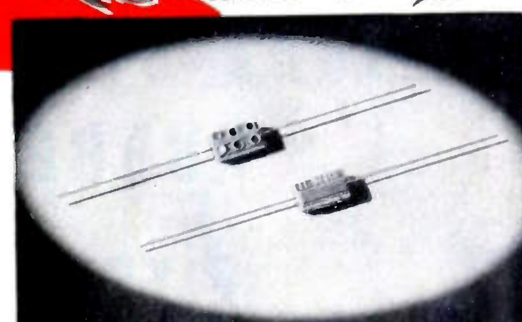
Instruments of war must be unerringly dependable, and every part used in their construction must contribute to this standard. That is why El-Menco Capacitors have won such wide recognition in their particular field . . . Because of their margin of extra wide safety factor they are absolutely reliable.

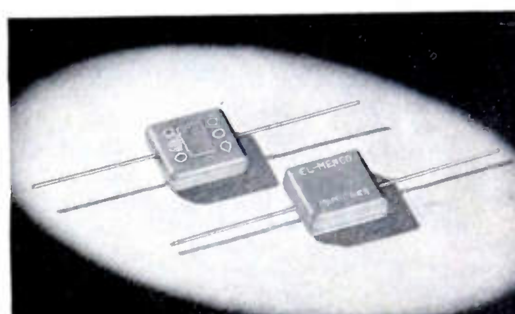
For higher capacity values, which require extreme temperature and time stabilization, there are no substitutes for El-Menco Silvered Mica Capacitors. El-Menco Capacitors are made in all capacities and voltages in accordance with military specifications.



CM-15

From the smallest to the largest each is paramount in the performance field.

Write on your business letterhead for catalog and samples.



CM-35

Jobbers and distributors are requested to write for information to Arco Electronics, Inc., 103 Lafayette St., New York, N. Y. — Sole Agent for Jobbers and Distributors in U. S. and Canada.

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CAPACITORS

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

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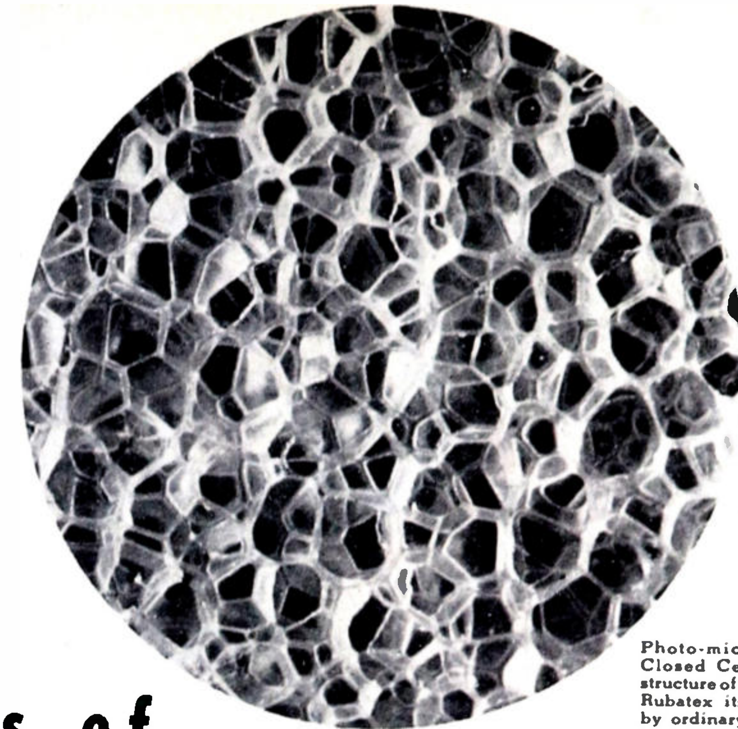

Photo-micrograph of a section of Rubatex Closed Cellular Rubber shows the dense structure of individually sealed cells which give Rubatex its superior qualities not possessed by ordinary sponge rubber having open cells.

Millions of "RUBBER BALLOONS" give you these 8 Advantages!

SOFT — Nitrogen-filled cells plus rubber's resiliency gives RUBATEX unusual workability.

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LIGHT WEIGHT — Its millions of nitrogen-filled "balloons" make RUBATEX surprisingly light. Rubber cell walls are the only significant weight factor.

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If you have a gasketing, sealing, shock absorbing or vibration damping application — you'll find RUBATEX is control for your product and profit!

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Rubatex engineers will be glad to work with you. Write and tell us your problem. Send for catalog RBS-4-51. Dept. IRE-7. Rubatex Division, Great American Industries, Inc., Bedford, Virginia.

NEWS and NEW PRODUCTS

By-Pass Capacitors

Erie Resistor Corp., Erie, Pa., is offering two new high voltage "Ceramicon" TV by-pass capacitors, which have been designed primarily to give high-voltage power-supply filtering for television receivers. Style 412 is rated at 20 kv and Style 414 at 10 kv.

The case insulation is of low-loss, molded thermosetting plastic which, it is claimed, provides a superior moisture seal. Ring convolutions are molded into the surface of the capacitor to prevent surface leakages caused by ordinary handling and a consequent deposit of conductive materials.

According to the manufacturer the convoluted design increases the effective surface creepage path by more than 14 per cent. Write to Erie for catalog and samples.

Preamplifier

The Brush Development Co., Instrument Div. 31, 3405 Perkins Ave., Cleveland 14, Ohio, now has available for delivery a new high-gain ac preamplifier designed to permit the extension of the measurable range of graphic recording instruments or cathode-ray oscilloscopes into the microvolt region, and maintain a relatively flat frequency response from 0.2 to 400 cps.

Designated as Model BL-954, the instrument was made for use with Brush magnetic direct-writing oscillographs, and medium or low-gain amplifiers to permit immediately available permanent records of measurements of electrical phenomena in the microvolt region.

Voltage gain of the instrument is 200,000 times when operated push-pull, and when used with Brush direct-writing oscillographs and amplifiers is sufficient to give one millimeter of pen deflection on the oscillograph chart per 2.5 microvolts input. Noise and hum interference have been reduced by the use of a battery power supply. Dry "B" batteries are mounted in the carrying case while either external-storage or dry batteries may be used to furnish "A" power. A standard cell furnishes voltage for an internal calibration circuit.

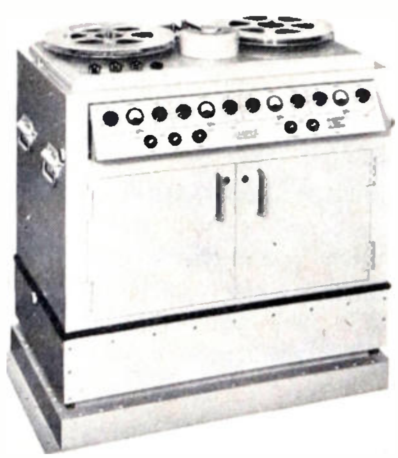
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Recorder

The **Ampex Electric Corp.,** Redwood City, Calif., announces a new magnetic tape recorder (Model 500) to meet the standards of high-frequency telemetering and applications in the use of rockets, guided missiles and computors.

This is a precision instrument reproducing 100 to 100,000 cps with the lowest possible flutter and wow, with less than 0.1 per cent peak-to-peak limit. The new drive system accomplishes more than 5 to 1 improvement over previous recorders.

Fabricated to conform to JAN specifications, provisions are made to record four individual data tracks. A signal-to-noise ratio is attained that is over 40 db below 1 per cent harmonic distortion when measured in 15 per cent band widths.



Most of the causes of flutter and wow in conventional tape recorder drives are eliminated in Model 500. The tape is rigidly held to a capstan by vacuum, providing extremely reliable tape motion without introducing any flutter or wow by pressure rollers or slippage. Since both the record and playback heads contact the tape at the capstan, tape scrape and vibration are reduced to a minimum. The capstan is coupled directly to a damped flywheel which is belt driven by a high-speed hysteresis synchronous motor. The tape is fed to the capstan under constant tension to eliminate any effect of differences in diameter of the tape reel.

The tape capacity of Model 500 is up to 5000 feet on 14 inch reels providing a playing time of 16 minutes at 60 inches per second, or 32 minutes at 30 inches per second.

Vacuum Tube Voltmeter

The Freed Transformer Co., Inc., 1718 Weirfield St., Brooklyn (Ridgewood) 27, N. Y., has introduced a new voltmeter, model No. 1040.

This instrument is composed of a high-impedance precision five-step attenuator, an rc coupled multistage amplifier, a balanced rectifier, a balanced dc amplifier, and a special meter in which the deflection is proportional to the logarithm of the current through it. A high amount of degeneration is used in both the ac and dc amplifiers. The switching from one voltage range to another is accomplished in input circuit.

Type 1040 is particularly recommended for vibration studies involving very low frequencies; frequency characteristics and gain measurements on amplifiers; transmission losses on telephone circuits; filter and carrier systems up to 200,000 cps; and acoustic measurements such as determination of frequency response of microphones and loudspeakers. Because of the high sensitivity of the instrument, this voltmeter can be advantageously used as a null detector in dc bridge measurements. In addition to its use as a voltmeter, the instrument can be used as an ammeter to measure a wide range of currents by connecting it across suitable resistors.

Servo Stabilizer

A device for stabilizing ac servo systems was recently announced by **Kalbfell Laboratories, Inc.,** P. O. Box 1578, San Diego 10, Calif. It contains a variable damping adjustment, is equipped with cathode followers for input and output, and plugs into an octal socket. The Twin-T Servo Stabilizer contains a phase shifting network to compensate for the lag caused by motors or other inertial elements. It overcomes hunting while maintaining the fastest possible response time. The twin-T network is made up of silver mica capacitors, and deposited carbon resistors to give stability with respect to time and temperature.

(Continued on page 48A)

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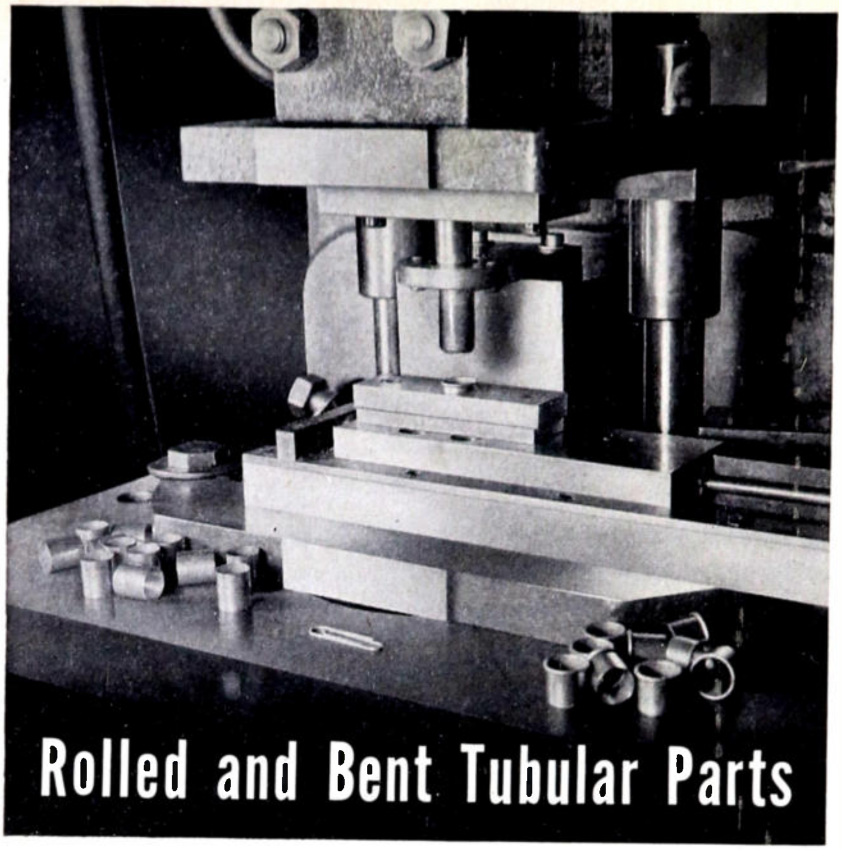

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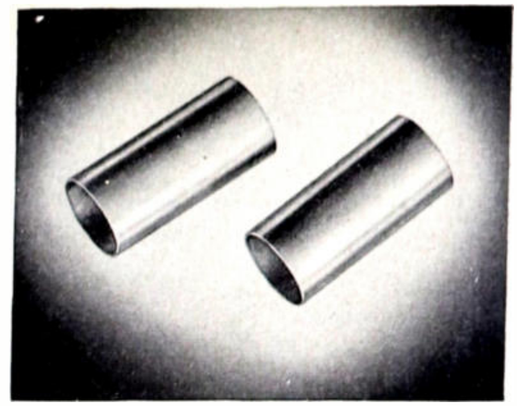
Men, experience, and machines-that-do-everything-but-talk, are generally the answer to a problem of obtaining parts of complex shape and precise dimension.

Here at Superior, customers for parts of this kind get a particularly good answer. We have the experienced men with a solid background of tubular parts production who are willing and able to take the time and care required for top-quality products. And we have the machines.

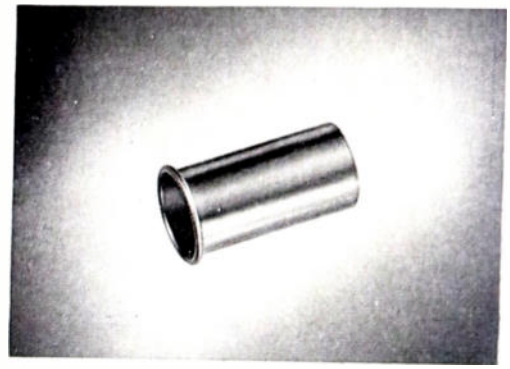
The delivery end of one of them is shown above. The part coming out came into our plant as a 2" tube, went through several redraw and annealing operations, was finally cut to exact length, tumbled to remove cutting burrs, then rolled by a controlled process to the precise dimensions established by customer specifications.

There's nothing spectacular in the story . . . it's just the outline of one of the many jobs that we know how to do well. Behind the story, however, is a thought for you.

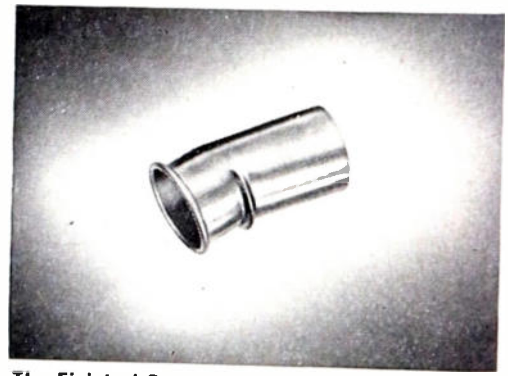
Our production story is backed by our ability, facility and desire to help you. If you are an experimenter in electronics or a manufacturer of electronic equipment and you need a tubular part to do a tough job well, better check with us. We'll be glad to assist with research, development, and design aid toward the solution of your problems. Tell us about them by writing Superior Tube Company, 2506 Germantown Ave., Norristown, Pennsylvania.



Cutting and Tumbling. Cutting machines and jigs of many types and sizes are combined with extensive tumbling equipment to permit fast, accurate production of quantities of parts at Superior.



Fabrication. Parts can be readily rolled at either or both ends, flared, flanged, expanded, or beaded (embossed) as required. The anode above is one of many such parts we produce at high speed and low cost.



The Finished Part. Final stage in the fabrication of the part, shown above at three stages of production, is a bend nicely controlled for both precise angle and freedom from other, unwanted distortion.

This Belongs in Your Reference File

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NICKEL ALLOYS FOR OXIDE-COATED CATHODES: This reprint describes the manufacturing of the cathode sleeve—from the refining of the base metal; includes the action of the small percentage impurities upon the vapor pressure and sublimation rate of the nickel base. Future trends of cathode materials are also evaluated.

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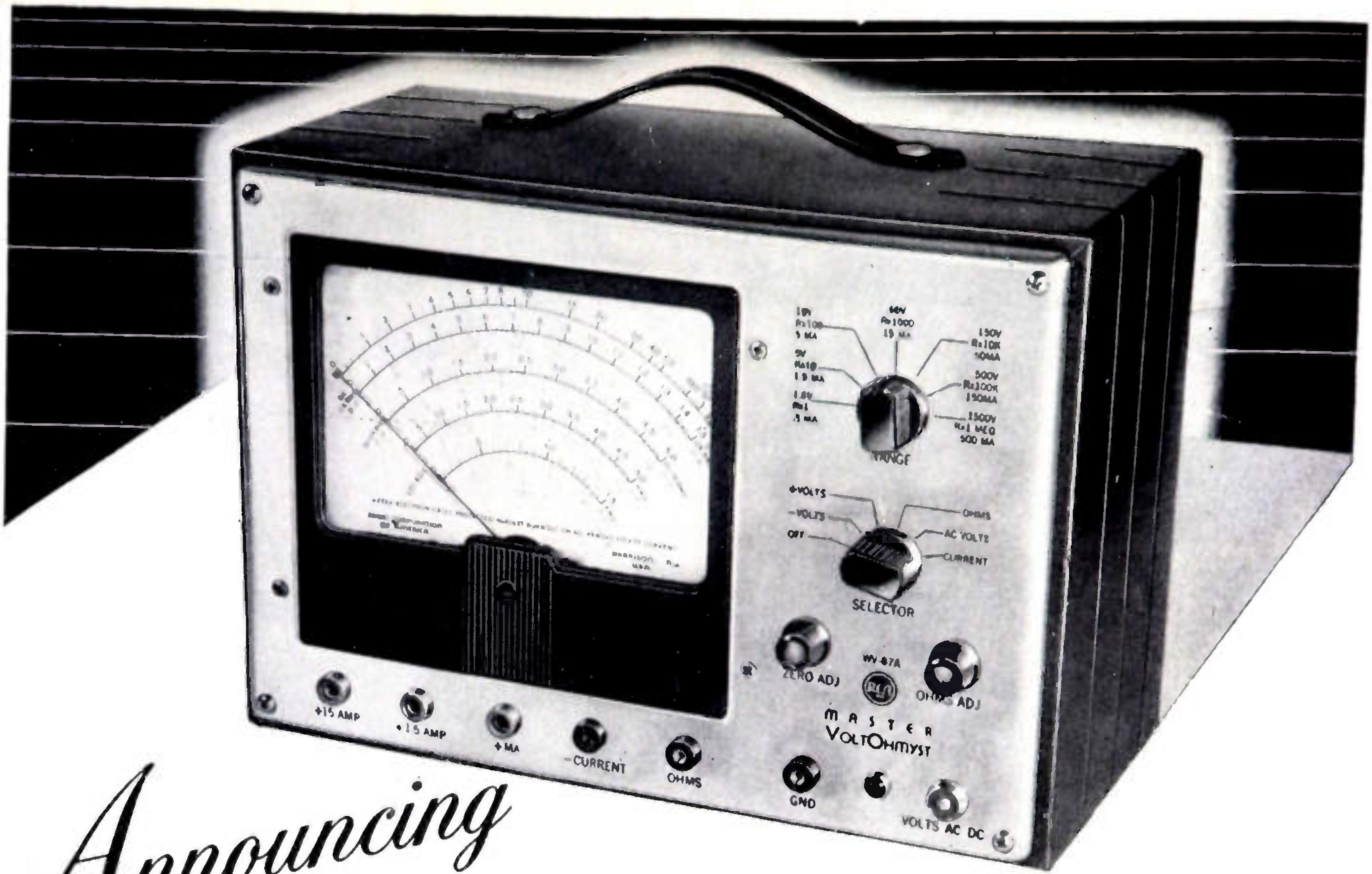

Announcing

The new RCA WV-87A *Master* VoltOhmyst*

$112.50 Suggested User Price

Measures . . . (Full-scale ranges)

DC VOLTAGE: 0 to 1.5, 5, 15, 50, 150, 500, 1500 volts
PEAK-TO-PEAK VOLTAGE: 0 to 4, 14, 42, 140, 420, 1400, 4200 volts
RMS VOLTAGE: 0 to 1.5, 5, 15, 50, 150, 500, 1500 volts
RESISTANCE: 0 to 1000 megohms in seven overlapping ranges
DC CURRENT: 0 to 0.5, 1.5, 5, 15, 50, 150, 500 milliamperes; 0 to 1.5, 15 amperes

Sold Complete — with the following Probes and Cables

- Direct Probe and Cable
- DC Probe
- Ohms Cable and Probe
- + Current Cable (Red)
- — Current Cable (Black)
- Ground (Case) Cable

Accessory Probes Available on Separate Order

√ WG-264 Crystal-Diode Probe for measuring ac voltages at frequencies up to 250 Mc.

√ WG-289 High-Voltage Probe, with WG-206 Multiplier Resistor, for increasing dc-voltage range to 50,000 volts and input resistance to 1100 megohms.

FEATURING an 8½" meter, the new WV-87A Master VoltOhmyst is really the master of every testing application. Its peak-to-peak scales are particularly useful for television, radar, and other types of pulse work.

The WV-87A measures dc voltages accurately in high-impedance circuits, even with ac present. It also reads rms values of sine waves and the peak-to-peak values of complex waves or recurrent pulses, even in the presence of dc.

Like all RCA VoltOhmysts, the WV-87A features ±1% multiplier and shunt resistors, a ±2% meter movement, high-input resistance, zero-center scale adjustment for discriminator alignment, dc polarity-reversing switch, and a sturdy metal case for good rf shielding.

On direct-current measurements, extremely low-meter resistance gives an average voltage drop of only 0.3 volt for full-scale readings on all ranges. Nine overlapping ranges provide dc readings from 10 microamperes to 15 amperes.

An outstanding feature is its usefulness as a television signal tracer . . . made possible by its high ac input resistance, wide frequency range, and direct reading of peak-to-peak voltages.

The RCA WV-87A *Master* VoltOhmyst has the accuracy and stability for laboratory work. Its large, easy-to-read meter also makes it especially desirable as a permanently mounted instrument in the factory and repair shop.

For complete information on the WV-87A, see your RCA Test Equipment Distributor or write RCA, Commercial Engineering, Section GX47, Harrison, New Jersey.

*Reg. U. S. Pat. Off.

Get complete details today from your RCA Test Equipment Distributor.

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HARRISON, N. J.

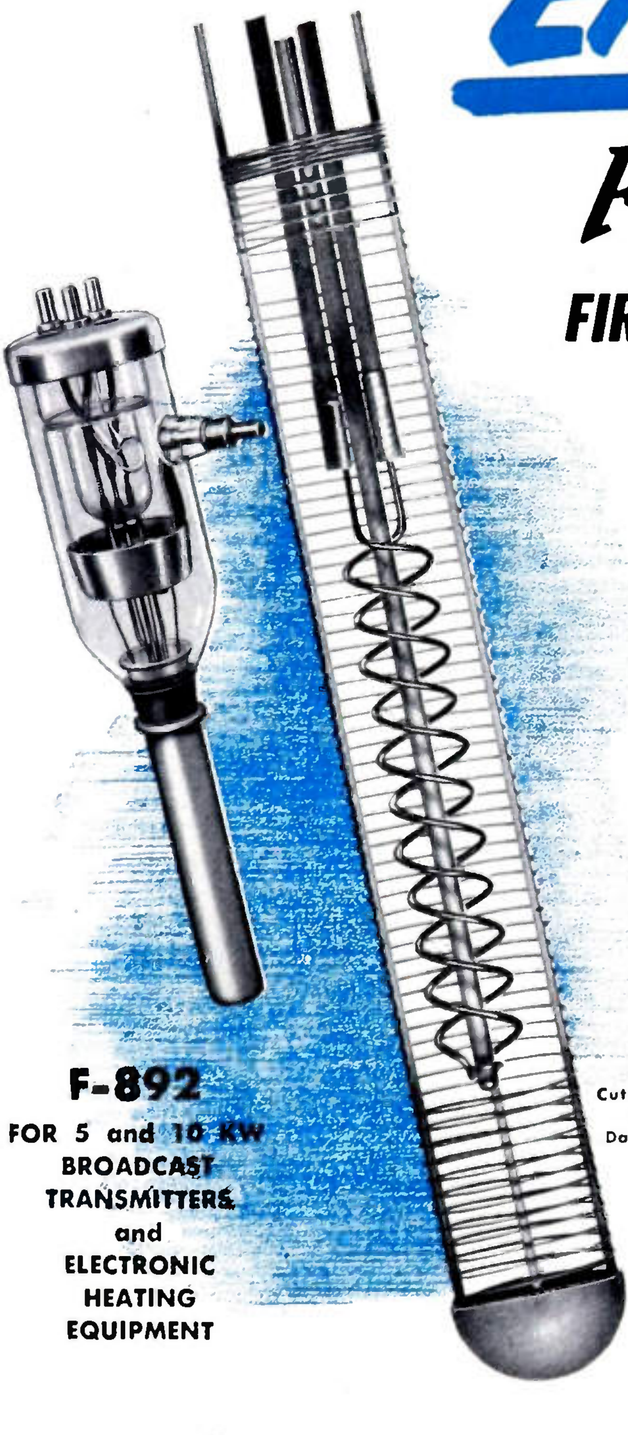
NOW... Filament-to-Grid Shorts *Eliminated!*

Federal's F-892

FIRST with the new, proved design

Double Helical Filament

- ***Does away with BOWING!***
- ***Greatly increases tube LIFE!***



F-892

FOR 5 and 10 KW BROADCAST TRANSMITTERS and ELECTRONIC HEATING EQUIPMENT

Cut-away View of Federal's Double Helical Filament

ANOTHER important Federal *"First"* is the new Double Helical Filament—for Federal's re-designed F-892!

The design has been completely checked and subjected to numerous filament cycling tests equivalent to two years' operation in normal broadcast equipment.

One tube filament was cycled 1500 times—1000 times with the starting current *twice* rated, and 500 times with the starting current *four* times rated—*without movement or distortion!*

Wound through 360° for mechanical stability and carrying opposing electrical fields which provide improved electrical stability, the F-892's Double Helical Filament definitely eliminates *bowing*—one of the primary causes of filament-to-grid shorts. For proof, all F-892's now in the field are still in service!

For full information on Federal's sturdier, longer-life, more dependable F-892, write to Vacuum Tube Division, Dept. K-537.

"Federal always has made better tubes"

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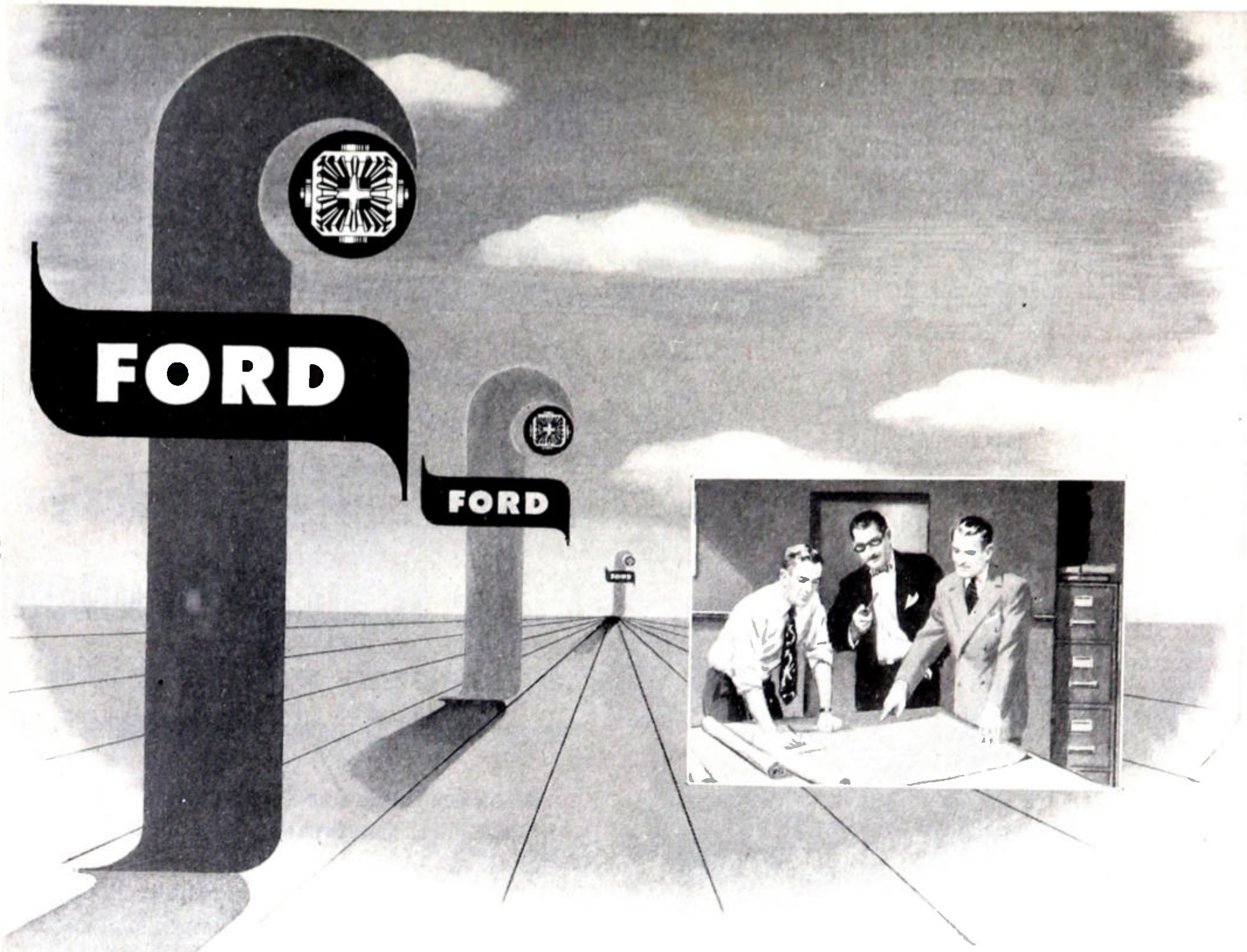

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Tomorrows and tomorrows—five . . . ten . . . twenty years from now—are being shaped at Ford today. For the Ford Instrument Company continues to pioneer the future as it has for the past thirty-seven years.

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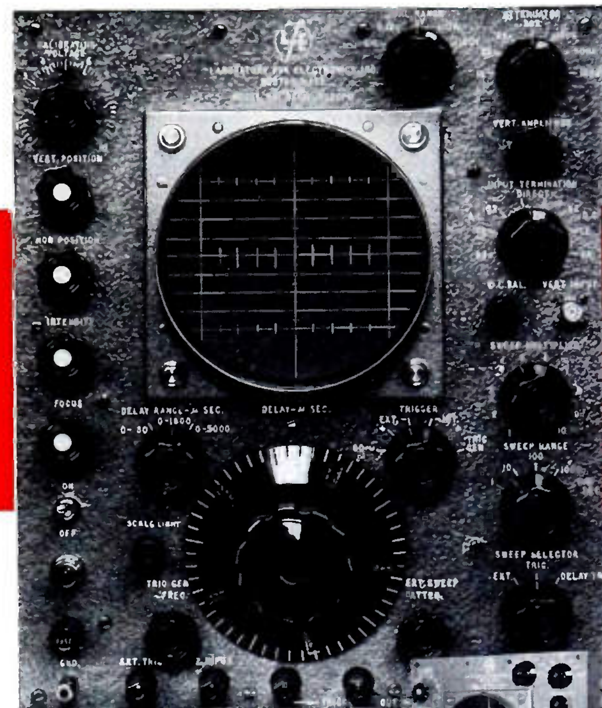
NEW, 10 Mc Wide Band Oscilloscope...

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



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

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

SPECIFICATIONS

Y-Axis

Deflection Sensitivity—15 millivolts peak-to-peak/cm
Frequency Response—DC to 10Mc
Transient Response—Rise Time—0.035 microseconds
Signal Delay—0.25 microseconds
Input line terminations—52, 72, or 93 ohms, or no termination, for either AC or DC input
Calibrating Voltage—60 cycle square wave.
Input Imp.—1 megohm, 30 mmf.

X-Axis

Sweep Range—0.01 sec/cm to 0.1 microseconds/cm
Delay Sweep Range—5-5000 microseconds in three ranges—continuously adjustable
Triggers—Internal or External, + and −, or 60 cycles, or delayed trigger outputs are available at suitable binding posts.
Built-in trigger generator for triggering external circuits and sweeps.

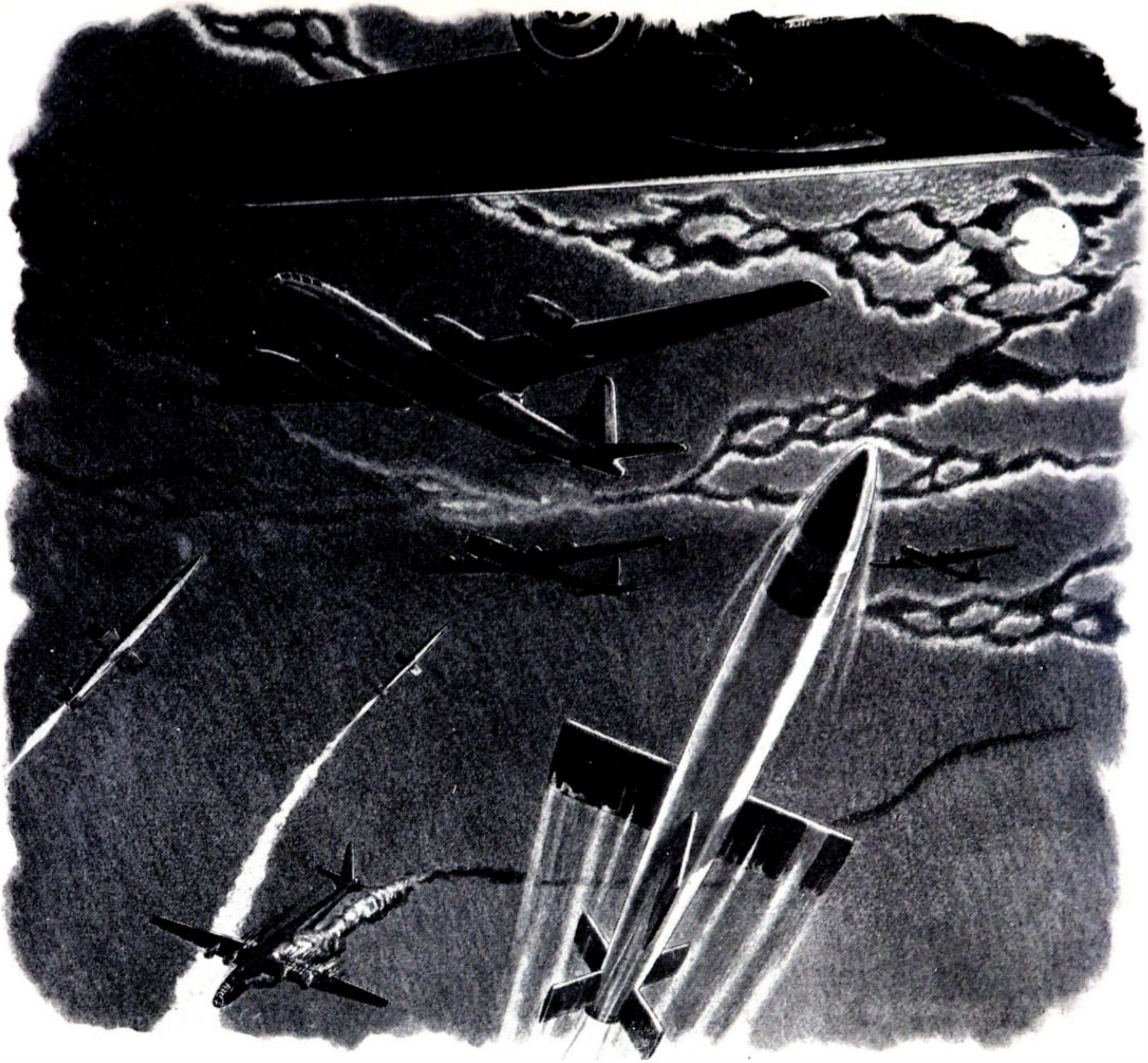
General

Low capacity probe
Functionally colored control knobs conveniently grouped
Folding stand for better viewing
Adjustable scale lighting
Facilities for mounting oscilloscope cameras
Dimensions—12½" wide, 15" high, 19" deep
Weight—55 lbs.
Price—$895. F. O. B., Boston

Write for full color booklet with complete information.

L/E

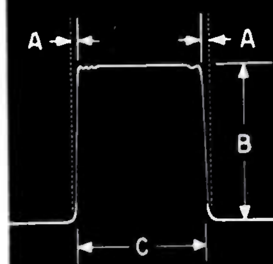
LABORATORY for ELECTRONICS, INC. 75 PITTS STREET BOSTON 14, MASS.



GENERAL PURPOSE PULSE GENERATOR

-hp- MODEL 212A

TYPICAL 1 MICROSECOND
PULSE INTO 50-OHM LOAD



A. 0.02 μsec rise and decay time. Minimum overshoot.

B. 50 watt peak power. (50 v. to 50 Ω load.)

C. Pulse length variable 0.07 to 10 μsec.

SPECIFICATIONS

PULSE LENGTH: Continuously variable, 0.07 to 10 μsec. Direct reading panel control.

PULSE AMPLITUDE: 50 v. into 50 Ω load. Pos. & neg. pulses. 100 v. open circuit.

AMPLITUDE CONTROL: Continuous control throughout range. 50 db in 10 db steps. 10 db fine adjustment.

INTERNAL IMPEDANCE: 50 Ω or less.

PULSE SHAPE: Rise and decay time approx. 0.02 μsec. (10% to 90% amplitude.)

REPETITION RATE: 50 pps to 5,000 pps. Internally or externally controlled.

SYNC IN: May be triggered by pos. or neg. pulse of 5 v. at rates up to 5,000 pps.

SYNC OUT: 50 v. into 200 Ω load. Approx. 2 μsec long. Approx. 0.25 μsec rise time.

PULSE DELAY: Main pulse delayable 0 to 100 μsec from sync output pulse.

PULSE ADVANCE: Main pulse can be advanced 0 to 10 μsec from sync output pulse.

POWER SUPPLY: 110/220 v.; 50/60 cps.

SIZE: Panel 10½" high, 19" wide. Depth 12".

PRICE: $550.00 f.o.b. Palo Alto.

Data Subject to Change Without Notice

CONTINUOUSLY VARIABLE, HIGH POWER PULSES OF SUPERIOR WAVE FORM!

THIS NEW -*hp*- 212A PULSE GENERATOR saves you time and work testing "fast" circuits as well as making everyday laboratory checks of other generators, rf circuits, peak-measuring equipment, etc. It is the first commercial pulse generator to successfully combine broad laboratory usefulness with the fast rise time, high power, variable pulsing and other features demanded in radar, television and nuclear work.

ACCURATE PULSES AT END OF LONG TRANSMISSION LINE

The pulse length is continuously variable from 0.07 μsec to 10 μsec, and is varied by a direct reading panel control. Extremely fast rise and decay time, together with freedom from ringing or overshoot provide a virtually distortion-free pulse. A low internal impedance (50 ohms or less) insures a pulse shape virtually independent of load. This low impedance also makes it possible to deliver accurate pulses at a distance from the instrument, if the transmission lines are correctly terminated.

The Model 212A's repetition rate is continuously variable from 50 to 5,000 pps. It can be controlled internally, or from an external synchronizing source. Synchronizing pulses are available from the instrument either in advance of or following the output pulse. An amplifier-attenuator output system gives a low source impedance, and makes possible continuously variable pulse amplitude, positive or negative.

Brief specifications of this new -hp- instrument are shown in the adjoining column. For complete details... see your local -hp- representative... or write to the factory.

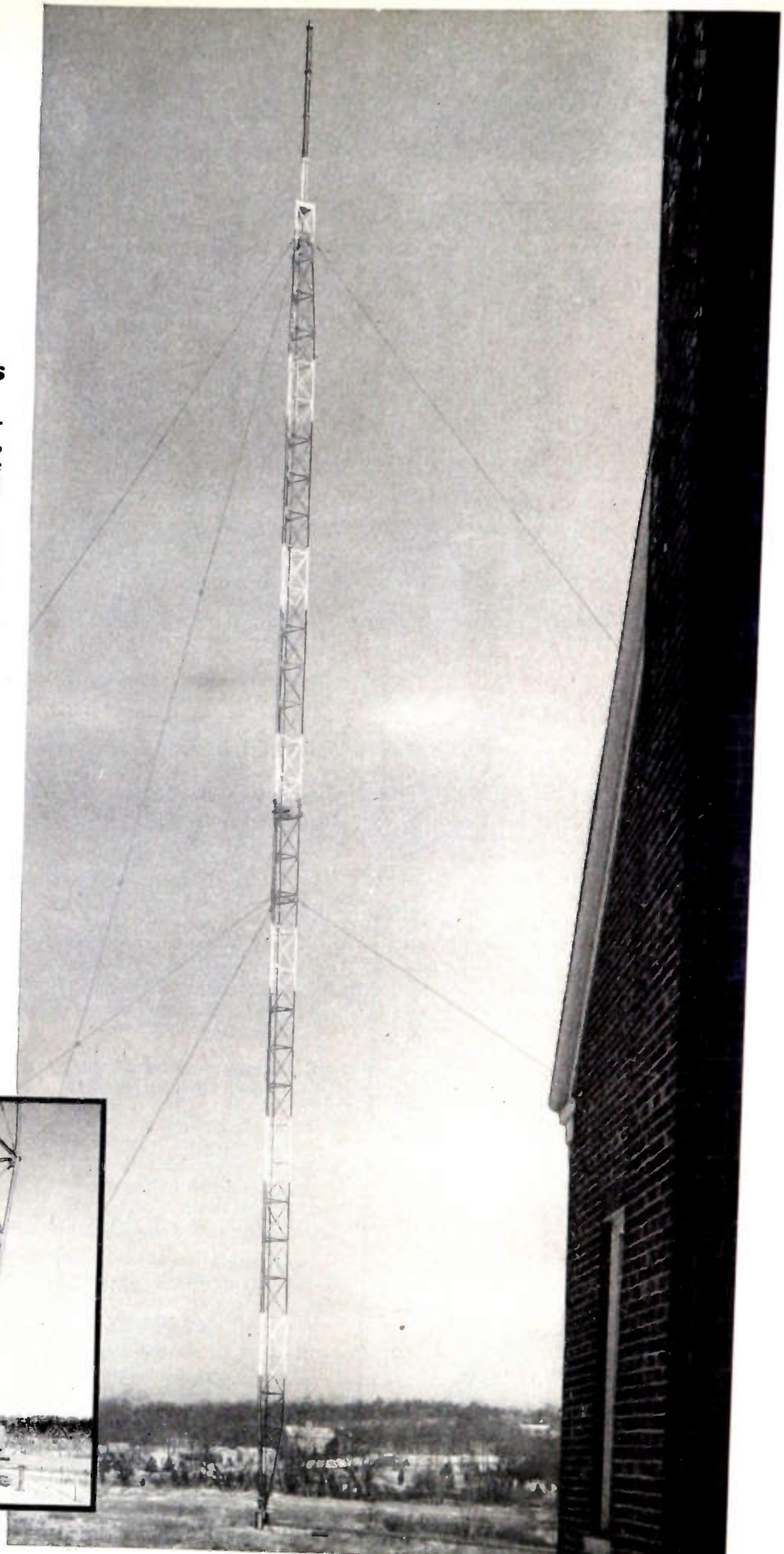
HEWLETT-PACKARD COMPANY

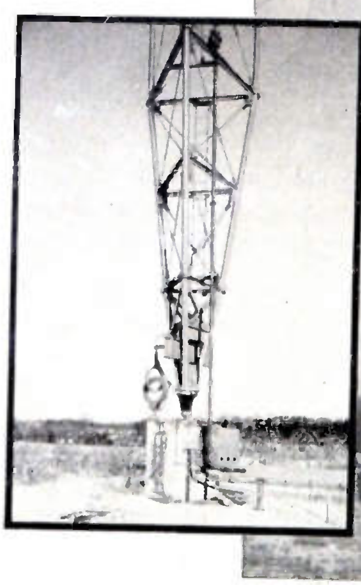
2040-D Page Mill Road • Palo Alto, California

Export: FRAZAR & HANSEN, Ltd., 301 Clay St., San Francisco, Calif., U. S. A. Offices: New York, N.Y. and Los Angeles, Calif.

2040

ELECTRONIC MEASURING INSTRUMENTS





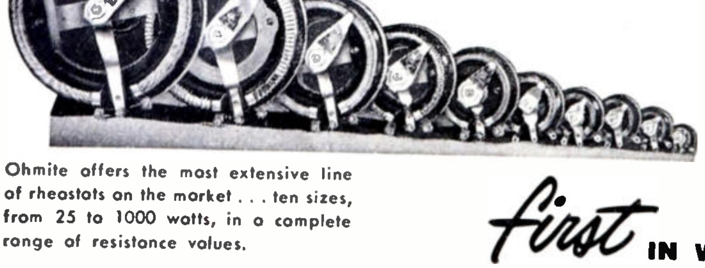
FOR SMOOTH, DEPENDABLE ELECTRICAL CONTROL

• S-m-o-o-t-h action . . . that's an indication of the close, evenly graduated electrical control provided by every Ohmite rheostat. This smoothness is a result of their distinctive design, incorporating a wobble-free brass bushing for effortless shaft rotation, tempered steel contact arm assuring uniform brush pressure, and smoothly gliding metal-graphite brush.

Ohmite rheostats not only perform better, they *last longer*. All-ceramic and metal construction . . . windings locked in place by vitreous enamel . . . wear-resistant bearings—all these quality features mean *extra* years of dependable, trouble-free service. Investigate these fine units for your product.

Write on company letterhead for catalog and engineering manual No. 40

Ohmite Manufacturing Co.
4860 Flournoy St., Chicago 44, Illinois

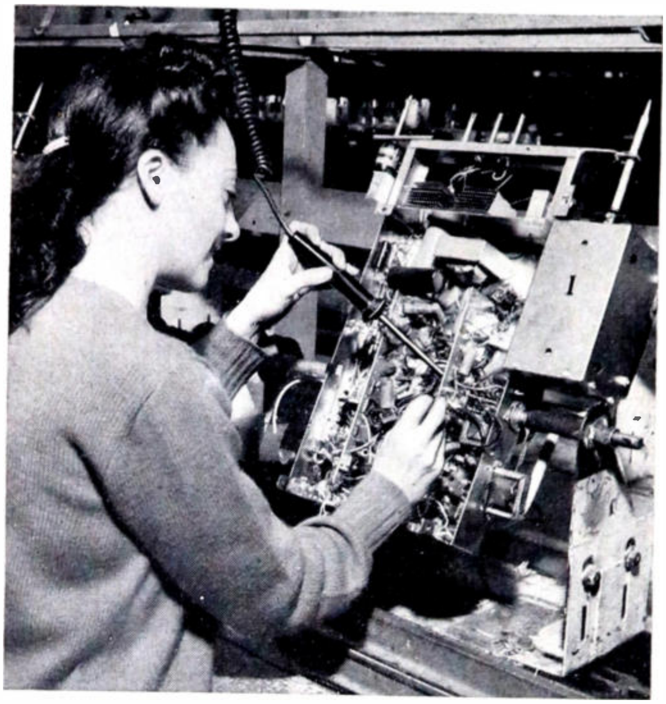


Ohmite offers the most extensive line of rheostats on the market . . . ten sizes, from 25 to 1000 watts, in a complete range of resistance values.

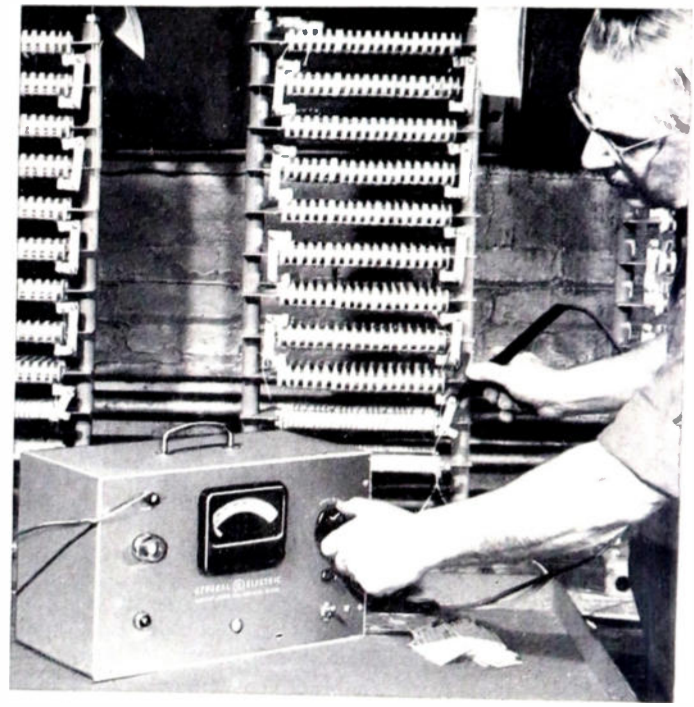
OHMITE®

first **IN WIRE-WOUND RHEOSTATS AND RESISTORS**

DESIGNER'S



1. New, fast-heating G-E iron weighs only $8\frac{1}{2}$-oz.



2. New G-E portable hi-pot tester is easy to operate.

Two ways to speed your production

Reach hard-to-solder places with this new thin-shank iron

"As easy to use as a pencil," say operators who use General Electric's new lightweight soldering iron.

Its thin, $\frac{5}{16}$-inch-diameter shank lets the $\frac{1}{4}$-inch tip into places a regular iron can't touch. Operators can solder more joints per minute—and with fewer rejects—because the iron's lightness, balanced design and comfortable handle all reduce fatigue.

Long-lasting G-E Calrod* heater provides quick heat-recovery properties, gives plenty of heat for uniformly strong soldered joints. Maintenance of this 60-watt, 120-volt iron is low because the long-life Ironclad tip need not be filed or dressed. Send for Bulletin GED-1583.

*Reg. Trade-mark

Eliminate cages and barriers with this new insulation tester

Now you can perform high-potential tests on your equipment with minimum danger to personnel. That's because the current output of General Electric's new high-potential insulation tester is limited to 5 milliamperes—well below the "let go" value.

Testing time is cut, too—no need to set up cages, barriers, or tape. Tester is portable, weighs only 22 lbs. Simply plug it into any 115-volt a-c outlet and start testing.

Line surges are virtually eliminated in output. Flash-overs can't burn insulation. Neon light on panel gives warning *before* insulation breaks down. Output is adjustable from 0 to 3500 volts, with test capacitance up to .006 muf. Bulletin GEC-700.

GENERAL ELECTRIC

DIGEST

TIMELY HIGHLIGHTS ON G-E COMPONENTS

Four ways G-E selenium rectifiers meet your d-c power requirements

Selenium rectifiers provide the electrical designer with versatile and flexible means of getting the right quantity of d-c power. But not all selenium rectifiers are alike. Here are four important "quality points" you'll find in G-E units in comparison with competitive equipment:

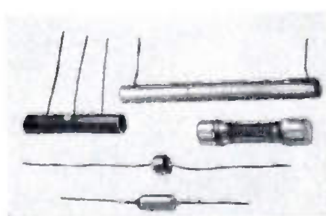
1. Lower forward resistance—for higher output and cooler operation—plus lower costs in other circuit components.
2. Less back leakage—for higher efficiency as well as higher output.
3. Cooler operation—the result of the above characteristics—since there is less heat to dissipate, less ventilation is needed.
4. Slower aging—which extends expected life at rated output to over 60,000 hours.

And of course the G-E line is complete, to meet all your design needs.

For a complete refresher on rectifier fundamentals, circuits, and applications, send for the new 28-page G-E booklet prepared to aid the design engineer. Check Bulletin GET-2350.

Standard stack construction

Tube-mounted construction



Miniature cell assemblies

Dual-rated capacitors simplify design problems

Meet your design needs, standardize, and cut inventories with these G-E fixed paper-dielectric capacitors. Equally applicable to a-c and d-c, they come in many case styles, with ratings from 236 through 660 volts a-c and 400 through 1500 volts d-c. All units are treated with Pyranol* and hermetically sealed to prevent leakage or contamination. Check Bulletin GEC-809.

Current-sensitive relays stand severe vibrations

G-E current sensitive d-c relays are available with d-c pickup ratings in steps from 4 to 1500 ma. They are especially applicable to circuits using limited power for energizing coils—as in aircraft. Lightweight and corrosion-proof, these relays withstand severe vibration and operate at rated current through a wide range of altitudes. See Bulletin GEC-834.

EQUIPMENT FOR ELECTRONIC MANUFACTURERS

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

Components

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

Development and Production Equipment

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

*Reg. trade-mark of General Electric Co.

General Electric Company, Section B667-21
Schenectady 5, New York

Please send me the following bulletins:

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

- ☐ GEC-700 High-Potential Tester
- ☐ GEC-809 Paper-Dielectric Capacitors
- ☐ GEC-834 Current-Sensitive D-C Relays
- ☐ GED-1583 Lightweight Soldering Iron
- ☐ GET-2350 Selenium Rectifiers

Name__________

Company__________

City__________ State__________

Besto-Wire®

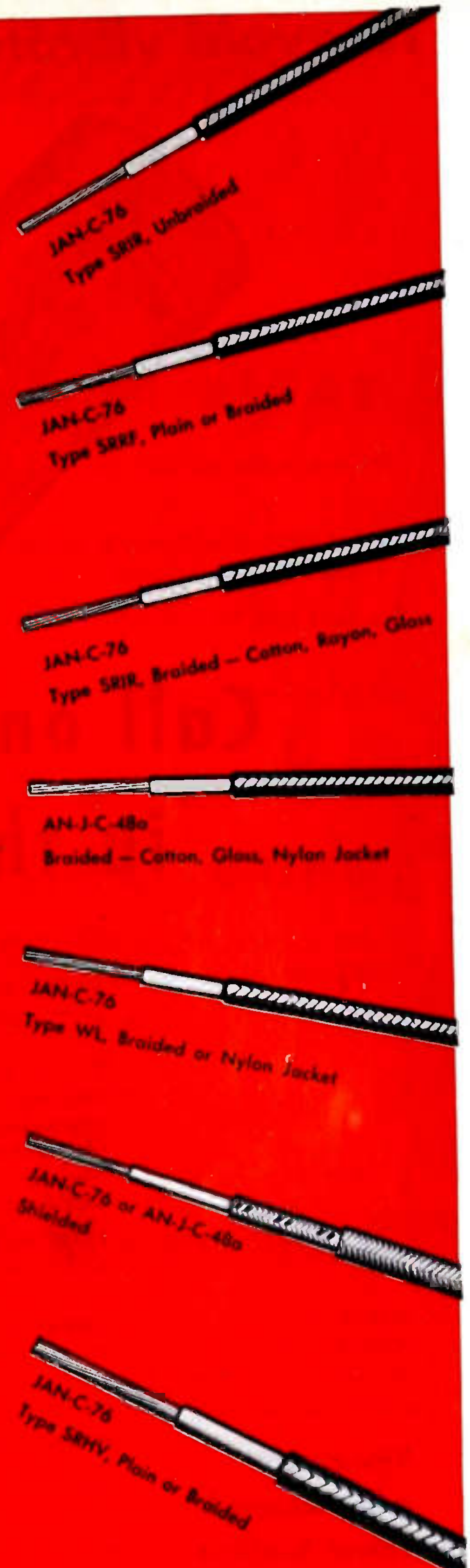
WIRE AND CABLE MEETS OR EXCEEDS GOVERNMENT SPECIFICATIONS

In our constantly expanding lines are scores of wires and cables that meet government approval specifications or exceed them.

In many cases you will find the wire that meets your requirements is one of our regular stock numbers — or our engineers will be glad to work with you to develop wires especially for your needs.

We have had years of experience in designing and manufacturing wire and cable for a wide range of applications in electronics, aviation and industry. Put this experience to work to help solve your problems wherever they concern wire or cable. We welcome your inquiries.

**Ask us for our new complete catalog.
We'll be glad to send you a copy.**



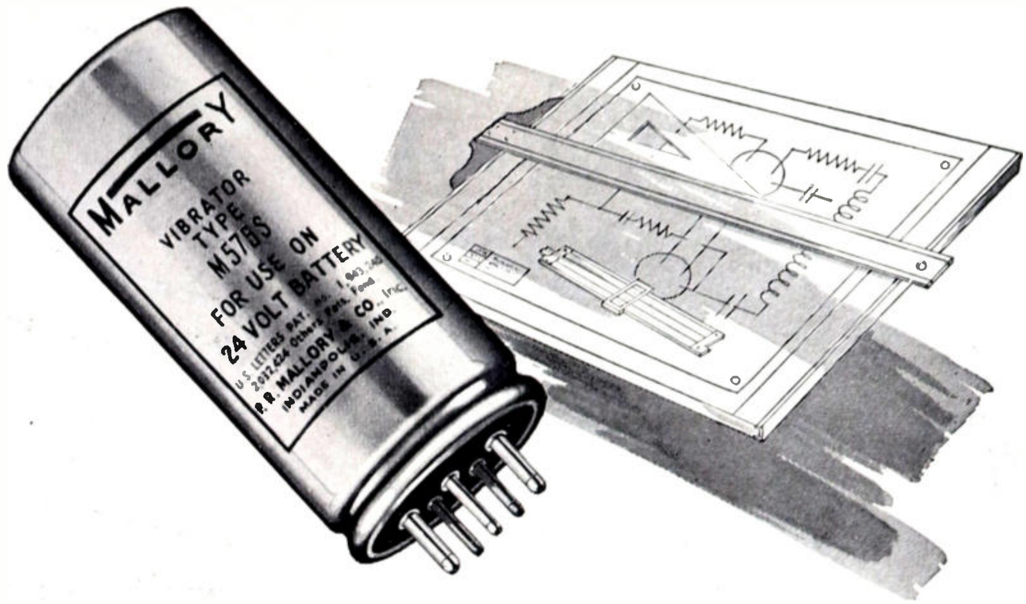
Besto-Wire

RHODE ISLAND INSULATED WIRE CO., Inc.

50 BURNHAM AVENUE CRANSTON, RHODE ISLAND

NATIONAL SALES OFFICES: 624 S. MICHIGAN BLVD., CHICAGO, ILL.

To avoid vibrator power supply troubles...



Call on MALLORY in the design stage

Vibrator power supplies need the best in design to meet the tough performance requirements of battery-operated electronic equipment. Experienced engineering must go into the selection of each element . . . transformer, vibrator, buffer capacitor . . . so each will perform its own function and operate with the others to provide maximum performance.

You can end your power supply worries by calling on Mallory to translate your power needs into a precisely engineered design. You'll save engineering time and reduce production costs. And you'll be assured of getting the power performance your equipment needs.

In vibrator design and production, Mallory offers you . . .

EXPERIENCE gained through years of working with leading manufacturers on a broad variety of vibrator applications . . . an unmatched fund of engineering skills in this field.

VIBRATORS for civilian and military use . . . produced by the organization which pioneered the development of commercial vibrators over 20 years ago and now supplies more vibrators for original equipment than all other makes combined.

ENGINEERING AND PRODUCTION FACILITIES that can design complete vibrator power supplies and manufacture them from start to finish . . . to your exact requirements.

Investigate Mallory's facilities now, whether your equipment is in production or still on the drawing board. Write or telephone today.

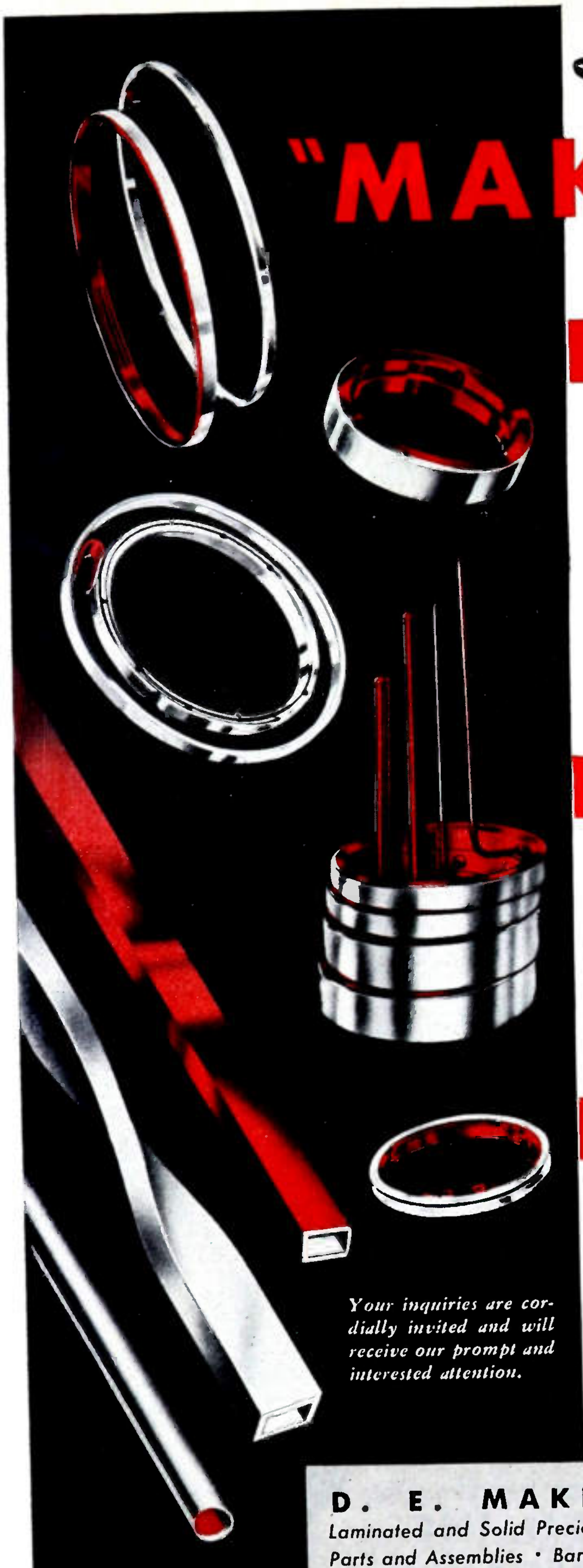
Vibrators and Vibrapack® Power Supplies

P.R. MALLORY & CO. Inc.
MALLORY

SERVING INDUSTRY WITH THESE PRODUCTS:
Electromechanical—Resistors • Switches • Television Tuners • Vibrators
Electrochemical—Capacitors • Rectifiers • Mercury Dry Batteries
Metallurgical—Contacts • Special Metals and Ceramics • Welding Materials

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

Specify "MAKEPEACE" *for*



PRECISION RINGS

DEMCO collector rings and ring assemblies give you these advantages . . .

ECONOMY . . . laminated construction provides contact metal where required . . . base metal for strength.

HIGH FINISH . . . on contact surface for long wear and noise-free operation.

PRECISION MADE . . . it is unnecessary to add further machine operations.

COMPLETE . . . rotor or pancake type multi-ring and ring-and-brush assemblies supplied.

PRECISION TUBING

PRECISION DRAWN . . . ID held as close as ±.001. Solid coin silver, brass or aluminum Rev. MIL-T-85.

LAMINATED SILVER . . . on ID for

- Low attenuation
- Corrosion resistance
- Highest mirror finish

Laminated silver ID and OD for round tuned lines.

PRECISION "KNOW-HOW"

With nearly sixty years of experience in the production of both laminated and solid precious metals, MAKEPEACE is today an accepted "headquarters" for the many special precious metal products and assemblies called for in the electronic field.

Our staff of thoroughly experienced design and production engineers and metallurgists . . . as well as our research and testing laboratory . . . are all at your service.

D. E. MAKEPEACE COMPANY

Laminated and Solid Precious Metals for Industrial Use • Fabricated Parts and Assemblies • Bar Contact Material • Precious Metal Solders

MAIN OFFICE AND PLANT, ATTLEBORO, MASSACHUSETTS
NEW YORK OFFICE, 30 CHURCH STREET
CHICAGO OFFICE, 55 EAST WASHINGTON STREET

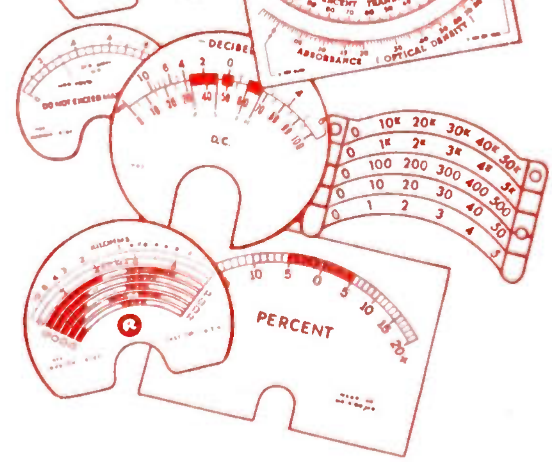
marion methods

die-cut, process and print each dial individually

Because the accuracy of an indicating instrument is completely dependent upon the accuracy of its dial, Marion takes extraordinary care with each dial . . . from the simplest black and white type to a fluorescent type of seven colors.

Marion dials are never printed in sheets and then stamped out, as are ordinary dials. Each Marion dial is die-cut, prepared and printed *separately*. This individual handling guarantees finished painted edges, which reduces high voltage corona; it also assures accurate mechanical registration of the dial with the pivot center of the instrument.

Precision and dependable performance are built into every Marion dial . . . in each step of manufacture.



Preliminary Operations . . . Dial data furnished by the customer is carefully checked by Marion's Engineering Department. After Engineering OK, data and suggested layout are sent to the Art Department.

The dial scale is drawn 4 to 6 times "life size," then the drawing is photographed and reduced to the proper size.

Preparation of Plates . . . After photography, a positive print is made. Color separations are made by hand, and deep-etched zinc plates for offset lithography are produced. The offset process is used to assure sharp printing definition and good color.

Preparation of Dial Blank . . . Each metal dial blank is thoroughly rinsed and vapor de-greased. Then, three separate coats of special fume and age-proof eggshell-white lacquer are applied. This lacquering technique gives a surface that will not chip, flake, fade or discolor.

Registering . . . After careful inspection, dial blanks are securely mounted on the printing press. Each blank is individually adjusted, and the printing plate is positioned exactly. This step ensures perfect registration for multi-color printing.

Printing and Drying . . . Each dial is then printed separately. After special inks of each color are applied, dials are baked for 15 to 20 minutes to set the ink. This process eliminates smudging, and minimizes the amount of lint picked up during drying. Dials are thoroughly inspected again before they are mounted on Marion instruments.

Other Marion Methods. Marion's method of assuring the top accuracy and service of each dial by individual handling is only one of a number of methods which Marion is presenting in the hope that some of them will help you as they have helped us. We will be pleased to send more information if desired.

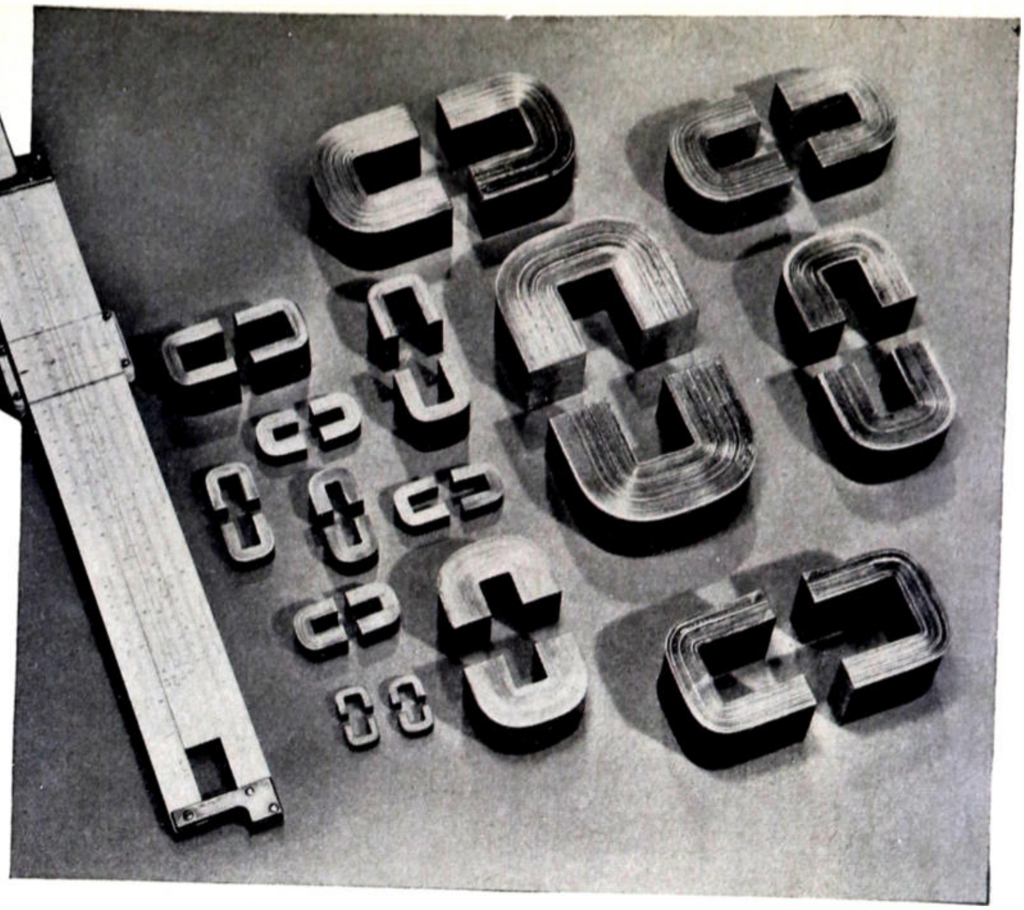
MARION ELECTRICAL INSTRUMENT CO., 407 CANAL ST., MANCHESTER, N. H.

marion meters

MANUFACTURERS OF MARION *Ruggedized* PANEL METERS


AN INVITATION
to
Development and Design Engineers
Working on Highly Specialized
Magnetic Recording Equipment
Shure is engaged in intensive research on special recording heads for use in fields far removed from conventional recording applications. This research might be of practical value to you.
We invite inquiries from your company if you have a specialized recording application. Possibly we can assist you—by providing a solution based upon our extensive development and production facilities.
Write on company letterhead to:
Manufacturing Division
SHURE BROTHERS, INCORPORATED
225 W. Huron Street Chicago 10, Illinois
SHURE

SILECTRON "C" CORES *for quick delivery* IN PRODUCTION QUANTITIES



. . . wound from strip as thin as 0.00025"

Quality-Tested and Proved

★ Arnold "C" Cores are made to highly exacting standards of quality and uniformity. Physical dimensions are held to close tolerances, and each core is tested as follows:

★ 29-gauge Silectron cut cores are tested for watt loss and excitation volt-amperes at 60 cycles, at a peak flux density of 15 kg.

★ 4-mil cores are tested for watt loss and excitation volt-amperes at 400 cycles, at a peak flux density of 15 kg.

★ 2-mil cores are tested for pulse permeability at 2 microseconds, 400 pulses per second, at a peak flux density of 10 kg.

★ 1-mil cores are tested for pulse permeability at 0.25 microseconds, 1000 pulses per second, at a peak flux density of 2500 gauss.

★ ½ and ¼-mil core tests by special arrangement with the customer.

Now available—"C" Cores made from Silectron (oriented silicon steel) thin-gauge strip to the highest standards of quality.

Arnold is now producing these cores in a full range of sizes wound from ¼, ½, 1, 2 and 4-mil strip, also 29-gauge strip, with the entire output scheduled for end use by the U. S. Government. The oriented silicon steel strip from which they are wound is made to a tolerance of plus nothing and minus mill tolerance, to assure designers and users of the lowest core losses and the highest quality in the respective gauges. Butt joints are accurately made to a high standard of precision, and careful processing of these joints eliminates short-circuiting of the laminations.

Cores with "RIBBED CONSTRUCTION"* can be supplied where desirable.

Ultra thin-gauge oriented silicon steel strip for Arnold "C" Cores is rolled in our own plant on our new micro-gauge 20-high Sendzimir cold-rolling mill. For the cores in current production, standard tests are conducted as noted in the box at left—and special electrical tests may be made to meet specific operating conditions.

● **We invite your inquiries.**

*Manufactured under license arrangements with Westinghouse Electric Corp.

WSD 4211

THE ARNOLD ENGINEERING COMPANY

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

General Office & Plant: Marengo, Illinois

electronic wire and cables for standard and special applications

Whether your particular requirements are for standard or special application, choose *LENZ* for the *finest* in precision-manufactured electronic wire and cable.

GOVERNMENT PURPOSE RADIO AND INSTRUMENT HOOK-UP WIRE,
plastic or braided type, conforming to Government Specification JAN-C-76, etc., for radio and instruments. Solid or flexible conductors, in a variety of sizes and colors.

SPECIAL HARNESSES,
cords and cables, conforming to Government and civilian requirements.

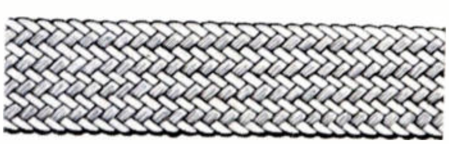
RADIO AND INSTRUMENT HOOK-UP WIRE,
Underwriters Approved, for 80° C., 90° C. and 105° C. temperature requirements. Plastic Insulated, with or without braids.

SHIELDED JACKETED MICROPHONE CABLE
Conductors: Multiple—2 to 7 or more conductors of stranded tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. Tough, durable jacket overall.

RF CIRCUIT HOOK-UP AND LEAD WIRE
for VHF and UHF, AM, FM and TV high frequency circuits. LENZ Low-Loss RF wire, solid or stranded tinned copper conductors, braided, with color-coded insulation, waxed impregnation.

JACKETED MICROPHONE CABLE
Conductors: Extra-flexible tinned copper. Polythene Insulation. Shield: #36 tinned copper, closely braided, with tough durable jacket overall. Capacity per foot: 29MMF.

SHIELDED MULTIPLE CONDUCTOR CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. For: Auto radio, indoor PA systems and sound recording equipment.



TINNED COPPER SHIELDING AND BONDING BRAIDS
Construction: #34 tinned copper braid, flattened to various widths. Bonding Braids conforming to Federal Spec. QQ-B-S75 or Air Force Spec. 94-40229.

SHIELDED COTTON BRAIDED CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Cable concentrically formed. Closely braided tinned copper shield plus brown overall cotton braid.

PA AND INTERCOMMUNICATION CABLE
Conductors: #22 stranded tinned copper. Insulation: textile or plastic insulated conductors. Cable formed of Twisted Pairs, color-coded. Cotton braid or plastic jacket overall. Furnished in 2, 5, 7, 13 and 25 paired, or to specific requirements.

CHECK LENZ FIRST!

Lenz Electric Manufacturing Co.

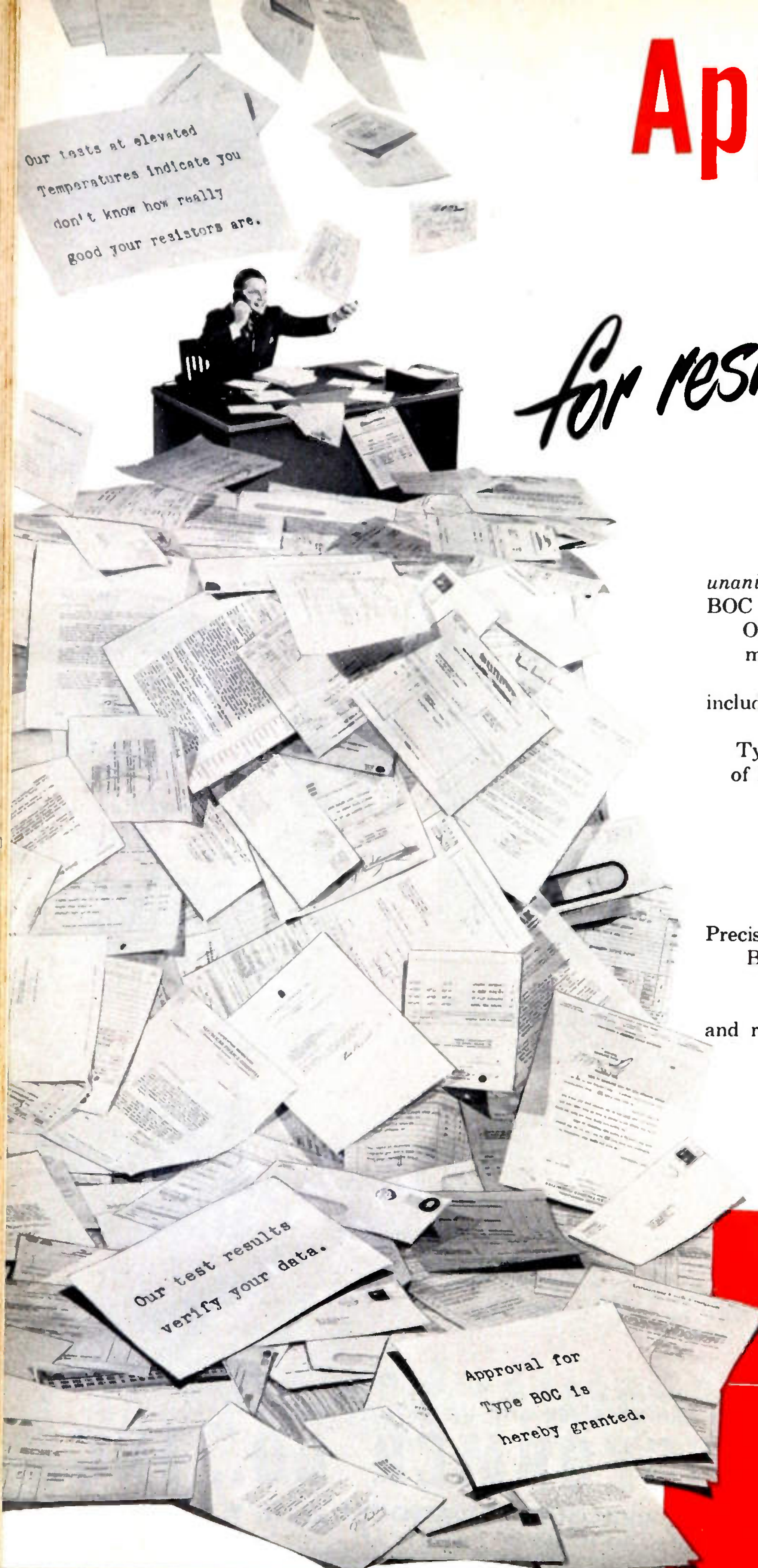
1751 N. Western Ave., Chicago 47, Illinois

cords, cable and wire for radio ◆ p. a. ◆ test instruments ◆ component parts


FREED Instruments & Transformers
"PRODUCTS OF EXTENSIVE RESEARCH"
MINIATURE TRANSFORMERS
MINIATURE TOROID INDUCTORS
PULSE MODULATORS
No. 1150 UNIVERSAL BRIDGE
No. 1030 LOW FREQUENCY "Q" INDICATOR
SLUG TUNED DISCRIMINATORS
HIGH FIDELITY TRANSFORMERS
A.C. VOLTS
A.C.-V.T. VOLTMETER MODEL 1040 FREED TRANSFORMER CO
No. 1010 COMPARISON BRIDGE
FREEDSEAL TREATMENT
No. 1020B MEGOHMMETER
FILTERS
No. 1180 A.C. SUPPLY
NO. 1040 VOLTMETER
VOLTAGE RANGES: .001 volts to 100 volts in five ranges (.01, .1, 1, 10, and 100 volts full scale).
ACCURACY: 2% on full scale on all five ranges, on sinusoidal voltages.
FREQUENCY RANGES: 10 to 200,000 cycles, .1 db. variation from 20 cycles to 150,000 cycles; .50 db. variation from 10 cycles to 200,000 cycles.
INPUT IMPEDANCE: Equivalent to 500,000 ohm resistance in parallel with a 15 MMF. condenser.
STABILITY: Effect of variation in line voltage from 100 volts to 125 volts is 1%. Effect in changes of tubes is less than .5%.
METER: 4" suppressed zero 1 MA meter protected against overloads.
POWER SUPPLY: The instrument is entirely self-contained and operates on 100-125 volts, 50-60 cycles. Total consumption, 40 Watts.
DIMENSIONS: 4⅞" High, 5⅛" Wide, 9⅞" Long.
WEIGHT: 12 pounds.
No. 1110A INCREMENTAL INDUCTANCE BRIDGE
No. 1170 D.C. SUPPLY
SEND FOR COMPLETE TRANSFORMER AND INSTRUMENT CATALOGS
FREED TRANSFORMER CO., INC.
1720-B WEIRFIELD ST., BROOKLYN (RIDGWOOD) 27, N. Y.
EXPORT DIVISION:—458 BROADWAY N.Y.C. 13, N.Y.

Approval is

for resistors too!



In all our experience, no resistor has been so extensively tested—*and so unanimously approved*—as IRC's new Type BOC Boron-Carbon ½-watt PRECISTOR. Of the 3,000,000 already manufactured, more than 100,000 were given the most stringent tests-in-production, including critical temperature cycling and 500-hour load-life tests. Result:—Type BOC conforms to *all* requirements of MIL-R-10509A! Also, customers have conducted their own laboratory and field tests—and they express their approval of Type BOC in letters like those shown here.

In the case of IRC's new JAN Type Precision Wire Wounds and Advanced Type BT Resistors, too, rigid quality control and continued testing have won industry-wide approval. Most stable and reliable of all precision wire wounds, Type WW's far surpass JAN-R-93 Characteristic B Specifications. And Type BT's continue to meet and beat JAN-R-11 Specifications.

IRC

News—New Products

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

(Continued from page 12A)

Power Supply Bulletin

The Superior Electric Co., Bristol, Conn., manufacturers of voltage control equipment, have recently released a new bulletin describing the Varicell dc power supply.

Bulletin V1051 covers in detail the Varicell, a source of variable stabilized-regulated dc voltages operating from ac power lines. The bulletin is complete with photographs, circuit diagrams, outline dimensions, ratings and descriptive data. Also included in the bulletin are descriptions on the workings of the Varicell. This data is fully illustrated with circuitry and photographs.

Variable Inductance Coils

A new series of variable inductance coils covering the 2-180 microhenry range has been announced by **North Hills Electric Co.,** P. O. Box 427, Great Neck, N. Y.

Design for such applications as video peaking, rf and IF amplifiers, and filter networks, these coils feature compact plastic forms, four terminals (two of which may be used as separate tiepoints), and durable windings.

For development use, these coils (Series 102) are individually boxed and labelled. For complete data, write, North Hills Electric Co., or telephone Bayside 4-0540.

Amplifier and Filter Literature

Spencer-Kennedy Laboratories, Inc., 186 Massachusetts Ave., Cambridge 39, Mass., has several new bulletins to aid in solving amplifier and filter problems.

The first is a 200 mc bandwidth untuned amplifier having a gain of 20 db, and a response characteristic that is flat within 1½ db (Model 202P Wide-Band Chain Amplifier).

The second is a 100 mc bandwidth untuned pulse amplifier having a gain of 30 db, and a maximum output voltage of 125 volts (Model 214 Chain Pulse Amplifier).

The third is a 20 db gain untuned television amplifier with a bandwidth of from 40 to 220 mc that will not fail because of a tube failure (Model 212TV Chain Amplifier).

The fourth is a high-pass, low-pass, band-pass electronic filter that has a continuously variable cut off from 20 cps to 200 kc and no insertion loss (Series 300 Variable Electronic Filter).

Winding Tester

A newly-developed portable electronic winding tester that can detect a single shorted turn of #40 AWG wire was announced today by **Columbia Technical Corp.** 5 E. 57 St., New York 22, N. Y.

Known as the PMD Tester, this instrument detects the location as well as the nature of an electrical fault in any type of winding. The presence and location of open

(Continued on page 52A)

ANOTHER CBS-HYTRON FIRST YOU'LL BE BUYING SOON

CBS-HYTRON 1AX2

NEW HEAVY-DUTY TV HIGH-VOLTAGE RECTIFIER CAN TAKE IT!

TV high-voltage rectifiers take a beating: Terrific variations occur in applied filament voltage . . . 0.8 to 2.4 volts! Sudden arcs in the rectifying system place destructive electromechanical stresses on the filament. And the increasingly larger TV picture tubes demand peak emission and peak inverse voltage simultaneously. The new CBS-Hytron 1AX2 was especially designed to take such rough treatment and come up smiling.

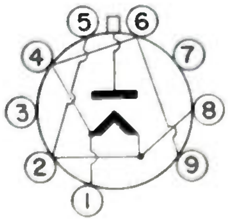
1AX2 DATA

The CBS-Hytron 1AX2 is a compact, 9-pin miniature TV pulse rectifier. Plate is brought out to top cap and filament is oxide-coated. Absolute maximum ratings are: peak inverse plate voltage, 25,000 volts; d-c load current, 1.0 ma.; and steady-state peak plate current, 11.0 ma.

Typical Operation — TV Pulse Rectifier

Filament voltage	1.4 v ± 10%
Filament current	650 ma
Positive-pulse plate voltage	20,000 v
Negative-pulse plate voltage	5,000 v
Peak inverse plate voltage	25,000 v
D-c output voltage	20,000 v
D-c load current	300 μa

BOTTOM VIEW OF SOCKET



ADVANTAGES OF NEW CBS-HYTRON 1AX2

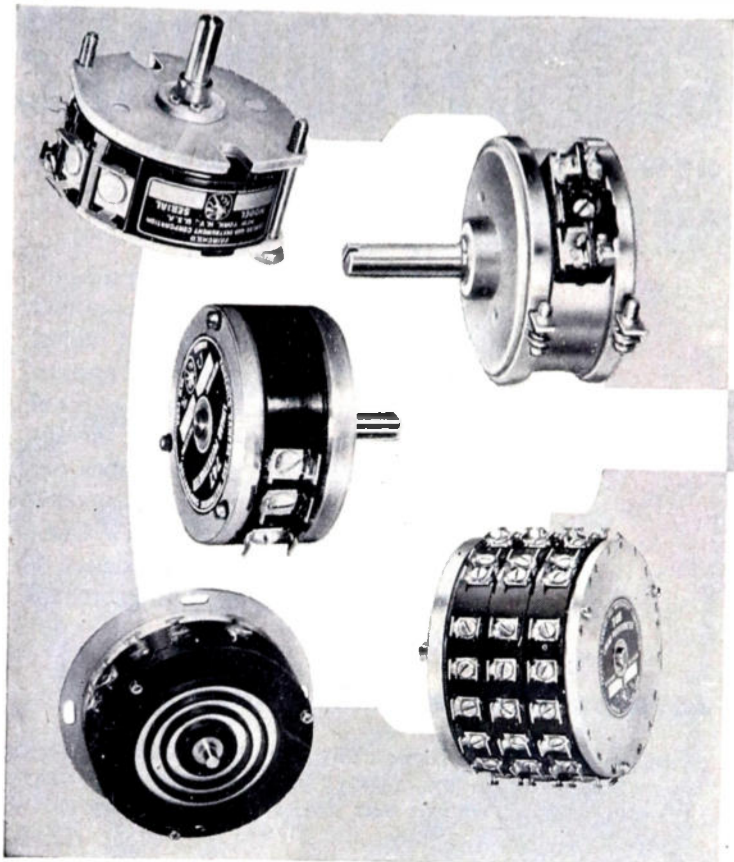
1. Rugged, high-wattage filament of CBS-Hytron 1AX2 has adequate peak emission for the new, larger TV picture tubes. 1AX2 may be run simultaneously at both its peak inverse voltage and maximum d-c current.
2. Higher load of 1AX2 filament on transformer tends to regulate filament voltage. Eliminates need for limiting resistor. Yet lower plate-to-filament capacitance (0.7 μμf) of 1AX2 prevents loss of high voltage.
3. Insulated tension bar (patent applied for) through center of 1AX2 coiled filament limits destructive movement of filament by electromechanical stresses.
4. Filament of 1AX2 is located in base and shielded to eliminate bombardment of cool ends of filament by gas molecules.
5. An overloaded 1X2A may be replaced with its big brother, the CBS-Hytron 1AX2, by simply removing the limiting resistor. In rare cases, it may be necessary to add another turn to the secondary of the filament transformer to obtain the required 1.4 volts for the 1AX2.

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

CBS HYTRON RADIO TUBES

MAIN OFFICE: SALEM, MASSACHUSETTS

Where control accuracy is important... use Fairchild Precision Potentiometers



In many control systems, accuracy is the all-important factor. But the accuracy of an entire system depends on the proper functioning of each component.

Fairchild precision potentiometers are designed and built to meet extreme accuracy requirements. That's why they are being selected more and more for control systems where accuracy, limited weight and space, and simplicity are paramount.

These potentiometers perform mathematical computations in electrical computing systems for machine-tool controls, aircraft instruments, aerial-camera controls, radar, guided missiles, gunsights, and analog computers of all types. They are available in non-linear and linear types and in ganged combinations of either or both, with windings to meet your requirements. If you have computer or control problems, ask for our recommendations about using Fairchild Precision Potentiometers to solve them.

HOW PRECISION IS DESIGNED AND BUILT INTO FAIRCHILD POTENTIOMETERS

1. Shaft is centerless-ground from stainless steel to a tolerance of +0.0000, −0.0002 in. which, together with precision-bored bearings, results in radial shaft play of less than 0.0009 in.

2. Mounting plate has all critical surfaces accurately machined at one setting to insure shaft-to-mounting squareness of 0.001 in./in. and concentricity of shaft to pilot bushing within 0.001 in. FIR.

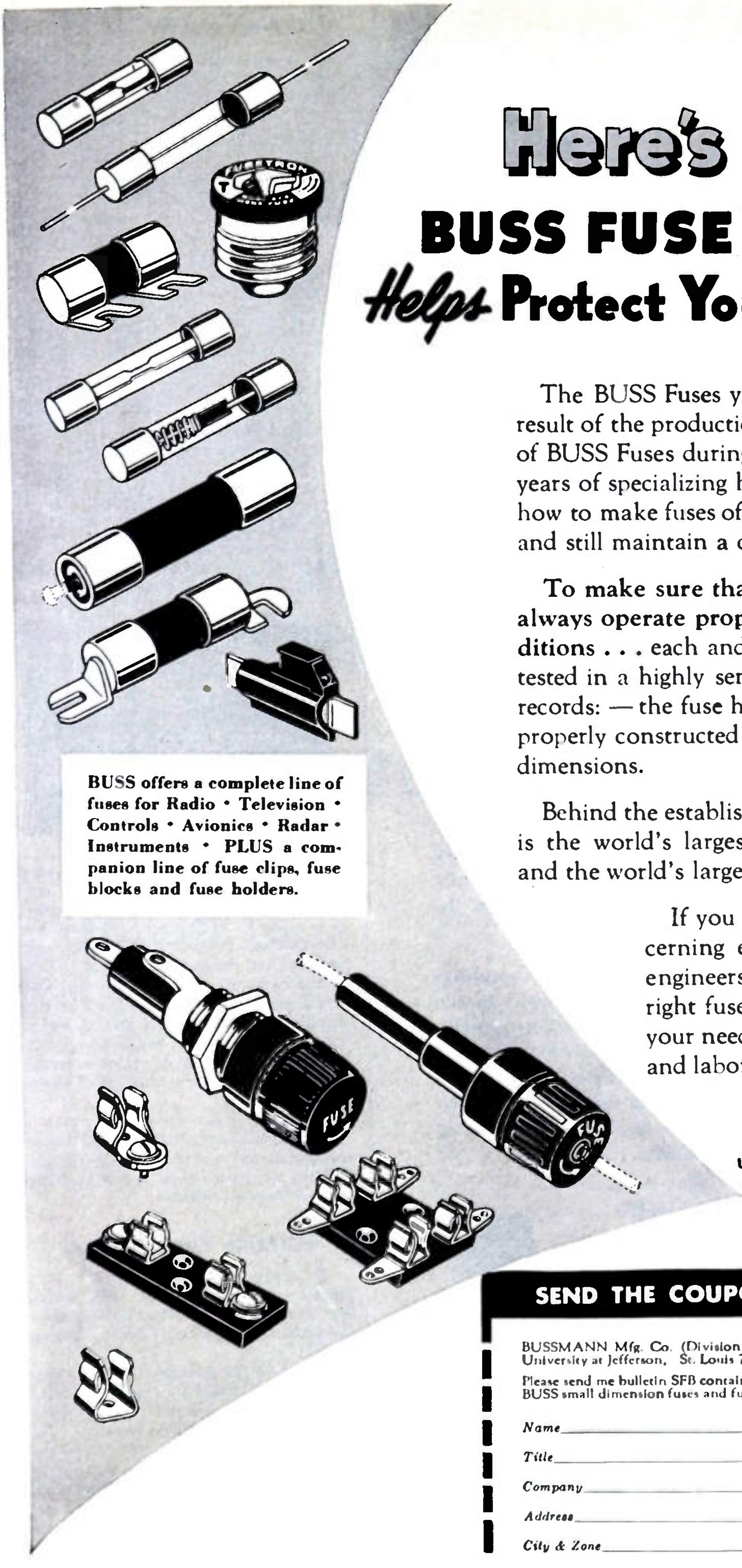
3. Housing is precision-machined from aluminum bar stock. Close tolerance of this construction permits ganging up to 20 units on a single shaft with no eccentricity of the center cups, even though only two bearings are used for the entire gang.

4. Windings are custom-made by an exclusive technique. Guaranteed accuracy of windings in standard types is: linear, 0.5%; non-linear, 1.0%. Higher accuracies (to 0.05%) are available in other types. Guaranteed service life is 1,000,000 cycles.

DO YOU HAVE CONTROL PROBLEMS?

Fairchild Sample-Laboratory engineers are available to help you with potentiometer problems. To get the benefit of their knowledge and experience write today, giving complete details, to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, New York, Department 140-26HI.*

FAIRCHILD
PRECISION POTENTIOMETERS



Here's How BUSS FUSE "Know How" Helps Protect Your Reputation

The BUSS Fuses you buy today are the result of the production of millions upon millions of BUSS Fuses during the past 37 years. These years of specializing have taught BUSS engineers how to make fuses of unquestioned high quality — and still maintain a competitive price.

To make sure that a BUSS Fuse will always operate properly under service conditions . . . each and every BUSS Fuse is tested in a highly sensitive electronic device that records: — the fuse has the right capacity, is properly constructed and right in all physical dimensions.

Behind the established reputation of BUSS Fuses is the world's largest fuse research laboratory and the world's largest fuse production capacity.

If you have a special problem concerning electrical protection, let our engineers help you select or design the right fuse, or fuse mounting, to meet your needs. Our staff of engineers and laboratory are at your service.

BUSS offers a complete line of fuses for Radio • Television • Controls • Avionics • Radar • Instruments • PLUS a companion line of fuse clips, fuse blocks and fuse holders.

BUSSMANN MFG. CO.
University at Jefferson St. Louis 7, Mo.
Division of McGraw Electric Company

SEND THE COUPON FOR COMPLETE FACTS

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

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

Name ________________

Title ________________

Company ________________

Address ________________

City & Zone __________ *State* ______ IRE-752

News—New Products

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

(Continued from page 48A)

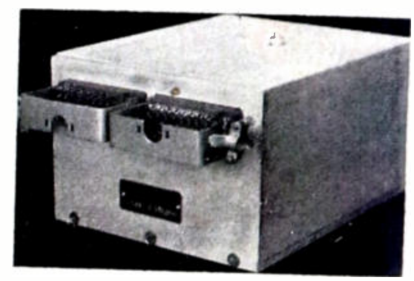
circuits, shorts to ground, shorts between phases or between turns, reversed terminal connections, wire breakage, and material defects can be found in a minimum of time even with unskilled labor.

The tester can be used with equal sensitivity on both ac and dc devices. Unlike ordinary testers, which require that the windings be energized for testing purposes, the PMD test probes generate their own field in the probes. The field oscillates at 800 cps, in contrast to 60 cps which accounts for the high sensitivity.

The PMD tester consists of a power supply unit, a vacuum tube oscillator, a a regulator and rectifier circuit.

Sampling Switch

A new type of sampling switch for zero drift correction of dc amplifiers in analog computers is announced by **Applied Science Corp.**, P. O. Box 44, Princeton, N. J.



Motor-driven, the switch now makes possible the use of one ac amplifier alone for zero correction and gain improvement of as many as 30 dc computing amplifiers. The unit has two poles, with 60 contacts per pole, and the sampling rate is $3\frac{1}{2}$ RPS. It has inter-contact resistance over 1,000 megohms. The design is compact for easy installation and weighs $7\frac{1}{2}$ pounds including the 110 volt 60 cps motor. In tests the switch has proved to have an extremely low noise level and a service life of several thousand hours.

The same design and circuitry may be used for special sampling rates and contact configurations as ordered. For detailed bulletin or inquiries on special switches write to Applied Science.

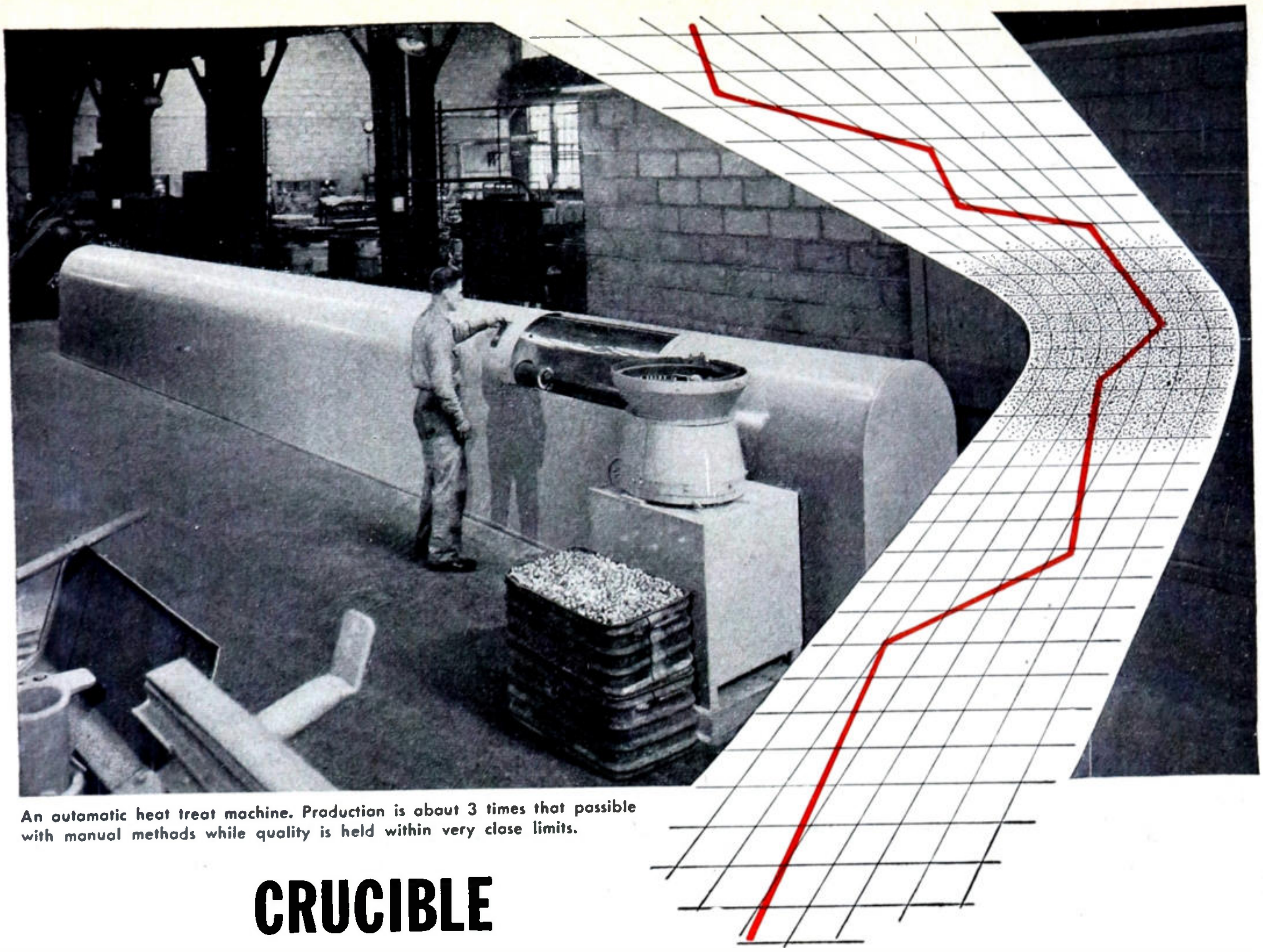
Portable Survey Meter

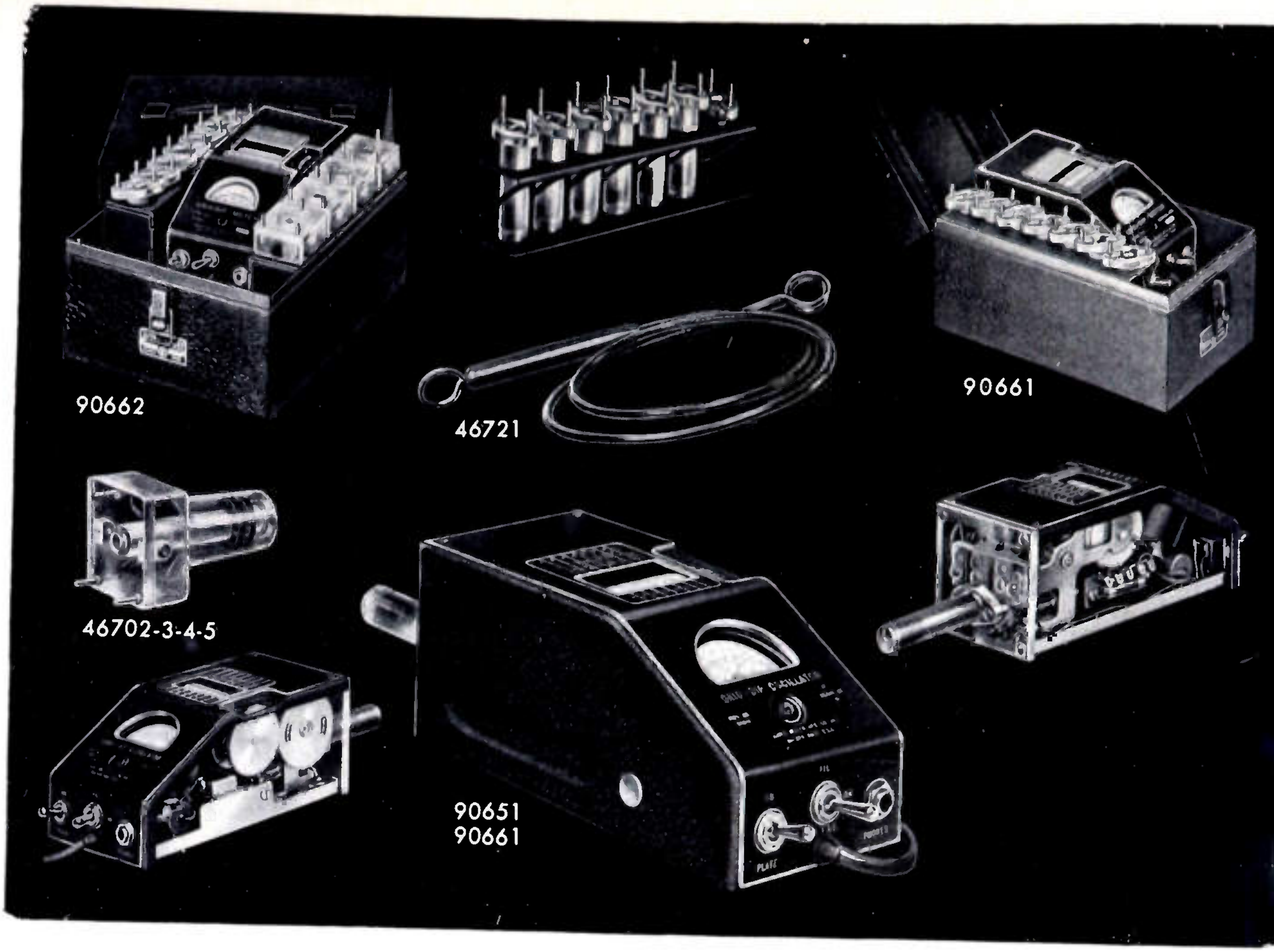
A newly designed wide-range beta-gamma survey meter, Model 2585, has just been announced by **Nuclear Instrument & Chemical Corp.**, 229 W. Erie St., Chicago 10, Ill.

Based on the principle of the wartime "Cutie Pie," Nuclear's new type offers many features which eliminate the disadvantages of the previous instrument.

Model 2585 is so designed that it can be set upright for continuous monitoring. It is light weight, with the circuit and battery housing ahead of and close to the fingers instead of being balanced above the

(Continued on page 60A)





Designed for Application

Grid Dip Meters

Millen Grid Dip Meters are available to meet all various laboratory and servicing requirements.

The 90662 Industrial Grid Dip Meter completely calibrated for laboratory use with a range from 225 kc. to 300 mc. incorporates features desired for both industrial and laboratory application, including three wire grounding type power cord and suitable carrying case.

The 90661 Industrial Grid Dip Meter is similar to the 90662 except for a reduced range of 1.7 to 300 mc. It likewise incorporates the three wire grounding type cord and metal carrying case.

The 90651 Standard Grid Dip Meter is a somewhat less expensive version of the grid dip meter. The calibration while adequate for general usage is not as complete as in the case of the industrial model. It is supplied without grounding lead and without carrying case. The range is 1.7 to 300 mc. Extra inductors available extends range to 220 kc.

The Millen Grid Dip Meter is a calibrated stable RF oscillator unit with a meter to read grid current. The frequency determining coil is plugged into the unit so that it may be used as a probe.

These instruments are complete with a built-in transformer type A.C. power supply and interminal terminal board to provide connections for battery operation where it is desirable to use the unit on antenna measurements and other usages where A.C. power is not available. Compactness has been achieved without loss of performance or convenience of usage. The incorporation of the power supply, oscillator and probe into a single unit provides a convenient device for checking all types of circuits. The indicating instrument is a standard 2 inch General Electric instrument with an easy to read scale. The calibrated dial is a large 270° drum dial which provides seven direct reading scales, plus an additional universal scale, all with the same length and readability. Each range has its individual plug-in probe completely enclosed in a contour fitting polystyrene case for assurance of permanence of calibration as well as to prevent any possibility of mechanical damage or of unintentional contact with the components of the circuit being tested.

The Grid Dip Meters may be used as:

1. A Grid Dip Oscillator
2. An Oscillating Detector
3. A Signal Generator
4. An Indicating Absorption Wavemeter

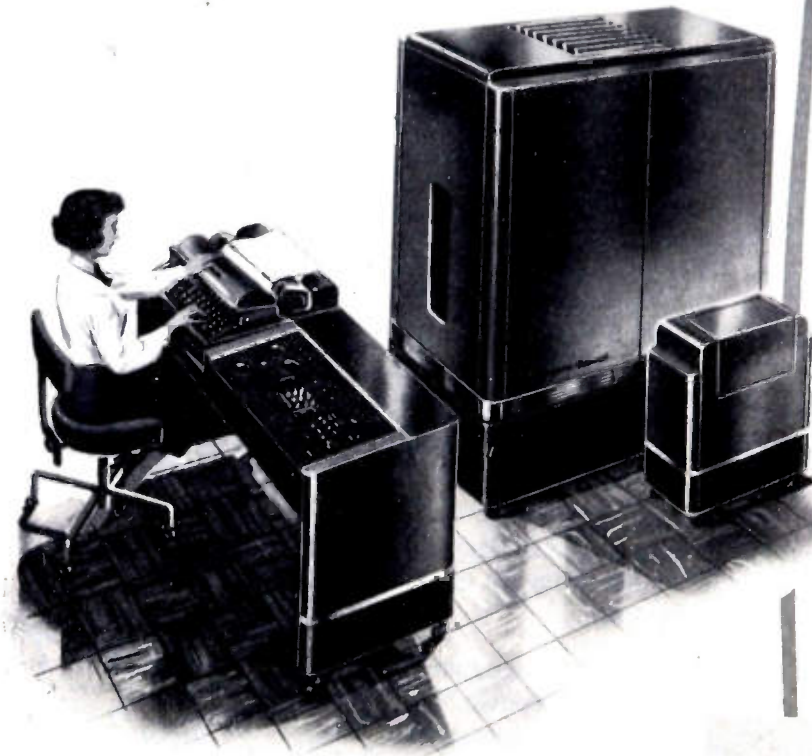
The most common usage of the Grid Dip Meter is as an oscillating frequency meter to determine the resonant frequencies of de-energized tuned circuits.

Size of Grid Dip Meter only (less probe): 7 in. x 3⅛ in. x 3⅝ in.

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.



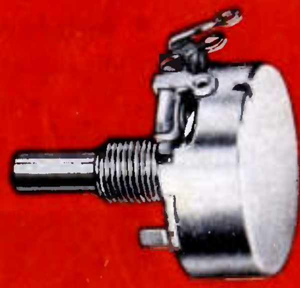
TYPE 65

½ watt 70° C, ¾" diameter miniaturized variable composition resistor.

Resistance	CTS Part CTS Shaft Type RE	CTS Part Locking Bushing CTS Shaft Type LT-2
250±10%	X3516	X3530
500±10%	X3517	X3531
1000±10%	X3518	X3532
2500±10%	X3519	X3533
5000±10%	X3520	X3534
10,000±10%	X3521	X3535
25,000±10%	X3522	X3536
50,000±10%	X3523	X3537
100,000±10%	X3524	X3538
250,000±10%	X3525	X3539
500,000±10%	X3526	X3540
1 Meg±20%	X3527	X3541
2.5 Meg±25%	X3528	X3542

TYPE 95, JAN-R-94, Type RV4

2 watt 70°C, 1⅛" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.



Resistance	JAN-R-94 TYPE RV4 JAN Shaft Type SD	JAN-R-94 TYPE RV4 JAN Shaft Type RJ	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	RV4ATSD101A	RV4ATRJ101A	W3160
250±10%	RV4ATSD251A	RV4ATRJ251A	W3161
500±10%	RV4ATSD501A	RV4ATRJ501A	W3162
1000±10%	RV4ATSD102A	RV4ATRJ102A	W3166
2500±10%	RV4ATSD252A	RV4ATRJ252A	W3163
5000±10%	RV4ATSD502A	RV4ATRJ502A	W3164
10,000±10%	RV4ATSD103A	RV4ATRJ103A	W3167
25,000±10%	RV4ATSD253A	RV4ATRJ253A	W3168
50,000±10%	RV4ATSD503A	RV4ATRJ503A	W3169
100,000±10%	RV4ATSD104A	RV4ATRJ104A	W3170
250,000±10%	RV4ATSD254A	RV4ATRJ254A	W3171
500,000±10%	RV4ATSD504A	RV4ATRJ504A	W3172
1 Meg±20%	RV4ATSD105B	RV4ATRJ105B	W3173
2.5 Meg±20%	RV4ATSD255B	RV4ATRJ255B	W3165
5 Meg±20%	RV4ATSD505B	RV4ATRJ505B	W3159

TYPE 45, JAN-R-94, Type RV2

¼ watt, 15/16" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.

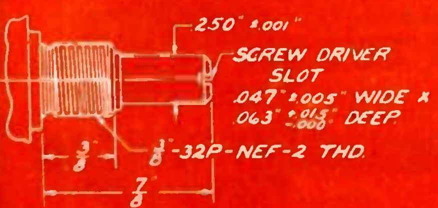
Resistance	RV2, JAN Shaft Type SD CTS Part	JAN-R-94 TYPE	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	A5876	RV2ATSD101A	A5922
250±10%	A5877	RV2ATSD251A	A5923
500±10%	A5878	RV2ATSD501A	A5924
1000±10%	A5879	RV2ATSD102A	A5925
2500±10%	A5880	RV2ATSD252A	A5926
5000±10%	A5881	RV2ATSD502A	A5927
10,000±10%	A5882	RV2ATSD103A	A5928
25,000±10%	A5883	RV2ATSD253A	A5929
50,000±10%	A5884	RV2ATSD503A	A5930
100,000±10%	A5885	RV2ATSD104A	A5931
250,000±10%	A5886	RV2ATSD254A	A5932
500,000±10%	A5887	RV2ATSD504A	A5933
1 Meg±20%	A5888	RV2ATSD105B	A5934
2.5 Meg±20%	A5889	RV2ATSD255B	A5935

TYPE 35, JAN-R-94, Type RV3

½ watt, 1⅛" diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.

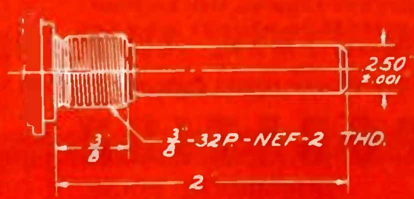
Resistance	RV3, JAN Shaft Type SD CTS Part	JAN-R-94 TYPE	CTS Part Non-JAN Locking Bushing CTS Shaft Type LT-1
100±10%	A5861	RV3ATSD101A	A5907
250±10%	A5862	RV3ATSD251A	A5908
500±10%	A5863	RV3ATSD501A	A5909
1000±10%	A5864	RV3ATSD102A	A5910
2500±10%	A5865	RV3ATSD252A	A5911
5000±10%	A5866	RV3ATSD502A	A5912
10,000±10%	A5867	RV3ATSD103A	A5913
25,000±10%	A5868	RV3ATSD253A	A5914
50,000±10%	A5869	RV3ATSD503A	A5915
100,000±10%	A5870	RV3ATSD104A	A5916
250,000±10%	A5871	RV3ATSD254A	A5917
500,000±10%	A5872	RV3ATSD504A	A5918
1 Meg±20%	A5873	RV3ATSD105B	A5919
2.5 Meg±20%	A5874	RV3ATSD255B	A5920
5 Meg±20%	A5875	RV3ATSD505B	A5921

JAN SHAFT TYPE SD



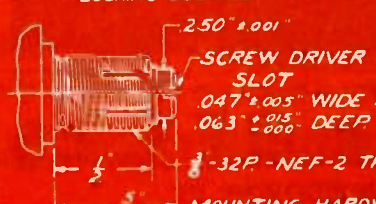
MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 9/16 HEX. x 1/32
LOCK WASHER #1920A

JAN SHAFT TYPE RJ



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 9/16 HEX. x 1/32
LOCK WASHER #1920A

CTS SHAFT TYPE LT-1
LOCKING BUSHING



MOUNTING HARDWARE ASSEMBLED
MOUNTING NUT 9/16 HEX. x 1/32
LOCK NUT 1/2 HEX. x 3/32
LOCK WASHER #1920A

News—New Products

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

(Continued from page 52A)

handle. The five position switch, providing zero setting and three ranges up to 5000 mr/hr, is located at the top of the pistol-grip handle to allow one-hand operation of the instrument. Zero adjust and calibrate controls are on top, ahead of the 2½ inch meter which slopes toward the user for easy reading. The instrument is finished in smooth gray baked enamel and weighs 3 pounds, complete with batteries.

Projection Kinescope

Tube Dept., Radio Corp. of America, Harrison, N. J., announces the 7WP4, a new projection kinescope for use in theater-television equipment. It is capable of providing a picture 20 feet by 15 feet when used with a suitable reflective optical system.

The 7WP4 is designed with a faceplate curvature intended for use in an optical system having an 80-foot throw. The tube employs electrostatic focus and magnetic deflection.

Operating with a maximum ultor voltage of 80,000 volts and a maximum focusing-electrode voltage of 20,000 volts, the 7WP4 incorporates such high-voltage design features as a bulb having corrugated side walls with insulating coating to provide a long leakage path over its external surface, an inner cone-neck section to provide adequate vacuum insulation between internal ultor coating and outer neck section, and one high-voltage envelope connection. All other connections are made through a plastic-filled diheptal 14-pin base.

New Catalog

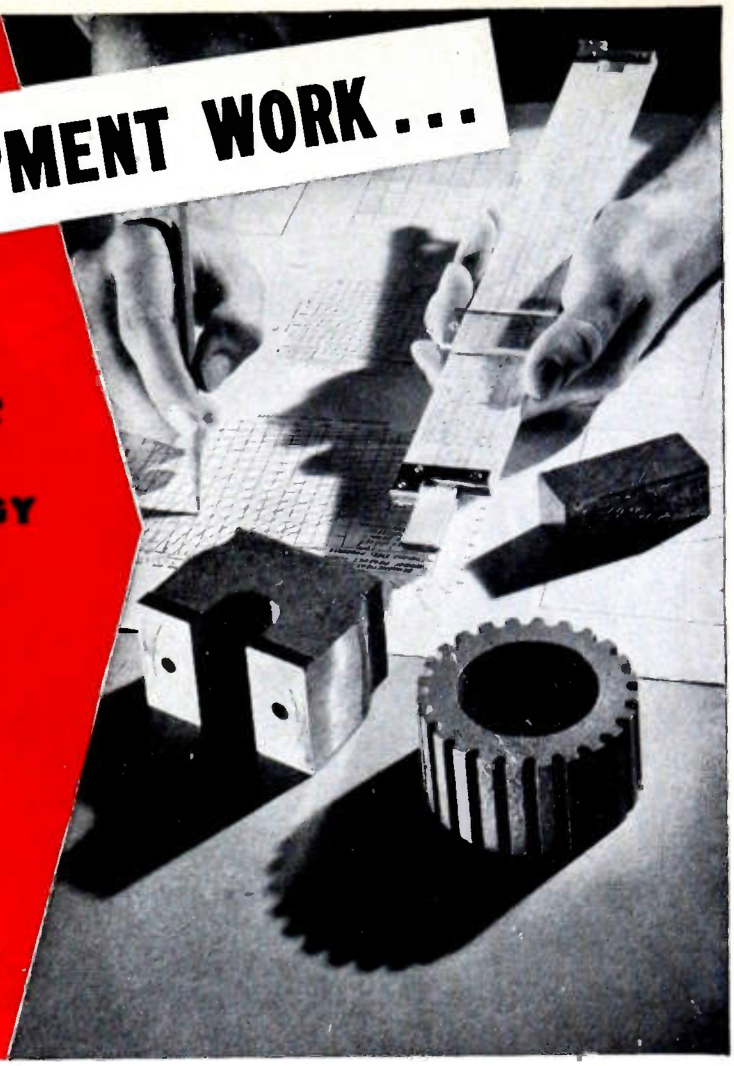
P. R. Mallory & Co., Inc., 3029 E. Washington St., Indianapolis 6, Ind., announces a Product Index is now available providing specific information in condensed form concerning Mallory electrochemical, electromechanical, electronic and metallurgical products.

The catalog is not highly technical in nature but is designed to acquaint engineer and layman alike with basic data on available products. It includes brief descriptions of the specifications, features and applications of the complete line of Mallory batteries, capacitors, contacts, rectifiers, resistors, switches, vibrators, metals and ceramics, tuners, and resistance welding supplies.

Copies of the Product Index can be obtained by writing P. R. Mallory & Co., Inc.

(Continued on page 106A)

FOR YOUR DEVELOPMENT WORK . . .



Permanent Magnets

WITH 16% GREATER ENERGY

INDIANA HYFLUX* ALNICO V

*The permanent magnet material that offers an energy product averaging 5⅓ million BH max or more, with 5¼ million guaranteed.

Whether your problem is new design or product improvement, take advantage of the greater energy product INDIANA HYFLUX Alnico V offers!

These exclusive, new, super strength permanent magnets mean lower production costs, more compact design and higher efficiency for your products.

What's more, INDIANA HYFLUX — with its 16% greater energy product — costs not a penny more than regular Alnico V!

Here's still another bonus you'll enjoy! THE INDIANA STEEL PRODUCTS COMPANY, world's largest producer of permanent magnets, offers free of charge its wealth of experience and "know-how" that has developed more than 30,000 permanent magnet applications.

Let INDIANA engineers help you with your design problems. They can supply — out of stock — many types and sizes of INDIANA HYFLUX Alnico V for your experiments, can suggest those best suited to your product.

INDIANA is the only manufacturer furnishing all commercial grades of permanent magnet alloys. You have a choice of cast, sintered, formed or ductile materials.

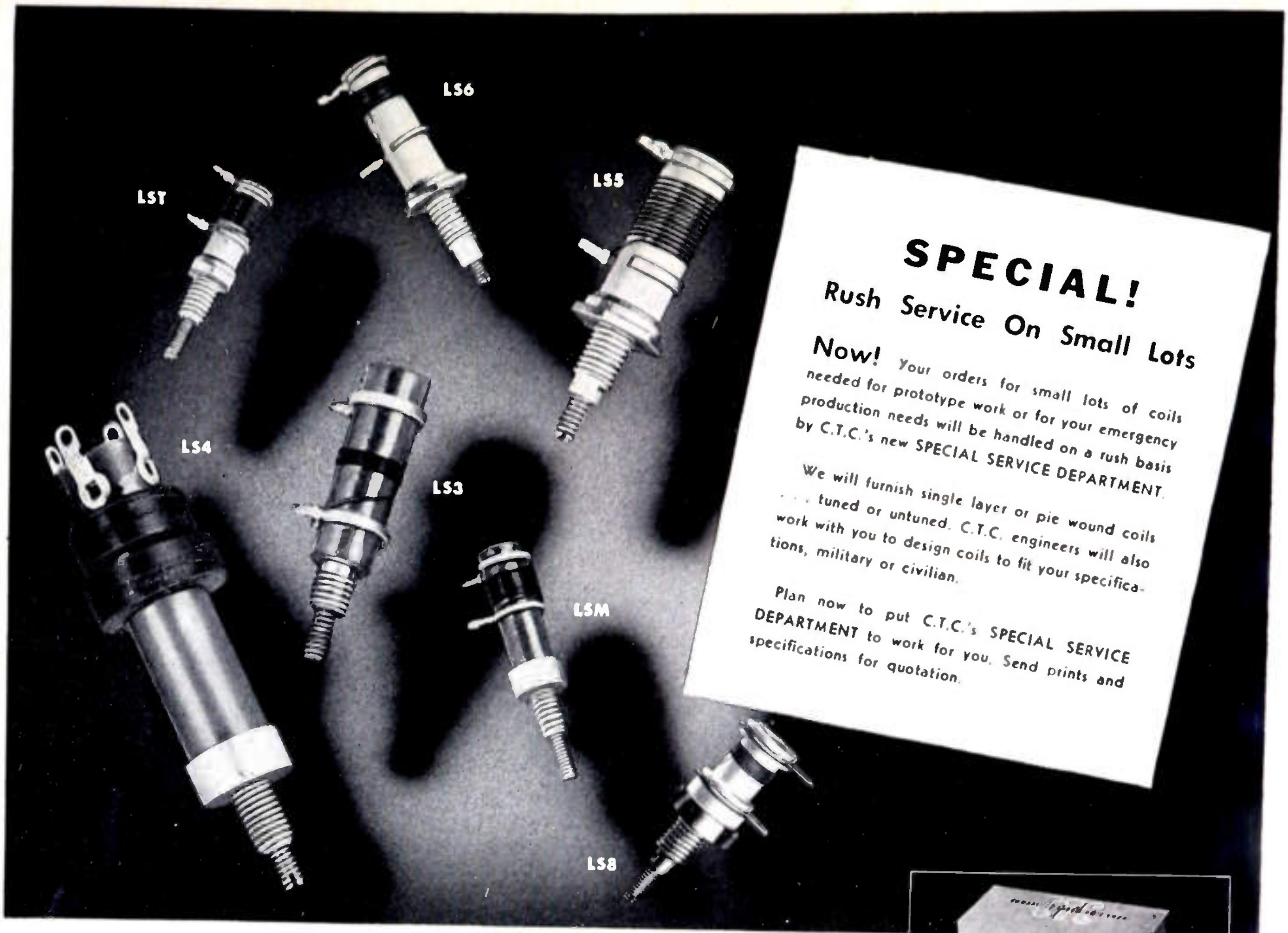
Why delay — write or phone INDIANA today. Ask for Catalog No. 11G-5 that describes stock experimental magnets.

INDIANA PERMANENT MAGNETS

THE INDIANA STEEL PRODUCTS COMPANY

VALPARAISO, INDIANA • • • Sales Offices Coast to Coast

SPECIALISTS IN "PACKAGED ENERGY" SINCE 1908



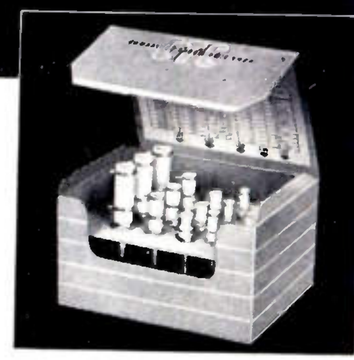
Here are the coils you want ...the way you want them!

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

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

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

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



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

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

CAMBRIDGE THERMIONIC CORPORATION

CTC

custom or standard . . . the guaranteed components

New catalog! Send for your copy now.

PROCEEDINGS OF THE I.R.E.®

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 40 — *July, 1952* — NUMBER 7

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

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

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

Irving Wolff

DIRECTOR, 1952-1953

Irving Wolff was born on July 6, 1894, in New York, N. Y. He received the B.S. degree in physics at Dartmouth College in 1916, and the Ph.D. degree at Cornell University in 1923.

Dr. Wolff was an instructor in physics at Iowa State College during 1919. From 1920 to 1923, he filled the same post at Cornell University, becoming the Heckscher Research Fellow the next year.

In October, of 1924, Dr. Wolff joined the Radio Corporation of America, where he has remained. From his first year at RCA through 1931, his research was devoted to the field of acoustics. In 1932, he initiated a program of microwave research. Following this with a program of radar research in 1934, he supervised various groups at RCA who were conducting research for military agencies during World War II. Immediately following the war, Dr. Wolff and some of his RCA associates developed the basis for the Teleran System of Air Navigation. In 1946, he was appointed director of RCA's Radio Tube Research Laboratory, and in 1951, Director of Research of RCA Laboratories.

Dr. Wolff received the IRE Fellow Award in 1942 for his, "basic research in centimeter-wave radio and application of it to the development of navigation instruments," and the Distinguished Public Service Award of the U. S. Navy in 1949, for pioneer research in radar.

Dr. Wolff joined the Institute in 1927, and has been active on a number of IRE technical committees concerned with acoustics, electronics, and publication. He has been a member of the IRE Board of Editors and the IRE Education Committee since 1945. He is now serving on the IRE Appointments and Nominations Committee, and is the IRE Regional Director of Region 3. He was Chairman of the Philadelphia Section, 1936-1937.

Dr. Wolff is a fellow of the American Association for the Advancement of Science and the Acoustical Society of America, and a member of the Physical Society and Sigma Xi.

Universal Engineering Shorthand

ALLEN F. POMEROY

It has long been evident that, for maximum utility and convenience, graphical symbols should be uniform throughout any field of engineering and its related industries.

The underlying considerations leading to this conclusion are admirably summarized in the following guest editorial, prepared by a Senior Member of the Institute, who is a co-chairman of the A.S.A. Sectional Committee on Graphical Symbols and Designations for Use on Drawings, and a member of the Technical Staff of the Bell Telephone Laboratories.—*The Editor.*

Graphical symbols for electrical and communications engineering can be a shorthand that all engineers, technicians, and draftsmen can understand, or they can amount to nothing more than a means of cramming a considerable amount of information into a small space on a given drawing. When every organization has its own symbols, the latter case is true. An engineer reading a schematic which originated in another organization cannot be sure that he understands each symbol. In order to be sure, a legend must accompany every diagram stating the significance of each symbol.

Engineers in many companies must draw and explain schematics used in their own particular organizations, and, in addition, those used by their customers, by other organizations, by the Armed Services, and also those used on contracts. Sometimes the contracting organization can persuade its customers to accept schematics and the corresponding instruction manuals for symbols unique to the contracting organization. In this case, the customer is forced to train its technicians in two symbol languages: first, the symbol language it uses normally; and second, the symbol language unique to the contracting organization.

It seems logical that in the best interests of all concerned, there should be only one symbol language.

Can such a language be agreed upon? Great progress has been made in the past year in creating confidence among the workers in the electrical and communications fields toward the attainment of a single language. An industry-wide task group of the American Standards Association has already prepared a draft of a standard on graphical symbols applicable to the fields of telecommunications, control engineering, and power engineering. Similar work on standards for military use is being carried forward by the Munitions Board Standards Agency in co-ordination with ASA. On the international level standards are being prepared by International Electrotechnical Commission.

All persons tend to cling to that with which they are familiar. This is a good tendency. Change for the sake of change is not necessarily an improvement; but the change that is adopted by mutual agreement, is progress. An engineer, by the nature of his training, looks at problems objectively, reviews all the facts, and proposes a solution likely to be accepted, provisionally, by all concerned.

It has been found by experience that conflicts in standards between strong groups in any one organization are often difficult to resolve within that organization. However, when representatives of these groups meet in a task group with representatives of other organizations, mutual and satisfactory compromises can usually be evolved.

A new day is dawning for electrical and communications engineers. There is hope that a uniform graphical-symbols shorthand that all can understand will be agreed upon universally.

Research and Development for National Defense*

EDWIN A. SPEAKMAN†, SENIOR MEMBER, IRE

Summary—This article describes the duties, composition, and major operations of the Research and Development Board of the Department of Defense. This Board is the agency charged with coordination of the research and development programs carried on by the Army, Navy, and Air Force. In the article, particular attention is given to the activities and scope of the RDB's Committee on Electronics and to new developments, both military and nonmilitary, in the field of electronics.

The author uses examples to illustrate the point that the application of science and technology to warfare is not a recent development. He warns against easy acceptance of the idea that new weapons and techniques will provide the answer to all the problems involved in our national defense, although the principal aim of the RDB and the research and development agencies of the armed forces is to provide our fighting men with the best possible weapons.

Research and Development Board

Before the dawn of history, man turned periodically to warfare with his neighbor, and from the first battle with clubs down to the present time there has been a constant race for new weapons and countermeasures. Furthermore, the application of science and technology to warfare is not new. We think of biological warfare, for example, as a revolutionary and untried weapon, but early historical accounts tell of the ancients catapulting the bodies of their dead over the walls of besieged cities to spread the disease which they knew would follow, although they did not know why. And thirteen centuries later, Genghis Khan, famous ruler of the Mongols, used chemicals in the form of huge balls of pitch and sulfur thrown at the enemy to produce a combination of screening smoke, choking fumes, and incendiary effects, thus ushering in the dread chemical warfare that was used with such telling effect by the Germans in World War I. A third example is the machine gun. It has been generally supposed that this weapon developed from the Gatling gun, invented during the American Civil War; yet a device embodying the principle of the modern machine gun was invented by the Greek engineer Dionysius of Alexandria in the third century B.C. This weapon fired a succession of arrows supplied by a magazine or hopper, and at fairly close ranges gave a fire power unchallenged by any other light ordnance piece until the invention of the longbow in the late Middle Ages.

It is therefore not unusual that the race for scientific weapons continues—it has had a long and eventful history—but it is startling to realize that today the Department of Defense and the Atomic Energy Commission engage the services of almost two thirds of the nation's scientists and engineers and that the program for military preparedness represents an annual expenditure of one and one-half billion dollars.

The size of the military research and development effort is not only startling, but sobering and thought-provoking. We are confronted by the fact that the greatest portion of the creative thinking and effort of our scientists and engineers is, of necessity, being concentrated on weapons, devices, and techniques of warfare, and on countermeasures. Not only is the country, as a whole, spending a vastly larger sum on research and development than ever in its history, but most of this is going for military purposes.

Obviously, this phenomenon is having a profound effect on our culture. There have been diversions and disruptions in the fields of teaching and of academic and industrial research, and basic research, the lifeblood of scientific progress, has been subordinated to applied research and to development. As Dr. Vannevar Bush stated in the most recent report of the Carnegie Institution of Washington, there may be grave dangers inherent in increased governmental regimentation of scientific research, although there appears to be no alternative method of administration.

It is not my intent, however, to contemplate philosophically the possible effects of these things, no matter how strongly we may feel about them. We must adjust ourselves to the realities of the times if we are to accomplish the enormous tasks ahead of us. We want to make sure that if our men are called to the defense of the nation again they will be equipped as well as it is possible to equip them. The responsibility of the RDB lies in this area, our primary duty being to guide military research and development toward the achievement of maximum weapons effectiveness for our Armed Forces. To see how the RDB performs this vital function, let us first examine briefly the structure and responsibilities of the Board and then consider the work of the Committee on Electronics, with specific examples of how research and development are related to military plans and strategy.

Duties of the Board

In the preceding paragraph I have set forth the over-all mission of the RDB. More specifically, the Board is directed by law:

(1) to prepare a complete and integrated plan of research and development for military purposes;

(2) to co-ordinate research and development among the military departments and allocate responsibility for joint programs; and

(3) to consider the interaction of research and development and strategy, and to advise the Joint Chiefs of Staff in connection therewith.

I want to emphasize here that the actual research and development work is done by the Services themselves; the RDB is not an operating organization. Part of the Services' research is carried out in their own installations, but most of it is handled on contract with industry and academic institutions. The RDB supplies leadership and integration of this total program to assure that the available money and talent are expended on the most vital areas and not wasted on less essential things.

Composition of the RDB

The Board proper is composed of a civilian chairman and two representatives from each of the Military Departments. Mr. Walter G. Whitman, currently on leave from the Massachusetts Institute of Technology, where he is head of the Chemical Engineering Department, is chairman. The Army representatives are Assistant Secretary Earl D. Johnson and Major General Kenneth D. Nichols, Chief of Research and Development. Representating the Navy are the Honorable John F. Floberg, Assistant Secretary for Air, and Rear Admiral M. E. Curts, Assistant Chief of Naval Operations (Readiness). The Air Force members are Under Secretary Roswell L. Gilpatric and Major General Laurence C. Craigie, Deputy Chief of Staff, Development. The seven-man Board makes broad policy decisions affecting the entire military research and development program.

The Board itself is a part-time group, meeting, ordinarily, about once a month. Serving under the Board and preparing material for its consideration is a staff organization consisting of approximately 300 civilian and military personnel and a committee and panel organization of 2,400 part-time military and civilian experts. The civilian experts, most of whom serve without compensation, are drawn from industry, the universities, and other government agencies. All are recognized leaders and authorities in their fields of science or engineering.

Each of the fifteen committees of the Board has cognizance over a specific field or certain type weapon. At present we have committees on aeronautics, atomic energy, biological warfare, chemical warfare, electronics, equipment and supplies, fuels and lubricants, geophysics and geography, guided missiles, human resources, materials, medical sciences, navigation, ordnance, and technical information.

The area of responsibility of a committee is normally subdivided and assigned to panels. As an example of a structure of this type, Fig. 1 shows the panels of the Committee on Electronics.

Through the occasional rotation of committee and panel members, new ideas and viewpoints are introduced while, at the same time, continuity of operations is preserved. The committees and panels meet six or eight times a year, as required by their various duties. Meetings are held at Washington or at points where military research work is in progress. Each committee is served by a full-time staff of technical experts.

The duties of the committees are several: They review, in detail, the military research and development program in their assigned field and make recommendations concerning modifications of the program. They periodically assess the state of weapons development in their field. A particularly valuable product of their activity lies in providing an effective forum for the cross exchange of information and ideas. Representatives

* Decimal classification: R010×R560. Original manuscript received by the Institute, March 4, 1952.
† Research and Development Board, Washington 25, D. C.

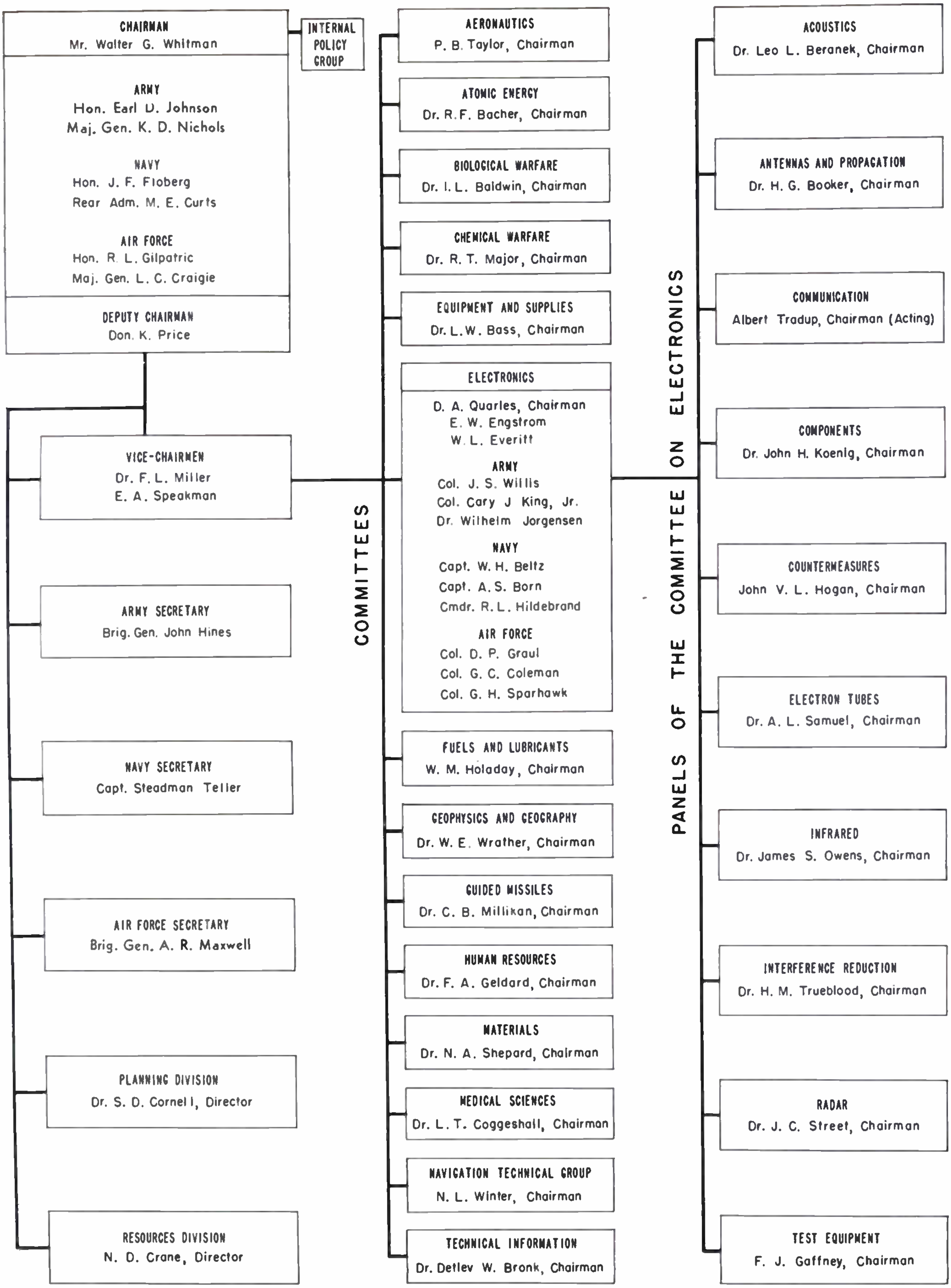

Fig. 1—Research and Development Board, Department of Defense.

from the Services, which actually administer the research, report their plans and the progress of their work, and the men from outside the Departments ask questions, offer advice, and make other valuable contributions toward the improvement of the programs. The committees and panels therefore provide a vital connection between the military research and development program, on the one hand, and industry, the universities, and other government agencies on the other. Many important problems are solved at the committee or panel level.

Committee on Electronics

A brief explanation of the activities and scope of the RDB's Committee on Electronics will probably be of particular interest to readers of this publication. From a standpoint of expenditures, the field of electronics comprises between 20 and 25 per cent of the total military research and development program, or about 250 million dollars. This includes programs under the sole cognizance of the Committee on Electronics and, in addition, those co-ordinated jointly with other committees, such as Guided Missiles, Aeronautics, and Ordnance.

Under the terms of its directive, which is similar in scope and intent to those of other RDB committees, the principal objectives of the Committee on Electronics are co-ordination, guidance, and integration of the Military Departments' efforts toward developing, through electronics research, the best possible weapons and supporting systems for the Armed Forces of the United States.

The Committee's area of interest includes about 3,100 projects which lie in widely divergent technical fields. This poses difficult problems of Committee administration, and the Committee has accordingly divided its field among ten panels encompassing more homogenous areas. These are radar, communication, electronic countermeasures, acoustics (sonar), infra-red, antennas and propagation, components, electron tubes, test equipment, and interference reduction.

Each of these panels consists of a civilian chairman and one or more civilian members, plus a member representing each of the Military Departments. Several of the committee and panel experts are members of the Institute of Radio Engineers, and we are happy to have the resources of this excellent organization as a fountainhead of competent and experienced electronics specialists to serve as consultants and replacements for members.

Periodically, the members of the committee and panels visit electronics installations and contractors engaged in development throughout the country in order to acquaint themselves as fully as possible with work being done in the field.

Main Operations of RDB

The RDB maintains close working relationships with other agencies of the Department of Defense, such as the Joint Chiefs of Staff, the Weapons Systems Evaluation Group, and the Munitions Board, and also with the Central Intelligence Agency, Atomic Energy Commission, National Science Foundation, and National Academy of Sciences.

In order for the Board and its committees to evaluate scientific programs, it is necessary to have some idea of the operational and strategic needs of the Military Departments. This information is supplied by the Joint Chiefs of Staff in the form of strategic plans and military requirements which are used by the committees of the Board in their examination of scientific programs. In this way each important area of research is studied in relation to military needs in order to assure that new weapons and techniques meet the requirements of the Armed Forces.

The Board, on the other hand, advises the Joint Chiefs periodically with reference to new weapons and countermeasures. This information is in the form of reports, or technical estimates, of what new weapons and techniques will result from research, and an estimate of the date when these new weapons will become available. Promoting an even closer tie between the two agencies, the chairman of RDB attends meetings of Joint Chiefs of Staff at which research and development matters are discussed.

In addition, the Board, through its committees, also provides technical guidance to Army, Navy, and Air Force to assist them in formulation of research programs.

The Weapons Systems Evaluation Group, sponsored jointly by the JCS and the RDB, performs operations research in certain broad areas, such as land combat.

The RDB advises the Munitions Board of new developments that will require critical and strategic materials so that the Munitions Board can anticipate the need for such materials in production. On the other hand, the Munitions Board informs the RDB as to those materials which are in short supply and where it is necessary that research effort on substitutes be increased.

Information on the military research and development activities of other countries is, of course, of interest to us, and the Central Intelligence Agency supplies us with pertinent facts on various areas. This information is used for comparative purposes to assess our effort in general and to aid in a prediction of our weapons' performance against any which other countries may now have under development. As is well known, the supply of technical information from certain countries is very tight; the Iron Curtain makes such intelligence most difficult to obtain.

With reference to the National Academy of Sciences, we find that they have certain groups and specialists in various technical fields that can be utilized in the solution of important problems. For example, with the strong backing of the RDB, the Minerals and Metals Advisory Board[1] was established, less than a year ago, under the National Academy of Sciences to undertake the responsibility for examining programs and making recommendations in this field. The MMAB has recently submitted a number of excellent reports, including one on titanium, a new metal which has great significance to American industry.

New Developments—Electronics

Now, after my discussion of the organization of the research program, I should like to mention a few results of that program.

Everyone is aware of the importance of radar in bringing World War II to a successful conclusion and of its many commercial applications today in the navigation of both ships and aircraft. The military program is aimed at increasing the range on small targets, such as aircraft and guided missiles, and at developing high-powered transmitters that will enable us to pick out and track enemy missiles or bombers. Electronics has so many commercial applications that it is only natural that we find it an integral part of most weapons. The electronic equipment of one of our new planes accounts for 40 per cent of its cost. The electronic equipment of one of our large bombers contains some 1,200 vacuum tubes. In fact, we look upon certain type aircraft as flying laboratories, with all their intricate systems for control of the plane, navigation, communication, gun firing, and radar location of targets. As a result of this wide application of electronics we have encountered many new problems that must be solved in order to assure reliable weapons. One urgent problem concerns the development of more reliable electron tubes, but we think that the development of the transistor, largely through the efforts of Bell Telephone Laboratories, offers great promise as a possible solution.

The field of communication is another important branch of electronics research. Recent developments sponsored by the Signal Corps make it possible to send almost 100 voice messages or several television or radar presentations simultaneously over one radio channel. This has great significance to our air defense, since it would be necessary, in the event of war, to have high-speed communication links throughout the continent in order to alert our air-defense centers to the approach of enemy planes.

The requirement for effective electronic countermeasures can hardly be overemphasized. You may remember how, in February, 1941, the German naval vessels, *Gneisenau*, *Scharnhorst*, and *Prinz Eugen*, made their escape through the English Channel from Brest. The British had the Channel completely covered by radar, but their signals were so successfully jammed by the Germans that no radar echoes were obtained from the ships. This was the first significant operational use of countermeasures in World War II.

For repelling Allied bomber attacks, the Germans relied in great measure on antiaircraft artillery—some 15,000 heavy guns controlled by about 3,000 fire-control radars. This presented a formidable obstacle, particularly since our bomber losses at that time were rather high. An Allied countermeasure program was begun in 1944, and the second important operational use of countermeasures occurred late in that year when jammers were installed in our bombers. This operation was most successful in reducing the effectiveness of the German guns.

In World War II our radar-countermeasures units also used small pieces of tinfoil, called "chaff," to produce false signals on enemy radar screens. When a package containing several hundred thousand pieces of this material was dispersed by aircraft the echo produced looked exactly like that of a bomber and frequently diverted the Germans' antiaircraft fire from the real target Our entire program on counter-

[1] Formerly, Metallurgical Advisory Board

measures continues much of the research aimed at jamming and destroying the usefulness of enemy radio signals. This will enable our planes to operate more freely under radar-directed antiaircraft fire.

I have mentioned the importance of electronic devices as components of aircraft; the effectiveness of guided missiles also depends to a large extent on electronic equipment. Our guided-missile development shows promise of providing weapons which can be launched from the ground against enemy aircraft or missiles and others which can be launched from our own aircraft against various surface targets. Most of these missiles depend upon electronic systems of guidance and are extremely complicated. An important aspect of this program is devoted to preserving flexibility of design so that various type warheads, including atomic warheads, can be carried by these missiles.

Peacetime Applications

And now, lest I leave the impression that the research and development program is a kind of military Juggernaut under whose wheels peacetime projects will be ruthlessly sacrificed, I should like to emphasize that not all of the money spent for this purpose goes up in the smoke of battle. Military research, no matter how deadly its intent, nearly always produces something with a humanitarian application. Many would argue, for instance, that the medical and biological aspects of atomic-energy research have already canceled out the havoc wrought at Hiroshima and Nagasaki.

Studies of the action of nitrogen mustards, developed as a toxic chemical agent, reveal their remarkable destructive effect upon white blood cells. These findings suggest the use of the mustards in the treatment of leukemia, and have led to the initiation of studies that have spread to more than 120 clinics, with hundreds of cases under treatment. More recent experiments have indicated that these mustards may also prove effective as a remedy for rheumatoid arthritis and for asthma.

In the field of electronics, many of the manufacturing techniques now being used in the manufacture of picture tubes for television were born of World War II research on cathode-ray tubes used for radar.

No Easy Answers

With a program of research and development as large as I have described, one might assume that future wars will be fought at great distances where human beings seldom come in contact with each other. We must be careful, however, not to convince ourselves that science will provide an easy answer to all our problems. New and fantastic weapons may be of some assistance but we must guard against trusting our future entirely to such devices. Great strides have been made and will continue, but I want to stress the fact that no new weapon which we now know anything about will eliminate the need for powerful, well-trained armed forces in the Army, Navy, and Air Force. We will need just as much hard work and unselfish devotion on the part of each individual as we have ever had in the past if our nation and liberties are to survive.

We believe that the RDB has contributed a great deal to the integrating of these programs which I have described, and to an over-all program of military research and development. If the task of co-ordination is not yet all that we should like it to be, at least we can feel that we have helped to bring about a fruitful cross exchange of information. Not only are the services aware of what each other is doing, but through our civilian Committee and Panel members, they are aware of relevant work which is being done by industry and universities.

Our primary aim is to assure our fighting men the superior weapons necessary to overcome the numerical superiority of the Communist forces. And as Secretary Lovett recently said, "We don't have to match the Communists man for man; we aren't going to dance with them."

Measuring the Mean Power of Varying-Amplitude Complex Audio Waves*

HUNTINGTON W. CURTIS†, MEMBER, IRE

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

Summary—**A new instrument which is capable of determining and recording, over successive short time intervals, the mean-power level of complex audio waves of rapidly varying amplitude is presented. While designed primarily for research work with speech sounds, where variations of as much as 35 db in level within 1/100 of a second are not uncommon, the unit could be applied to the power-level analysis of any voltage wave varying with components fully within the audio range. The apparatus provides a representation of power level in decibels for the brief phonetic elements of speech or for other similar elements of widely varying amplitude, without confusion of data or ambiguity of interpretation.**

Introduction

In audio-frequency research, where rates of signal-level change up to 3,000 db per second may be encountered, experimenters have needed instrumentation capable of presenting this rapidly varying level accurately enough to permit analysis of power level "fine structure." Values of mean power for short time intervals would be highly desirable for the analysis of individual phonetic elements in speech transmission studies where audibility and comprehension are being investigated.

Specifically, an instrument for such mean-power studies should meet four requirements: (1) For ac inputs of fixed level the unit output should be constant, showing no variation during individual input cycles. (2) Its output should be independent of input signal wave shape and should read nothing but average power for all complex waves having fixed amplitude components, regardless of phase relation between the components. (3) The value read for power level should follow closely abrupt variations in the signal so that at any instant during intervals of signal change the curve obtained by joining successive "per-input-cycle" values of mean power will be closely represented by the output of the instrument. (Note that transient impulses such as sibilants, lacking a fundamental frequency, must also be represented by some kind of sampling technique). (4) It should present its output in decibels to conveniently represent the wide

* Decimal classification: R245. Original manuscript received by the Institute, March 26, 1951; revised manuscript received, February 18, 1952.

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range of power levels to be analyzed. If these requirements are met, the data recorded will represent accurately even such extreme cases of power-level variation as those in speech where a period of large amplitude may be immediately preceded or followed by a period during which the wave has an amplitude close to zero.

The methods, and therefore equipment, previously applied to the study of rapid variations in sound power level have been of two general types: (a) graphical and mathematical analysis of the instantaneous sound-pressure plot obtained from a recording cathode-ray or string oscillograph, and (b) observation of the "power-level" values obtained from units which attempt to give as an output an accurate, continuous mean-power curve. Method (a) is laborious, in that from the instantaneous amplitude curve an instantaneous power curve must be derived by plotting the squares of the instantaneous amplitude values as a function of time, and then graphically averaging the resultant amplitude-squared wave over successive desired units of time—per cycle, or per syllable, or over an arbitrary unit of time—thus obtaining for the chosen average intervals a set of discrete mean-power values which could then be plotted and joined by a smooth curve to form the desired "mean-power" curve. The original oscillogram of instantaneous amplitude contains all of the desired power data, subject to limitations in the accuracy of the recording technique employed; but it is evident that the preparation of desired power-level data from the oscillogram is tedious and inconvenient. The general method (b) results from the understandable desire to let some type of instrument perform the indicated squaring-averaging functions, or perhaps in some other manner to derive a time-varying mean-power wave from an input time-varying amplitude wave.

To illustrate the problem, Fig. 1 is an oscillogram of instantaneous sound pressure as a function of time for the sample word *recording*. Examination will show that the wave amplitude during the (*e*) and (*o*) vowel sounds is twenty to thirty times greater than that of the intervening consonant (*c*), and that the change in level is almost instantaneous. With this word *recording* as input, the output variation of two power-level-computing instruments is sketched in Fig. 2. The time scale is identical with that of Fig. 1.

In (*a*) the curve approximates the successive pointer positions which would be obtained with a vu meter: This is shown merely to illustrate an extreme case of the masking of short-duration, low-level intervals which are immediately preceded by periods at a much higher level. The vu meter values, of course, could not apply to the mean-power, "fine-structure" problem because of the slow response (long time constant) inherent in the vu meter design.

An improvement is immediately apparent in (*b*), where an output typical of a graphic level recorder is sketched. This type of instrument, when used with an averaging circuit having a time constant much shorter than that of the vu meter, can begin to follow some of the rapid mean-power variations; however, some of the information of brief low-power portions of the input wave is still masked by the carry-over of preceding,

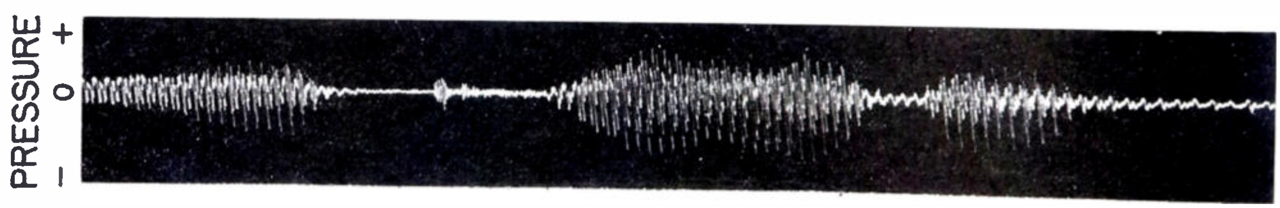

Fig. 1—Instantaneous pressure oscillogram, word *recording*.

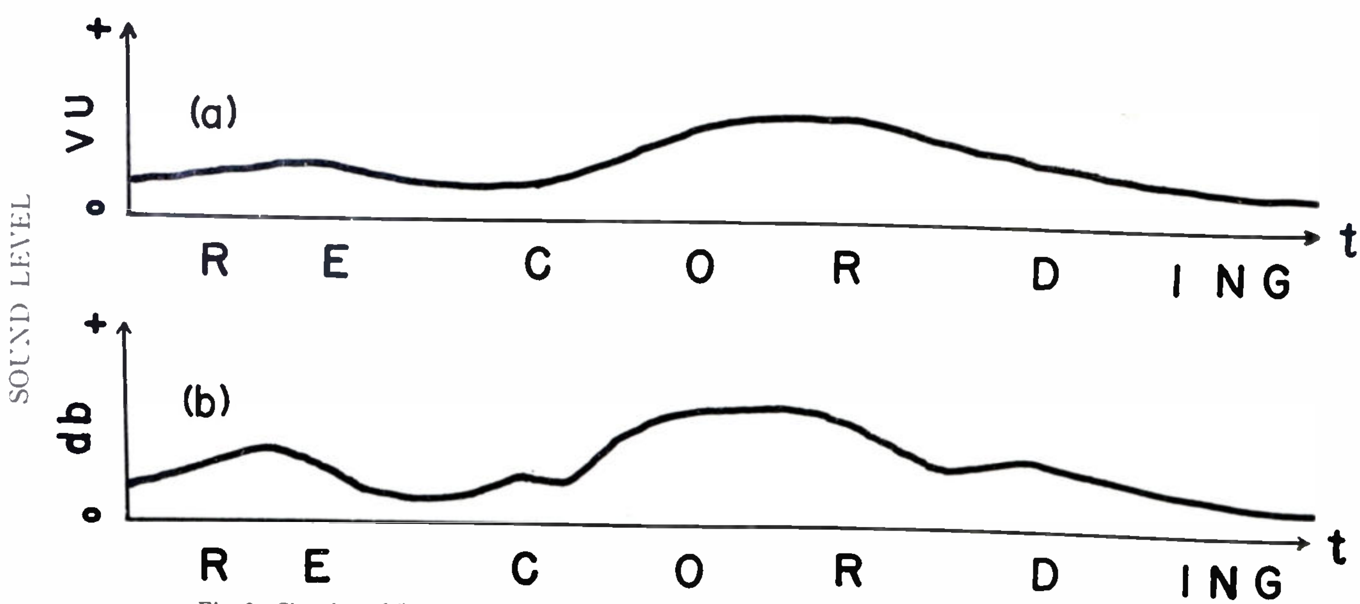

Fig. 2—Sketches of "mean-power" indicating system response with the wave of Fig. 1 as input.

much higher power periods. Attempts to improve on the response of power-level recorders by further shortening the time constant, thus increasing the maximum rate of response, introduce, however, the additional defect that while finer variations in mean power may be followed, the instantaneous power variations within each fundamental cycle begin to be followed also. The result is that steady-state input waves will show misleading mean-power variations at twice the fundamental frequency.

Basis of New Approach

As was suggested earlier, a curve formed by joining the values of average power of each input cycle might well be used to represent the signal level of a wave of rapidly changing amplitude. Rather than attempting the construction of a device to perform this function, it was decided that a more practical approach would be to divide the input signal into short equal time intervals—intervals only slightly longer than the period of the lowest generally occurring frequency to be analyzed (which is on the order of 130 or 140 cps for speech)—and to present as an output the successive values of mean power over each of these intervals. As a design base the time of each interval was set at 1/100 of a second.

Before further considering the actual instrument design, a brief development follows showing the mathematical expressions upon which its operation is based. The mean-power $\overline{P}$ over a ΔT second interval beginning at time t_0, for a wave $e(t)$ proportional to instantaneous sound pressure at a microphone, is given by

$$\overline{P}_{\Delta T} = \frac{K}{\Delta T}\int_{t_0}^{t_0+\Delta T} [e(t)]^2 dt, \qquad (1)$$

where K is a constant. This follows from the relation $\overline{P}_{\Delta T} = \Delta W/\Delta T$, which defines average power over an interval ΔT, ΔW being the energy summed during the interval, and also the fact that

$$\Delta W = \int_{t_0}^{t_0+\Delta T} p(t)dt,$$

where $p(t)$ is instantaneous power, itself proportional to $[e(t)]^2$.

Thus for an input signal the mean-power level during each of a succession of short intervals may be found by squaring the input wave and performing a separate integration of the squared wave for each interval. The result is a sequence of separate mean-power values spaced uniformly in time. In order to convert these values to decibels a circuit which will take the logarithm of each of the mean power values is required, as indicated by the following relation in which $\overline{N}_{db}$ is the mean power expressed in decibels above an arbitrary reference P_0, and k_1 and k_2 are constants.

$$\overline{N}_{db} = 10 \log \frac{\overline{P}_{\Delta T}}{P_0} = k_1 + k_2 \log \int_{t_0}^{t_0+\Delta T} [e(t)]^2 dt. \qquad (2)$$

It must be recognized that the values $\overline{P}_{\Delta T}$ and $\overline{N}_{db}$ in (1) and (2) are not continuous functions but separate values occurring at the end of each interval ΔT.

Block Diagram

The unit being described is a computer in which the mathematical operations of squaring, integrating, putting limits on the integral, and taking a logarithm are performed. Fig. 3 presents the relation of the functional units performing these operations.

Within each block of this figure an indication is given of the relation of electrical output to input for the block; in addition, equations are given between certain blocks to show the relation of the signal at that point to the unit input. The functional blocks of this figure will now be discussed separately.

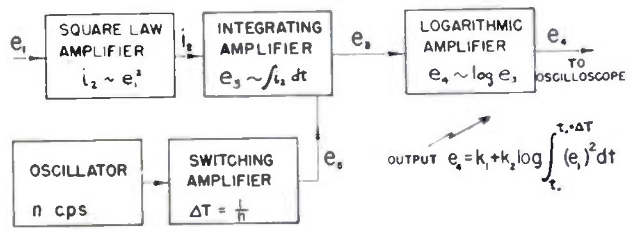

Fig. 3—Block diagram for "short-time mean-power" computer.

Square Law Amplifier.

Here the input signal undergoes essentially linear rectification, giving an output $|e_1|$ from the input e_1; then this $|e_1|$ is applied to a vacuum-tube circuit for which $i_2 = k_1 + k_2|e_1|^2$, k_1 and k_2 being constants.

The linear rectification is accomplished in a full-wave circuit using germanium diodes operating into a high-resistance load. This circuit was chosen to make use of a single-ended square-law circuit rather than a push-pull circuit (in which odd powers of the g_m expansion are cancelled). The necessity for obtaining matched tubes was thus avoided. The square-law relation was obtained with a selected 6AU6 pentode biased essentially at cut-off, its plate current closely exhibiting desired square-law characteristic over operating range of the tube.

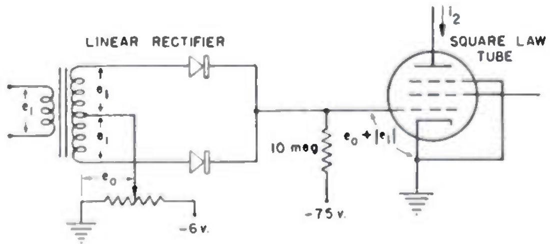

Fig. 4—Square-law amplifier schematic diagram.

Integrating Circuit.

The plate current of the preceding (square-law) circuit is used to alter the voltage at a capacitor. If the voltage e_3 at the beginning of a time interval ΔT has the value k, then at the end of the interval

$$e_3 = k + \frac{1}{C}\int_{t_0}^{t_0+\Delta T} i_2 dt.$$

At the end of the interval the voltage must be returned to the value to give identical initial conditions for each integration. (This is the function of the switching circuit.) Selection of a capacitor having extremely low dielectric absorption is important in this circuit; for the 1.0 microfarad unit employed here, a Teflon dielectric capacitor was needed to minimize the apparent drift of k which followed a high power-level interval.

Switching Amplifier.

In this circuit the value of the recurring interval ΔT is set at 0.01 second by a 100 cps oscillator. Ideally, the function of the switching amplifier is to return the

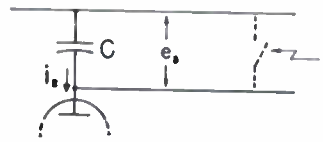

Fig. 5—Integrating capacitor in square-law amplifier plate circuit.

integrating capacitor voltage to a fixed reference level at the end of each 0.01-second interval; an instantaneous closure after each 0.01 second of the switch indicated in the dotted portion of Fig. 5 theoretically should accomplish this. Practically, however, this return of the integrating capacitor to a fixed level for a new start for each integration is accomplished by the arrangement of Fig. 6.

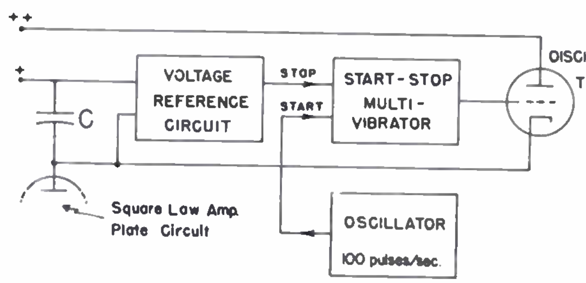

Fig. 6—Switching amplifier and integrating capacitor.

During the 0.01-second interval over which charge is integrated, the discharge tube is kept cut off by the start-stop multivibrator. At the end of each successive interval a pulse from the oscillator causes the multivibrator to change conduction state; this in turn drives the grid of the discharge tube strongly positive. The integrating condenser voltage is rapidly altered by the high current drawn from it, and this condenser voltage continues to change until it reaches a fixed value determined in the reference circuit. The reference unit compares the voltage existing across the condenser with a fixed standard; when the condenser has discharged to the voltage of the standard, the reference unit terminates the condenser discharge by sending a "stop" pulse to the start-stop multivibrator. Since the condenser cannot integrate the incoming instantaneous power wave during the discharge period, this period must be kept short to avoid error. For a switching frequency of 100 cps the total integration-discharge time is 10,000 microseconds; the circuit constants were chosen to keep the discharge time below 200 microseconds for all but the highest amplitude input signals, thus producing only a negligible error in the final db output.

Logarithmic Circuit.

A circuit with an output proportional at every instant to the logarithm of its input is needed to convert the peak values of integrating condenser voltage to an output amplitude proportional, at these peaks, to power level in decibels. The nonlinear unit chosen for the logarithmic element was a Varistor maintained at constant temperature. Over limited ranges of current through this unit, the voltage across the terminals is proportional to the logarithm of the current. By employing two compensating circuits to the Varistor, one for high and one for low currents, an output voltage may be obtained which closely approximates the logarithm of the input current for almost three decades of the input. The Varistor approximate characteristic is given by $i = k_1[\epsilon^{k_2(V-iR)}]$, where i and V are the Varistor current and terminal voltage and k_1, k_2, and R are constants of the unit. The compensating circuits in effect alter this relation to the desired form $i = k\epsilon^V$.

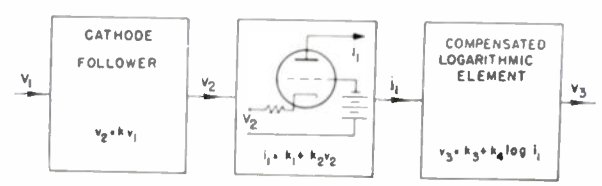

Fig. 7—Logarithmic amplifier block diagram.

In the logarithmic circuit diagrammed above the first block cathode follower is used mainly to provide a voltage from a low-impedance source to operate the following circuits. The second block might be termed a "cathode-driven cathode follower." Employing high-current feedback, it produces a plate current which varies in a linear manner with changes of input voltage. This circuit was needed to serve as a phase inverter since the voltage most conveniently obtained from the integrating capacitor is increasingly negative with increasing power level, whereas the logarithmic element requires increasing current for higher output voltage. The final output from this logarithmic circuit is fed to a recording oscillograph.

RESULTS

The operation of the system just described is illustrated by the oscillogram of Fig. 8. The word *record-*

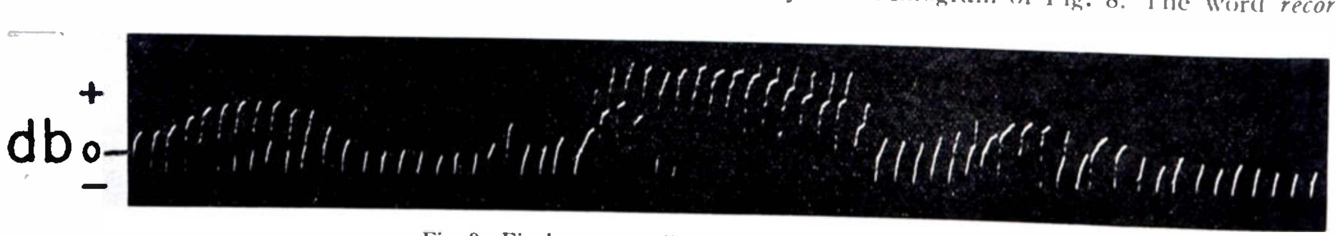

Fig. 8—Final output oscillogram, word *recording*.

ing was spoken as an input signal for comparison with Fig. 2. From the derivation of the equation for power level in decibels (2) it should be evident that only the amplitude of the separate points occurring at the end of each integrating interval (i.e., the peaks of the wave shown in Fig. 8)—indicate mean power for the entire interval. In analyzing the data contained in this figure for mean-power values, it is important to consider only these peaks, not the form of the wave during the integrating interval.

The "zero-decibel" level of this oscillogram is the level produced by the circuits with no input signal in the microphone and the amplifier. Note that there is a completely quiet interval (i.e., a return to this level) just before and after the *c* sound in *recording*; the time of these brief silent periods is easily counted in hundredths of a second; furthermore, the mean amplitude and the duration of the brief, low-intensity, consonant is readily observed.

Conclusion

The design of a new "short-time-mean-power" measuring instrument has been discussed. The unit output is best represented as a continuous strip recording of oscilloscope deflections; the expected accuracy of any single mean-power reading in decibels above "zero signal" as taken from this record is less than ±1 db for all outputs throughout a 25-db range. This accuracy could be further improved by redesign, mainly of the squaring circuit; at present, however, this accuracy seems sufficient for the use to which the data will be put.

Acknowledgments

This report is based on author's work at the University of Iowa. He wishes to acknowledge the encouragement and advice of E. M. Lonsdale, Dr. James F. Curtis, Dr. L. A. Ware, and T. A. Hunter of the university staff.

Video Recording, Film Considerations*

GEORGE H. GORDON†

This paper is published with the approval of the IRE Technical Committee on Video Techniques and its Subcommittee on Video Utilization.—*Editor*

Summary—**The basic method of specifying the performance of photographic materials is described. Some variables which affect the film performance and their relationship to video recording are discussed. Some common photographic terms used in video recording are defined.**

Introduction

VIDEO RECORDING represents the combination of television and motion-picture arts. Each alone has developed a certain vocabulary, control methods, and performance specifications. Inasmuch as each has the common aim of presenting a visual image, which in its simplest form should subjectively resemble the original subject, it is not surprising that certain properties and requirements should be common to both arts. However, the fundamental differences both in the arts themselves and in the technological backgrounds of the men responsible for the development of each art lead to certain differences in the methods of measurement and testing and in the manner of presenting the data so obtained. The blending of television and motion-picture photography into video recording requires a corresponding blending of definitions, control methods, and performance specifications. The purpose of this paper is to relate the methods and specifications of the photographic art so that over-all performance characteristics of the video-recording system may be determined.

* Decimal classification: 770.2836. Original manuscript received by the Institute, June 13, 1951; revised manuscript received, January 30, 1952.

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Film Performance Methods

The performance of photographic materials is almost universally described by methods which relate the exposure to which the film is subjected to the resultant developed image. This information is presented as a well-standardized form of graph called a sensitometric curve or *H* and *D* characteristic (after Hurter and Driffield, early experimenters in the field). Either explicitly stated or implied are all the factors affecting the result. The exposure data are conventionally plotted as the abscissa and the result of the photographic process, i.e., density, is plotted as the ordinate, as shown in Fig. 1, which

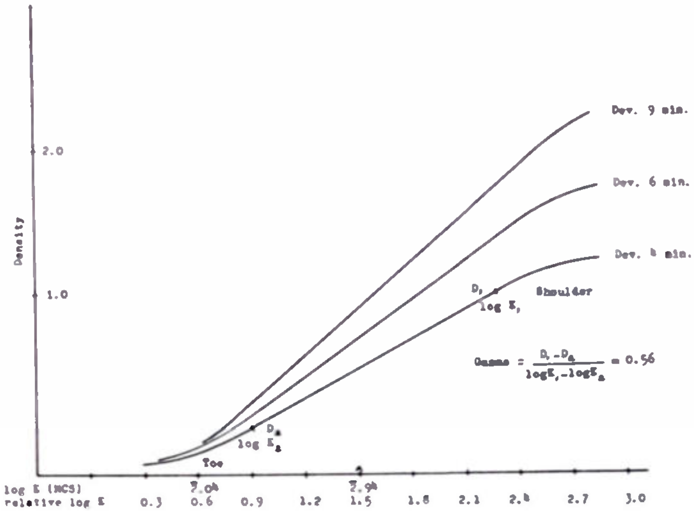

Fig. 1—Typical sensitometric curves.

represents a hypothetical photographic material that has received three times of development.

For convenience in handling the wide range of values which may be encountered in photography and because the eye responds logarithmically to brightnesses, it is customary to use the logarithm of the total energy of the exposure as the abscissa units. If the exposures have been made on a standardized instrument, the absolute value of the exposure intensity in units such as the meter-candle is known, the time of exposure is known, and the spectral energy distribution of the light source is known. Thus E or log E may have a definite value in meter-candle-seconds or the logarithm of this number. For processing control purposes there are some sensitometers which give reproducible exposures, but are not calibrated in absolute units. In such cases any convenient exposure step may be regarded as a reference point and the other exposures assigned values relative to it. Thus two exposures differing by a factor of 2 would be labeled as differing by 0.3 log E without attempting to evaluate the absolute level of either.

In order to represent the process in question the sensitometric exposures should approximate as nearly as possible the actual operating conditions. Thus, in conventional photography, negative materials are exposed to varying intensities of light of average daylight quality for a time comparable to usual camera exposures, for instance 1/50 second. For video recording, the film in question should be exposed by light of quality, intensity, and duration approximating actual exposure to a cathode-ray tube. In the absence of mechanical devices meeting these requirements the actual recording cathode-ray tube may be used. For processing control, sensitometric conditions widely different from operating conditions might be used, but such a procedure may lead to unjustified conclusions concerning the actual results obtained. In other words, the relationship between operating conditions and the sensitometric conditions may not be constant for all film products or processing conditions.

Measurement of Density

The amount of image produced is most conveniently expressed as density which is defined as the logarithm of the reciprocal of the transmission.[1]

If T = transmission = ratio of the transmitted light to the incident light

O = opacity = reciprocal of transmission,

then D = density = log O = log $1/T$.

Thus a material which transmits half the incident light has $T=0.5$, $O=2$, and $D=0.3$. This is convenient because dark images have a high value which is subjectively satisfying and the range of values is kept relatively small. Of more importance is the fact that the normal eye sees a scale of equal density increments as an approximately linear scale of brightness. Density units are in some respects analogous to db in mathematical properties. The use of density as ordinate and log E or relative log E as abscissa permits the use of arithmetic paper for plotting sensitometric curves. Density values may vary, depending on whether diffuse or specular light is used in the measurement. If the image is not neutral in color, or if the film has tinted support, density depends on the color of the light. Briefly, diffuse density is the value obtained if, with specular or parallel light rays incident on the film perpendicular to the film plane, all transmitted light is collected and measured. Specular density is the value obtained if, with specular or parallel light rays incident on the film perpendicular to the film plane, only specular transmitted light is collected and measured. Specular transmission has a lower value and specular density a correspondingly higher value than those obtained for diffuse light. ASA standard Z38.2.5—1946 describes in detail the measurement of photographic density.

Diffuse density which is generally applicable for contact printing has a lower value than specular density which is generally applicable for projection purposes. The ratio of specular to diffuse density is known as the Callier coefficient of the film, and is usually greater for coarse-grained than for fine-grained films. Most commercial densitometers in use in the photographic industry read very nearly diffuse density, and are sufficiently accurate and precise for routine usage even if they depart slightly from ASA specifications.

Values of the ratio of specular to diffuse density for various film materials as a function of density and processing condition are available. Usually the ratio is taken into account in specifying the sensitometric characteristics desired rather than in attempting to read different types of density or in directly applying a conversion factor.

Exposure

Some factors of the exposure which influence the image produced other than the gross amount of energy falling on the film are the time duration and intensity of the exposure, and the spectral energy distribution of the light.[2]

The spectral energy distribution of the light may affect both the shape and slope of the sensitometric curve. In general, with fixed processing, exposure to the shorter wavelengths of light produces a lower slope of the sensitometric curve than does exposure to longer wavelengths. Since adjustment of processing conditions can usually compensate for these differences, and the choice of phosphors is relatively limited, this factor does not appear to be of much concern in video recording.

The time-intensity relationships of the exposure, however, may be of considerable importance.[3] With few ex-

[1] C. E. K. Mees, "Theory of the Photographic Process," Macmillan Co., New York, N. Y., pt. IV; 1942.

[2] *Ibid.*, chap. V.

[3] *Ibid.*, chap. VI.

ceptions, in normal photographic usage the exposure required to produce a specified density is a constant equal to the product of the exposure intensity and time. Thus doubling or halving the intensity with a corresponding halving or doubling of the time produces the same resulting effective exposure. This has led to the formulation of the reciprocity law which can be expressed as $E = It$, where E is effective exposure, I is intensity, and t is time. This relation holds over a limited

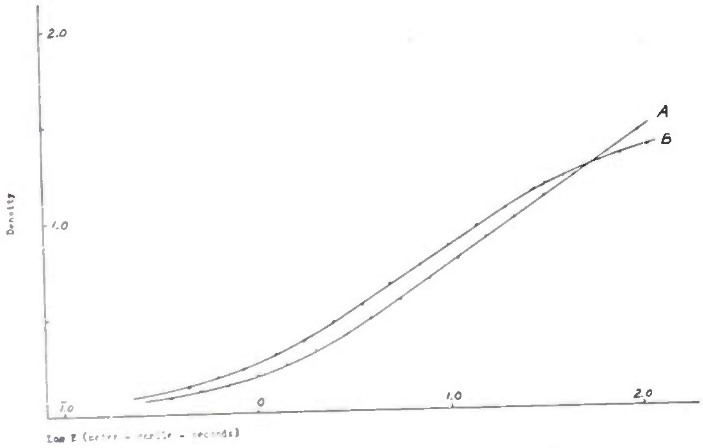

Fig. 2—Reciprocity failure of positive film in negative developer. Equal total exposure. A—constant time (0.02 sec.) variable intensity B—constant intensity variable time (4.16 sec. max.)

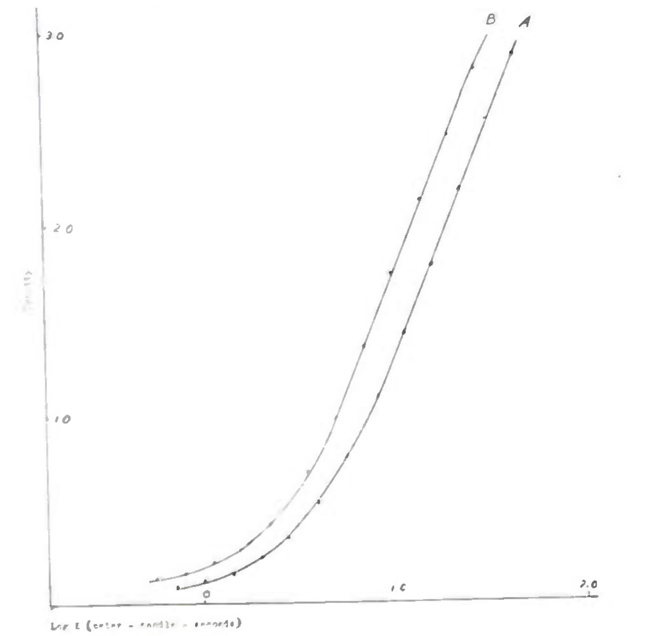

Fig. 3—Reciprocity failure of positive film in positive developer. Equal total exposure. A—constant time (0.02 sec.) variable intensity B—constant intensity variable time (4.16 sec. max.)

range, but at extremes of high and low intensity and correspondingly short and long times the product It usually must be increased to maintain constant density of the image. The extent of failure of the reciprocity law is also a function of the development process. These effects are illustrated in Figs. 2 and 3 for a commonly used fine-grain release positive type film. Departure from the reciprocity law is most serious for this film in the high-density region when processed with the negative-type developer.

An effort is made to manufacture most photographic products so that the range of intensities and times over which the reciprocity law is valid covers the normal usage of the film. Unfortunately, the intensities and times encountered in video recording are far beyond the extremes usually encountered in photographic practice, and there is little valid information of a basic nature regarding film behavior under these conditions. In normal photographic practice the time of exposure used in sensitometric investigations is rarely shorter than one one-thousandth of a second with intensities such that normal acceptable film densities are obtained. In video-recording practice the time of exposure is probably on the order of a few microseconds, depending upon the decay characteristics and the degree of excitation of the phosphor. So, reliance must be placed on empirical data pertaining to the conditions of use and the possible influence of reciprocity law failure must be borne in mind.

Processing

The processing which the film receives has considerable effect on the resultant image. Not only the slope of the sensitometric curve, but its shape, effective film speed, graininess, and color of image are functions of the development process. The minimum specification for processing is the time and temperature of treatment and chemical composition of the developing solution. Within rather wide limits, these three minimum factors may be varied to produce variations in the sensitometric curve or to maintain a specified standard. The term "negative" or "positive" developer refers only to a general pattern of chemical composition which has been found to be convenient for producing negative images suitable for original film exposed in a camera, or positive images on prints from such negatives (see Figs. 2 and 3).

Interpretation of Data

If all the above factors affecting exposure, processing, and measurement of density produced have been selected to correspond with the operation under consideration, the resulting sensitometric curves are a quite complete specification of film performance. The curve commonly has curved ends and a straight portion between. The concave section associated with the region of low exposure is called the "toe" and the convex section associated with the region of high exposure is called the "shoulder." In normal photographic practice, and video recording seems to be no exception, the toe of the characteristic plays an important part in the reproduction of tone values. For processing-control purposes, the slope of the straight-line portion of the sensitometric curve, which is called "gamma," is a convenient characteristic of the curve. The value of gamma should not be regarded as an all-embracing description of a photographic process. The shape of the toe, the densities at which the curve departs from linearity, and the effective speed or log E required to produce a significant density should all be considered. If consideration is directed at a given film material, with comparable exposures and processes, constancy of gamma does correlate well with the uniformity of the resulting product.

The term gamma refers to the slope of a plot of two logarithmic functions, density and log E. Thus it is analogous to the statement that an electronic circuit or component device has a power law output. For instance, most cathode-ray tubes have output brightness which is some power function of the grid voltage. If such a tube followed a 2.5 power law, a plot of log brightness versus log grid voltage with proper operating potentials would be a straight line with a slope of 2.5. In photography, gamma and contrast are often improperly regarded as synonymous. Actually, they are separate concepts which have some interaction. In television practice, gain controls are often referred to as "contrast controls." From the similarity of names and lack of distinction between gamma and contrast, there occasionally arises the erroneous impression that photographic gamma is equivalent to electronic gain. This is not the case.

Cascading photographic processes is similar in many respects to cascading electronic units. Thus there is a loss of definition (bandwidth) in a printing process and the graininess (noise) usually comes from the early stages. The complete transfer characteristic can only be calculated by a specific plot of the characteristics involved because nonlinear portions of each film characteristic are almost inevitably used. Such a cross plot may be constructed by using the density differences in the original film as log E differences in the exposure of the print stock and the print-film characteristic, thus determining the resultant density. Exposure level of the print may be taken into consideration by assuming that a certain portion of the original film should be printed to a certain density and by adjusting the coordinates of the plots in such a way that this specific point is so plotted. For instance, a scene highlight may be recorded as a density in the negative of 1.10, and this may be printed to a density in the positive of 0.3. Other objects of lesser brightness may be recorded as 0.9 density in the negative and these would be reproduced at the density called for by the print characteristic corresponding to a log exposure 0.2 greater than that required to produce 0.3 density.

The deliberate use of curved-film characteristics is often of apparent advantage, but is a difficult operation calling for extreme care in adjusting exposure. Otherwise the desired advantage from the distorted characteristic may not only be lost but actually result in a very poor product indeed. Chiefly for this reason, the so-called toe recording of sound, that is, using the curved toe of the film characteristic, has rather erratic results. Also, processing of film for consistent characteristics is more difficult if curved portion is of primary interest.

Film Selection

For the purpose of video recording, the choice of film materials should be guided by certain fairly simple criteria. The graininess of the film should be as low as possible and the resolving power as high as possible. These two factors have a general correlation, but are not necessarily equivalent terms. The low graininess roughly corresponds to a low electronic noise level and high resolving power roughly corresponds to a wide frequency response in electronics. The film chosen should have a sensitometric curve under video-recording conditions which, with commercial processing conditions, produces a gamma approximating the desired gamma and which also has a long straight-line portion with a rather sharp toe and does not develop a shoulder until rather high densities are reached. It is conceivable, of course, that a curved characteristic may be desired for special problems concerning cancellation of some reproduction errors in the video image, but so far this has been of only theoretical interest.

The color sensitivity of the film at the present time is of little concern as long as the film has adequate sensitivity to the radiation from the cathode-ray-tube phosphor. The image has already been reduced to monochrome by the television system and panchromatization has no significance in terms of image tone reproduction. It might be borne in mind that panchromatization usually extends the spectral sensitivity of the film, but does not increase its inherent sensitivity. Thus, with a blue phosphor such as P11, a blue-sensitive, low-speed release positive or sound recording film may be as good as high-speed negative film with respect to sensitivity. The turbidity of the emulsion and halation properties of the film may also influence the choice of materials by their effect on image spread and the density gradient across the boundary between a dense and clear area.

Conclusion

The performance of the photographic or film stage in a system is specified by sensitometric data which relate the brightnesses to which the film is exposed to the image produced by the processing of the film. For precise work all of the factors of exposure and processing must be considered. However, in regular photographic practice as well as video recording the usual aim is to establish certain desirable operating conditions and maintain them relatively constant. This means that in spite of the complex photochemical mechanism of the actual exposure and complicated chemical reactions of processing, fairly simple control methods may be utilized once operating conditions have been established.

A Multichannel Single-Sideband Radio Transmitter*

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Summary—This paper describes a new single-sideband radio transmitter for transoceanic service which represents a substantial improvement over past design. Its important features include: (a) a frequency band which permits deriving four telephone channels, if desired; (b) a push-button method for changing frequencies within a matter of seconds; (c) an increase in power over its predecessor; and (d) all-around improved transmission performance.

Introduction

In transoceanic radiotelephone service, extensive use is made of single-sideband transmission to permit a more efficient use of the frequency spectrum and to obtain an improvement in the received signal over that of a double-sideband system.[1,2] Development has also progressed to the point where multichannel systems are now being employed to utilize the transmitter equipment more fully.[3] An important factor in this type of service is the necessity for changing operating frequencies to provide maximum reliability. These frequencies may be changed as many as four or five times a day. One of the chief aims in the development of the new transmitter which is described below has been to reduce the length of time required for setting the different frequencies. This, of necessity, requires some rapid method which will yield an adjusting accuracy comparable to that obtained by the usual manual procedure. The method chosen is that of employing a servo system in which motors are used to switch coil taps and adjust condensers to pretuned positions.[4] Apparatus is provided which will store the tuning information so that by simply operating one of a given number of keys the proper settings of the coils and condensers will automatically be made. In this way the normal time for changing a transmitter frequency is cut down to a matter of seconds. Other improvements, brought about from experience and knowledge gained from previous designs, are included to facilitate operation and maintenance.

General Performance

The transmitter provides for operation in the range 4 to 23 mc. It will accept two independent frequency bands each from 100 to 6,000 cycles. These bands appear in the radio-frequency output signal as upper and lower sidebands of a carrier whose amplitude is normally reduced 20 db below reference amplitude. In turn, each of these two sidebands may be divided into two 250–2,750 cycle voice bands, making available a total of four voice channels. Obviously, each of these two sidebands may be divided in other ways, such as for voice-frequency telegraph.

This transmitter will operate satisfactorily in areas having ambient temperatures between 15 and 50 degrees C and at altitudes up to 5,000 feet. Its automatic frequency-changing feature by push-button control at the transmitter permits operation on any one of ten preselected frequencies. This control may be extended to a remote location by appropriate dialing methods. The time required to change from one frequency to another is about 15 seconds.

The required input is 10 kva 3-phase, 230 volt, 50–60 cycle ac. A separate single-phase, 115-volt, 50–60-cycle ac branch circuit of 1-kva capacity is required for illumination and convenience outlets within the transmitter.

The peak envelope power output of this transmitter is 4 kw when operating into a balanced resistive load over the range 190 to 900 ohms. This means that the transmitter adjustments will accommodate the impedance encountered when connected to balanced open-wire transmission lines having a characteristic impedance of 600 or 300 ohms where the standing-wave ratio does not exceed 1.5. It also means that the transmitter adjustments will accommodate two 100-ohm coaxial lines connected as a balanced circuit where the standing-wave ratio does not exceed 1.05.

From past experience with transmitters of this type it has been determined that the distortion generated by two equal tones which load the transmitter to rated capacity should be 25 db or more below either one of the two tones. It has also been determined that the signal-to-noise ratio should be at least 45 db and preferably 50 db. From actual measurements made on a model of this transmitter it has been found that both the distortion and the signal-to-noise ratio are about 10 db better than the requirements stated above. Measurements of adjacent band radiation on this model indicate that an appreciable improvement has been obtained as compared to an earlier transmitter of similar type.[3] Harmonic radiation has been found to be less than 200 mw.

Circuit Description

The schematic, Fig. 1, shows the principal circuit blocks as well as the two types of servos used. On this diagram the two voice-frequency inputs are shown at the upper left as group A and group B, with provision for switching group A to double-sideband operation if desired. Each block or unit is identified with the particu-

* Decimal classification: R423.5. Original manuscript received by the Institute, April 5, 1951; revised manuscript received, March 3, 1952.

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

[1] A. A. Oswald and J. C. Schelleng, "Power amplifiers in transatlantic radio telephony," Proc. I.R.E., vol. 13, pp. 313–362; June 1925.

[2] F. A. Polkinghorn and N. F. Schlaack, "A single sideband short-wave system for transatlantic telephony," Proc. I.R.E., vol. 23, pp. 701–718; July, 1935.

[3] A. A. Oswald, "A short-wave single-sideband radiotelephone system," Proc. I.R.E., vol. 26, pp. 1431–1454; December, 1938.

[4] J. C. Lozier, "Carrier-controlled relay servos," *Elec. Eng.*, vol. 69, pp. 1052–1056; December, 1950.

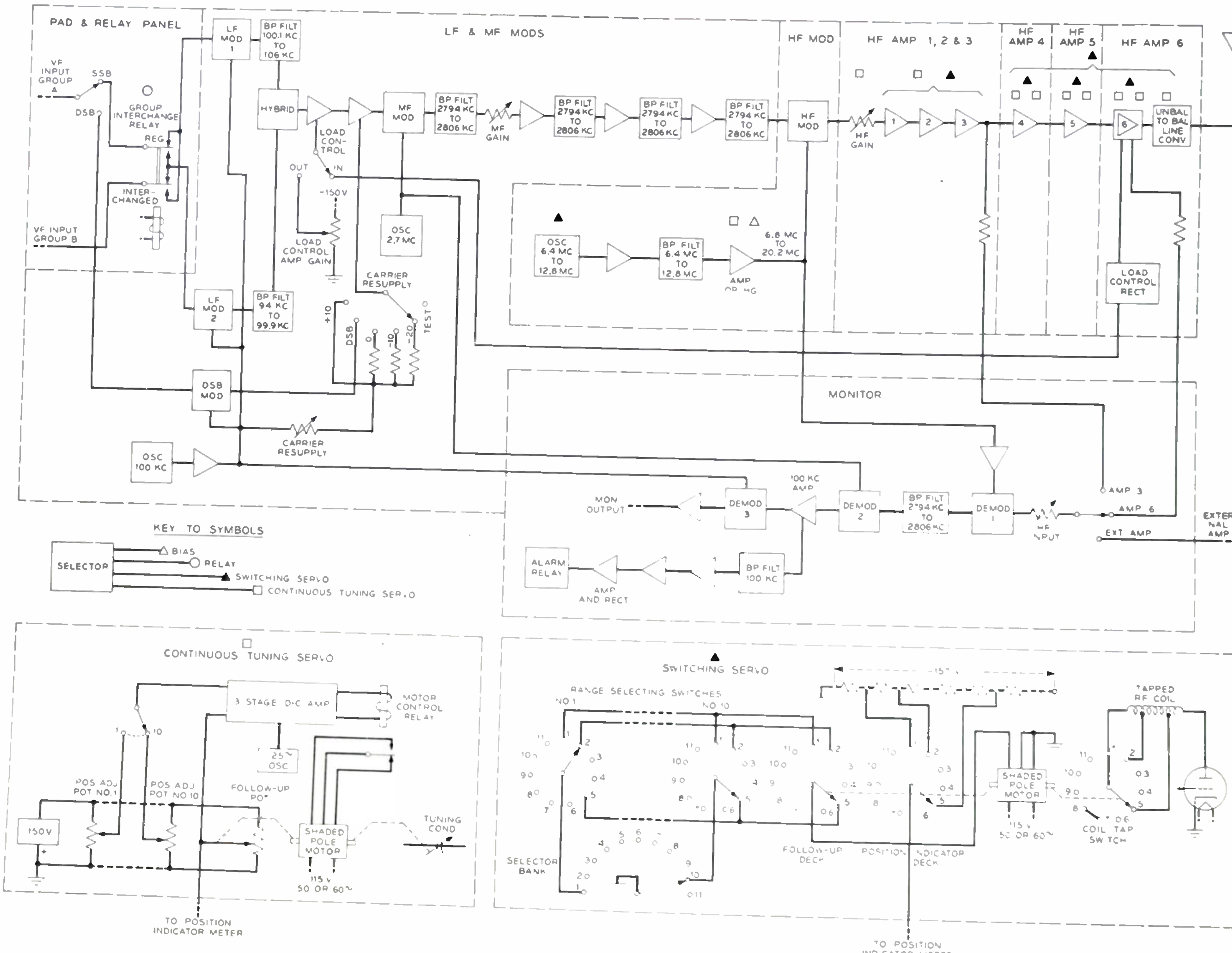

Fig. 1—Schematic showing principal circuit blocks.

lar function so that the message paths may be traced through from the input at the left of the diagram to the antenna at the upper right.

Starting with the pad and relay panel, the vf input in group A encounters first a switch (ssb-dsb) which connects it either to low-frequency (lf) modulator 1 or to the double sideband modulator (dsb mod). Both vf inputs group A and group B are connected through the contacts of a group-interchange relay. Besides the relays indicated, this unit contains input pads to each of the lf modulators and the dsb modulator which provide individual gain adjustments in the separate branches to compensate for differences in component losses. Other relays are provided for the application of test tone. The group-interchange relay serves to interchange group A and group B inputs to the lf modulators so as to abide by an operating convention. The reason for the convention is that the frequency range of the final conversion-frequency oscillator in both the transmitter and the receiver can be reduced a considerable amount if the lower sideband from the final modulator is utilized at the lower frequencies and the upper sideband is used at the higher frequencies. A change from one to the other obviously has to be made at both the transmitter and the receiver at some point in the frequency range; this has to be followed rigorously, otherwise the groups will be interchanged at the receiver output terminals and in the connecting wire lines. The operating convention which has been widely accepted requires the change-over at 10 mc. In this new transmitter, however, in order to eliminate some undesirable modulation products, it is essential that the actual change-over be made at 18 mc. Accordingly, in order to make the transmitted signal conform to the convention, the channel inputs to the lf modulators are interchanged between 10 and 18 mc.

When the switch and relay are in the positions shown in the block diagram, group A audio signals modulate the 100-kc carrier supplied to the copper-oxide varistor of lf mod 1. The same process occurs in an identical varistor with group B in lf mod 2. The 100-kc carrier employed in these modulators and in the double-sideband modulators which also uses an identical varistor is obtained from a low-temperature coefficient quartz oscillator which is followed by an amplifier. No temperature control is applied to this oscillator because the frequency change over the operating temperature range is negligible. The sideband outputs are filtered by sharply selective band-pass quartz-crystal filters. From lf mod 1 the upper sideband is taken by a filter which passes from 100.1 to 106 kc and from lf mod 2 the lower sideband is taken by a filter which passes from 99.9 to 94 kc. Because of the carrier balance in the lf mods and the attenuation in the crystal filters, the carrier is essentially suppressed at these points. The outputs of these filters are combined in a hybrid coil and impressed on the load-control amplifier.

The load-control amplifier is a variable gain device which can be used with its gain controlled by the magnitude of the combined sideband peaks or with its gain fixed. When the switch designated "load control" is thrown to the position "in," the amplifier bias is obtained from the load-control rectifier which samples the transmitter output signal and the bias varies with it. When the switch is thrown to "out," the bias is fixed. With the switch in the "in" position the gain of the amplifier is about 8 db greater than when it is in the "out" position. The gain remains at this higher value for all signal amplitudes up to those corresponding to almost the peak capacity of the transmitter. As the amplitude nears the peak capacity the load-control rectifier begins to increase the bias and reduces the gain at such a rate that the transmitter is never loaded beyond its peak capacity. The gain recovery rate is much slower than the reduction rate. This feature is of considerable advantage from the standpoint of cross-modulation and adjacent-band radiation. Its effect on quality and intelligibility has been found to be negligible by alternate AB tests.

The signal is then further amplified and combined with the desired amounts of carrier. Attenuators and a carrier resupply switch are provided so that the carrier may be inserted at +10, 0, −10, and −20 db from reference value. It is at this point that the double-sideband modulator is switched into the circuit and appears on one of the contacts on the carrier resupply switch. In this way the dsb signal is fed to the combining amplifier in place of the carrier resupply. It should be noted that there are no crystal filters in the output of the dsb mod so that the transmission band is not limited at this point. It should also be noted that the carrier is introduced into the signal after the load-control amplifier so that the action of the latter has no direct effect on the reduced carrier.

The output of the combining amplifier is next fed to the medium-frequency modulator (mf mod). This uses germanium varistors and is supplied with a 2,700-kc conversion frequency which is obtained from a temperature-controlled, quartz-crystal oscillator. The output of this modulator comprises two sidebands, each of which contains group A, group B, and a carrier. Unless the group-interchange relay is operated, the upper sideband contains the carrier at 2,800 kc with group A above and group B below it. This sideband is selected and the conversion frequency and the other sideband centered at 2,600 kc are eliminated by a series of band-pass filters which are placed in tandem with three medium-frequency amplifiers. A gain adjustment is provided at the input of this string of amplifiers. All of the equipment up to this point in the circuit remains fixed regardless of the final operating frequency.

The signal centered at 2,800 kc and the final conversion frequency are then impressed on the high-frequency modulator (hf mod), which also uses germanium varistors. The conversion frequency in the range 6.8 to 20.2 mc is obtained from a temperature-controlled, quartz-crystal oscillator followed by an amplifier and a harmonic generator. The transmitted carrier frequency is determined by the final conversion frequency supplied to the hf mod which must be 2,800 kc above or below it, depending upon whether the lower or upper sideband from the hf mod is chosen for transmission. Positions for ten quartz crystals in the range from 6.4 to 12.8 mc are arranged with a switch so that the transmitter may be tuned up on ten different frequencies. This switch is positioned by a servo switching control. A wide-band filter is located in the output of the amplifier following the crystal oscillator to avoid tuning at this point in the circuit. The second stage is operated as an amplifier for conversion frequencies below 12.8 mc and as a harmonic generator above this value. The output tuning of this stage is controlled by a continuous tuning servo. Again, as explained for the mf mod, the output of this hf mod contains duplicate signals as upper and lower sidebands as well as the final conversion frequency. The selection of the signal to be transmitted depends entirely upon the tuned circuits of the succeeding amplifiers.

The hf mod output is at low impedance suitable for connection, by means of coaxial cable, to the hf amplifiers 1, 2, and 3. The coaxial cable is terminated by a hf gain-control potentiometer which is positioned by a continuous tuning servo. The amplifier unit consists of three cascaded stages operated class A. The plate-tuning condensers are ganged and are positioned by a continuous tuning servo. The plate coils are tapped and the tap switches are ganged and positioned by a switching servo. A single 25-watt pentode is used in each stage.

The signal is next fed to hf amplifier 4 which consists of two 75-watt beam-power pentodes connected in parallel and operated class A. The output circuit is a pi-type network which performs an impedance transformation. It requires two independent variable condensers, each of which is controlled by a continuous tuning servo and a tapped coil controlled by a switching servo. Deposited carbon resistors are used at both ends of this network for circuit loading, the major portion of the power being dissipated by the second resistor which is at the grids of hf amplifier 5.

This amplifier has two 400-watt tetrodes connected in parallel and operated class AB. The output circuit and its controls are basically the same as those in hf amplifier 4. No resistance loading is required for normal operation because the following stage is grounded grid.

HF amplifier 6 employs a single 2.5-kw forced-air-cooled triode operating class B with its grid grounded. The output circuit is in two parts. First, a pi-network similar to those in hf amplifiers 4 and 5 performs an impedance transformation. Next, a tuned auto transformer is used to match the unbalanced pi-network to a balanced transmission line. Three continuous tuning servos and one switching servo are necessary. Through a transfer switch the auto transformer may be connected directly to an open-wire line leading to the antenna, to the terminals of a pair of coaxial lines which may be installed to drive an external amplifier, or to a dummy load mounted within the transmitter.

The monitor unit is a simple form of single-sideband receiver consisting of three demodulators which utilize the conversion frequencies and carrier source in the transmitter to recover the original voice frequencies. Its principal purpose is to serve as a means for checking the transmitter performance in respect to distortion and nonlinearity by the transmission of signals applied in one voice-frequency group. Accordingly, no group filters are provided in the output to separate the groups. It makes possible an alarm circuit which functions if the amplitude of the carrier falls below a predetermined value. This is accomplished by routing a sample of the demodulated carrier at 100 kc through a narrow-band crystal filter after which it is amplified and rectified to operate a relay. Provision is made for connecting the monitor to the output of hf amplifier 3 or to an external amplifier, if one is being excited by the transmitter, instead of the normal operating connection to hf amplifier 6.

There are ten continuous tuning and six switching servos. Local control is governed by a mechanically interlocked switch having one push-button for each of the ten operating frequencies and an eleventh button for establishing a standby condition. The operation of any one of the eleven push-buttons first causes a stepping-type selector switch to function. The selector together with a system of relays, called the servo director circuit, removes the high voltage from the last three amplifier stages, de-energizes the carrier alarm, energizes all servos for about 15 seconds, and at the end of the cycle reapplies the high voltage and restores the carrier alarm.

In the lower left-hand corner of the block diagram is shown a typical continuous tuning servo schematic which is used wherever it is necessary to adjust a shaft to any position within 270°. The shaft of a tuning condenser is shown mechanically connected to a shaded pole motor which, in turn, is mechanically connected to a so-called follow-up potentiometer through an anti-backlash gear to maintain positive angular relationship. The main winding of the motor is connected to 115 volts ac. The two shaded pole windings are brought out by three wires, one of which is common. If one winding is shorted, the motor runs in one direction. If the other winding is shorted, the motor runs in the opposite direction. If both windings are shorted or both are open, no action ensues. The three wires from the shaded poles are connected to the terminals of a two-position relay as shown. A position-adjusting potentiometer is required for each of the ten operating frequencies. During a frequency-changing operation one of these potentiometers is connected in parallel with the follow-up potentiometer across a regulated power supply. Relay contacts shown in the diagram as a switch connect the arm of the positioning potentiometer to one of the input terminals of a

3-stage dc amplifier, and the arm of the follow-up potentiometer is always connected to the other input terminal. The amplifier plate current flows through the winding of the two-position relay. If the setting of the positioning potentiometer is not the same as that of the follow-up potentiometer, when power is applied, a dc error voltage will be impressed on the amplifier of such polarity that the relay will hold in one position and the motor will run in the direction to reduce the error voltage. Since the relay is a two-position device as the error voltage approaches a small value, the system would start to hunt or oscillate if no other voltage were applied to the amplifier. The hunting frequency depends on the mechanical and electrical time constants involved, and is around 4 cycles per second. To overcome this tendency a small 25-cycle voltage is applied to the amplifier so that for small error values the 25-cycle voltage causes the relay to oscillate at that frequency but to remain on one contact longer than on the other. When the potentiometer settings are the same, the relay still oscillates at 25 cycles; but since the relay then stays on each contact the same length of time and the frequency is above the natural period, the shaft does not turn. The 25-cycle voltage thus has the desirable effect of making the torque developed by the motor proportional to the difference in settings of the potentiometers, for small differences. Since there are 10 positioning potentiometers for each continuous tuning control and there are 10 continuous tuning controls, there are a total of 100 positioning potentiometers on the front of the transmitter to store the tuning information. In addition there are 10 potentiometers used to provide different bias potentials to the harmonic generator and these are switched directly by a bank of contacts on the selector.

In the lower right-hand corner of the block diagram is shown a typical switching servo which is used to move a switch to one of several fixed positions. The example shown has eleven positions and is used to control a tapped inductance. The switch shaft is driven by a shaded pole motor identical with those previously described. The motor also drives the shaft of a two-deck switch which serves a follow-up function and provides a way to read the switch position on a meter. The contacts of the follow-up deck are connected to similar contacts on ten range-selecting switches which are set manually to any desired position and are located on the front of the transmitter near the 110 potentiometers. One of these range switches is used for each of the ten frequencies to control the position of all inductance switches. The arm of each of the range switches is connected to a contact in one bank on the selector. The arm on this bank of the selector is connected to ground. One shaded-pole winding of the motor is shorted and connected to the ground so that the motor will always turn in one direction when power is applied unless the other winding is also connected to ground. If, for example, the inductance switch is on contact No. 2 and range switch No. 10 is set at position No. 5, when the No. 10 button is pushed, the selector arm will move to position No. 10 which imposes a ground on No. 5 contact of range switch No. 10, and this ground then appears on position No. 5 of the follow-up switch. The motor then will turn until it reaches this position and stop.

Not shown on the block schematic are the rectifiers which supply the various circuit units. The plate, bias, and cathode heater supplies of all vacuum tubes, with the exception of those in the last three hf amplifier stages, are regulated in order to obtain gain stability through the circuit. The voltages of two of the three rectifiers used to supply these hf amplifiers are manually adjustable. A 4,000-2,000-volt, three-phase, full-wave rectifier supplies the plate voltage for the final two amplifier stages. A solenoid-operated grounding device serves to discharge both filter condensers in this rectifier whenever it is necessary to enter the cabinet. Another similar grounding device shorts and grounds the open-wire transmission lines. Located at convenient places in the front and rear of the transmitter are grounding sticks which can be used for grounding circuit elements under inspection.

All metering necessary in normal operation can be done from the front of the transmitter without interrupting the circuit. Most of the vacuum-tube space currents are measured by the use of small metering resistors in the cathode circuits which are shunted with a suitable meter by means of a transfer switch. In this way the number of meters is kept at a minimum. For the final three hf amplifiers individual meters are provided to measure the space currents. Other meters are used to measure the ac and dc voltages, and the position of the various servo-controlled elements. Two additional meters measure the RF current in the load resistors of hf amplifiers 5 and 6 during lineup tests. Jacks make available test points for voltage and frequency measurements during maintenance operations.

Equipment Description

The transmitter shown in Figs. 2, 3, 4, and 5 is a single floor-supported unit containing equipment mounted on an iron framework which is enclosed by a sheet-metal housing. The housing consists of two end panels, a top panel, four doors on the front with a meter panel above them, and four doors on the rear. The left-hand front door has a section which can be let down to form a horizontal writing shelf. The doors are treated with sound-deadening material. The exterior of the housing has a light aluminum gray lacquer finish and the interior a light gray enamel finish. The outside dimensions of the assembly are 84 inches long, $42\frac{1}{2}$ inches deep, and 84 inches high, exclusive of the air-intake hood. The completed transmitter weighs approximately 5,800 pounds. The frame is built of channel and angle iron with steel plates welded into a single structure. There are four bays accommodating 19-inch wide panels and rectangular steel boxes that shield the hf amplifier stages. These boxes are supported on horizontal angle-

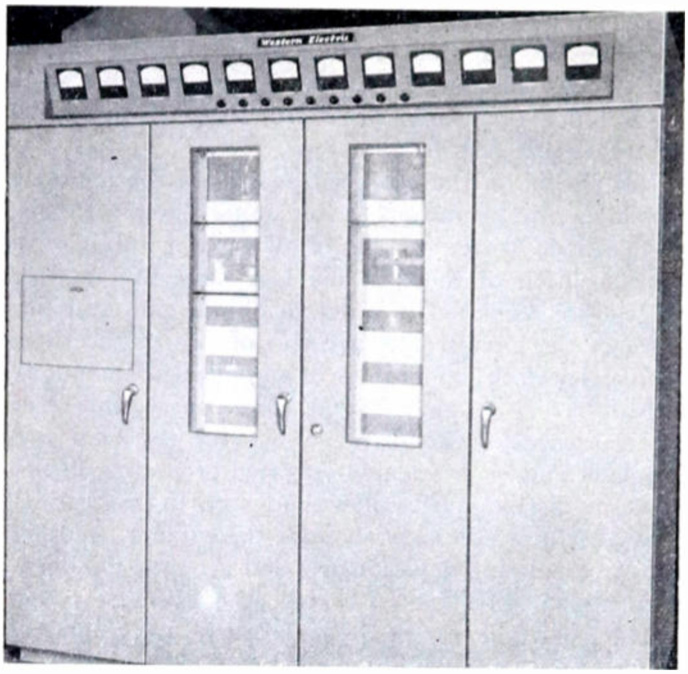

Fig. 2—Front view of transmitter with all doors closed.

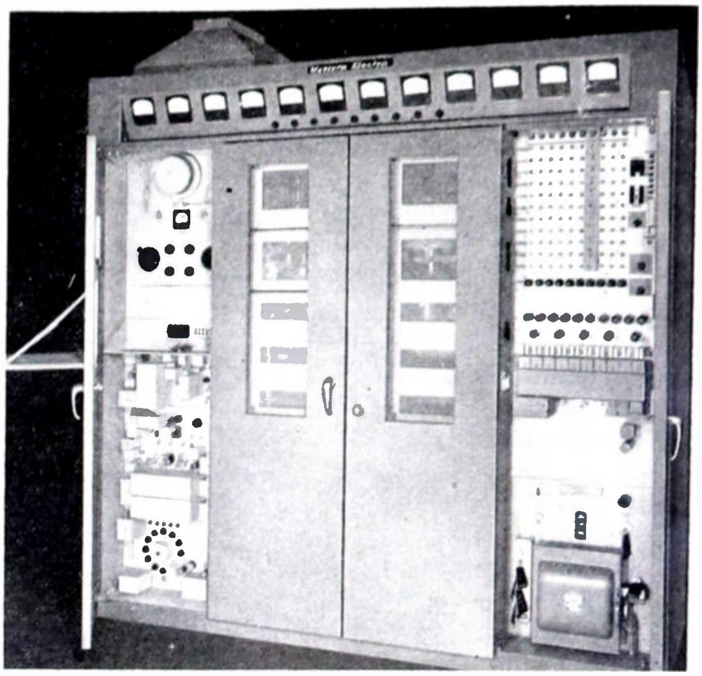

Fig. 3—Front view of transmitter with cabinet doors, to control bays, open.

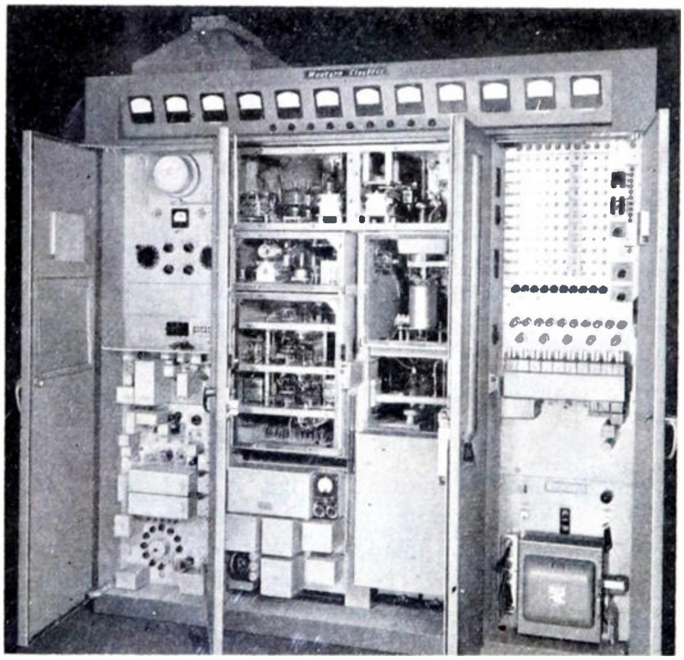

Fig. 4—Front view of transmitter with doors of cabinet and shielded compartments open.

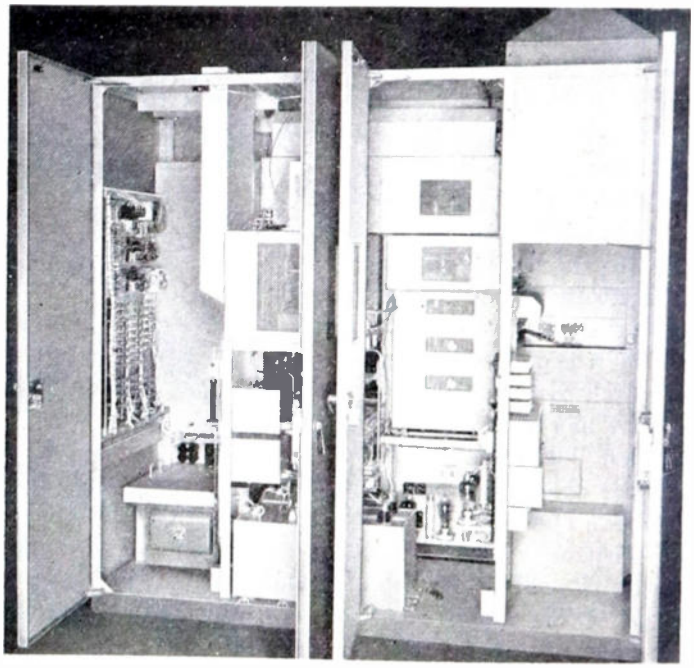

Fig. 5—Rear view of transmitter with cabinet doors open.

iron rails and have doors at front and rear for good accessibility. The transmitter can be assembled and completely wired before the housing is attached which is quite advantageous from a manufacturing standpoint.

A plenum chamber occupies the top third of the left-hand bay and contains a blower mounted on vibration mounts. This blower is driven by a 230-volt, 50/60-cycle, 3-phase motor. Dual belts and variable dual-sheave pulleys allow changes to be made in blower speed to accommodate 50/60-cycle operation from sea level up to 5,000-feet altitude. Air from the blower is used to cool the tubes in the hf amplifier stages and other parts of the transmitter. The cooling air enters through a 12-inch high hood on the top of the left-hand bay, and is exhausted through a flue at the top of the right-hand bay. The hood contains a dust filter and the combination reduces the air-intake noise quite noticeably. If desired, the intake hood and exhaust flue may be connected to an external duct system. This reduces the air noise still further by a noticeable amount.

A pressure-controlled switch associated with hf amplifier 5 compartment prevents filament voltage from being present on the final amplifier stages whenever the air pressure is below a predetermined value.

The transmitter is equipped with key interlocks so that access to the inside of the transmitter through the two middle front doors, the four rear doors, or the meter compartment door on the top cannot be obtained until the main power switch has been thrown and locked in the off position. The general objective is to prevent exposure of parts at potentials higher than 150 volts. The right and left front doors may be opened at any time since they only give access to operating controls. The key interlocked doors and some of the interior shielded compartment doors are equipped with additional protection in the form of electric door switches. An interlocked test switch permits power to be restored to the modulators and other low-voltage units with the cabinet doors open.

In order to insure adjusting only those potentiometers on the servo panel that are associated with a specific frequency, a magnetically held mask is available which may be placed over the desired row to outline the ends of the potentiometer shafts.

Performance

In general, multichannel transmitters are required to meet more rigid transmission performance requirements than are single-channel transmitters. Some of the considerations involved in meeting requirements for multichannel transmitters are discussed in a companion paper, entitled "Amplifiers for Multichannel Single-Sideband Radio Transmitters." The methods for obtaining high standards of performance described in that paper have been applied in the design of this transmitter and rather extensive measurements have been made on one model. The results give some indication of the expected performance. The following characteristics will be discussed:

1. Signal-to distortion ratio.
2. Speech-to-intermodulation noise ratio.
3. Adjacent-band radiation.
4. Linearity.

The signal-to-distortion ratio was measured using the two-tone method. Fig. 6 shows the curves obtained from measurements made with the load control "out." The third-order, fifth-order, and seventh-order distortion curves are plotted against the relative input tone. In obtaining data of this nature in a high-frequency radio transmitter where the bandwidth is small compared to the operating frequency only odd-order modulation products need be considered because even-order products produce noise which lies far outside the band of the system. From Fig. 6 it is apparent that the third-order distortion is the highest, and therefore is taken as

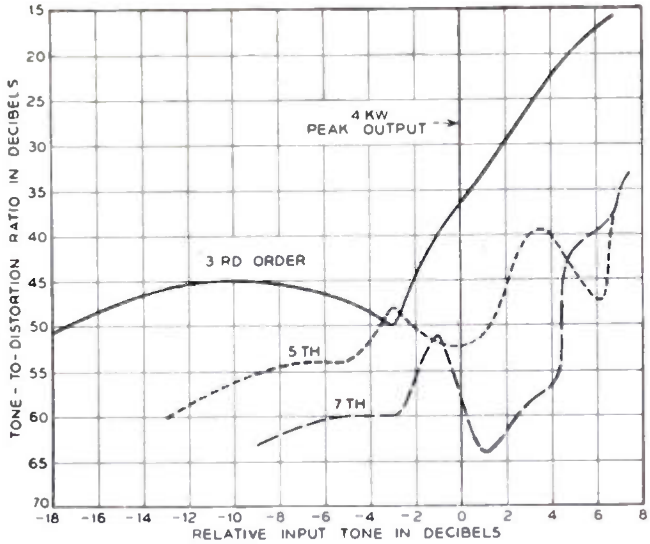

Fig. 6—Tone-to-distortion ratios plotted against input tone amplitude.

a measure of the signal-to-distortion ratio. It may be noted that at the peak output of 4 kw the signal-to-distortion ratio is about 36.5 db.

The speech-to-intermodulation noise curve is shown in Fig. 7. These data are plotted against the transmitter

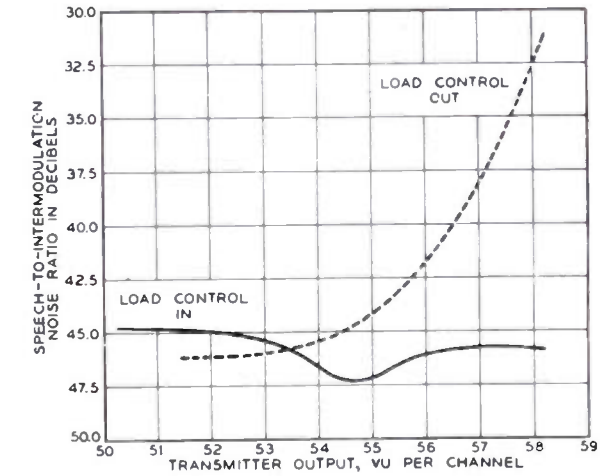

Fig. 7—Measured speech-to-intermodulation noise ratios plotted against transmitter vu output per channel

vu output per channel for conditions of load control "in" and load control "out." These curves indicate that the load control prevents excessive intermodulation as the transmitter output per channel increases. They also show that the transmitter can be operated at an output of 56.7 vu per channel when using the load control and produce no more interchannel modulation noise than when operating at 55-vu output per channel without load control. It is evident, then, that the load control allows transmitter to be operated at 1.7 vu greater output per channel on basis of interchannel modulation noise.

The present crowded condition of the high-frequency spectrum in which transmitters of this type operate demands that the adjacent-band radiation should be kept to a minimum to reduce possibilities of interference.

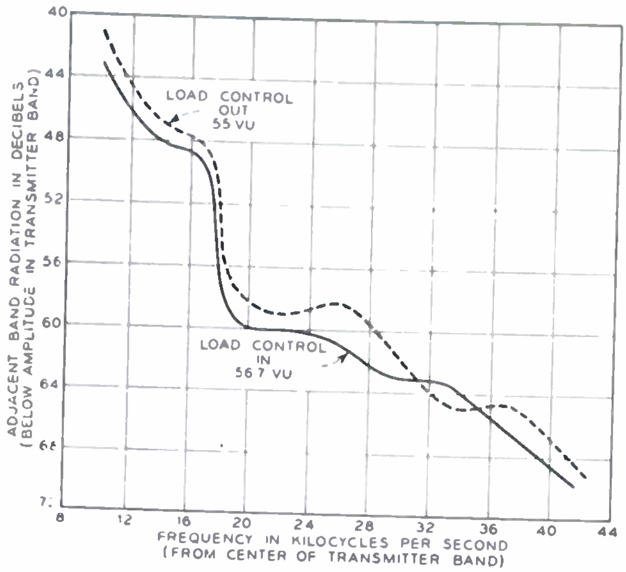

Fig. 8—Adjacent-band radiation plotted as a function of frequency measured from the center of the transmission band.

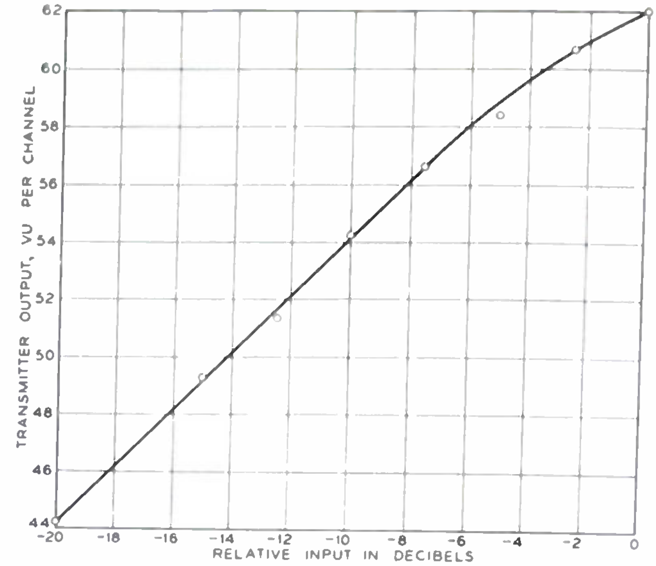

Fig. 9 Linearity curve.

Measurements of the adjacent-band radiation[5] were made both with the load control "in" and with the load control "out." The results are shown in Fig. 8, where the adjacent-band radiation is plotted as a function of frequency measured from the center of the transmission band. The ratios were obtained by measuring the speech amplitude in a 2.7-kc transmitter band and relating it to the amplitude of modulation noise in a 2.7-kc receiver band centered at the frequency indicated. The data show that with the load control "in" the adjacent-band radiation is about 1 to 2 db lower than with the load control "out." It may be noted that the transmitter output per speech channel with the load control "in" is 1.7 vu greater than with the load control "out." This supports the observation made in regard to Fig. 7, and shows that the advantage of the load control is also about 1.7 vu in transmitter output per channel on the basis of equal adjacent-band radiation.

Fig. 9 shows the linearity curve. The over-all input-output characteristic is shown using speech loading with the load control "out." This curve shows the relationship between the transmitter vu output and the relative input in decibels. It will be noted curve is linear up to an output of about 57 vu where it begins to fall off slightly.

The model of this radio transmitter was installed at the Lawrenceville, N. J. overseas radio station, operated by the American Telephone and Telegraph Company in December, 1949. It has been used for the transmission of four independent 250–2,750-cycle speech channels in regular overseas service.

[5] N. Lund, "Methods of measuring adjacent-band radiation from radio transmitters," Proc. I.R.E., vol. 39, pp. 653–656; June, 1951.

Amplifiers for Multichannel Single-Sideband Radio Transmitters*

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Summary—**Considerations are given for designing high-frequency amplifiers whose performance will meet the high standards required for amplifying multichannel signals. A relationship between tone and speech data is presented to show how the tone rating of the amplifier can be determined from the speech rating and interchannel modulation noise requirements. Some of the factors which influence distortion and some methods of reducing such distortion are also discussed.**

* Decimal classification: R355.7×R423.5. Original manuscript received by the Institute, April 5, 1951; revised manuscript received, March 3, 1952.

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IN THE DESIGN of carrier telephone systems, the goal has been to carry as many channels as possible over one pair of wires to reduce the cost of service per channel. Similarly, large economies can be obtained by multichannel operation of high-frequency radio systems. These economies result because the cost of the additional equipment required in a transmitter and receiver to obtain multichannel operation is far less than the cost of individual transmitters, receivers, and antennas. In addition, it is possible to place a greater number of channels in a given frequency space when using multichannel transmitters rather than single-channel systems. This is true because only one band is required to take care of frequency instability of the multichannel transmitter, whereas an equal bandwidth is required for each single-channel transmitter. Some of the early high-frequency single-sideband systems were of the twin-channel type.[1] In those twin-channel systems, two 3-kc channels were transmitted over the same radio transmitter with a 3-kc guard band between the channels. This guard band was used to prevent the relatively strong third-order modulation products, generated by one channel, from falling into the other channel and causing excessive interference. In recent years the design objective has been to use all available frequency space by placing additional channels in the guard bands. Consequently, the third-order modulation products which fall into adjacent channels must be minimized. Any transmitter which is modulated by a complex wave, such as a speech signal, will generate some adjacent-band distortion products which may fall into the channels of other services and cause interference. Therefore, it is necessary that the transmitting amplifier used in such a transmitter shall be capable of amplifying the multiplex signal without generating either excessive interchannel modulation noise or excessive adjacent-band radiation.

In the conventional single-sideband system the output of the final modulator is at low amplitude and contains some unwanted modulation products. This requires that the high-frequency amplifier have high gain to obtain the desired power output and have high selectivity in order to suppress the unwanted modulation products. In addition, the shielding of the amplifier and the selectivity of its circuits must be adequate from a harmonic radiation standpoint.

In order to obtain satisfactory and continuous operation of the long-distance overseas services, it is often necessary to change the transmitter frequency three or four times during a day over a frequency range of four to one. These frequency changes should be made as rapidly as possible so as to make the greatest amount of circuit time available. Therefore, it must be possible to set rapidly and accurately the tuning of the high-frequency circuits so as to obtain the same high standards of distortion and gain stability for all operating frequencies.

For single-sideband transmitters, it has been convenient to rate the transmitting amplifier in terms of the maximum envelope peak-power output that can be produced without generating excessive distortion. This maximum envelope peak-power has been defined as the peak-power output that results when the transmitter is loaded with two equal tones such that the third-order distortion is a definite number of db (usually 25) below the amplitude of one of the two tones. From a radio-transmission standpoint, a more significant rating is the equivalent vu output of the transmitter per channel, which is defined as the reading of a vu meter at the audio input increased by the gain between that point and the single-sideband transmitter output. The gain is taken as the maximum value which can be used without generating excessive interference to channels either internal or external to the transmitted band. The vu rating is more significant than the tone rating because it is the equivalent vu output of the transmitter which determines how effectively the signal will overcome noise and interference at the distant radio receiver. The designer, then, is given the required vu output per channel and number of channels to be transmitted.

However, before the designer can select the proper tube complement for the amplifier, he needs to determine an equivalent tone rating for the amplifier. This is necessary because power tubes are usually rated on the basis of tone output. Therefore, the designer needs to know the correlation between tone rating and speech vu rating for the particular number of channels to be used. Such a relation has been determined empirically, and the method for determining it will be presented.

It will be assumed that a radio transmitter is to be designed to have certain characteristics. The particular specifications of the system that must be known before solution of the problem can begin are (1) number of channels and frequency arrangement of the channels, (2) permissible maximum interchannel cross talk and adjacent-band radiation, and (3) the desired radio transmitter output in vu per channel.

Correlation of Tone and Speech Tests

The problem of correlating the tone and speech operation of a radio transmitter is somewhat similar to the problem of correlating the amplitude of the modulation products generated by tones with the interchannel modulation noise generated by speech in carrier telephone systems. This problem in carrier systems has received considerable attention. For amplifiers used in such systems, load rating and modulation requirements are handled separately.[2,3] It is assumed that the amplifier uses a large amount of negative feedback, and consequently, shows low modulation below the overload point and high values of modulation above the overload point. The load rating is then established on the basis of input-amplitude distribution; that is, the amplifier is loaded so that the maximum amplitudes for the combined channel distribution will exceed the overload point for only a small percentage of the time. Modulation requirements are then established separately for amplitudes which do not exceed the overload point.

Telephone carrier systems have many channels with unregulated speech volume on each channel, so that a

[1] A. A. Oswald, "A short-wave single sideband radio telephone system," Proc. I.R.E., vol. 26, pp. 1431–1454; December, 1938.

[2] B. D. Holbrook and J. T. Dixon, "Load rating theory for multi-channel amplifiers," *Bell Sys. Tech. Jour.*; October, 1939.

[3] W. R. Bennett, "Cross-modulation requirements on multi-channel amplifiers below overload," *Bell Sys. Tech. Jour.*; October, 1940.

considerable allowance in load rating must be made to take care of the large range of talking volumes on each channel. Since these carrier systems use feedback amplifiers operated well below overload, the modulation requirements can be fairly well approximated by assuming that only second- and third-order products are important.[3] A high-frequency radio system, using a small number of channels, differs from a carrier system in that the speech volume input is controlled, that is, maintained on the several channels at a predetermined level. Low levels extremely susceptible to interchannel modulation noise are thus avoided, and the transmitter may be operated near the overload point. One other difference between the radio transmitter and the carrier amplifier is that the modulation products are generated in a class B amplifier of the radio transmitter, where at low amplitudes, the magnitudes of the modulation products do not change rapidly with input amplitude; while the modulation products generated in the class A carrier amplifier vary as the exponent of the order of modulation product. That is, the class A carrier amplifier input-output characteristic may be approximated by a simple power series using constant coefficients, but a similar analysis for the class B amplifier as usually operated is more involved.

In general, interchannel modulation noise is thought of as resulting from the harmonics of single frequencies or the beating of two or more discrete frequencies in the speech band. In a system using a high-frequency transmitter where the bandwidth is small compared to the operating frequency, only odd-order modulation products of two or more frequencies need be considered. The even-order products and the products involving only a single frequency produce noise which lies far outside the pass band of the amplifier.

A correlation between tone and speech measurements based on experimental data for three types of single-sideband transmitters has been obtained. This correlation shows that there is a fairly definite relation between two-tone third-order distortion measurements and modulation noise measurements where the modulation noise is the result of third-order modulation between speech frequencies. The results are plotted in such a manner that it is possible to answer the following design question for a new single-sideband system. If the speech-to-modulation noise ratio for the new system is to be X decibels, what should the corresponding tone-to-distortion ratio be for that system?

Interchannel modulation noise data have been obtained for two-, three-, and four-channel systems when using three different single-sideband reduced carrier radio transmitters. In each of these systems the speech bands were 2,500 cps wide with 500-cps bands separating the speech bands. The reduced carrier was transmitted at an amplitude of 26 db below the rated envelope peak-power output and was placed at the center of the over-all band. For the three- and four-channel systems the maximum interchannel modulation noise was found on the channels adjacent to the carrier, and this modulation noise was mainly the result of third-order modulation products. The interchannel modulation noise data were obtained as speech-to-modulation noise ratios for each channel combination. For each channel combination, a series of measurements was made for different input speech amplitudes. These speech measurements for each transmitter were then compared with the two-tone third-order distortion measurements for that transmitter. All the tone and speech data have been plotted in Fig. 1 using certain factors relating the transmitter speech input in vu and the transmitter tone input in dbm. These factors for the three different channel combinations are

Number of Channels	*DBM-VU Difference Factor*
4	6
3	5.5
2	4

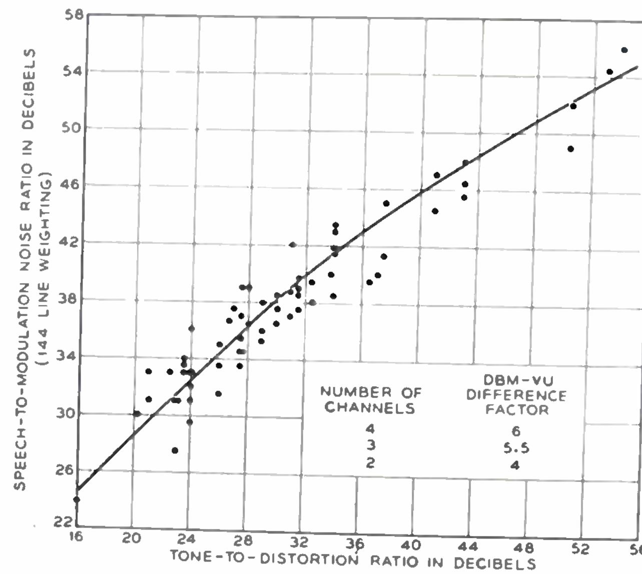

Fig. 1—Combined plot of measured two-tone third-order distortion ratios and speech-to-modulation noise ratios for three different types of transmitters.

These data show, for example, that in a four-channel system, if the tone input is numerically 6 db greater than the speech in vu, the speech-to-modulation noise ratio will be related to the tone-to-distortion ratio by the curve plotted in Fig. 1. The data plotted in Fig. 1 have been obtained for three different transmitters having widely different distortion characteristics. The measured third-order distortion characteristics for these transmitters is shown in Fig. 2. Transmitter A had an envelope peak-power rating of 7.5 kw and incorporated radio-frequency feedback. Two curves are shown for this transmitter one with 20 db of negative feedback and one with no feedback. Transmitter B had an envelope peak-power rating of 4 kw and had the negative feedback inherent to a grounded-grid final amplifier stage. Transmitter C had an envelope peak-power rating of 2 kw and used no feedback. The two-tone curves for transmitters B and C show some sharp variations from a smooth curve. These variations are usually the results of the distortion product generated in one part of the amplifier partially cancelling or adding to the distortion product generated in another part of the amplifier. With tone input, it would be expected that this cancellation or addition process would occur at a particular amplitude of the cyclic input wave. When speech is applied, the input amplitudes of the component speech frequencies have a fairly large range of values, and the cancellation effect takes place over only a small range of amplitudes. For component amplitudes slightly greater or less than the critical amplitude, the modulation noise is much higher than the cancellation value and the effect of low-amplitude components is minimized. Therefore, when interpreted for speech, the two-tone distortion curves are graphically smoothed as shown by the dotted curves in Fig. 2. It is the smoothed data which are used to obtain the relationship between speech and tone measurements as shown in Fig. 1.

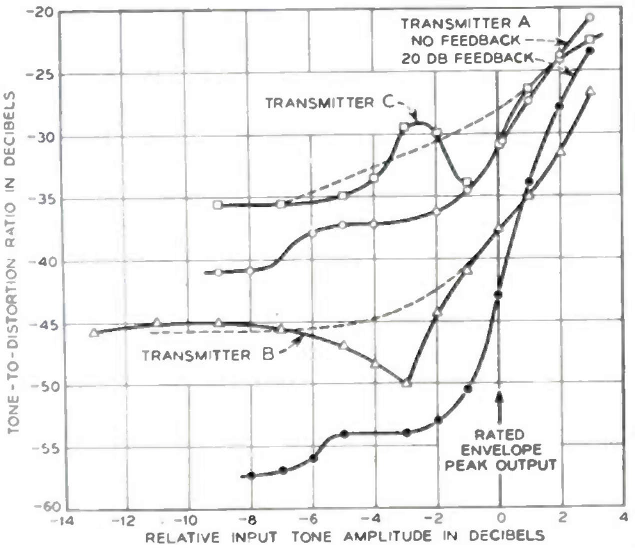

Fig. 2—Measured variation of two-tone third-order distortion ratios with input amplitude for three different types of transmitters.

By using Fig. 1 the tone power output and two-tone distortion requirement for the amplifier can be established. For example, if it is required to have a speech-to-modulation noise ratio of 35 db in a four-channel system, Fig. 1 shows that the two-tone third-order distortion ratio should be 27 db when each of the two tones is 6 db greater than the speech output in vu. If the required speech output from the transmitter is 54 vu per channel, then the corresponding single-tone output of the transmitter should be 60 dbm, and the envelope peak-power should be 66 dbm when the two-tone distortion ratio is 27 db.

The correlation gives a method for determining what the tone rating of the transmitter should be in order to obtain a certain speech-to-modulation noise ratio in the channel nearest the center of the transmitted band. In some cases it may be desirable to design for a certain speech-to-modulation noise in a channel just outside the transmitted band. In a four-channel system occupying a 12-kc band, it has been found that the modulation noise in the next 3-kc channel just outside the transmitted band is normally about 4 db less than the modulation noise in a channel adjacent to the carrier which is at the center of the transmitted band.

Automatic Load Control

Normally, each telephone channel in a multiplex system is busy for only a small percentage of time. This means that all four channels will be busy at the same time for an even smaller percentage of time. These facts suggest that it would be advantageous to use some device which causes the transmitter to be loaded to its peak capacity regardless of the number of channels that are busy. Such a device, which is termed an automatic load control, can be used to give an average increase in speech power output. This device is essentially a backward acting automatic volume control with the time constants so adjusted that there is little clipping of the speech signals. The attack time constant of the device is about 2 milliseconds so that the transmitter amplifier will not be overloaded for any appreciable length of time. The recovery time constant is about 100 milliseconds so that there is little distortion of the speech syllables. A biased rectifier is used to rectify a sample of the radio-frequency envelope at the output of the amplifier. The control voltage so obtained is used to control the gain of a low-level amplifier stage. Measurements on a system using this device indicate that the transmitter output can be increased by about 1.5 vu in a four-channel system without increasing the adjacent-band radiation.[4] These results were obtained when each channel was loaded with speech for about 50 per cent of the time. If the duty cycle for each channel is less, the advantage of the load control will be greater. This load-control device is also effective in preventing accidental overload of the transmitter, and also tends to reduce the bad effects of variations in the transmitter gain.

Causes of Distortion and Its Reduction

After determining the tone-to-distortion ratio and tone power output required, power amplifier tubes capable of meeting these requirements can be selected. It would be desirable to select tubes for the driver stages so that these stages would not contribute to the over-all distortion of the amplifier. In practice this is uneconomical, and a compromise must be made to keep the size of the driver stage within reasonable limits.

A method of reducing distortion is to use radio-frequency feedback between the output stage and some preceding stage. Experimental feedback amplifiers having power outputs up to 150 kw have been built and tested in the laboratory to evaluate the advantages of feedback. These experiments showed that substantial distortion reduction can be obtained when the transmitter output volume is below the usual rating of the transmitter, but that these advantages rapidly disappear as the output volume approaches this rating. This,

[4] Performance data for this device as used in the Western Electric Type LD-T2 Transmitter is given in a companion paper by L. M. Klenk, A. J. Munn and J. Nedelka, "A multichannel single-sideband radio transmitter," Proc. I.R.E., vol. 40, pp. 783–790; this issue.

together with the added complexity of over-all radio-frequency feedback, has made its use appear unwarranted for the type of application discussed here.

A typical third-order two-tone distortion characteristic for a complete amplifier is shown in Fig. 3. This

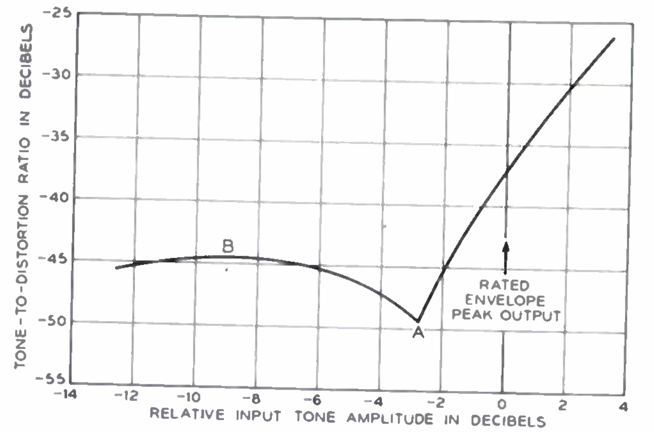

Fig. 3—Typical two-tone third-order distortion characteristic for a transmitter amplifier.

characteristic shows that the distortion has a sharp null or cancellation at point A which corresponds to a power output of about 3 db less than the rated envelope peak power. Also this characteristic shows that the distortion is relatively constant in the B region which corresponds to a power output of about 6 to 10 db less than rated peak output. For still lower power output the distortion gradually decreases.

A typical dynamic characteristic for the same amplifier, exaggerated for illustrative purposes, is shown in Fig. 4. This characteristic is a plot of the input to the amplifier against the output of the final stage. For very small inputs in the region A the amplifier is operating class A, and for higher inputs in this region the operation is class AB. The curvature of the dynamic characteristic is concave upward in region A, and such curvature is typical of a class A or AB amplifier. The amount of curvature in this region is small so that the distortion is relatively small as shown in the region to the left of B in Fig. 3. At point X the final stage begins to draw grid current and the amplifier begins to operate as a class B amplifier. As soon as the final stage draws grid current, an additional load is placed on the driver stage, and, as seen on the dynamic characteristic, this prevents the output from increasing proportionately as fast as the input. When the grid of a typical power tube is driven positive, there is a region where the grid current becomes constant or actually may decrease and then increase again. This effect is known to be caused by secondary emission from the grid. In this region C on the dynamic characteristic, the grid of the final stage is not loading the driver stage as heavily as in the region B, and consequently the dynamic characteristic tends to straighten out again. In region D the dynamic characteristic is again concave upward because the plate current-grid voltage characteristic of the final amplifier has this curvature and is the controlling factor on distortion. Region E represents an overload condition and shows a leveling off which is the result of the peak plate voltage approaching the value of the dc plate supply voltage for the output stage.

The dynamic characteristic shown in Fig. 4 may be used to explain the shape of the two-tone third-order distortion curve shown in Fig. 3. The amount of distortion generated when operating up to any particular point on the dynamic characteristic is roughly proportional to the maximum distance between the dynamic and a straight line joining that point to the origin. For low amplitudes in region A of Fig. 4 the departure from linearity is small so that the distortion is also small. As the amplitude is increased, the departure from linearity increases until at point W the departure is a maximum. Operating up to point W then corresponds to region B shown in Fig. 3. As the amplitude is increased beyond point W, the departure from linearity decreases until at point Y, the departure is a minimum. Operating up to this point Y on the dynamic characteristic then corresponds to the null in the distortion curve at point A in Fig. 3. When the amplitude is increased beyond point Y, the departure from linearity increases again and the distortion increases.

It may be noted that the cancellation effect is the result of a change in curvature of the dynamic characteristic, and this change in curvature is a result of grid current drawn by the final amplifier stage. This cancellation effect normally does not occur in a class A amplifier because the dynamic characteristic does not have such a point of inflection. For a class A amplifier the distortion increases steadily with input and usually follows a simple law.

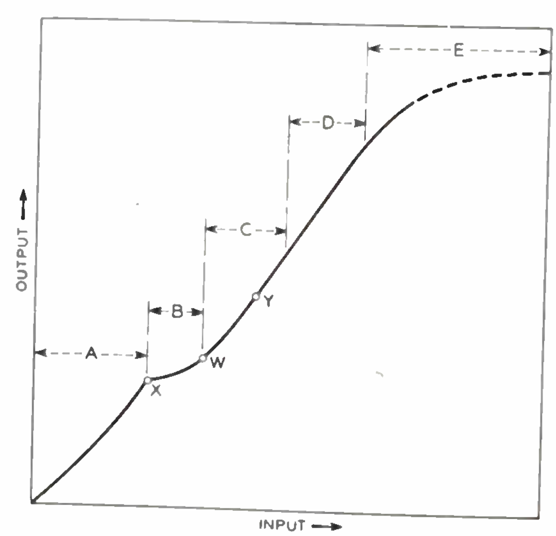

Fig. 4—Typical dynamic characteristic for transmitter amplifier.

In region C of the dynamic characteristic shown in Fig. 4, it is noted that secondary emission has a tendency to straighten out the dynamic characteristic and thereby give lower distortion. Therefore, tubes which show no secondary emission would give more distortion for the same driving power. Consequently, tubes which do show some secondary emission effect are preferred for the power amplifier.

From the above discussion it is seen that the manner in which the dynamic characteristic for one stage adds to another may either increase or decrease the over-all distortion. In a multistage amplifier such effects can be quite complex if many stages are operated near overload. It is usually desirable that not more than the final two-stages contribute appreciably to the over-all distortion. The following general guides have been found useful in designing amplifiers for low-distortion output:

1. Operate as many stages class A as economically feasible.
2. Where grid current is drawn, calculate the fundamental component under full drive conditions and shunt the grid with a swamping resistor such that the current supplied to this resistor by the previous stage is at least 10 times that due to the fundamental component of the grid current.
3. Where tubes are operated other than class A, select a bias so that under full drive conditions, plate current will flow through at least 220 degrees of the cycle. Self bias, either in whole or in part, should not be used for stages operated other than class A.
4. Select the value of impedance presented to the plate of the final stage so that under full drive conditions the radio-frequency voltage developed at the plate will be 75 to 80 per cent of the dc plate voltage. Therefore, the efficiency under such peak drive conditions will be of the order of 60 per cent.
5. Select any stage of the amplifier as a reference. Preceding stage which drives it should be of sufficient size that at least a 3-db margin above full drive requirements can be obtained without undue distortion contribution from driving stage.
6. Plate, screen, and bias supplies for those stages other than class A should have excellent regulation characteristics. The impedance of such supplies to the audio-frequency currents generated in grid circuits or class B plate circuits should be kept low. Under two-tone full-load test conditions, the audio voltage developed across such impedances should be less than 2 per cent of the dc voltage at audio frequencies above 100 cycles. At the resonant frequency of the rectifier filters, usually about 35 cycles, the audio voltage should be less than 10 per cent.
7. The shielding of the radio-frequency circuits and the filtering of the power-supply circuits should be designed so that the spurious feedback will not affect the voltage at any grid by more than 10 per cent at any operating frequency.
8. In general, it is the feedback path within the tube itself which determines what values of stable gain can be obtained. In the majority of cases, using present-day tetrodes, a gain of 15 to 20 db per stage can be realized with satisfactory stability. The gain per stage for higher power stages where grids are driven positive is rarely in excess of 12 db on account of the power which must be wasted in grid swamping resistors to minimize distortion. In the case of grounded-grid amplifiers the gain is even less than this, but the power is not wasted since it appears in plate circuit of driver stage.

Circuit Arrangements

After selecting a tube complement which will meet the power output and distortion requirements, the designer may consider some of the other requirements of the amplifier. These requirements may be outlined as follows.

Frequency Range

In general, about a 6-to-1 frequency range will be specified, and this will be within the limits of 3 to 30 mc; for example, a 4-to-23 mc range. The tuning facilities must be such that the transmitter can be tuned quickly and accurately for optimum operation into a wide range of antenna impedances at any frequency in this range.

Selectivity

Since the output of the modulator contains, in addition to the sideband to be amplified, the other sideband and a conversion frequency, the tuned circuits in the amplifier must be selective enough to reject the unwanted frequencies to a satisfactory degree. An overall check is required to verify that the spurious frequency radiation is well below either the F.C.C. or the customers' requirements.

Gain

In order to obtain low values of distortion in the final modulator of the transmitter, it is necessary to operate the modulator at low signal amplitude. The output of such a modulator may be of the order of a tenth mw when the amplifier is delivering its peak envelope power. If, for example, the amplifier is to deliver 4-kw output, the required gain is of the order of 76 db. Obviously, the amplifier must consist of a number of stages with good shielding between the low- and high-level portions of the amplifier circuits.

Power Frequency Noise

Unwanted modulation of the single sideband or carrier by power-supply frequencies is classed as noise. The sum total of these products should be at least 45 db less than the peak capacity of the amplifier. It is preferable that they be 50 db less. In some high-power amplifiers, it may be necessary to quarter phase the filament circuits to meet this requirement.

For a specific amplifier, there is probably no particular circuit arrangement which can be defended as being optimum. In the past, balanced or push-pull circuits have been widely used. This type of circuit has advantages where triode amplifier tubes are utilized, inasmuch as bridge neutralization can be employed and made effective over a wide frequency range with relatively minor adjustment of neutralization over the range. Since the load connected to the output of the amplifier is usually balanced, a simple circuit can be used to couple the load to the power output tubes of the amplifier. In addition, when high powers are involved, some of the heavy current blocking capacitors used for single-sided circuits are eliminated in the push-pull type of circuit. However, the push-pull circuit requires more tuning elements and in general costs more. Furthermore, the circuit elements require rather precise control to obtain a high degree of symmetry, and the tubes must be selected to have reasonably well-matched characteristics. Harmonic suppression in the bridge-neutralized push-pull amplifier is sometimes inadequate because a center ground, which would be desirable from a harmonic suppression standpoint, must be omitted to keep the amplifier in stable balance.

Tube developments in recent years, particularly as regards moderately high-power tetrodes and high-transconductance high-frequency triodes, have made possible the construction of multistage high-gain amplifiers of substantial power output requiring no neutralization. The general scheme is to utilize tetrodes throughout or else a grounded-grid triode as a final output stage and incorporate a large enough tetrode in the preceding stage to drive it, smaller tetrodes being used as preliminary stages. The elimination of the neutralization problem practically nullifies the major advantage possessed by a push-pull circuit, hence single-sided circuits are generally employed in present-day practice. The principal disadvantage of a single-sided circuit is the necessity for converting the output of the final stage into a balanced form for connection to the balanced antenna system. However, the problem of providing such a conversion circuit is relatively minor as regards cost or technical difficulty.

For amplifiers which have a power rating of 20 to 60 kw, it has been necessary to use triodes since high-power tetrodes have not been available.[5] In this case, a bridge-neutralized, single-stage amplifier may fit the particular circumstances most economically, particularly if the driving power available is insufficient to excite a grounded-grid type of amplifier. Sufficient neutralization must be provided so as to prevent gain variations which will cause poor distortion characteristics. Experience shows that a variation in gain of plus or minus one db can usually be tolerated when working with a triode amplifier which affords 7 to 12 db of power gain. The degree of neutralization usually must be such that not more than 5-per cent variation in input voltage results when the output circuit is detuned or decoupled sufficiently to produce a 50-per cent change in radio-frequency output voltage.

From a circuit standpoint, single-side circuits present three distinct advantages when compared with push-pull circuits. First, they require fewer circuit elements, which minimizes costs. Secondly, single-side circuits present fewer circuit paths by which parasitic oscillations may occur. Third, harmonic suppression can be achieved with simpler circuits.

Regardless of the type of circuit employed, the interstage tuning and antenna coupling facilities incorporated into the amplifier design should be such that each tube operates over the same portion of its characteristic at all frequencies in the operating range of the amplifier. This means that under any particular condition, the plate-load resistance and the successive grid load impedance are constant with frequency, and therefore the gain is independent of frequency.

No special difficulty is encountered in arranging conventional types of interstage circuits to achieve this constancy of gain per stage. Obviously, the inductors or capacitors of the circuit or both must be variable over a wide range of values to be tunable over a 6-to-1 frequency range. Inductive coupling does not lend itself well for this application of constant gain per stage and ordinarily has no compensating advantage to warrant its use.

Plate and Grid Supply Rectifiers

The plate and grid currents drawn by single-sideband class B amplifiers contain audio-frequency components. These audio currents flowing through the rectifier filter impedance develop audio voltages which cause the amplifier output signal to be distorted. Ideally, the impedance of the rectifier supply system should be zero to these currents over the entire audio scale. In actual practice, it is feasible to keep this impedance low over most of the frequency range, but it is necessary to tolerate an increase in impedance at the frequency where the inductive and capacitive elements of the smoothing filter antiresonate. Filters designed to antiresonate in the region of 35 cycles have been found satisfactory since this is well above the range of syllabic frequency (0–20) and below most of the difference frequencies associated with speech. To minimize the impedance over a wide range it is necessary to make the capacitive element large and the inductive element small. The power rating of the plate rectifier should be such that it can carry continuously the plate current which flows when the normal signals are applied to the transmitter. Also, the rectifier should be capable of carrying overload currents for short periods during which test tones are applied to the transmitter.

Conclusions

Considerations involved in designing high-frequency amplifiers which will meet the high standards required for amplifying multichannel signals have been given. A relationship between tone and speech data has been presented to show how the tone rating of the amplifier can be determined from the speech rating and interchannel modulation noise requirements. Some of the factors which influence distortion and methods of reducing such distortion have been presented. The advantages and limitations of balanced and unbalanced amplifier circuits have been discussed, and it is concluded that unbalanced circuits are preferred for low- and medium-power amplifiers; for some high-power amplifiers, a balanced circuit may have an advantage.

[5] C. F. P. Rose, "A 60-kilowatt high-frequency transoceanic-radiotelephone amplifier," Proc. I.R.E., vol. 33, pp. 657–662; October, 1945.

Design of Modulation Equipment for Modern Single-Sideband Transmitters*

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Summary—**This paper deals with considerations that go into the design of modulation equipment for a single-sideband radiotelephone transmitter in which filters are used for sideband suppression. Balance requirements, frequency stability, the choice of intermediate frequencies, and methods of avoiding transmission of spurious frequencies are among the factors which are discussed.**

Introduction

An IMPORTANT PART of the modern single-sideband radio-telephone transmitter is the part in which the single-sideband signal is generated and placed in its proper frequency relationship to the assigned carrier frequency. Previous authors have given this part of the equipment various names, such as "exciter," "generator," or "drive," as distinguished from the part of the transmitter the chief function of which is to amplify the power. Its design is usually not independent of the amplifier design, however, and the selectivity and gain of the amplifier stages are factors of considerable importance in the design of a single-sideband generator.

Two fundamental methods for deriving a single-sideband signal are known. In both of these the sidebands are produced by modulating the amplitude of a carrier with the signal. In one, the elimination of one sideband is accomplished by filtering. This more common method has been in use for about thirty years since its first application in carrier telephone systems.

The second method of generation, which was invented by Hartley,[1] depends on phase cancellation for the elimination of one sideband, and has been referred to as the "phase rotation method"[2] or as the "balanced-sideband system."[3] A typical form of the generator uses two balanced modulators which have common sources of carrier and signal input, but which are arranged so that the carrier and signal voltages are supplied to one modulator 90 degrees displaced in phase from the carrier and signal voltages of the other modulator. As a result, in the combined modulator outputs, one sideband is cancelled. The use of balanced modulators serves to suppress the carrier.

This latter method is somewhat newer and has been used relatively little, largely because it suffers the disadvantages which are characteristic of any balancing device involving active elements. There are problems in obtaining, and still greater problems in maintaining, the high degree of balance which would be required for satisfactory suppression of the unwanted sideband. It is attractive because, theoretically, it offers the possibility of generating a single sideband directly at the final frequency. In this case, however, the preservation of exact phase and amplitude balance of the modulator outputs becomes especially difficult in a transmitter whose operating frequency must be changed rapidly from one to another of several frequency assignments.

The degree of sideband suppression which may be obtained in practice by the balanced-sideband method is far less than can be achieved with filters. The balanced-sideband method involves power-consuming and expendable tubes, and components requiring adjustment and maintenance. Filters, on the other hand, give a higher degree of selectivity and greater stability of performance, yet are passive networks which need no attention. It is believed, therefore, that for high standards of performance and quality of parts, a single-sideband generator incorporating balancing methods will not be less expensive than one using filters.

For these reasons, the conventional filter method has been used in all of the designs of transmitting equipment for Bell System overseas radiotelephone service. Only this method will be treated in the following paragraphs.

Before undertaking a detailed analysis of the individual problems and requirements of the various modulators and filters involved in a typical system, it seems advisable to review the way that an over-all system works. For this purpose, the frequency block schematic shown in Fig. 1 will be used. This was drawn for a modern transmitter which transmits two independent sidebands arranged symmetrically with respect to the final carrier. Each of these sidebands occupies a space between 0.1 and 6 kc measured from the carrier frequency. By means of suitable terminal equipment which is not included as a part of the transmitter, each of these bands may be subdivided into two speech channels, each approximately 2,500 cycles wide, or they may carry speech inversion or privacy as well as multiplex-tone telegraph circuits.

For illustration, the signal entering the Group A input is shown to consist of two speech bands, A1 and A2, lying in a band of frequencies from 0.1 to 6 kc. Similarly, the Group B input consists of another group of two speech bands, designated B1 and B2, which occupy a similar band of frequencies. These input frequencies modulate a 100-kc carrier which is supplied to both low-frequency modulators from a crystal oscillator. As a result of the modulation, each modulator produces double-sideband output with greatly reduced carrier,

* Decimal classification: R355.8×R148.16. Original manuscript received by the Institute, April 5, 1951; revised manuscript received, March 3, 1952.

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[1] R. V. L. Hartley U.S. Patent No. 1,666,206; granted April 17, 1928.

[2] O. G. Villard, Jr., "A high-level single-sideband transmitter," Proc. I.R.E., vol. 36, pp. 1419–1426; November, 1948.

[3] A. H. Reeves, "The single sideband system applied to short-wave telephone links," *Jour. IEE* (London), vol. 73, pp. 245–280; September, 1933.

as shown by the blocks on the spectral diagrams. These outputs are filtered by paired crystal filters having conjugate characteristics. Thus, the output of Low-Frequency Modulator 1 Carrying Group A is passed through a filter which transmits only frequencies lying above 100.1 kc and not greater than 106 kc. At the same time, the output of Low-Frequency Modulator 2 Carrying Group B is passed through a filter which transmits only frequencies below 99.9 kc and not lower than 94 kc. The outputs of the filters are then combined through a hybrid and yield a signal in whose spectrum the blocks A1 and A2 from Group A lie above the carrier and the blocks B1 and B2 from Group B lie below the carrier. At this point in the system, the carrier which had been balanced to a very low amplitude in the modulators is resupplied at a desired value, either 10 or 20 db below reference amplitude.

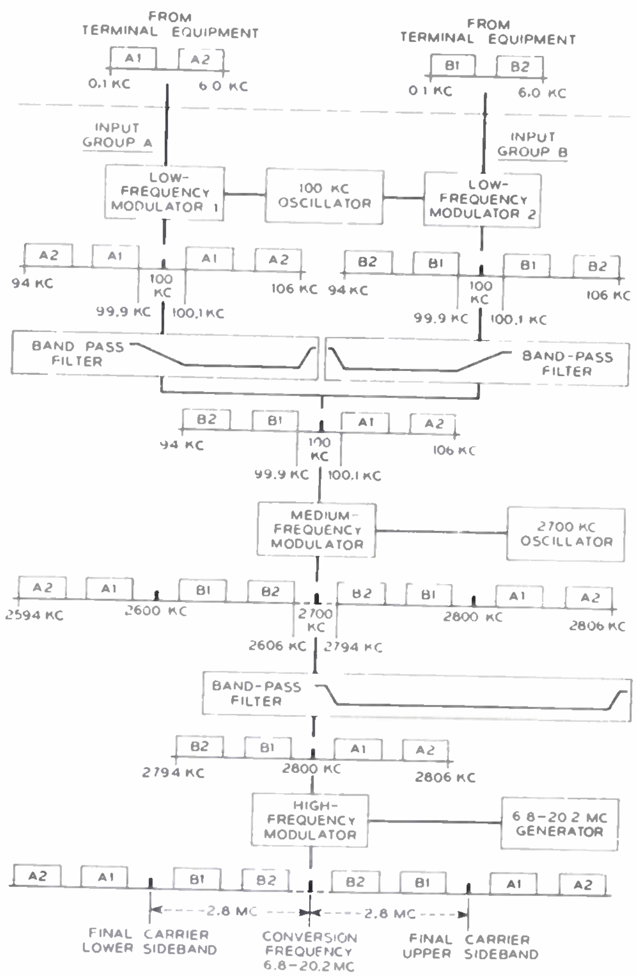

Fig. 1—Frequency block schematic of a modern single-sideband generator.

By virtue of the crystal filters and the high ratio of carrier-to-signal used in the modulators, the signal at this point is clean, i.e., free of unwanted sideband and extra products. It occupies a band of frequencies 12 kc wide centered at 100 kc.

The next modulation step translates this group of sidebands to a medium frequency, using a conversion frequency of 2,700 kc. The output of the second modulator, Medium-Frequency Modulator, contains the 2,700-kc conversion frequency at an amplitude, which is reduced because of balance, and a sideband centered 100 kc above the conversion frequency, with carrier at 2,800 kc and another sideband 100 kc below the conversion frequency with carrier at 2,600 kc. The upper sideband is selected by a series of electric filters for transmission. Examination of the spectrum of this sideband shows that it consists of the blocks B2, B1, A1, and A2 occupying a 12-kc band centered on the carrier frequency, which is now 2,800 kc. By conservative operation of the modulator and by adequate filtering, this group of sidebands can also be made clean and free of spurious and extra-band products. This signal is now ready for the final modulation.

In the frequency block diagram that we have been following, the final modulator (High-Frequency Modulator) is shown simply as a means for combining the signal group centered at 2,800 kc with a high frequency in the range 6.8 to 20.2 mc. The exact conversion frequency to be used depends on the frequency which is assigned to the final carrier. In the output of the High-Frequency Modulator, the conversion frequency appears at an amplitude which has been reduced by balance. Two sidebands of this modulation also appear, one centered at a frequency 2.8 mc above the conversion frequency and one centered at a frequency 2.8 mc below the conversion frequency. Either of these may be transmitted by simply tuning the circuits of the following amplifier stages to its centered carrier. But it will be observed that the lower sideband is inverted with respect to the upper sideband. In other words, A1 appears below the carrier in the lower sideband group, whereas it appears above the carrier in the upper sideband group.

This makes it necessary for the distant receiver to know which sideband is being transmitted in order to be able to route the intelligence in each band to its proper destination. The easiest rule to follow would be always to transmit the same sideband, either upper or lower, but there are economies to be gained both at the transmitter and at the receiver by changing from one sideband to the other in order to reduce the range of conversion frequency required for the final modulator.

The convention was established in earlier transmitters[4] that for all radiated frequencies less than 10 mc the carrier will appear in the output spectrum above the sidebands of the Group A input, and that for frequencies of 10 mc or above, the carrier will appear below the Group A sidebands. This rule is now observed widely, and therefore any transmitter which is designed to operate in systems where correspondence with various receivers is required should be designed and operated in accordance with the rule. Observance of the rule is effected simply by transmitting the lower sideband of the final modulation for carrier frequencies below 10 mc

[4] A. A. Oswald, "A short-wave single-sideband radiotelephone system," Proc. I.R.E., vol. 26, pp. 1431–1455; December, 1938.

and by transmitting the upper sideband at 10 mc and above.

In recent years more attention has been given to the problem of keeping spurious modulation products from appearing in or near the sideband frequencies which are transmitted. Many spurious products result from the final modulation process, and their frequencies are easily predicted although their amplitudes depend on the order of the product as well as on the operating characteristics of the modulator. A strong, undesired product which falls close to a desired sideband can be avoided, sometimes by transmitting the other sideband. Thus, in order to dodge spurious frequencies it was found that 10 mc is not the best frequency at which to change over from one sideband to another. Instead, a higher frequency such as 18 mc was found to be better. A discussion of spurious products and methods used to avoid them will be given later in this paper.

Nothing in the convention cited above dictates which sideband is to be transmitted. The rule merely defines the relation between groups and the carrier which must exist in the output. In a case where the opposite sideband is chosen, this relation can be re-established by simply interchanging the input lines coming into the transmitter. Such an interchange facility has been built into a recent transmitter design, and is used when the carrier frequency lies between 10 and 18 mc.

Choice of Intermediate Frequencies

In general, sideband selection must be accomplished at frequencies where suitable filters can be built economically. In an earlier paper,[4] a description was given of a system which used 125 kc for the first conversion frequency of the transmitter and 100 kc for the corresponding demodulation in the receiver. This choice of frequencies required two independent designs of the crystal filters. In all later systems, 100 kc has been used in both the transmitter and the receiver to permit economies which naturally result when the same filters are used in both places. This has been carried still further by building into the common filter only the selectivity needed at the transmitter. Additional selectivity for the receiver is then obtained by adding a single filter having a 12-kc pass band.

The choice of the second intermediate frequency and the conversion frequency for the second modulator should be made with consideration of the selectivity of the final frequency amplifier circuits. Ordinarily, the over-all selectivity of these circuits is not the same at all parts of the frequency range. The way that it varies with frequency depends on a number of factors. It is not necessarily a function of the Q of the coils or condensers because the circuits are usually deliberately loaded with shunt resistance. The most probable trend is for selectivity to be greatest at the lowest frequency because of the added capacitance used in tuning to low frequencies and the increase of circuit losses at high frequencies. At the highest assigned frequency (f_h), the selectivity should be sufficient to suppress the conversion frequency which differs from the desired sideband by the intermediate signal frequency (f_m). The requirement depends on the ratio f_m/f_h. At the lowest assigned frequency (f_l), the closest unwanted component is the intermediate signal frequency itself, (f_m). Therefore, the selectivity needed at this frequency depends on the ratio $(f_l-f_m)/f_l$. These two selectivity requirements may be adjusted by selection of the intermediate frequency to be in proportion to the over-all selectivity figures of the circuit, Q_h and Q_l. This gives a relation

$$\frac{Q_l(f_l - f_m)}{f_l} = \frac{Q_h f_m}{f_h},$$

from which the optimum value of intermediate frequency may be obtained:

$$f_m = \frac{f_h \cdot f_l}{f_h + (Q_h/Q_l) f_l}.$$

On the basis of this formula, if uniform selectivity is assumed and the frequency range is taken as 4 to 23 mc, the "optimum" intermediate frequency is found to be 3.4 mc. If the trend of selectivity were such as to make it better at lower frequencies, then a higher value of f_m would be obtained. Actually, there are a number of other factors to be considered in making the choice. For one, if there is to be a tie-in of the second conversion frequency with the first through a series of frequency multipliers or dividers, then the two frequencies will need to be multiples of each other, and that will affect the choice. Also, in order to use the same filters in the receiver, it is necessary to give consideration to the receiver problems. Furthermore, it will be seen from later considerations that the choice of the value of f_m also affects the frequency distribution of spurious modulation products, and is thus important in maintaining discrimination against these.

In the system which was described using the block schematic, Fig. 1, the frequency (f_m) was 2.8 mc. Later a chart of spurious products will be shown. It will be found by those who choose to work it out that if the value 3.4 mc, which was obtained from the formula above, were used there would have to be changes elsewhere to avoid unwanted products in the output.

Low-Frequency Modulators

An earlier transmitter[4] used vacuum tubes in a balanced circuit for the first modulator. A trimmer condenser was supplied to compensate for tube capacitance differences and a differential screen bias adjustment to compensate for inequality of transconductance. In spite of these adjustment possibilities, it has been difficult to maintain a balance in this first stage where the carrier suppression is effected. This is the common fault of tube modulators.

In carrier telephone service, where tube modulators were employed earlier, the copper oxide varistor has been found to possess a stability which makes it superior to tubes in balanced-modulator circuits. The reader is referred to a paper which gives an excellent discussion of "Copper Oxide Modulators in Carrier Telephone Systems."[5] The similarity between radio-telephone and carrier-telephone systems in respect to methods of single-sideband generation makes it natural to adopt copper oxide varistors for use in the first modulation stage of a modern transmitter. Consideration must be given, however, to the fact that a varistor modulator introduces a loss in the signal path, while a tube modulator may introduce a gain. Stability, however, is worth this price, and the gain may be made up by amplification elsewhere.

The first modulators of the generator system shown in Fig. 1 are of the ring type in which each arm of the ring contains 16 copper-oxide discs connected in series parallel. By this means the average characteristics which result for each arm may be very well matched and a good balance (about 45 db) results.

In order to ascertain the degree of balance required for the copper-oxide modulators, the example given in Fig. 2 will be considered. This is a simplified circuit of a balanced, ring-type modulator on which the applied

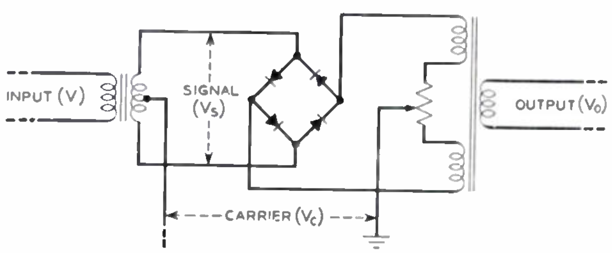

Fig. 2—Simplified circuit of a varistor-type low-frequency modulator.

signal voltage (V_s) and carrier voltage (V_c) have been indicated. In order to maintain a satisfactory ratio of signal-to-distortion in the output voltage (V_o) of such a modulator, it has been found necessary to keep the ratio of carrier voltage to signal voltage (V_c/V_s) high. In other words, the percentage modulation must be low. On each arm of the ring, this ratio is $2V_c/V_s$. For a signal-to-distortion ratio of about 50 db, this ratio should have a value of about 15. Across the output terminals the voltage of each sideband appearing in V_o will be lower than the signal applied, (V_s). This "loss" in the modulator is assumed to be 6 db, which means that the sideband voltage is $V_s/2$. From this figure we can establish a requirement on the degree of balance required to keep the carrier leak at a satisfactorily low value. The carrier will later be resupplied at an amplitude 20 db below the sideband reference amplitude. In order that the carrier leak will have a negligible effect in determining this amplitude, the leak should be about 40 db below reference. Thus, if the reference is $V_s/2$ as found, the carrier leak voltage must be not more than $V_s/200$.

[5] R. S. Caruthers, "Copper oxide modulators in carrier telephone systems," *Bell. Sys. Tech. Jour.*, vol. XVIII, pp. 315–338; April, 1939.

The degree of balance against carrier is normally stated as the power ratio of the applied carrier to the carrier leak, and is expressed in db. If the carrier voltage source sees an impedance Z_c in the varistor, and carrier leak appears as part of V_o across an impedance Z_o, the degree of balance may be expressed as

$$\text{balance} = 10 \log_{10}\left[\frac{W_{\text{applied}}}{W_{\text{leak}}}\right]$$

$$= 10 \log_{10}\left[\frac{4 \times 10^4 V_c^2 Z_o}{V_s^2 Z_c}\right].$$

Using figures found in a recent design, this gives a requirement on balance of 71 db. Even with the statistical balance obtained by use of multiple element varistors and great care in the manufacture of coils, this figure cannot be obtained without adjustment. For this reason the ground point is made adjustable by means of a small potentiometer in series at the center of the output coil as shown in Fig. 2. Because of the stability of copper-oxide varistors, the balance, once adjusted, holds for long periods without readjustment.

Medium- and High-Frequency Modulators

In recent years, varistors of germanium have been found to have characteristics which make them useful as high-frequency modulators. In balanced circuits, their stability has been found to be greater than that of the best tubes, with the result that a useful degree of balance can be established and maintained over long periods of time. This is a valuable aid in the suppression of unwanted modulation products.

Two modulator circuit arrangements, which have been used successfully with germanium varistors in the second or third modulation stages of single-sideband transmitters, are shown in Figs. 3 and 4. These are adaptations of two of the circuit types which are described and analyzed in the paper on copper-oxide modulators which has already been cited.[5] Both are balanced and have the same properties in regard to suppression of certain classes of modulation products at the output. Under conditions of perfect balance, only odd multiples of the signal frequency v are involved in the expression for the frequencies of products which are impressed on the output filter. This is indicated by the subscript of n_o, which therefore assumes only the values 1, 3, 5, 7, and so on. The conversion frequency c, however, is involved in all of its multiples m, where m may assume any integral value, or may be zero. It can be seen that the output will contain the signal frequency v which must be filtered off, but will not contain the harmonic $2v$ because n_o has only odd values. Also, the output will not contain the conversion frequency c because n may not be zero. Thus, the conversion frequency and the signal harmonic, both important products, are suppressed by the balance. The selectivity requirements of the output filter are simply that it should pass the desired sideband and suppress all other components which are delivered to its input terminals.

Operation of these modulator circuits is also dependent on the impedances which the filters introduce into the modulator circuit. For example, in Fig. 3 the input and output filters are seen to be effectively in parallel.

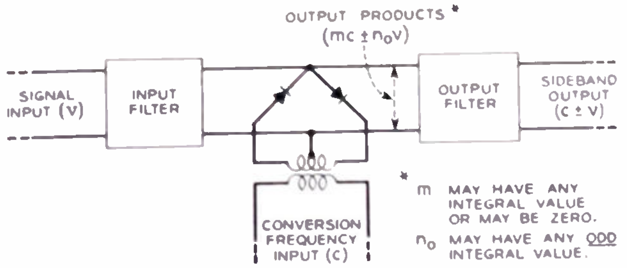

Fig. 3—Simplified circuit of a parallel-type, varistor modulator for high frequencies.

To assure deliverance of the incoming signal energy to the varistors and not to the load, the input filter should match the varistor impedance and the output filter should appear as a high shunt impedance to the signal frequency v. Following the same line of reasoning, at the desired sideband frequency $(c \pm v)$, the impedance of the output filter should match the varistor impedance while the input filter should appear as a high shunt impedance.

By similar reasoning, the requirements for the filters in the configuration of Fig. 4 will be seen to be different.

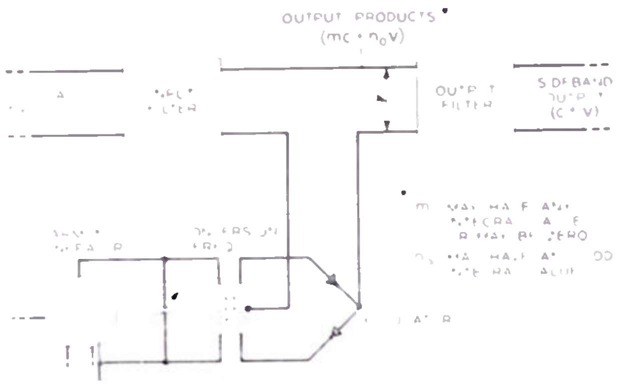

Fig. 4—Simplified circuit of a series-type, varistor modulator for high frequencies.

Here the filters are effectively in series instead of shunt. Therefore, the input filter impedance again must match the impedance of the varistors at input frequencies, but the output filter is in series, and must therefore present low impedance to input frequencies. In the same way, the output filter impedance must match the impedance of the varistors at the output frequencies, but the impedance of the input filter must be low at these frequencies. A failure to satisfy these impedance requirements can result in transmission loss, the loss being dependent in the usual way on the degree of mismatch and the amount of energy absorbed in an unwanted manner or diverted into wrong channels. The over-all loss measured in terms of the decibel difference between signal input energy and sideband output energy is termed "conversion loss." Using a modulator of the type shown in Fig. 4, conversion losses ranging from 6 to 11 db have been measured in the range from 4 to 23 mc.

Spurious Modulation Products

Let us consider once more a high-frequency modulator of the type shown in Fig. 4. When a conversion frequency c is combined with a signal frequency v, the output contains a whole series of frequencies symbolized by the formula $(mc \pm n_o v)$, where m may take on any integral value including zero while the subscript o is used to show that n_o is restricted to odd integers. If the frequency sources supplying the modulator were perfect, these would be the only sidebands present, but the conversion frequency source in practice may contain harmonics and/or subharmonics in considerable strength. This is illustrated in Fig. 4, where the conversion frequency is seen to be obtained from a tuned transformer in the output of a harmonic generator stage. The tuned circuit is not likely to have a Q better than 50 because of the loading introduced by the modulator. This means that if the second harmonic is being selected the attenuation of the fundamental will only be about 37 db while the attenuation of third harmonic will be only about 32 db. From an analysis which was made of the current of a typical harmonic generator tube, the fundamental was observed to be only 4 db stronger than the second harmonic, while the third harmonic was 10 db weaker than the second harmonic. According to these figures then, the energy supplied to the modulator at the desired second harmonic would be only about 33 db greater than the undesired fundamental energy and 42 db greater than the undesired third harmonic energy. Therefore, sidebands of these two spurious conversion frequencies will also appear. In the case of the fundamental, the frequency is $c/2$, while in the case of the third harmonic, the frequency is $3c/2$. So, products whose frequencies are $[(mc/2) \pm n_o v]$ and $[(3mc/2) \pm n_o v]$ must also be considered unless the output filter of the modulator will give enough additional filtering to make these products tolerably weak.

In practice, it is well to chart all the significant products to determine where they will fall in the spectrum. Then by carefully choosing the point where the changeover is made from lower to upper sideband and by being careful also of the effect of harmonics and subharmonics of the conversion frequency, it is possible to keep the significant products outside of the band of transmission which is defined by the discrimination of the amplifier circuits.

The transmission band of an amplifier depends greatly on the number of tuned circuits which are effective in the transmission path as well as on the selectivity of the individual tuned circuits. In the case of a transmitter of only four channels, the amplifier bandwidth is likely to be at least ten times the bandwidth occupied by the transmitted signal. Measurements on a transmitter of recent design indicated a bandwidth in the amplifier as great as 0.5 mc at some frequencies. This means that unwanted products must be kept from appearing within this band unless they are of negligible amplitude.

The chart shown in Fig. 5 applies to a recent transmitter in which the signal frequency v is 2.8 mc. The chart is plotted with output frequency of the final carrier as abscissa. The vertical scale is also frequency in megacycles, and on it are plotted all of the significant frequencies, spurious or desired, which may exist for each output carrier frequency. The desired output is plotted as a 45-degree straight line designated $(c-v)$ at the lower frequencies where lower sideband is used and designated $(c+v)$ at the higher frequencies where upper sideband is used. The conversion frequency c is also plotted as a 45-degree line which is displaced 2.8 mc upward in the region below 18-mc output, and a second line which is displaced 2.8 mc downward for output frequencies, above 18 mc.

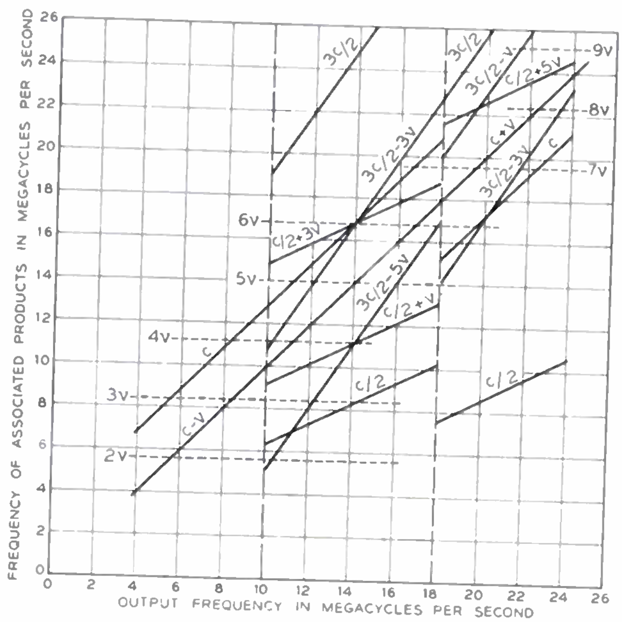

Fig. 5—Chart of spurious and desired frequencies in the output of a high-frequency modulator.

Output frequencies below 10 mc are relatively free from spurious frequencies except for harmonics of the signal input v. The second harmonic $2v$ is seen to coincide with the desired sideband at 5.6 mc, and the third harmonic $3v$ coincides at 8.4 mc. In these regions, then, these harmonics will appear as spurious frequencies. They must be kept low enough in amplitude so that they will not cause harmful interference. The second harmonic is reduced by the balance of the modulator, and this balance is therefore important. The third harmonic is not reduced by balance, and can be maintained at a low amplitude only by care in the operation of the modulator to keep the conversion frequency amplitude great compared with the signal.

Above 10 mc and below 18 mc the lower sideband is still being used, but the conversion supply is now the second harmonic of the crystal. Therefore, the fundamental of the crystal as well as its third harmonic contribute sidebands such as $(c/2+v)$, $(c/2+3v)$, $(3c/2-3v)$, and $(3c/2-5v)$. The first of these is the most fearsome because its amplitude, according to the estimate made earlier in this paper, is likely to be only 33 db lower than the desired sideband. It lies closest to the transmitted sideband at the output frequency 10 mc, where it is seen to be only 0.8 mc away. The others are successively weaker in the order named because they involve higher multiples of v. If the product $3v$ is tolerable, as it must be to avoid trouble near 8.4 mc, then $(c/2+3v)$ will probably be still weaker because it is a higher-order product. Nevertheless, conflict with this product is completely avoided by changing to upper sideband at 18 mc before the line $(c/2+3v)$ intersects the line $(c-v)$.

Above 18 mc, the desired sideband is $(c+v)$ which is again the 45-degree line. Products which fall nearby are seen to be $(3c/2-v)$, $(3c/2-3v)$, and $(c/2+5v)$. Conflict with the first was avoided by use of lower sideband below 18 mc. Conflict with the second will have to be tolerated near the frequency which is obtained from the equation

$$3c/2 - 3v = c + v$$

or

$$c = 8v = 22.4 \text{ mc}$$

whence

$$c + v = 25.2 \text{ mc.}$$

By the reasoning that was applied for the product $(3c/2+3v)$ a little earlier in this discussion, it is believed that this product can also be tolerated; but if measurements or experience show it to be troublesome at 25.2 mc, it will have to be dodged. This could be done by turning to the use of the third harmonic for the supply of conversion frequency. The same is true of the product $(c/2+5v)$.

Most of the worry about harmonics and subharmonics in the conversion frequency supply could be relieved by adding an amplifier stage between the harmonic generator and modulator. This would provide the additional discrimination of another tuned circuit, but it has the disadvantage of being just one more circuit to tune each time the frequency of the transmitter is changed.

Testing for Spurious Outputs

Even after a careful design in which spurious outputs have been charted and avoided by the best means known, a test will be desired to prove that those which must be transmitted are really of insignificant magnitude. From the chart and the discussion given above, it appears that there are certain points at which the lines representing spurious products would, if extended, intersect the line representing the desired sideband. The frequencies at which they make the nearest approach to the line of the desired output sideband are points at which a measurement of the spurious output should be made. In general, the crossover points correspond to odd harmonics of the signal input to the final modulator. Even harmonics would also be found if the modulator balance were inadequate.

Frequency Stability

In the modulation system which has been described in this paper, three separate quartz-controlled oscillators were employed. The over-all frequency stability obtainable in this way will now be discussed.

The maximum over-all frequency departure contains the effects of deviations in all three oscillator frequencies. If f_1, f_2, and f_3 are the conversion frequencies obtained from the three oscillators, and t_1, t_2 and t_3 are the tolerance factors of the quartz plates of the three oscillators, then the maximum frequency departure of the output carrier is

$$\text{max frequency departure} = \frac{t_1f_1 + t_2f_2 + t_3f_3}{\text{output carrier freq}}.$$

Studies have shown that this departure figure is greatest at the lowest transmitted carrier frequency. Assuming $f_1 = 100$ kc, $f_2 = 2.7$ mc, and $f_3 = 6.8$ mc, while $t_1 = 0.007$ per cent, $t_2 = 0.001$ per cent and $t_3 = 0.001$ per cent, and letting the final carrier frequency be 4 mc, we find that

$$\text{max departure} = \frac{7 + 27 + 68}{4 \times 10^6} = 25.5 \times 10^{-6}$$

or

$$= 0.00255 \text{ per cent.}$$

Obviously, it is the high-frequency oscillator whose stability is the most important because it contributes most of the departure. The improvement to be gained by greater stability of the first oscillator is small.

Conclusion

It can be seen from the foregoing discussion that the design of single-sideband generators involves a great many considerations. The existing practices and the receivers in the field must be considered. The availability of component parts and the economic factors in the design of such things as varistors, filters, and transformers affect the choice of circuits and the details of fitting them into a system. Experience over many years, by a great number of people, has had immeasurable influence on the present art. The author wishes to express his appreciation to numerous colleagues who have contributed to the collection of ideas presented here.

Single-Sideband Transmission by Envelope Elimination and Restoration*

LEONARD R. KAHN†, MEMBER, IRE

Summary—**A new type of single-sideband transmitter is described which does not require the use of linear radio-frequency amplifiers. Amplification is accomplished by a process in which the phase-modulation component of the single-sideband wave is amplified by means of Class-C amplifiers, and the amplitude envelope is restored at the final amplifier. Experimental results show performance equal to or better than conventional linear radio-frequency amplifier practices. The over-all efficiency is approximately the same as that of a double-sideband amplitude-modulated transmitter. This system is especially suitable for high-power operation.**

Introduction

THIS PAPER DISCUSSES a new method of amplifying single-sideband signals and other forms of hybrid modulated waves.[1] The power and spectrum efficiency of single-sideband transmission is well known, and the reader has many fine papers[2,3] at his disposal. The main problem associated with single-sideband operation is the complexity and cost of the equipment involved.

The conventional system generates the desired single-sideband wave at very low power levels (a few watts) and amplifies this wave in a series of cascaded linear rf amplifiers. Linear rf amplifiers have comparatively low efficiency and are critical in adjustment. Due to their low efficiency, the cost of these transmitters is quite high.

Another characteristic of the conventional system is that each additional amplifier introduces its own distortion and attendant spurious output. Therefore, the higher the power output desired, the more difficult it is to maintain low spurious output.

There have been a few alternate systems proposed[4,5] which do not require linear radio-frequency amplifiers. These systems, however, require wide-band phase-rotation networks. The system to be proposed is a different approach to the problem.

The advantages of the proposed system are:

(a) High efficiency is obtained (comparable with standard double-sideband amplitude-modulation transmitters).

(b) Conventional telephone transmitters can be used for single-sideband operation. Auxiliary equipment and its installation are relatively inexpensive.

(c) Distortion and its accompanying spurious fre-

* Decimal classification: R423.5. Original manuscript received by the Institute, August 31, 1951; revised manuscript received, January 16, 1952. This work was done while the author was employed by RCA Communications, Inc., New York, N. Y.

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[1] A hybrid modulated wave has both amplitude and angular velocity modulation components. Examples of hybrid modulated waves are: carrier-suppressed-double-sideband, single-sideband, vestigial sideband, and quadrature-modulated.

[2] A. H. Reeves, "Single-sideband system applied to short wave telephone links," *Jour. IEE* (London), vol. 73, pp. 245–279; September, 1933.

[3] F. A. Polkinghorn and N. F. Schlaack, "A single-sideband short wave system for transatlantic telephony," Proc. I.R.E., vol. 23, pp. 701–718; July, 1935.

[4] E. S. Purington, U. S. Patent No. 1,994,048; filed September 6, 1930.

[5] O. G. Villard, Jr., "Composite amplitude and phase modulation," *Electronics*, vol. 21, pp. 86–89; November, 1948.

quency generation is independent of the transmitted power level. This is so because linear radio-frequency amplifiers are not required; therefore, the difficulty of maintaining low spurious output is not multiplied by each additional stage.

Before describing the new system, it may prove helpful to examine the structure of a single-sideband wave as shown on Fig. 1.

Fig. 1(a) shows a spectrum figure of a single-sideband wave having equal carrier and sideband amplitudes. The upper sideband is shown selected in this figure and it represents a signal frequency of 600 cycles.

Fig. 1(b) is the vector representation of the single-sideband wave shown in Fig. 1(a). The carrier frequency is the reference vector. The sideband revolves past the reference vector at a velocity equal to the tonal frequency of the signal. The resultant of the two frequency components varies both in amplitude and angular velocity. *Therefore, a single-sideband wave has both amplitude- and phase-modulation components.*

Fig. 1(c) shows the amplitude-modulation envelope of the wave. This envelope is identical with the wave shape derived from a full-wave resistance-loaded rectifier when fed a sine wave.

Figs. 1(a), 1(b), and 1(c) would also be obtained if a suppressed carrier single-sideband signal, modulated by two equal tones, was considered.

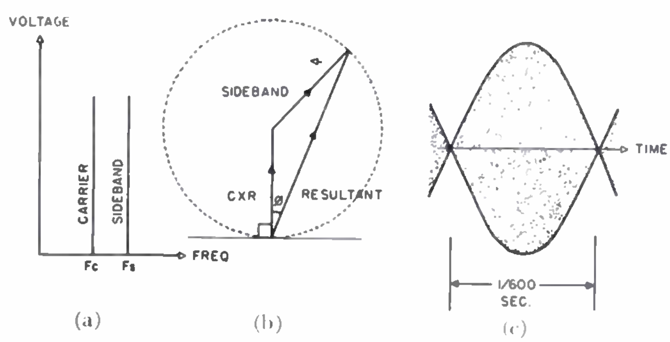

Fig. 1—(a) Spectrum diagram of single-sideband wave. (b) Revolving vector representation of single-sideband wave. (c) Envelope waveform of single-sideband wave.

Basic Concepts of System

The steps of the proposed system are:

(a) A portion of the single-sideband wave to be amplified is limited, thereby producing a pure phase-modulation signal.

(b) This phase-modulated wave is amplified to the desired output level by efficient, noncritical Class-C amplifiers.

(c) A portion of the original single-sideband wave amplitude envelope is detected and the resultant audio-frequency wave is amplified.

(d) The amplified detected, amplitude-modulated component remodulates the amplified phase-modulation component, resulting in an amplified copy of the original single-sideband wave.

Fig. 2 is a simplified block diagram which will be used to describe the fundamentals of this system. Waveforms at various points in the diagram are given in order to make this explanation more easily understood. Let us consider the case where two equal amplitude tones are fed to the input of the "single-sideband generator"[6] and the carrier is suppressed. The "single-sideband generator" produces a low-power (a few watts), single-sideband signal identical with the wave shown in Figs. 1(a), 1(b), and 1(c). This signal is represented at point *a* by a wave shape having an envelope corresponding to a full-wave rectified sine wave. This low-power single-

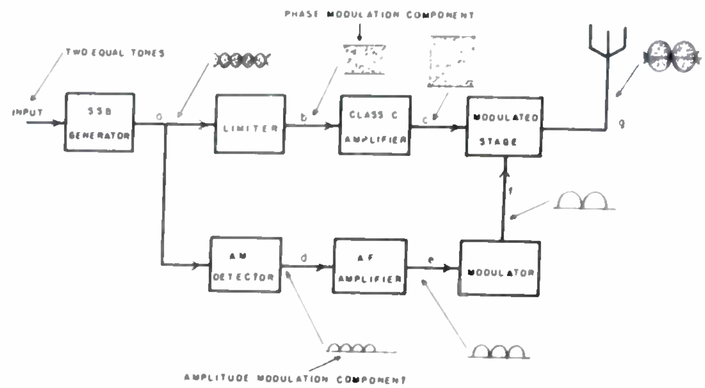

Fig. 2—Simplified block diagram of system.

sideband signal is fed to the "limiter." The "limiter" removes the amplitude-modulation component from the wave, so that the output is a pure phase-modulated wave. This phase-modulated signal can be conveniently amplified by the highly efficient "class-C amplifier." The "class-C amplifier" provides the phase-modulated driving power for the "modulated stage."

The amplitude-modulation component is detected and isolated from the phase-modulation component by the "AM detector." The output of the "AM detector" corresponds to the envelope wave shape of the single-sideband wave at point *a*; therefore, the wave shape at point *d* is the same as a full-wave rectified sine wave. This audio-frequency wave is amplified by the "af amplifier." The "modulator" modulates the phase-modulated signal in the "modulated stage." Hence, the envelope wave shape of the signal at point *g* is the same as the envelope wave shape at point *a*. Furthermore, the phase-modulation component at point *g* is identical with the phase-modulation component at point *a*. If time relationships between the phase- and amplitude-modulation components are properly maintained, the signal at point *g* will be a pure, high-power, single-sideband wave.

Practical Embodiment of System

Fig. 3 shows a possible working system. The additional blocks indicate circuitry required to produce a practical system from the theoretical system in Fig. 2.

[6] The following phrases in quotes refer to blocks of the referenced diagram.

The blocks marked "xtal osc" and "mixer" are used to change the frequency of the "ssb generator" into any required output frequency. Mixers should be carefully designed to maintain low intermodulation distortion.

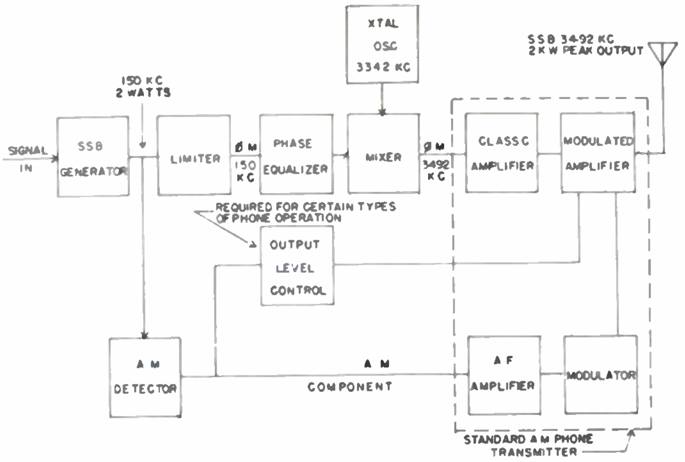

Fig. 3—Practical block diagram of system.

The "phase equalizer" is required to equalize the time delay between the phase-modulation channel and the amplitude-modulation channel. When the "phase equalizer" is correctly adjusted, the amplitude-modulation and the phase-modulation components retain the original phase relationship of the single-sideband wave.

The "output level control" is required in transmitting variable average amplitude signals. Since frequency-shifted, frequency-division multiplex signals are of constant amplitude, the "output level-control" circuit is not required. However, voice waves are not of constant average amplitude and the "output level control" will be required in this form of transmission. The "output level control" may be a series of dc amplifiers which control one of the electrode voltages of the "modulated stage."

There is, however, a different approach to this variable average amplitude problem which does not require the "output level control." This system would transform a single-sideband voice wave into a constant average amplitude wave, by varying the amount of suppression of the carrier. During silent periods the transmitter would produce large carrier levels. A squelch-operated carrier channel could be used in the receiving system to take advantage of these large-carrier level periods. In this manner, improved afc and avc operation may be obtained, and the entire receiver can be made less susceptible to jamming and other forms of interference.

Modulator Requirements

In order to maintain good spurious-radiation suppression, it is necessary that the "modulator" used have certain qualities. The appendix shows a method for determining the modulator requirements. Two answers to the problem are: For 35 db or more spurious-radiation suppression, the modulator must have sufficient high-frequency response to amplify the third harmonic of the difference between the highest-frequency and lowest-frequency appreciable energy signals; for 25-db spurious-output suppression, only the second harmonic is required.

Experimental Results

The quality of a single-sideband signal is dependent upon the amount of spurious output. The specification which is used by many companies is based on the amplitude of the third-order intermodulation distortion component. The ordinary test procedure is to feed two equal amplitude test tones into the transmitter, and then measure the amplitude of the spurious component $2f_1-f_2$ (third-order intermodulation product). The amplitude of this spurious component should be at least 25 db below one of the signal tones after the composite signal is detected in a linear detector.

Experiments were conducted at the Rocky Point Transmitter Laboratory, RCA Laboratories Division, with a one-kw telephone transmitter. A 2.5-kw peak-power single-sideband wave was produced utilizing the proposed system. The undesired sideband was reduced to 25 db below the desired test signal. These results meet current industry standards. At times, better than 30 db of spurious-radiation suppression were measured; and it appears that by careful design 35 db can be obtained. No negative feedback was used in these tests, and further improvement is possible by adding this expedient.

Acknowledgment

The author acknowledges his obligations to J. L. Finch of RCA Communications for guidance and encouragement; also to Nils Lindenblad who supervised the work done at Rocky Point. Others, whose help is appreciated are H. E. Goldstine, Walter Lyons, William Miller, and George Smith. L. O. Goldstone made suggestions in the preparation for this paper and a parallel thesis prepared for the Polytechnic Institute of Brooklyn.

The author is indebted to C. W. Latimer, H. H. Beverage, and C. W. Hansell for giving him the opportunity to carry out this project. The author also wishes to thank M. G. Crosby for his help in preparing this paper.

Appendix

This analysis will attempt to answer the following questions: (1) What are the power relationships of the system? (2) What is the frequency response requirement of the modulator for a specified allowable spurious output?

In order to answer these questions, *the severe case of a single-sideband, carrier-suppressed transmitter with two equal amplitude tones applied will be considered.* The analysis will be based upon the fact that the AM envelope of this wave has a full-wave rectified sine-wave form.

Power Relationships

Consider that this two-tone carrier-suppressed single-sideband wave has a peak voltage of $2E$ and that it is applied to a load resistance R. The peak power is $4E^2/R$. Single-sideband transmitters are rated according to their peak-power output.

The power corresponding to average amplitude of the wave is 1.62 E^2/R or $(2/\pi)^2 \times 4E^2/R$. This average value is the same as the amount of dc power obtained from a full-wave rectifier in the analogous case. If the amplitude modulation were removed from the single-sideband wave, this would be the power remaining. Thus, it is the power in the phase-modulation component of the single-sideband wave. *The ratio of peak power to average amplitude power (phase-modulation component) is* 4/1.62 *or* 2.47. In comparison, a 100-per cent modulated amplitude-modulation wave has a ratio of 4. If a 2-kw peak-power single-sideband signal is desired, the radio-frequency stages of the transmitter must supply 810 watts of phase-modulated power.

The total or effective power is the sum of the power in the phase-modulation and amplitude-modulation components. In the analogous case of a rectified sine wave, the dc power is equivalent to the phase modulation, and the ripple currents represent the amplitude modulation of the single-sideband wave. The effective power is $(0.707)^2 \times 4E^2/R$ or $2E^2/R$ watts. This is the power that a load resistor R would be required to dissipate.

The total or effective power is $2E^2/R$, and the average amplitude derived from the radio-frequency section of the transmitter is 1.62 E^2/R. Therefore, the difference power of 0.38 E^2/R must be supplied by the amplitude modulator. *The ratio of peak power to amplitude-modulation component power is* 4/.38 *or* 10.5. For a 2-kw peak-power single-sideband transmitter, the modulator must supply 190 watts of power to the output signal.

Modulator Frequency Requirements

To determine approximately the frequency-response requirements of the amplitude modulator, use will be made of the Fourier series expansion of the envelope wave shape of the single-sideband signal. Tabulation will be given of the power in each amplitude-modulation component harmonic. The amount of power in all the higher amplitude-modulation component frequencies that are beyond the response of the modulator will be totalled. Most of this power represents the spurious content of the signal.

If this "above modulator response power" figure is used as a rating of spurious output, an approximate rating on the pessimistic side will be obtained. The rating is pessimistic because a portion of this power is allocated to the desired output component instead of to the spurious output. Furthermore, the amount of "above modulator response power" is divided between a number of spurious components, whereas present ratings specify the power in a single spurious component. The situation is analogous to the measuring of total harmonic distortion when a specification calls for a certain maximum second-harmonic distortion figure.

The following equation is the expansion of the envelope of a two-tone carrier suppressed single-sideband signal:

$$e = 0.636E(1 + 0.667 \cos Wt - 0.133 \cos 2Wt + 0.057 \cos 3Wt - 0.032 \cos 4Wt \cdots),$$

where W is 2π times the difference frequency of the two applied tones.

Realizing that the unit dc term (average amplitude value) is equivalent to the phase-modulation component of the wave, and that the cos Wt term gives the peak value of the fundamental amplitude-modulation component in the above equation, the following table can be constructed (in which a 2-kw peak single-sideband signal is considered).

The first column is a function of the fidelity of the amplitude-modulation path and of the separation of the two tones being transmitted. For example, an amplitude modulator equalized for phase and amplitude up to 15 kc will respond to the fifth harmonic of a 3-kc tone separation.

The second and third columns were calculated by summating the power in the harmonics that were too high in frequency for the modulator to pass. Since most of this "above modulator response power" would have been used to neutralize the spurious signals, it is evident that this column gives the maximum possible total spurious-power output.

Highest Harmonic Modulator Equalized For	Power Due to All Harmonics Above Modulator Response	
	Absolute Power	DB Relative to PM Component
cos $0Wt$ (dc) PM component	190 watts	− 6.3 db
cos Wt	9.16 "	−18.9 "
cos $2Wt$	1.94 "	−26.2 "
cos $3Wt$	0.61 "	−31.2 "
cos $4Wt$	0.20 "	−36.0 "
cos $5Wt$	0.04 "	−42.0 "
cos $6Wt$		

The chart gives a somewhat *pessimistic* figure for spurious amplitude as has been previously pointed out.

Because practical modulators do not faithfully reproduce frequencies up to a certain point and then reject all higher frequencies, it becomes a very difficult problem to determine the spurious content of the signal. If it is necessary to calculate the exact spurious content, the designer must know the phase and amplitude characteristics of the transmitter and associated equalizers. With this information, he can work out graphically or analytically the sideband distribution of the phase-modulation component, and then vectorially add the sideband distribution caused by the reinsertion of the amplitude-modulation component.[5]

Direct-Reading Frequency Measuring Equipment for the Range of 30 CPS to 30 MC*

L. R. M. VOS DE WAEL†

Summary—A description is given of a direct-reading frequency-measuring equipment which enables precise frequency measurements between 30 cps and 30 mc to be carried out in one second.

The measuring device itself consists of an electronic counter ranging from 30 cps ÷ 1 mc, an instrument giving multiples of 1 mc in the range from 1 ÷ 29 mc and a combining part. The whole has only one knob, namely, that for choosing the desired multiple of 1 mc.

The result of a measurement is indicated on the electronic counter and on the knob.

When taking a series of measurements, the results are printed on a normal page printer, one measurement every two seconds. If desired, a continuous recording instrument may be used for the registration of every two out of the six figures of electronic counter.

The accuracy when making a measurement in one second equals the precision of the second impulse derived from the frequency standard (10^{-7} to 10^{-8}) plus or minus 1 cps.

To enable measurements on far-off telephone or telegraph transmitters to be made, the equipment is completed with a receiver, ranging from 5 ÷ 30 mc. In this case the accuracy is somewhat less. The maximum deviation is, however, always below plur or minus 5 cps.

I. Introduction

A. Some General Points of View

THE NUMBER OF radio telephony and telegraphy transmitters has very strongly increased during the last decade, owing to the great extension of the traffic.

The consequence of this increase is that each transmitter must work as precisely as possible in the frequencies assigned to it and that the tolerances permissible for deviation from the nominal value have regularly been diminished in agreement with the other countries. In order to ascertain that a transmitter is working in the frequency assigned to it, frequency-measuring equipment is needed, adapted to measure quickly, in a simple manner, and with great precision, the frequencies in which the transmitters work. Such measuring equipment is also needed (e.g., in a radio laboratory) for the precise determination of the frequency of generators.

Most frequency-measuring equipment is based on the principle of mixing the harmonics of a standard frequency with the unknown frequency. Fig. 1 shows a block diagram of a usual arrangement. A standard frequency of 100 kc synchronizes a multivibrator on 10 kc. The frequency spectrum thus generated, consisting of multiples of 10 kc, is supplied to a modulator, together with the frequency to be measured. The harmonic of 10 kc, which is nearest to the unknown frequency, forms a difference frequency lower than 5 kc in the modulator.

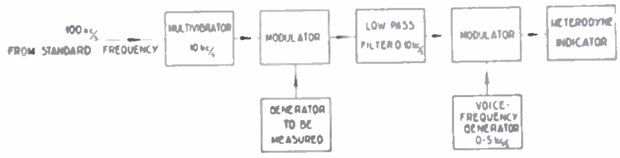

Fig. 1—Block diagram of a usual frequency measuring device.

This frequency is measured with a precisely calibrated voice-frequency generator by adjusting the latter, by means of a beat indicator, to zero beats with regard to the above-mentioned difference frequency. The unknown frequency is then equal to 10^4 times the number of order of the 10-kc harmonic used, increased or decreased (according as this harmonic lies below or above the frequency to be measured) by the frequency indicated by the voice-frequency generator. By somewhat altering the frequency to be measured, it can be ascertained whether the harmonic of the 10 kc is above or below the frequency to be measured.

If one wishes to measure the frequency of a transmitter instead of that of a generator, it is necessary to use a receiver. When telegraphy transmitters are being measured, some difficulties occur, especially if there is fading. These difficulties can be solved by using a so-called "take-over generator." This is a local oscillator whose frequency is equalized as precisely as possible to the frequency of the transmitter to be measured, after which the frequency of this take-over generator is determined by the frequency meter in the manner described. The extent to which the frequency of the take-over generator can be equalized to that of the transmitter has a direct influence on the measuring precision to be reached.

For the measuring range of, for example, 5 to 30 mc, the frequency range of the take-over generator is sometimes limited to the range of 1 to 2 mc, in which case an harmonic of the take-over generator is equalized to the transmitter frequency. This has the advantage that, to determine the frequency of the take-over generator, the necessary number of order of the harmonic of the 10-kc standard frequency need not be very high. On the other hand, the number of order of the take-over generator harmonic that has been used must be known.

The measuring method described allows for several variants and refinements by which the measuring precision may be augmented. In general, this involves a greater number of control and adjustment knobs. For a

* Decimal classification: R213.1. Original manuscript received by the Institute, July 26, 1951; revised manuscript received, February 29, 1952.

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correct interpretation of the various possibilities, well-trained personnel is required.

B. Objections to the Indicated Measuring Method

The objections to the measuring method described above follow:

1. When determining the number of order of the 10-kc standard-frequency harmonic that has been used, mistakes may be made (e.g., because the calibration of the local generator identifying the number of order of the 10-kc harmonic used is no longer valid, or because the auxiliary calculations are in error), especially if the frequency to be measured is not approximately known and a high number of order of the harmonic is necessary.
2. To obtain precise measurements the calibration of the voice-frequency generator must be checked regularly by means of the harmonics of a low standard frequency.
3. For the ultimate determination of the frequency, a few, though simple, computations must be made; the result cannot be read immediately.
4. With telegraphy transmitters especially, and if there is deep fading, it is difficult to adjust the take-over generator precisely to the transmitter frequency. This becomes even more difficult if the transmitter frequency is not constant. The precision to be reached is directly influenced by this.
5. A comparatively great number of knobs must be adjusted and regulated when a measurement is being made.

C. Requirements for New Measuring Equipment

In designing new frequency-measuring equipment, the following requirements should be met:

1. A measuring range of about 30 cps to 30 mc.
2. A measuring precision equal to the precision of the used frequency standard ± 5 cps or better.
3. Direct readability of results without auxiliary computations; preferably, it must be possible to print these results on paper.
4. Avoidance, as much as possible, of any equivocality when the measuring equipment is being adjusted.
5. Simplicity in operation, with as few control and adjustment knobs as possible.
6. Great measuring velocity, so that in a short time a great number of measurements can be made with great precision.

II. Principles of the Measuring Equipment

Frequencies up to 1 mc are measured by means of an electronic decimal counter, which directly indicates every sinusoidal frequency of about 20 cps to 1 mc with a precision equal to that of the second impulses derived from the frequency standard (i.e., 10^{-7} to 10^{-8}, depending on the precision of the frequency standard used and method of derivation of the second impulses) ± 1 cps.

The measurement with the decimal counter lasts only 1 second. The measuring, therefore, is not only effected very rapidly, but moreover a "momentary" frequency is obtained.

To obtain a correct measurement when measuring transmitters, the frequency of the take-over generator need be equal to the transmitter frequency only for a short time. This is important if the reception is weak or disturbed. If the measurement lasts 10 seconds, the precision of the decimal counter becomes ± 0.1 cps instead of ± 1 cps. Frequencies higher than 1 mc are transformed to a frequency lower than 1 mc. To effect this, standard frequencies being a multiple of 1 mc are required. They are obtained with the aid of an "harmonic selector."

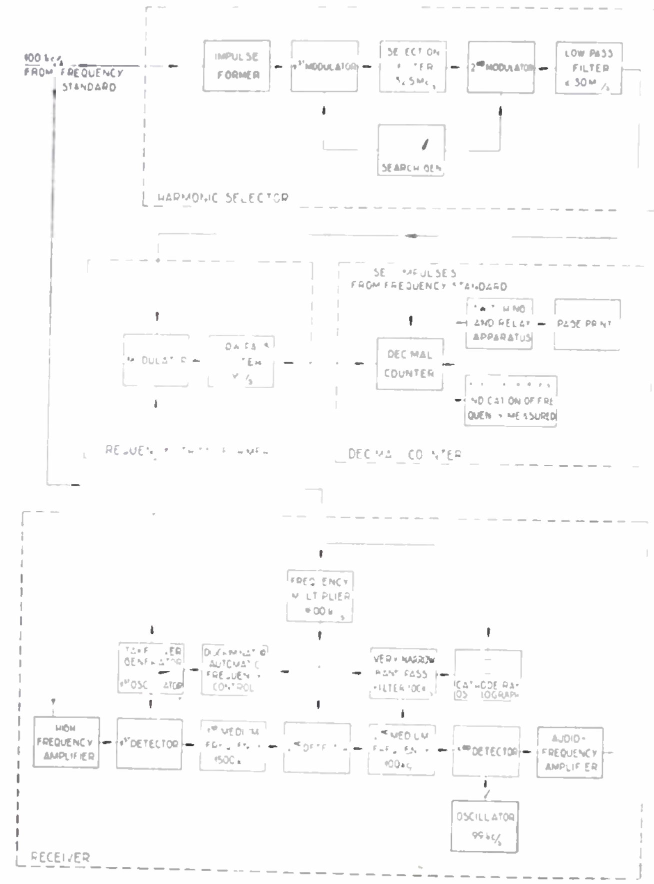

Fig. 2—Block diagram of the complete frequency measuring equipment.

With regard to the required equipment, a distinction should be made between the measuring of the frequency of a local generator (e.g., in a laboratory) and the measuring of the frequency of transmitters (e.g., in a frequency-measuring or control center). In the latter case, the frequency-measuring equipment, consisting of a harmonic selector, a frequency transformer, and a decimal counter, must be supplemented with a receiver. This receiver also determines the measurement precision that can be reached. A block diagram of the complete measuring equipment is given in Fig. 2.

The equipment consists of

1. the harmonic selector to obtain harmonics of 1 mc to be chosen arbitrarily, up to and including the 29th harmonic, starting from a standard frequency of 100 kc;
2. the frequency transformer in which the frequency to be measured is transformed to less than 1 mc by the aid of the nearest multiple of 1 mc obtained from the harmonic selector;
3. the decimal counter from which the frequency to be measured is read directly;
4. the receiver with take-over generator.

A. The Harmonic Selector

A 100-kc standard frequency is multiplied by ten in two stages (5×2) to 1 mc. After this, the sinusoidal 1-mc voltage is converted into short impulses having a repetition frequency of 1 mc. These impulses contain all harmonics that are multiples of 1 mc. The desired harmonic is selected with a circuit according to the method of double modulation, henceforth called "harmonic selector."

In this case, no adjustable filters are needed, but only a narrow band filter, though at a rather high frequency. This filter is henceforth called "selection filter." All harmonics up to and including the 29th can be selected by means of one knob without the range being changed.

The following indications are used:

f_s = repetition frequency of the impulse, derived from the frequency standard;

f_0 = central frequency of the selection filter;

$F = nf_s$ = desired frequency, n being the number of order of the harmonic;

F_{max} = highest desired harmonic;

f_g = frequency to which the search oscillator is adjusted.

In the first modulator (Fig. 2) the harmonic spectrum of the 1-mc impulses is mixed with the voltage of an adjustable oscillator, called "search oscillator," which is adjusted to the frequency $f_g = F + f_0$. A transformed frequency spectrum is therefore formed of which only the frequency f_0 is passed by the selection filter next to the first modulator. In the second modulator the same frequency f_g of the search oscillator is mixed with the frequency f_0 from the selection filter. After this second modulator comes a low pass filter whose cutoff frequency lies between f_0 and F_{max}. After passing this filter, the desired harmonic can then be taken off.

The central frequency f_0 of the selection filter, and consequently the frequency range of the search oscillator, were chosen in such a manner that they are above the highest desired frequency F_{max}, so as to ensure that all the remaining intermodulation products, except the required frequency F, shall lie above F_{max}. These intermodulation products can be stopped by a simple low pass filter. Moreover, the lowest range of the search oscillator is higher than the frequency f_0. It is clear that there can be no difficulties with the image frequency of the search oscillator, $f_g = F - f_0$, because it falls outside the range of this oscillator.

Furthermore, the central frequency f_0 of the selection filter was chosen at such a frequency that it was not a multiple of the impulse repetition frequency; otherwise, the harmonic of the impulse frequency, which is equal to f_0, will always lie in the pass band, independent of the frequency formed in the first modulator. Moreover, the pulse width was chosen in such a manner that the amplitudes of the harmonics of which the frequencies are nearest to the pass band of the selection filter are minimum.

On the one hand, the bandwidth of the selection filter must be as large as possible, thus facilitating the adjustment of the search oscillator. On the other hand, the frequencies from the frequency spectrum after the first modulator that are nearest to f_0 (which frequencies differ from f_0 by multiples of the impulse frequency) must already be sufficiently attenuated (viz., 50 or 60 db). In this connection, the possible deviation of the search oscillator from its nominal value, which reduces the permissible bandwidth, has also to be considered.

With the measuring equipment described here, $f_s = 1$ mc. The highest required frequency is the 29th, so that $F_{max} = 29$ mc. The pass band of the narrow selection filter was chosen in such a manner as to make $f_0 = 32.5$ mc. The frequency range of the search oscillator then ranges from $32.5 + 1 = 33.5$ mc to $32.5 + 29 = 61.5$ mc. This can be effected by using only one range. The cutoff frequency of the low pass filter is about 30 mc.

For example, supposing one wishes $F = 14$ mc; the search oscillator must then be adjusted to $f_g = 14 + 32.5 = 46.5$ mc. At the input of the first modulator there is now found . . . 13; 14; 15; . . . mc, together with 46.5 mc. Consequently, after the first modulator, we find at the input of the selection filter 46.5 ± (. . . 13; 14; 15; . . .) mc = . . . 33.5; 32.5; 31.5; . . . mc and . . . 59.5; 60.5; 61.5 . . . mc.

Of this frequency spectrum the selection filter only passes the frequency 32.5 mc, so that at the input of the second modulator we find 32.5 mc and 46.5 mc.

Thus we have after the second modulator 46.5 ± 32.5 mc = 14 mc and 79 mc. Consequently, after the low pass filter, only the 14-mc sinusoidal voltage to be found, is left.

Owing to the double modulation, a deviation from the nominal value of the frequency f_g of the search oscillator (in the above example 46.5 mc) does not influence the precision of the frequency ultimately obtained. Such a deviation may be caused by inaccurate adjustment of the search oscillator, or by changes in temperature. The frequency obtained remains always exactly harmonic with regard to the fundamental standard frequency. Only the amplitude may change owing to the fact that the frequency generated in the first modulator does not come to lie in the center of the pass band of the selection filter.

The adjustment of this generator, which ranges from 33.5 to 61.5 mc, is therefore not at all critical. Consequently, its scale could be calibrated directly from 1 to 29 mc. By applying a variable condenser of great capacity and having circular plates in series with a small fixed capacity, a fairly linear course has been obtained over 180° of the scale.

So with one knob, the adjustment of which is not critical, every frequency that is a multiple of 1 mc, lying between 1 and 29 mc, can be chosen. Every frequency thus obtained then has the same precision as the standard frequency used.

The selection filter at 32.5 kc consists of three amplifying stages having, in total, four circuits adjusted to 32.5 mc. To reduce the attenuation, the grid is connected every time to a tap of the coil.

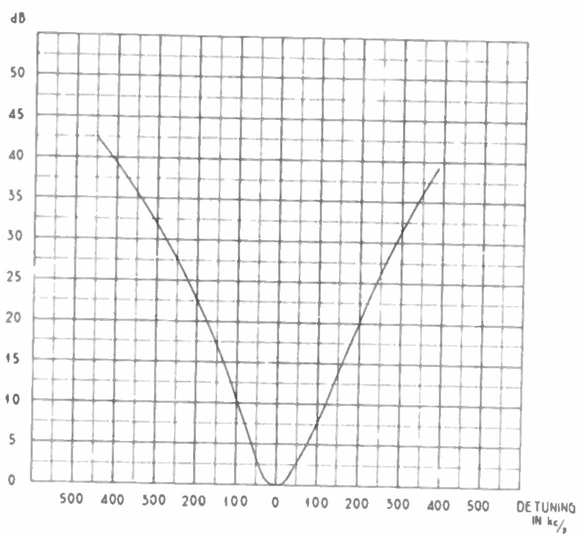

Fig. 3—Attenuation curve of selection filter 32.5 mcs.

Fig. 3 gives the attenuation curve of the filter; the bandwidth between the points of 3-db attenuation is 100 kc. When the frequency differs 0.5 mc from the central frequency f_0, the attenuation is about 45 db.

The low pass filter after the second modulator consists of four constant k sections and one derived section. It has a symmetrical input and output, adapted to a screened cable. The attenuation at 32 mc is about 45 db. Moreover, a wide-band amplifier for the range of 1 to 30 mc is added for use in laboratories. A 0.5-volt voltage at 80 ohms is available at the output. This wide-band amplifier as such is not required for the measuring apparatus proper.

B. The Frequency Transformer

To the frequency transformer are supplied the frequency to be measured and the nearest 1-mc harmonic, so that the difference frequency is lower than 1 mc and can be measured by the decimal counter. The modulator is followed by a low pass filter having a cutoff frequency of 1 mc. This filter stops the remaining intermodulation products.

An output meter after the low pass filter indicates when a difference frequency smaller than 1 mc is generated. When turning the tuning knob of the search oscillator, this output meter shows two deflections of the needle for every unknown frequency, viz., one for the nearest harmonic of the 1-mc impulses lower than the unknown frequency, and one for the next harmonic which is consequently higher. Also, if the frequency to be measured is not known approximately, it is very easy to determine between which two consecutive harmonics of the 1 mc it is situated. For this purpose only one knob need be turned.

If the unknown frequency is nearly a whole multiple of 1 mc, three consecutive positions of the search oscillator cause a deflection of the needle of the output meter. As a matter of fact, there are then three different frequencies which pass the low pass filter (viz., successively, 1° a frequency somewhat lower than 1 mc, 2° a very low frequency, and 3° a frequency somewhat higher than 1 mc). All these three frequencies can be measured by the decimal counter. The value of the unknown frequency follows unequivocally from the result so that this special case does not cause any difficulties either. Preferably, that nearest harmonic is chosen which is lower than the frequency to be measured, because then the indication of the decimal counter can be added to the chosen harmonic. If the higher harmonic is chosen, the complement of the decimal counter indication must be taken.

A difficulty occurs, however, if the unknown frequency is between 1 and 2 mc. In that case, 1 mc must be added in the frequency transformer. In addition to the desired difference frequency, there are then also generated the sum frequency, the unknown frequency, and the above-mentioned 1 mc. Besides the sum frequency, the last two mentioned frequencies must also be stopped by the 1-mc low pass filter.

This cannot be effected by reducing the cutoff frequency of this filter to less than 1 mc. The reason for this is that frequencies approximately equal to a multiple of 1 mc, and whose difference frequency after frequency transformation can therefore be nearly 1 mc, would also be stopped, and it would not be possible to measure them. Therefore, the modulator was effected as a balance modulator, the frequency to be measured and the chosen harmonic from the harmonic selector both being supplied in balance and the anodes of the two modulator tubes being connected in parallel before the output circuit. The suppression of the above-mentioned, undesired frequencies remains, however, partly insufficient. That is why, for the time being, for the measuring range of 1 to about 1.4 mc, the second harmonic of the 1-mc impulses equal to 2 mc is taken, the complement of the indication of the decimal counter having, therefore, to be considered.

A few other solutions of this problem are being investigated in order to remove this "aesthetic fault." The measuring precision, however, is not reduced by it.

In the meantime, it has proved possible to extend the range of the type of counter used here to about 1.5 mc. So one possible solution of the problem is to measure the frequencies up to about 1.5 mc with the counter itself. If the harmonic selector is used for measuring frequencies higher than 1.5 mc only, there are no difficulties whatever, and it is nowhere necessary to use the complement of the value indicated by the decimal counter.

C. The Decimal Counter with Auxiliary Equipment

The electronic decimal counter is an extension of a usual type;[1] it measures any sinusoidal frequency of about 30 cps to 1 mc, and indicates the result in decimal numbers directly on meters.

The incoming sinusoidal wave form is transformed into pulses, and the number of these pulses passed through the counter in a given time, usually one second, is measured. The pulse-forming circuit is switched on and off by a start and a stop circuit, each consisting of a trigger pair, which circuits are interconnected in such a way that, after the counter has been reset to zero, the stop trigger can be actuated only after the start trigger has been triggered by an incoming start impulse. So the inputs of start and stop triggers can be paralleled and fed by the same second impulses. If start and stop pulses are separated by 10 seconds, the measuring accuracy is raised by a factor 10, provided the frequency to be measured remains constant during this time interval.

The output of a (primary) frequency standard is transformed into short pulses and the second impulses are selected by means of gate circuits. The precision of the second impulses thus depends on the stability and the leading edge of the impulses only. The accuracy is of the order of ± 0.1 μsec or better. Details will be published later.

[1] S. S. West, "An electronic decimal counter and chronometer," *Elec. Eng.*, vol. 19, pp. 3–6; January; and pp. 58–61; February, 1947.

Furthermore, a number of relays has been included in the arrangement, which make it possible to print the measuring result indicated by the decimal counter in figures on paper by means of a normal page printer. The measurement with the decimal counter itself lasts exactly one second. In the next second the printing of the result of the measurement is commenced. One measurement can be effected every two seconds. In this way, in a short time the frequency variation of a generator can be measured with the precision of the second impulse derived from the standard frequency ± 1 cps.

Fig. 4 gives a record of a frequency measurement of a standard signal oscillator. The frequency was about 16.9 mc. The knob of the search oscillator of the harmonic selector was set on 16 mc (so the search oscillator was tuned to $32.5+16=48.5$ mc). After the frequency transformer, on the input of the decimal counter, appeared a frequency less than 1 mc, which was measured by the decimal counter. The first measurement gave 16 mc+912,394 cps.

912,394	912,395	912,393	912,393	912,396	912,399	912,401	912,402	912,402	912,401	123,976
912,402	912,400	912,405	912,396	912,394	912,394	912,395	912,395	912,393	912,398	123,972
912,393	912,393	912,393	912,392	912,390	912,392	912,393	912,394	912,393	912,393	123,926
912,393	912,394	912,393	912,392	912,392	912,394	912,394	912,393	912,391	912,390	123,926
912,395	912,397	912,396	912,397	912,396	912,394	912,403	912,401	912,397	912,392	123,968
	16.05	14/01								
912,353	912,353	912,352	912,351	912,351	912,351	912,349	912,347	912,347	912,348	123,502
912,346	912,346	912,345	912,345	912,344	912,346	912,347	912,347	912,349	912,350	123,465
912,349	912,350	912,350	912,349	912,349	912,347	912,347	912,346	912,346	912,347	123,480
912,350	912,351	912,350	912,350	912,349	912,349	912,349	912,351	912,351	912,353	123,503
912,352	912,352	912,352	912,351	912,350	912,348	912,347	912,347	912,345	912,344	123,488
	16.10	14/01								
912,318	912,318	912,317	912,318	912,318	912,316	912,316	912,315	912,316	912,314	123,166
912,319	912,319	912,317	912,317	912,317	912,317	912,316	912,315	912,314	912,314	123,165
912,312	912,312	912,312	912,312	912,312	912,313	912,308	912,305	912,303	912,300	123,089
912,299	912,297	912,297	912,295	912,294	912,294	912,294	912,296	912,296	912,296	122,958
912,295	912,295	912,294	912,294	912,294	912,293	912,290	912,290	912,290	912,289	122,924
	16.15	14/01								

Fig. 4—With a page printer printed results of frequency measurements of a standard signal oscillator.

Each line contains the results of ten measurements made. The eleventh number is the sum of the foregoing ten measurements, which sum is calculated simultaneously with a special relay circuit. As six digits only are available, the first digit of the sum, in our case figure 9, is omitted. This sum immediately gives the mean of ten measurements. In many cases, one has to deal with this sum only, which simplifies the arithmetic to be done afterwards. This equipment was originally developed for measurements of time signals and other purposes, but it is now used for frequency measurements too.

As a measurement is effected in 2 seconds, a very great number of measurements is obtained when the frequency of an oscillator is measured for a long time. To restrict the number of measurements, and thereby the work of interpreting them afterwards, without impairing the accuracy unduly, the equipment may be extended with a device to print automatically and at our own discretion a series with a fixed number of lines (e.g., 2, 5, or 10 lines each with ten measurements), the time of interval between these series being chosen at will (e.g., 5, 10, 15, 30, or 60 minutes). The series are separated by a number of open lines. Moreover, time and date are printed after each series (e.g., in Fig. 4 the first series was printed at 1605 o'clock on the 14th of January). Unattended measurements can then be made (e.g., during the night), the results being printed on a sheet of paper.

Moreover, there is another possibility. Of a long series of measurements, made on the same generator or transmitter, two figures of the measured frequency, to be chosen at will, e.g., the last two, or any other desired pair, can be registered by means of a self-recording meter. In this way a graph of the frequency course is directly obtained.

D. The Measuring Receiver

For the measuring of the frequency of transmitters, a receiver is requisite. In the case of telegraphy transmitters, and in general, with fast and deep fading, the signal of the transmitter is very hard to measure with the required precision. Therefore, a so-called "take-over generator" is used, i.e., a local generator whose frequency is equalized to that of the transmitter, the frequency of the "take-over generator" thereafter being measured in the manner described.

In the following description we confine ourselves to the 5- to 30-mc frequency range. In this range we find most of the intercontinental radio circuits, and here the equipment must meet the severest requirements.

Fundamentally, the receiver corresponds with the latest commercial traffic receiver developed by the Radio Laboratory of the Netherlands Postal and Telecommunications Services.[2] Since the receiver is now destined for the measuring of frequencies, some alterations were effected so as to obtain the required measuring precision.

The first medium frequency is at 1,500 kc. So the frequency of the first oscillator is 1,500 kc higher or lower than the transmitter frequency, but the higher frequency has been chosen. The first oscillator now also serves as "take-over generator" whose frequency is measured as described. The transmitter frequency is then 1,500 kc lower. In addition, by the measuring precision of the frequency meter itself, the precision with which the frequency of the transmitter can be measured is also determined by the deviation of the frequency of the first oscillator may have from the correct frequency, viz., transmitter frequency + 1,500 kc.

The first oscillator is provided with an automatic frequency control, the frequency deviations thus remaining less than about 5 cps. It is possible to narrow down this range even further, but a greater measuring precision is not required in the 5- to 30-mc range.

In order to increase the stability, the first oscillator is, moreover, placed in a simple thermostat, circuit and tube being placed in an aluminium box, surrounded by a second aluminium box, and separated from the first by an air layer of 1-cm thickness. There is no metal connection between inner and outer box. The temperature is regulated with a Fenwal switch, placed in an aluminium block, attached to the wall inside the inner box. The temperature is held constant within about $\pm\frac{1}{2}$°C.

The second medium frequency is 100 kc, so that the second intermodulation frequency has been chosen at 1,600 kc. The latter is derived from the 100-kc standard frequency, so that the second intermodulation frequency cannot create any measurement faults.

The automatic frequency-control circuit consists of a very narrow crystal band filter at 100 kc with a band width of about ± 35 cps, followed by a repeater, a limiter, and a discriminator. The control voltage energizes the tension coil of a Ferraris motor whose axis is coupled, via a cogwheel retardation, with a little variable condenser which is shunted parallel to the tuning condenser of the first oscillator. The regulation is effective in telephone transmitters both with or without suppressed carrier wave as in telegraph transmitters which are keyed in the carrier wave itself (so-called "on-off" keying), or transmitters which operate with "frequency shift."

For further particulars regarding the receiver reference is made to the article quoted above.

The output of the narrow crystal filter at 100 kc is likewise connected with the horizontal plates of a cathode-ray oscillograph, whereas the vertical plates are connected to a 100-kc standard frequency. Since the second intermodulation frequency has been derived from the frequency standard, a stationary Lissajous pattern (ellipse) is a clear indication that the frequency of the first oscillator (at the same time "take-over generator") precisely equals the transmitter frequency +1,500 kc. If the pattern is not stationary, it is an indication as to the frequency deviation that remains after automatic regulation has been carried out.

In case the automatic-frequency regulation cannot work properly, for example at weak reception, the first oscillator can be manually adjusted as well as possible by means of the cathode-ray oscillograph.

For a precise measuring, the first oscillator need be tuned correctly for a short period only, since the measurement with the decimal counter lasts only 1 second. Obviously, the adjustment without automatic control is more difficult. It takes more time, particularly if a number of consecutive measurements of the same transmitter are required, and the receiver must often be regulated afresh.

To determine, among other things, the name and country of origin of a transmitter, the signal after the second medium frequency at 100 kc is detected (for telephony) or mixed with an auxiliary frequency of 99 kc (for telegraphy). A discriminator for the reception of frequency-shift telegraph signals has also been built in, and provision has been made so that the shift can be

[2] C. T. F. Van Der Wijck, "Een moderne telegrafie-ontvanger" (A modern telegraph receiver), *Tijdschr. Ned. Radiogenoot.*, pt. XIV, nr. 2, pp. 28–40; March, 1949.

measured with the counter. Reception of single-sideband transmitters is also possible. The audible telegraph signals can also be translated into Morse signs or printed characters by means of a recorder or a printer.

The frequency ultimately measured is that of the "take-over generator," which is 1,500 kc higher than the frequency to be measured. This difference, which is a constant for every measurement, can be accounted for quite simply in the circuit arrangements, so that the real frequency of the transmitter is shown on the decimal counter, or is printed on a page printer. Consequently, auxiliary computations are not requisite.

III. Conclusion

In Fig. 5 the complete frequency-measuring equipment is illustrated. In the left bay is the decimal counter with the relay switching equipment. The center bay contains the harmonic selector, the frequency transformer, and the wide-band repeater. The large tuning scale is the search oscillator scale calibrated in mc. In this bay there are some vacant panels. The 5- to 30-mc measuring receiver has been fitted in the third bay. Since the latter is likewise used for various other laboratory purposes, it is more bulky and it has more knobs than are required for the frequency-measuring device proper.

The described frequency-measuring equipment has been developed and constructed by the Radio Laboratory staff of the Netherlands Postal and Telecommunications Services. With reference to developing the

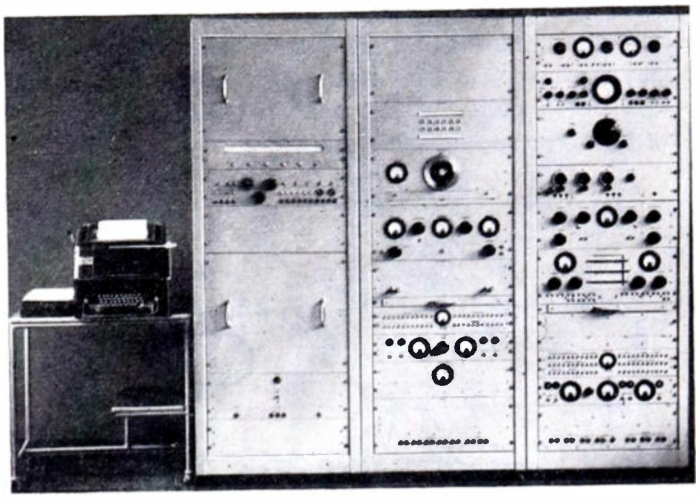

Fig. 5—Front view of complete frequency-measuring equipment.

equipment, the following persons should be particularly mentioned: Mr. J. G. Coster for the harmonic selector and frequency transformer, Mr. C. J. Sanders for the decimal counter with auxiliary switching equipment, and Mr. P. J. Hooymans for the measuring receiver.

A Wide-Range Oscillator in the Range from 8,000 to 15,000 MC*

R. W. WILMARTH†, MEMBER, IRE AND J. L. MOLL†, ASSOCIATE, IRE

Summary—**A retarding field oscillator which tunes over the range from 8,000 to 15,000 mc is described. Energy is taken from the tube by allowing the repeller, which is part of the resonant circuit, to radiate into a waveguide. Results for both inductive and capacitive tuning are discussed. With capacitive tuning, the efficiency of the oscillator is between 2 and 4 per cent over the entire range, with maximum efficiency occurring near the highest frequency.**

Introduction

In the construction of resonant cavities for oscillators at high frequencies, it is sometimes necessary to insulate part of the cavity from the rest of the cavity for static voltages, and at the same time to provide an rf short at the point of static insulation. This requirement dictates the use of a noncontacting rf short. Satisfactory noncontacting rf shorts can be designed to operate over a wide frequency range if the requirement of static insulation does not dictate too great spacings at the point of noncontacting rf short.

An example of a tube in which this problem is a primary consideration is the retarding field oscillator reported by Heil and Ebers.[1] An outline of this tube is shown in Fig. 1. The resonator is tuned by moving the repeller parallel to its axis of symmetry, thus changing the capacitive loading of the cavity. At the long wavelength end of operation, the repeller is close to the resonator nose. The oscillator works in a manner very similar to a reflex klystron in that the electrons receive velocity variations in the part of the trajectory near the nose and drift in a region relatively free of ac fields. The electrons are collected at the resonator nose on their return through the ac fields. The transit time is adjusted by adjusting the repeller voltage. At the short wavelength end of the range, the repeller is pulled away from the resonator nose and the electrons move in relatively uniform ac and dc fields. If the transit time is properly adjusted, the beam will have a negative conductance. Ebers has shown[2] that if $\frac{1}{4}$ cycle of the transit

* Decimal classification: R355.912. Original manuscript received by the Institute, September 17, 1951; revised manuscript received, February 20, 1952. The research described in this paper and in references 1 and 2 was sponsored by the Air Matériel Command, Wright-Patterson Air Force Base, Dayton, Ohio.

† Ohio State University, Columbus, Ohio.

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

[2] J. J. Ebers, "Retarding field oscillators," Proc. I.R.E., vol. 40, pp. 138–146; February, 1952.

is considered to be used for bunching, $\frac{3}{4}$ cycle drifting, and $\frac{1}{4}$ cycle for working, the operation of the oscillator is analogous to reflex klystron operation.

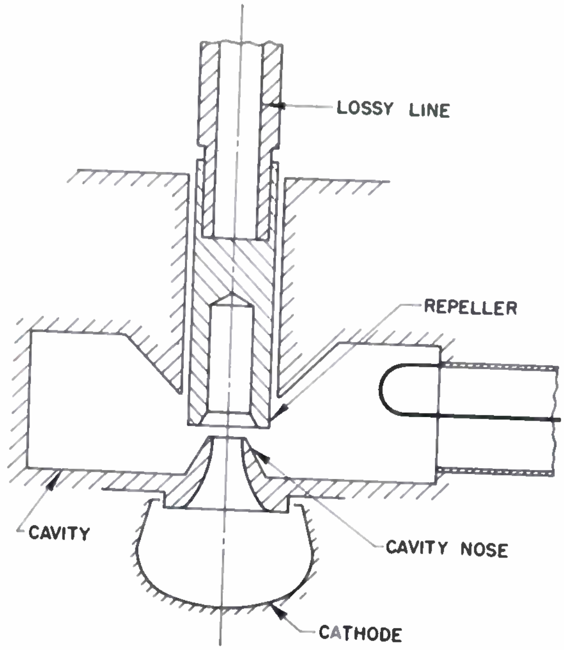

Fig. 1—Heil-Ebers tube.

In the oscillator of Fig. 1, the repeller is an integral part of the resonant cavity, and since there is a dc voltage between the repeller and the rest of the cavity, a noncontacting rf short is necessary. In the original tube this was obtained by using a capacitive by-pass in combination with a lossy line. Subsequent experience has shown that this resulted in a great loss of power in the noncontacting short.

In the oscillators described in this article, a noncontacting imperfect short is used in which energy is allowed to leak past the point of desired short to the useful load. The repeller forms the center conductor of a short coaxial line, the outer conductor being part of the cavity block. The center conductor (repeller) is used to excite a waveguide.

The oscillator in Fig. 2 uses this type of output to couple energy into the load. In this arrangement, we must provide a noncontacting rf short at the far side of the waveguide to prevent energy from going on up the repeller mechanism. The rf currents are much smaller in the waveguide than in the cavity, and a simple capacitive by-pass as shown in Fig. 2 is sufficient. The oscillator tunes over the entire range with no holes. (Tuning is accomplished, as in the tube in Fig. 1, by moving the repeller.) The section of low-impedance coaxial line between the cavity and the waveguide matches the impedance of the load to the tube.

The method of coupling energy from the reflex oscillator which we have described above is adapted to an inductively tuned tube in Fig. 3. The repeller is fixed in position, being held at the far side of the waveguide. Tuning is accomplished by moving the slug parallel to its axis of symmetry. The low-impedance section of line formed by the slug as inner conductor, and the cavity block as outer conductor, match the load impedance to the oscillator. Some design considerations and test results for both the capactively tuned tube and the inductively tuned tube are described in a later section.

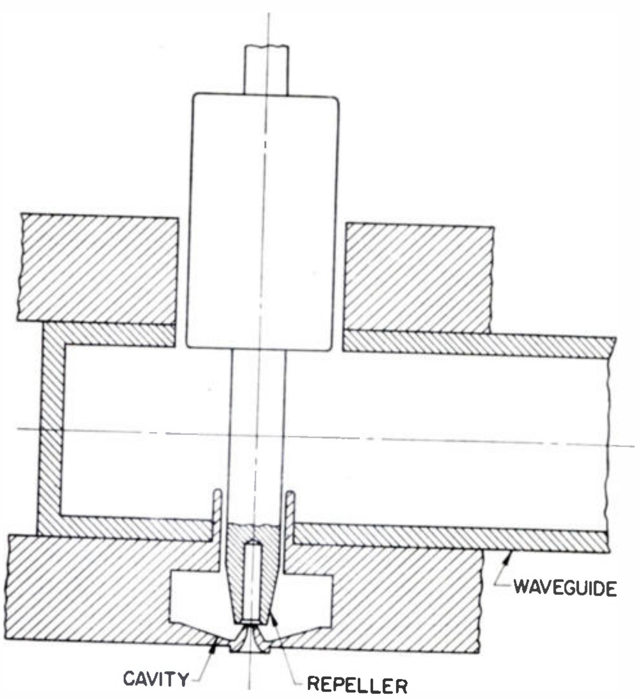

Fig. 2—Capacitively tuned retarding field oscillator with waveguide output.

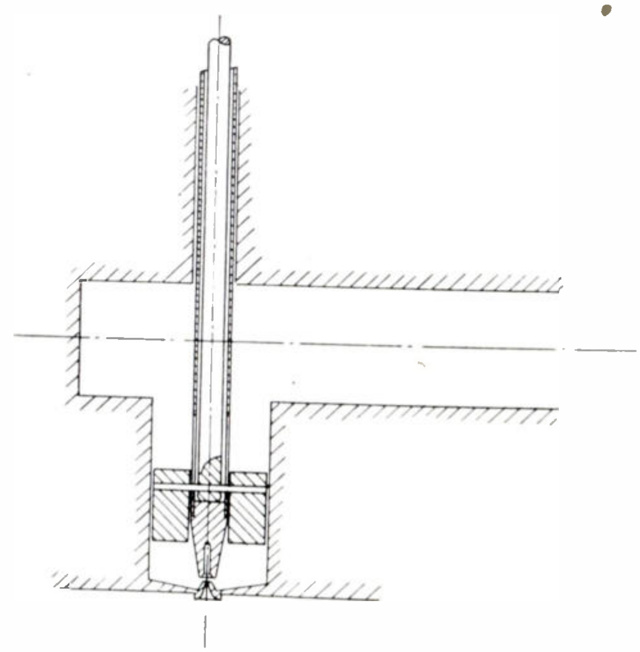

Fig. 3—Inductively tuned retarding field oscillator with waveguide output.

Capacitively Tuned Oscillator

Experimental models of the oscillator shown in Fig. 2 have tuned satisfactorily from 2- to 4-cm wavelength. A representative tuning curve is shown in Fig. 4. There were no holes in the power output curve, but the power output varies by a factor of about 2 to 1 across the tuning range. The tube tunes very rapidly at wavelengths above 3.6 cm because of the nature of capacitive tuning. The tuning range is limited at the high-frequency end when the repeller is pulled so far away from the cavity nose that the electron stream is not refocused back onto the cavity nose. At the low-frequency end, the

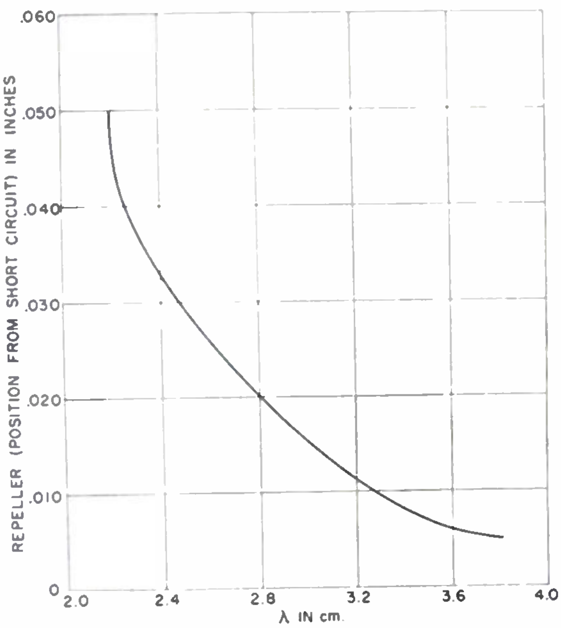

Fig. 4—Tuning curve for capacitively tuned retarding field oscillator.

limitation is the danger of shorting the repeller against the cavity nose. The useful range, however, is limited by the fact that the rate of tuning increases at low frequencies.

The performance of the tube was studied with beam aperture and repeller geometry as variables. The operation changed slowly with change in beam aperture, with maximum output over the tuning range occurring at 0.021-inch diameter aperture. This aperture is a compromise between the interception of current and coupling to the beam. The resonator intercepts approximately 10 per cent of the cathode current with this size aperture. The performance dropped off very rapidly when the hole in the repeller was increased beyond 0.040-inch diameter. It is believed that this was due to the failure of the electron stream to be properly refocused back onto the nose. Repeller hole diameters of less than 0.040 inch did not greatly affect the efficiency of operation of the tube, but did increase the rate of tuning of the oscillator at long wavelength.

The requirement of wide tuning range dictates the sharp angle of the resonator nose. This is a disadvantage from the standpoint of heat dissipation. However, experiments show that this sharp angle is not necessary for efficient operation of the tube.

Power output versus wavelength is shown in Fig. 5 for beam aperture of 0.021-inch diameter and repeller hole of 0.040-inch diameter. The dashed curve shows the variation of repeller voltage for optimum power output at each wavelength. The simple matching section, consisting of the short section of coaxial line, matches the oscillator to the waveguide sufficiently well over the entire tuning range. This is shown in the representative Rieke diagram of Fig. 6 (see following page). The sec-

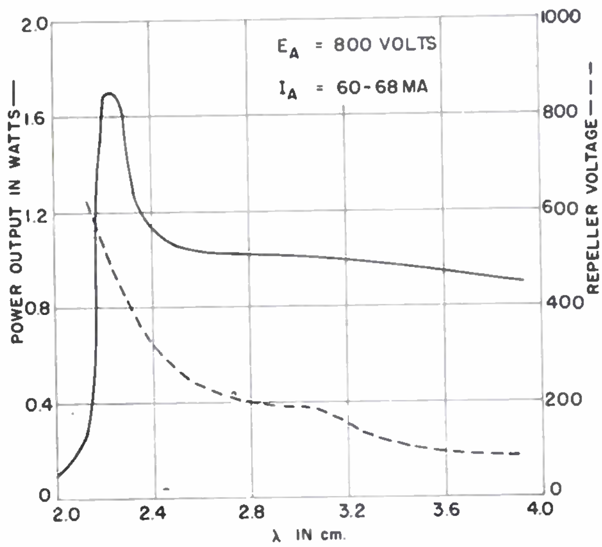

Fig. 5—Repeller characteristic and power output curve for capacitively tuned retarding field oscillator.

tion of line is a quarter wavelength long at the short wavelength end of the range. The R G 52U waveguide which this coaxial line feeds is shorted at a distance of 1 cm. from the coaxial line. This is a quarter wavelength in the middle of the range. The other end of the waveguide goes to a mica output window.

Inductively Tuned Tube

Fig. 3 shows a relatively primitive adaptation of the waveguide output oscillator to inductive tuning. The tuning slug is moved by a rod which moves inside the repeller and is connected to the tuning slug through a slot in the repeller. The repeller itself does not move. This tube oscillated over a range from 2 to 4 cm. The wavelength varies almost linearly with the position of the tuning slug as shown in Fig. 7. As yet, we have been unable to get the inductively tuned tube to oscillate as uniformly and efficiently as the capacitively tuned oscillator over a very wide range. This is due to the fact that we have not studied this type of tuning extensively on account of the more difficult mechanical problems involved.

Fig. 8 shows two geometries which result in efficient oscillator operation. The sharp angle is used for wide-range, capacitively tuned oscillators. If the tube were inductively tuned, the flat nozzle and flat repeller could be used. This would allow for greater heat dissipation

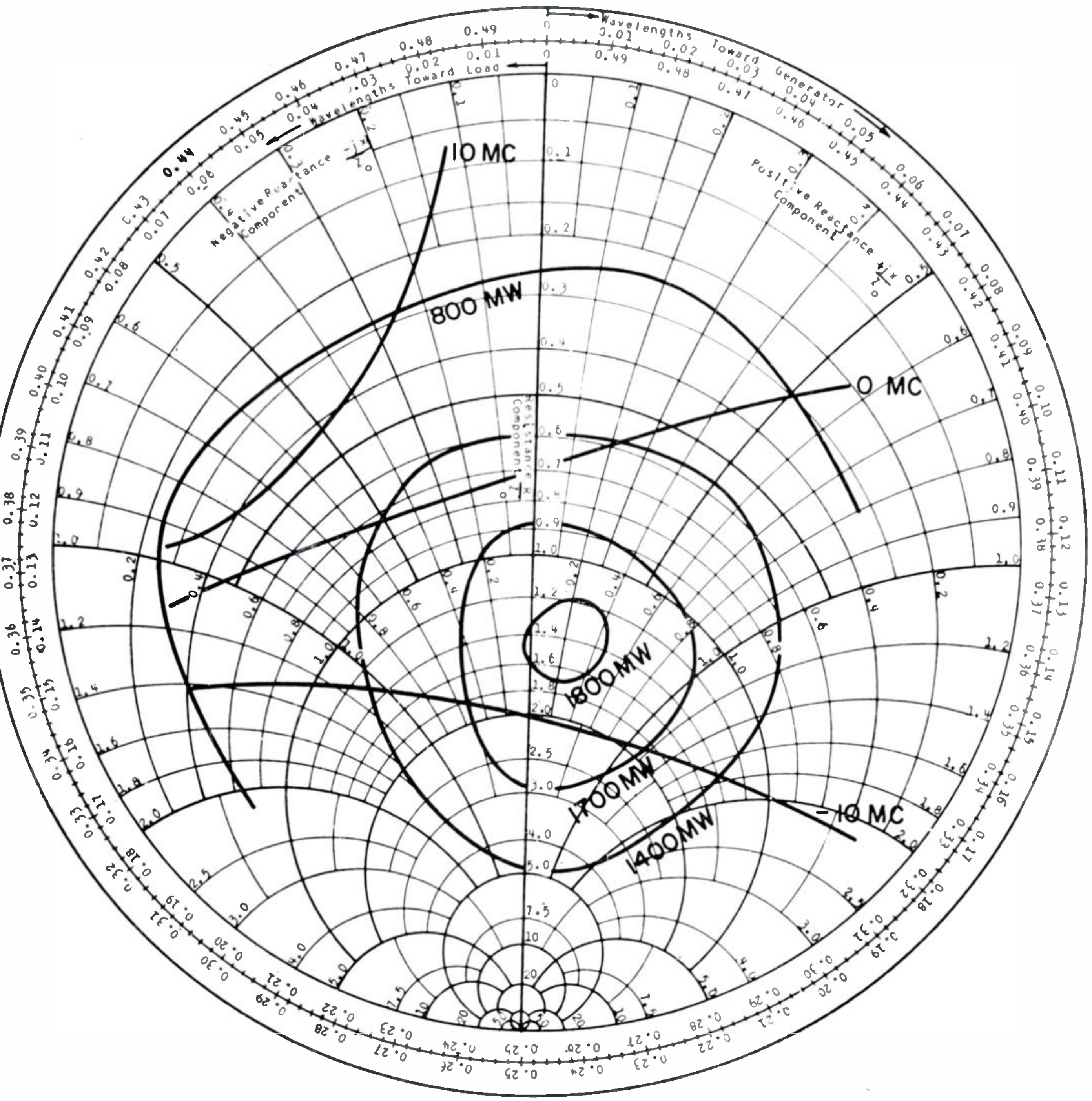

Fig. 6—Typical Rieke diagram for reflex oscillator with waveguide output. The diagram was made at a center wavelength of 2.25 cm. The repeller voltage was adjusted for maximum output into a matched load.

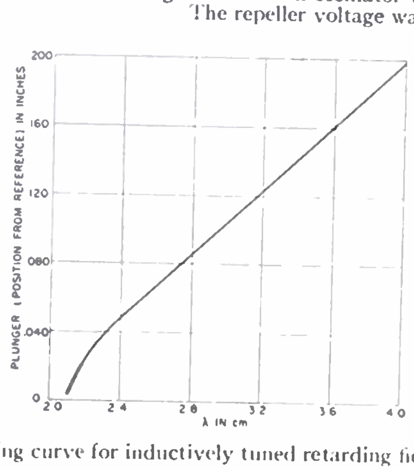

Fig. 7—Tuning curve for inductively tuned retarding field oscillator

at the nozzle. Inductive tuning has the intrinsic property that the wavelength changes at a slower rate with respect to position of tuner than in the case of capacitive tuning. If there are no losses in the tuner, the shunt resistance of the cavity is higher throughout the range. Since the repeller does not move, the electron mechanism is more uniform.

The Electron Gun

The gun design used in all of the tests described in this report is shown in Fig. 9. This gun was developed from a gun designed by Heil for use in power klystrons.[1] The geometry of the focusing elements of the

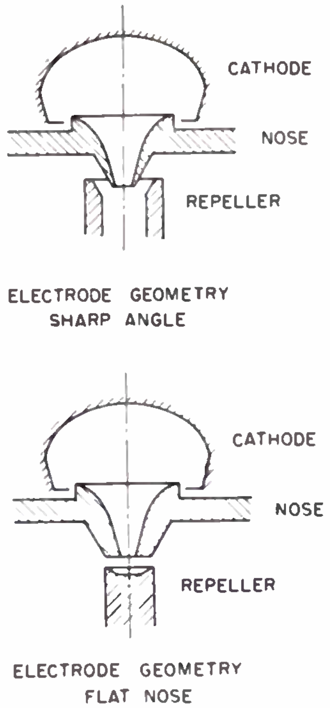

Fig. 8—Electrode geometries which result in efficient oscillation in retarding field oscillators.

new gun design is the same as in the original, with the exception that the projection on the beam-forming electrode is removed. This increases the spacings between gun elements. With this modification, the critical dimensions of the new design are scaled down by a factor of 5 from the original. No difference in the performance of the tube, or gun, could be detected as a result of the change in the beam-forming electrode. In the design as shown in Fig. 9, the details are assembled into the anode shell and are held in place by the mica retainer. The parts are properly aligned only after the cathode is up to temperature because of the differences in the coefficient of expansion. This gun design has the particular advantage that the parts are easily salvageable—an important consideration in experimental work on a demountable tube.

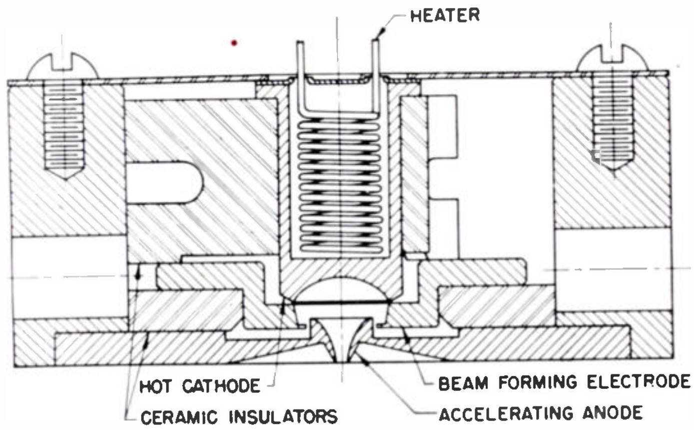

Fig. 9—Electron-gun design for demountable tubes.

With a 0.120-inch aperture, the original gun would deliver 140-ma beam current at 1,000-volts potential. All of the cathode current is focused through the aperture. According to linear scaling theory, the voltages and current relations should remain constant regardless of the scale to which the gun is built. Unfortunately, however, cathode activity and the difficulty of fabrication of the gun parts become limitations.

In the scaled gun, the cathode area is 0.18 cm^2. The current density is increased from the cathode surface to the minimum cross section of the beam by a factor of approximately 60. The first gun was built as nearly as possible to the exact scale of the original gun. Over 96 per cent of the cathode current was focused through a 0.024-inch aperture. For an opening of 0.0183 inch it was found that 80 per cent of the cathode current was focused into the beam. At 1,200 volts the beam current was 120 ma and the cathode current was 150 ma.

Conclusions and Recommendations

Inductive tuning offers what appear to be intrinsic advantages over capacitive tuning. These advantages have not been realized in the retarding field oscillator as yet. The problems to be solved before this type of tuning can be used to an advantage are the mechanical problems of moving the tuner and maintaining proper spacings in the tube. In addition to the mechanical problems of tuning, there is the electrical problem of maintaining the proper loaded shunt resistance, or loaded Q throughout the tuning range.

In addition to the problems of developing an inductively tuned tube, there is the question of extending the operating range of the oscillators to higher frequencies. Experiments indicate that the 0.020-inch diameter aperture is working at the limit of its frequency capability, at wavelengths of 2 cm. Thus, to go to higher frequencies would require a smaller aperture or higher voltage.

Another possibility is to use a gridded aperture and smaller beam voltages. No attempt has been made thus far to investigate this type of coupling to the beam.

With present techniques and knowledge, a wide-range-tunable retarding field oscillator has been produced to cover the range of wavelengths from approximately 2.2 to 4 cm. Maximum over-all efficiencies are of the order of 4 per cent and efficiencies of 2 per cent are maintained over the tuning range.

Acknowledgments

The authors wish to thank their associates in the Electron Tube Laboratory for help in devising and carrying out the experiments described in this paper. Mr. Stanley Taylor should be mentioned in particular for his help with machine work and with the mechanical design, as should Mr. Peter Whibley, who did all of the attendant glass work.

A High-Voltage, Cold-Cathode Rectifier*

E. G. LINDER†, SENIOR MEMBER, I.R.E., J. H. COLEMAN‡, ASSOCIATE, I.R.E., AND E. G. APGAR†

Summary—**A new cold-cathode, high-voltage rectifier is described. Both theory and extensive experimental data are included. The operation of the tube is based upon an improved type of electron trapping similar to that occurring in a cylindrical magnetron in the cutoff condition, but augmented because of the use of end plates on the cathode. Mercury vapor is used, which at normal operating temperatures, provides a gas at a pressure of from 10^{-3} to 10^{-2} mm of Hg. Because of this low pressure and the small dimensions of the tube, ionization is inhibited when the cathode is positive, and high-peak inverse voltages are possible. However, when the anode is positive, the improved electron trapping results in conduction in spite of the low pressure and small dimensions.**

Experimentally, this rectifier has been operated up to 40-kv peak inverse voltage. Life tests at 30-kv peak inverse voltage and a current of 0.4 ma, at 60 cps, have run to 3,000 hours. Satisfactory operation has been obtained over a temperature range from 20 to 80 degrees C. The upper frequency limit lies between 120 and 240 kcps.

This tube appears to offer, for the first time, a practical cold-cathode rectifier for output voltages up to 14 kv and currents up to one ma. Its advantages are particularly valuable in the case of voltage-multiplier rectifier circuits, since high-voltage insulated cathode-heater transformer windings are not necessary.

I. Introduction

THIS ARTICLE deals with a cold-cathode, high-voltage, rectifier tube, which was developed as a part of a project on electron-trapping devices. The general principle of electron trapping is herein discussed, as involved in the operation of this rectifier. It is a type of trapping which is produced by the combined action of electric and magnetic fields, and is similar to that which occurs in a cylindrical magnetron. It results in an electrical discharge resembling a glow discharge, but differing in that a magnetic field is required to maintain it.

The history of gas discharges in electric and magnetic fields goes back to 1898, when such a discharge was observed by Phillips.[1,2] On that occasion, Phillips, had reduced the pressure in a vessel sufficiently low to extinguish a glow discharge between two electrodes, and then switched on a magnetic field along the axis of the two electrodes. A luminous ring was observed to form between the electrodes and the glass wall, as indicated in Fig. 1(A). This glow was interpreted by Strutt[3] in 1913 as being caused by ionization of the gas in the vessel, by negative ions following long paths, on going transverse to the magnetic field, from the two center electrodes to the glass wall, which had been charged positive by the previous discharge. To check this explanation, Strutt placed a cylindrical electrode close to the inside of the glass wall and concentric with the center electrodes. When this outer electrode was held at certain positive potentials, a discharge was maintained, with appropriate magnetic field strengths, between this electrode and the center electrodes, which were at ground potential. A conducting cylindrical jacket over the two center electrodes, as shown in Fig. 1(B), gave the same result. The effect was investigated later by several other workers,[4] and was applied as a rectifier by Bush and Smith.[5]

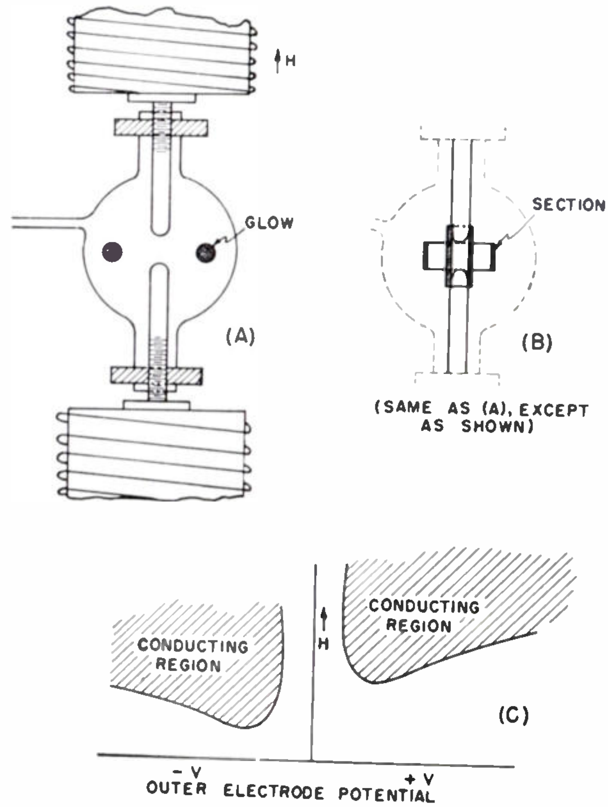

Fig. 1—(A) and (B), early types of tubes; (C) typical cutoff characteristics.

Bush and Smith used the structure of Strutt, shown in Fig. 1(B), with a magnetic field adjusted too low for conduction when the outer cylinder was positive, but sufficiently high for conduction when it was negative. These characteristics are illustrated in Fig. 1(C). The difference in magnetic field required by the different polarities was explained on the basis of Hull's[6] magnetron equations as being caused by the smaller magnetic field required for cutoff in the external-cathode magnetron compared to the same structure operating as an internal-cathode magnetron. This design was

* Decimal classification: R337. Original manuscript received by the Institute, June 25, 1951; revised manuscript received, February 4, 1952.

† Radio Corporation of America, RCA Laboratories Division, Princeton, N. J.

‡ Formerly with RCA Laboratories Division; now with the Radiation Research Corporation, West Palm Beach, Fla.

[1] C. E. S. Phillips, *Proc. Roy. Soc. A.*, vol. 64, p. 172; 1898.

[2] C. E. S. Phillips, *Phil. Trans.* (London), vol. 197A, p. 135; 1901.

[3] R. J. Strutt, *Proc. Roy. Soc. A.*, vol. 89A, p. 68; 1913.

[4] M. Wehrli, *Ann. der Phys.*, vol. 69, p. 289; 1922.

[5] V. Bush and C. G. Smith, Proc. I.R.E., vol. 10, p. 41; 1922.

[6] A. W. Hull, *Phys. Rev.*, vol. 18, p. 31; 1921.

found to have appreciable back current above a few thousand volts, and was replaced by the thermionic gas rectifier, which was more reliable with the same inverse voltage rating. An application of the effect, which is still in use, was made by Penning[7,8] as an ionization gauge.

Penning was first to make an analysis of the known cutoff curves (Fig. 1(C)) by applying Hull's equations to the case of electrons originating throughout the space between the electrodes. Using this procedure, he calculated the ranges of voltages and magnetic fields under which electrons could pick up enough energy for ionization. These calculations qualitatively fit the observed data of cutoff. No attempt was made to determine time-variable characteristics, either theoretically or experimentally. After the cutoff analysis, Penning used the structure of Strutt[3] (Fig. 1(B) without the center sleeve) as a pressure gauge. Some further use[9] of the effect was made in 1946 by pulsing the magnetic field to obtain conduction in radar modulator circuits.

II. Theory of Operation

The formulation of a complete theory of the operation of this rectifier has not as yet been carried out because of mathematical difficulties in meeting the exact boundary conditions. For example, the electron paths are dependent upon complicated electric and magnetic field distributions, upon space charge, and upon collisions with molecules and ions. However, a quite useful idea of the operation of this type of tube can be gained from a simplified theory in which the exact structure is replaced by one of more amenable geometry. This has been done in the following discussion. Although the results do not admit of a quantitative check by experiment on all points, yet the qualitative agreement is good. The theory has been very helpful in the understanding and development of the tube.

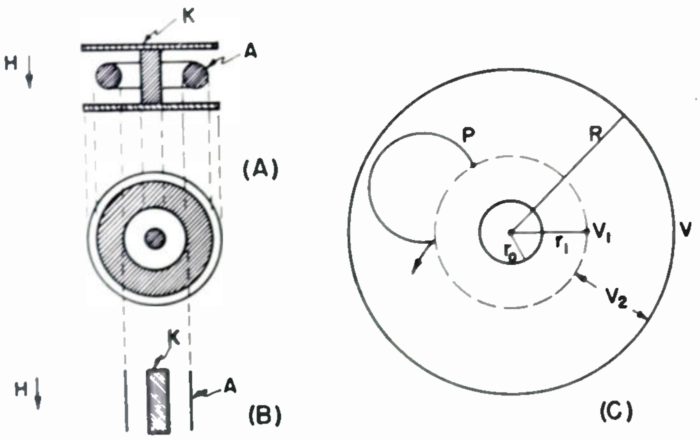

Fig. 2—Electrode structure and electron-path data.

[7] F. M. Penning, *Physica*, vol. 3, p. 873; 1936.
[8] F. M. Penning, *Physica*, vol. 4, p. 71; 1937.
[9] R. E. B. Makinson, J. M. Somerville, K. R. Makinson, and P. Thonemann, *Jour. Appl. Phys.*, vol. 17, p. 567; 1946.

The electrode geometry of the tube is shown schematically in Fig. 2(A), and in detail in Figs. 3 and 4. The tube consists of a spool-shaped cathode K, and a ring-shaped anode A, immersed in an axial magnetic field H. For the purpose of analysis, this structure is here replaced by the concentric cylinder arrangement shown in Fig. 2(B).

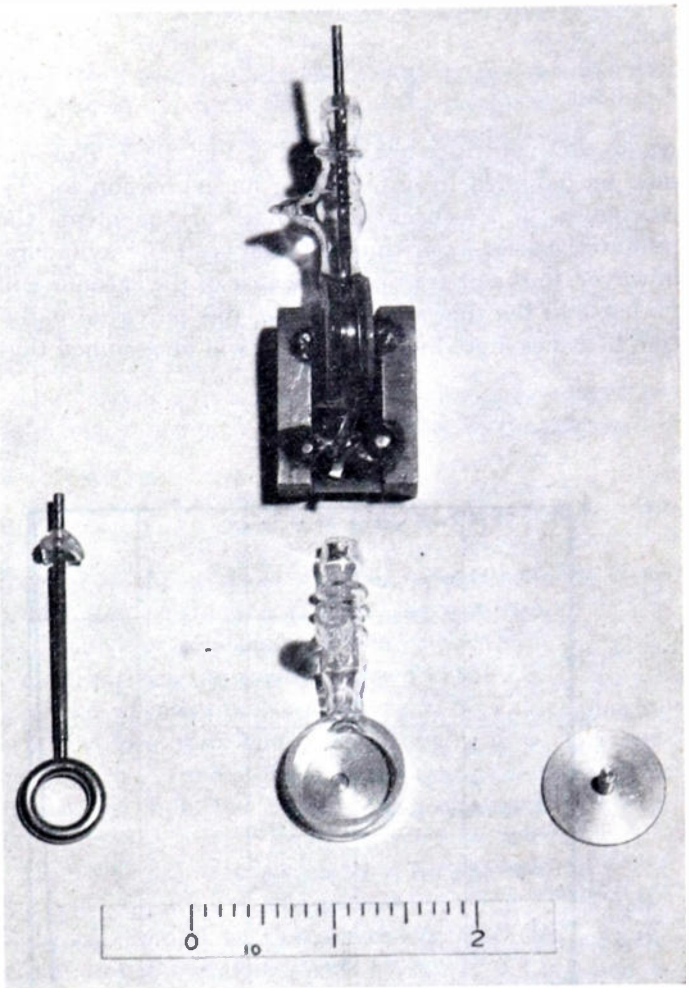

Fig. 3—Tube components and completed tube with magnet.

Electrons leaving cathode K (Fig. 2(B)) travel in radial paths to anode A in the absence of a magnetic field. However, if a field is present, the paths become curved, and at a critical field, the curvature is so great that they fail to reach the anode and return toward the cathode. The same is true of electrons originating at

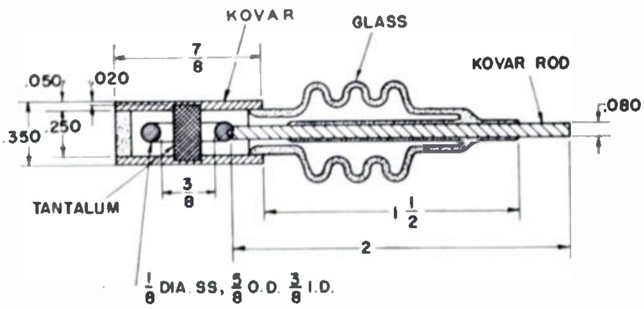

Fig. 4—Scale drawing.

any point in the interlectrode space, providing that the magnetic field is sufficiently large. Referring to Fig. 2(C), an electron originating at P, a distance r_1, from the center, will traverse a path such as shown, for fields such that $H > H_c$.

Fom Hull's[6] magnetron theory, H_c is given by

$$H_c = \sqrt{\frac{8mV_2}{e}} \frac{R}{R^2 - r_1^2}, \tag{1}$$

where the quantities are defined in Fig. 2(C). This will now be modified by introducing an expression for V_2. Normally, in a concentric cylinder arrangement, the potential varies logarithmically between the cylinders; however, in the present case, because of the cathode end plates, and the ring-shaped anode, this potential variation becomes more linear. Hence it will be assumed that

$$V_2 = V \frac{R - r_1}{R - r_0}.$$

Using this expression, and also letting the cutoff field for $r_1 = r_0$ be H_{c0}, yields

$$\left(\frac{H_c}{H_{c0}}\right)^2 = \frac{\left(1 + \frac{r_0}{R}\right)^2 \left(1 - \frac{r_0}{R}\right)}{\left(1 + \frac{r_1}{R}\right)^2 \left(1 - \frac{r_1}{R}\right)}. \tag{2}$$

This result is plotted in Fig. 5, for the case where $r_0/R = 0.25$. This graph shows that if $H_c = H_{c0}$ all electrons originating between the cathode surface, $r_1/R = 0.25$, and the surface, $r_1/R = 0.45$, are cut off and fail to reach the anode, as shown in Fig. 6(A), where the cross-hatched area represents the region of origin of all cutoff electrons. If $H_c = 2H_{c0}$, the cutoff region extends from the cathode to about $r = 0.92R$, as shown in Fig. 6(B). An electron originating at $r = 0.92R$ will just graze the anode, as shown at a. An electron originating in the central region ($r \approx R/2$) will have an orbit such as shown at b.

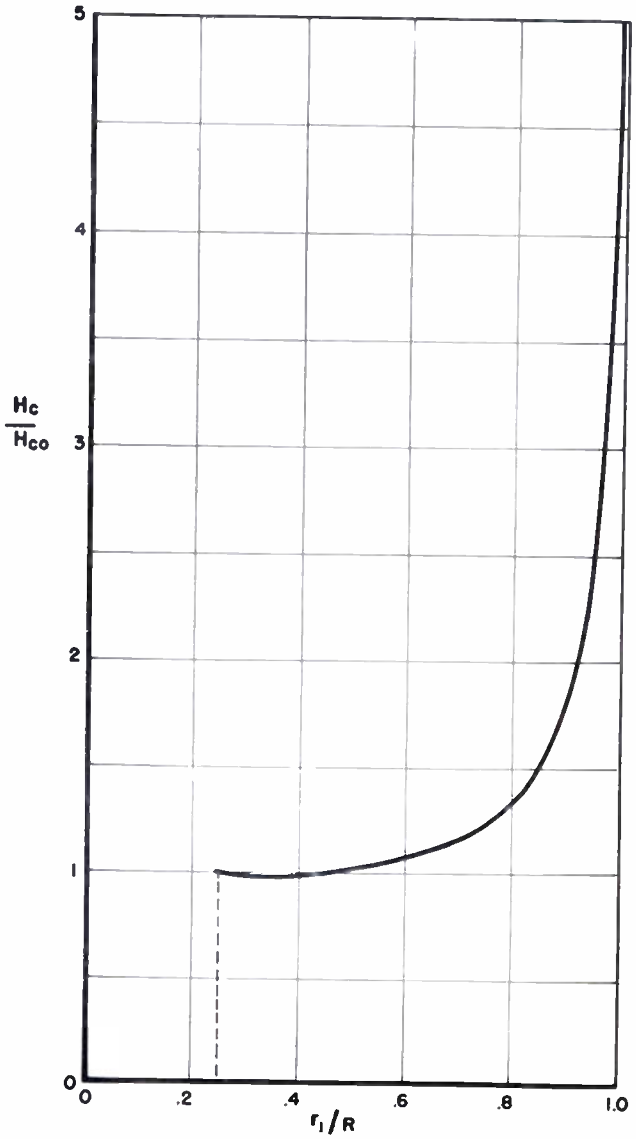

Fig. 5—Cutoff fields for electrons originating in the interelectrode space.

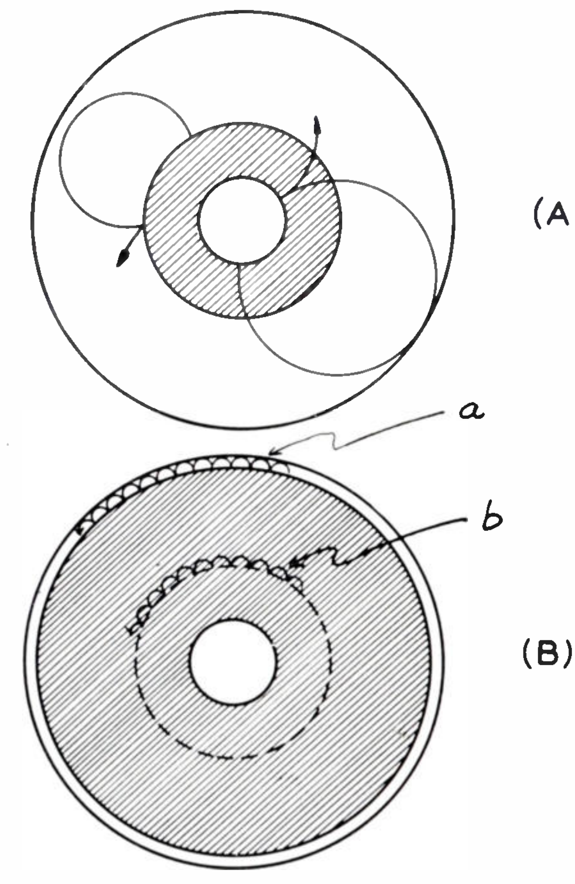

Fig. 6—Typical orbits with coaxial cylinders.

The orbits of Fig. 6 are those of the simple cylindrical arrangement of Fig. 2(B). In the structure of Fig. 2(A) they become more complicated in that motion in the axial direction also occurs.

In Fig. 7 side views of the tube are included. The case where the electron originates in the equatorial plane is shown at A. Here the orbit is planar, and has already been shown in Fig. 6. If the electron originates off the equatorial plane, it will have an axial component of motion. This will be oscillatory, back and forth through the equatorial plane, since the electron will be attracted by the anode and repelled by the end plates. This oscillatory motion will be accompanied by a curvilinear azimuthal motion, as shown in Fig. 7(B).

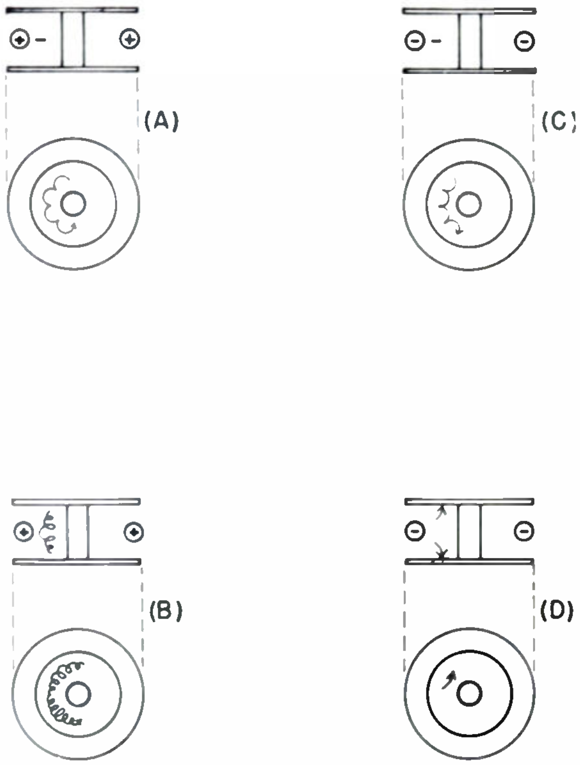

Fig. 7—Typical electron orbits.

The situation is significantly different when the anode is negative. The equatorial electrons then move as shown in Fig. 7(C); the nonequatorial electrons move in short spiral paths to the end plates, where they are collected, as shown at D.

These orbits illustrate the principle of "electron trapping." For example, in the orbits of Figs. 7(A) and 7(B), the electrons are unable to reach any electrode. The magnetic field curves their paths in such a way that no electrons reach the anode. The cathode exerts a repelling effect since its potential is negative with respect to the point of origin of the electrons, except for the small fraction which originate at the cathode itself. Hence, substantially all electrons are trapped. They remain in the interelectrode space until disturbances, such as collisions, deflect them into new paths, and eventually to the anode.

When the anode is negative, this trapping effect is greatly reduced. The paths are then as shown in Figs. 7(C) and 7(D). The equatorial electrons are trapped, as at (C), but the nonequatorial ones, which actually constitute almost the entire number, move in short paths directly to the cathode end plates, as shown at D.

Hence, when the anode is positive, there is a pronounced trapping effect so that a high density of long-path electrons is built up; whereas if the anode is negative, the electrons are quickly removed, the density is low, and the path length short. These conditions greatly affect the conduction through the tube, and are responsible for its rectifying action.

The relation of the trapping effect to conduction may be explained by use of the Townsend theory of conduction in gases.[10] Briefly, this theory assumes that electrons emitted from the cathode build up an electron avalanche by a process of cumulative ionization. Positive ions, formed in this process, return to the cathode and cause new electrons to be emitted. This is a feedback phenomenon. Under proper conditions the feedback may be large enough so that a continuous, self-sustaining discharge is formed.

The current i in a Townsend discharge is related to the primary current i_0 at the cathode by the relation

$$m = \frac{i}{i_0} = \frac{\epsilon^{\alpha x}}{1 - \gamma(\epsilon^{\alpha x} - 1)}, \tag{3}$$

where α is the average number of electrons, or ions, formed per centimeter of path per electron, γ is the number of electrons emitted from the cathode per positive ion striking its surface, and x is the electron-path length between cathode and anode. It is evident that for certain values of αx and γ, m will be infinite. This corresponds to the onset of a self-sustaining discharge. The quantity m is plotted as a function of αx and γ in Fig. 8(A).

From this figure it is seen that, for example, at point P_1 a self-sustaining discharge will not occur. This point might correspond to the present rectifier with anode negative since x is then small. However, if the anode is positive, electron trapping occurs, and x becomes large, corresponding to point P_2. A self-sustaining discharge will then start since conditions are such that the boundary $m = \infty$ is crossed. The tube will act as a rectifier whenever P_1 and P_2 are on opposite sides of this boundary.

It is clear that the shape of the orbit, especially its length, affects the value of x. The orbit also affects the value of α. This may be seen from Fig. 8(B), which shows a typical orbit of a trapped electron. Assuming that the initial velocity of the electron is negligible, the maximum energy that it may have in the illustrated orbit is ΔVe. A necessary condition for the formation of a discharge is that ΔV exceed the ionization potential. The amount of this excess is determined by the value of α required to satisfy the Townsend condition and yield a value of $m = \infty$.

As before, assuming a linear potential distribution between cathode and anode, the following relations are valid:

[10] J. D. Cobine, "Gaseous Conductors," McGraw-Hill Book Co., Inc., New York, N. Y., p. 156 and following; 1941.

$$\frac{R}{\Delta r} = \frac{V}{\Delta V} \tag{4}$$

$$E = \frac{V}{R} = \frac{\Delta V}{\Delta r}. \tag{5}$$

From the equation of electron motion,[11]

$$\Delta r = \frac{2mE}{H^2 e}. \tag{6}$$

From (4), (5), and (6),

$$H = \sqrt{\frac{2m}{e}}\,\frac{1}{R}\,\frac{V}{\sqrt{\Delta V}}$$
$$= \frac{3.375}{R}\,\frac{V}{\sqrt{\Delta V}}. \tag{7}$$

For magnetic fields greater than those given by (7) the loops of the orbit would be too small, so that ΔV would not exceed the ionization potential. Therefore, a discharge would not occur. Hence, it may be said that H controls the value of α because if H exceeds the value given (7) the number of ionizations per electron per centimeter will be small, i.e., α will be small.

The quantity γ is determined in an important manner by the voltage, since this controls the energy with which the positive ions strike the cathode. Below several hundred volts, γ is too small to permit a self-sustaining discharge.

From the foregoing discussion three regions may be outlined wherein a self-sustaining discharge cannot occur. These are as stated on the following page:

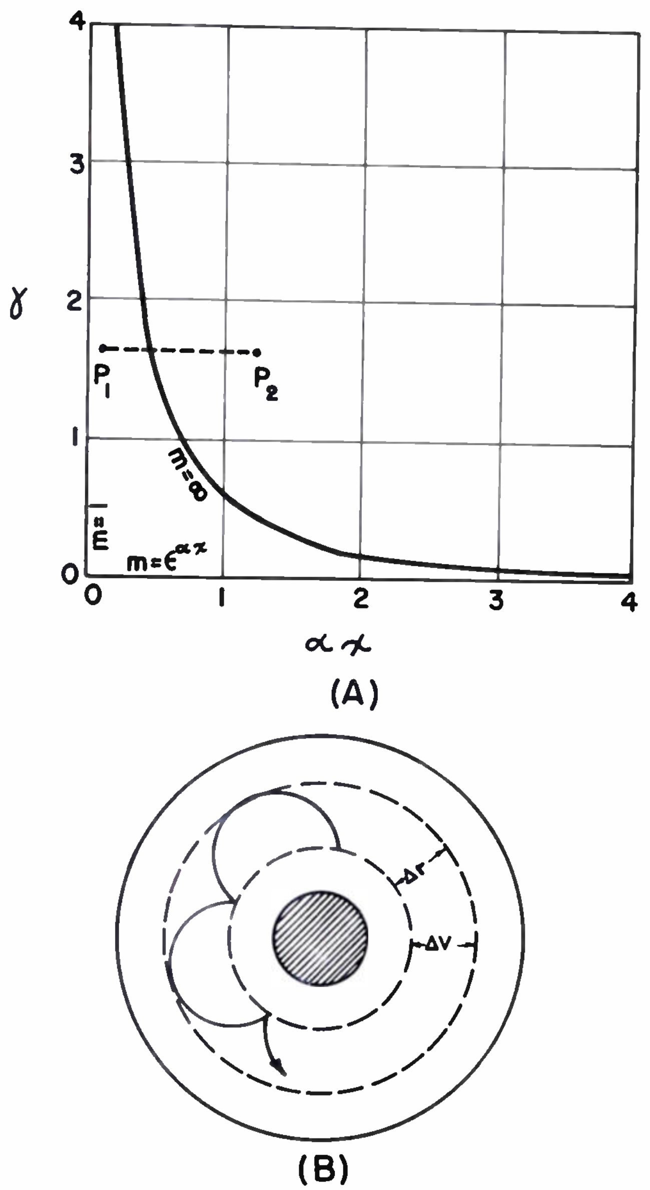

Fig. 8—Illustrating criteria for initiation of discharge.

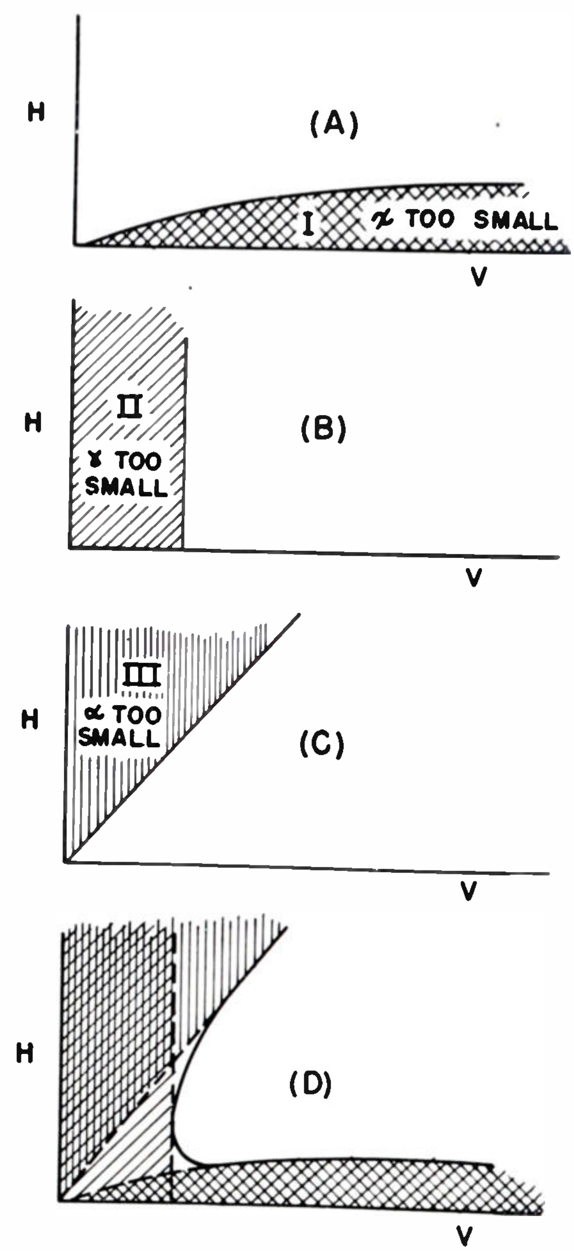

Fig. 9—Composition of cutoff curve.

[11] K. R. Spangenberg, "Vacuum Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., p. 116 and following; 1948.

1. The region below cutoff, where H is less than the value given by (1). Here no long paths are possible, and hence x is too small.
2. A region of small V, where γ is too small, since the cathode is bombarded by ions of insufficient energy.
3. A region where the magnetic field is too high, so that the path loops are too small to permit the electrons to gain enough energy for efficient ionization. Thus, α is too small.

These three regions are shown in Figs. 9(A), (B), and (C), and are combined in Fig. 9(D) to show the complete characteristic. Experimental curves of this type will be discussed later, and it will be seen that there is fair agreement with the above theory.

A somewhat more complete theoretical discussion will be given in a forthcoming paper on this tube in the *Proceedings of the National Electronics Conference* for 1951.[12]

III. Determination of Rectification Criteria

To make a preliminary study of various electrode geometries a tube was constructed as shown in Fig. 10(D). Separate leads were brought out from the end plates, inner cylinder, ic, and outer cylinder, oc, and their potential varied. A summary of the results obtained is given in Table I.

TABLE I
Electrode Connections for Tube of Fig. 10

Test no.	Inner cylinder potential	Outer cylinder potential	End-plate potential	Conduction
1	Ground	Positive	Ground	Yes
2	Positive	Ground	Ground	Yes
3	Positive	Positive	Ground	Yes
4	Positive	Ground	Positive	No
5	Ground	Positive	Positive	No
6	Ground	Ground	Positive	No

All combinations of electrode polarities were tried. It was found that Tests Nos. 1, 2, and 3 showed conduction in the normal manner; whereas Tests Nos. 4, 5, and 6 showed no conduction under any magnetic field, up to the flashover voltage point. These tests indicate that any of the structures of arrangements nos. 1, 2, and 3 will rectify. Arrangement no. 4 is the reversal of potentials on no. 1, 4 the reversal of 2, and 6 the reversal of 3. A sketch of structures based on these arrangements is shown in Fig. 10(A), (B), and (C), with the electrode representing the cathode indicated as K. The letters A, B, and C represent arrangements nos. 1, 2, and 3, respectively.

These tests indicated that the structure used in the present rectifier withstood higher inverse voltages than the rectifier of Bush and Smith[6] because of the presence of the end plates. Also, these tests showed that the structure of Strutt,[3] as used in a vacuum gage by Penning,[8] would rectify as in Fig. 10(B). The structure of Fig. 10(A) had not been tried previously as a rectifier. Tests were then made to determine the relative merits of the three structures.

In these tests, separate tubes were built, and tests were made at 60 cps. Structure 10(C) was consistently inferior in inverse voltage insulation to the other two. Thus it was decided to explore the various geometries of 10(A) and 10(B), as a practical rectifier.

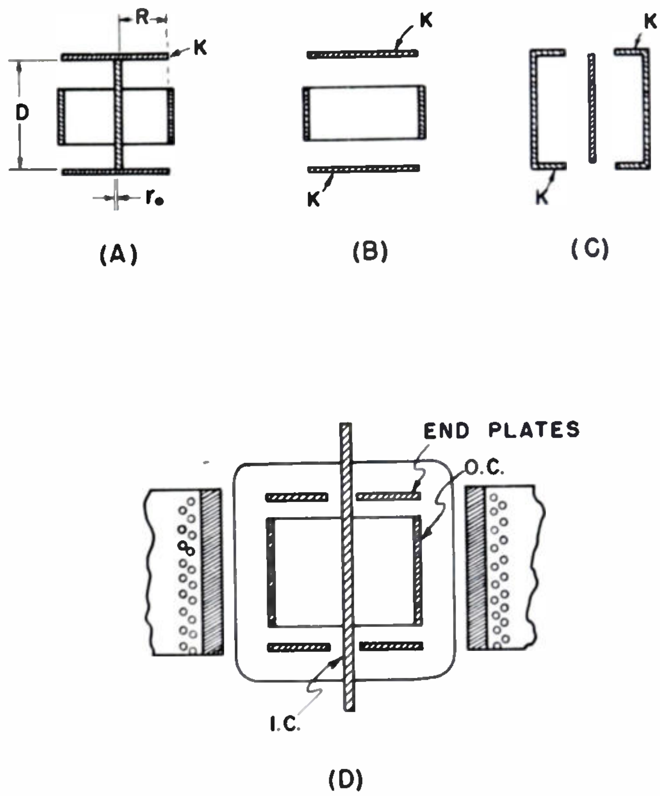

Fig. 10—Electrode geometries surveyed.

IV. DC Characteristics

Two types of dc characteristics, magnetic-cutoff and current-voltage, were taken for different geometries. Three geometrical dimensions were varied, anode radius (R), cathode radius (r_0), and end-plate separation (D), indicated in Fig. 10(A). The structure of Fig. 10(B) was treated as the limiting case of Fig. 10(A) with zero r_0.

A. Magnetic Cutoff

A range of anode radii, (R) from one inch to 3/32 inch, was used to determine the general variation in cutoff with geometry. Two anode radii (R), 3/16 and 3/32, were then selected in the practical range for a rectifier and their cathode radii (r_0), and the end-plate separation varied.

For the 3/16 inch R, D was made in three dimensions, $\frac{1}{2}$ inch, $\frac{3}{8}$ inch, and $\frac{1}{4}$ inch. For each D the r_0 was made $\frac{1}{8}$ inch, $\frac{1}{16}$ inch, and $\frac{1}{32}$ inch, giving a total of nine variations. Typical curves are plotted in Fig. 11. The tube with no center pin ($r_0=0$) is included. The large differ-

[12] E. G. Linder, J. H. Coleman, and E. G. Apgar, "A high-voltage electron-trap rectifier," *Proc. NEC* (Chicago); 1951.

ence between curves 1 and 2 is caused by the cathode radius approaching that of the anode. As is seen from (1), H_c increases rapidly as $r_0(=r_1)$ approaches R.

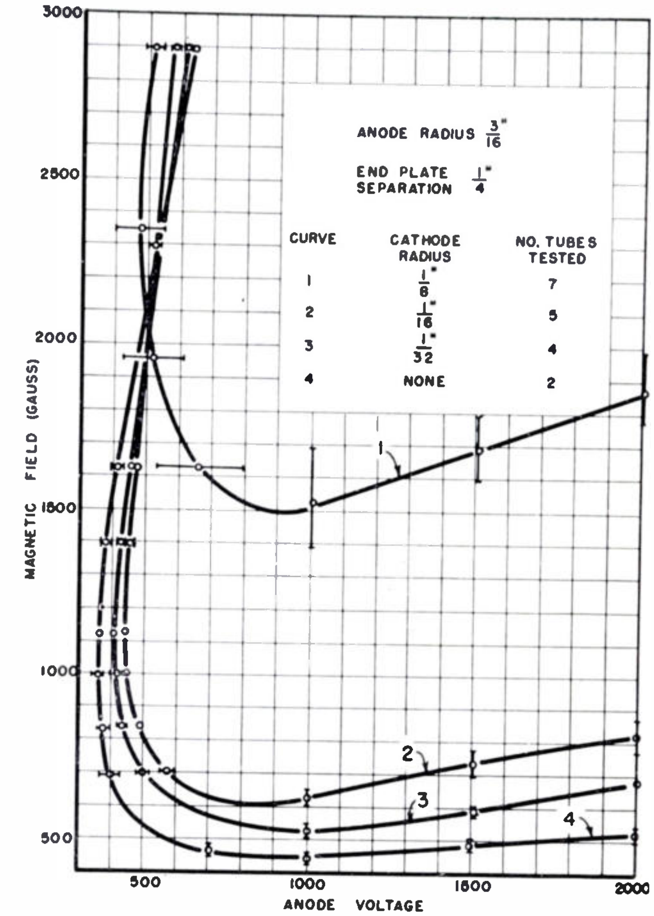

Fig. 11—Typical cutoff curves.

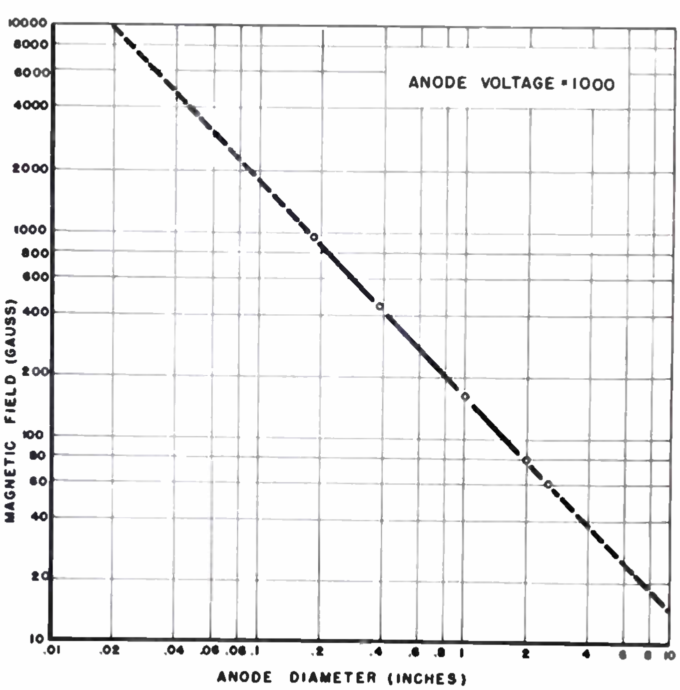

Fig. 12—Relation between anode diameter and cutoff field.

The almost vertical sections of these curves were taken with a fixed magnetic field and by noting the voltage at which current first began to flow. The sloping lower sections were taken with a fixed voltage and by noting the start of conduction as the magnetic field was increased.

The cutoff field at 1,000 volts is plotted in Fig. 12 for the series of tubes with different anode radii and small cathode radii. As the minimum magnetic field occurs near the 1,000-volt point, for all tubes in this series, this curve, including the extrapolated dashed sections, serves as a guide in predicting the operating magnetic field for any desired anode radius. The relationship shown here represents that of (1), for $R \gg r_1$ and $r_1 = r_0$.

To check the effect of cathode radius for a particular anode radius, the magnetic field was calculated from (1) and plotted as a function of r_0 in Fig. 13, for the two R,

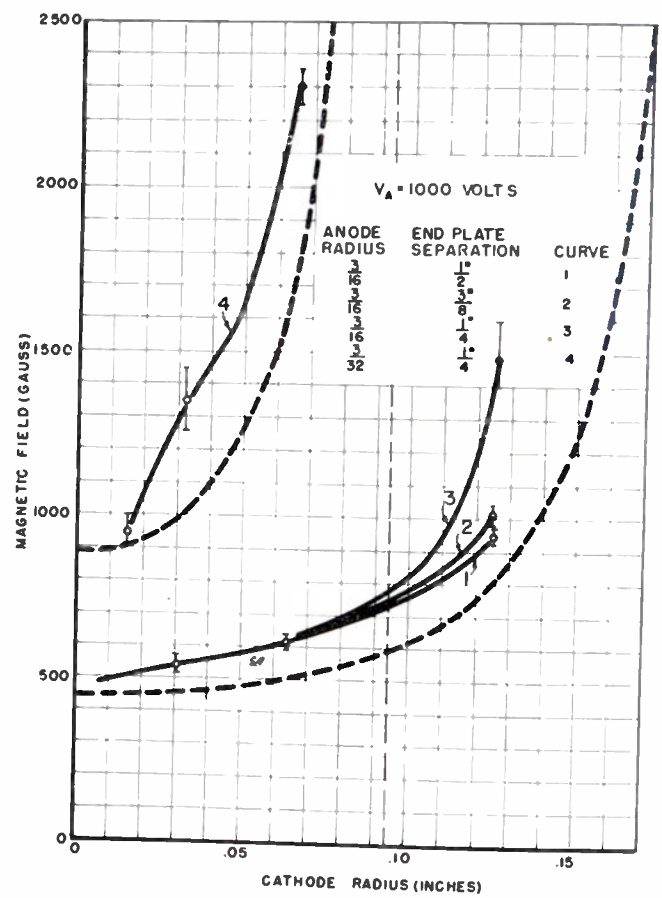

Fig. 13—Cutoff field versus cathode radius.

3/16 inch and 3/32 inch. It is seen that the tubes with the widest end-plate separation (D) approach closest to the theoretical curve. The experimental values are higher than the theoretical, probably because electrons originating in the interelectrode space, as well as at the cathode, must be cut off. See Fig. 5 and its discussion. Again these curves can be used to predict the minimum magnetic field for a particular r_0.

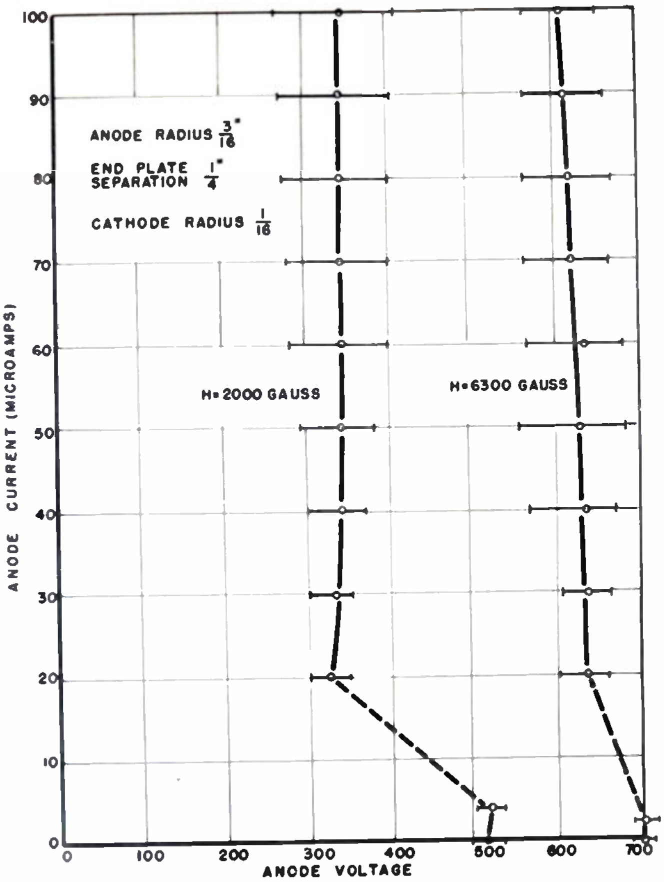

Fig. 14—Anode current versus anode voltage.

B. Current-Voltage

The anode current was measured at constant magnetic field as a function of voltage across the tube for currents up to 100 microamps. Higher currents were not used in these dc tests because of temperature changes arising from the increased power input. Typical results are shown in Fig. 14. It is seen that below 20 microamperes there is an unstable region. Between 20 and 100 microamperes the voltage drop varies but slightly with current.

V. AC Characteristics

The rectification efficiency at different frequencies is affected by two factors: starting time for conduction on the positive cycle; and deionization time, which may lead to conduction on the negative cycle.

A. Starting Time

The time required for the residual charges in the interelectrode space to build up to a steady discharge by cumulative ionization after the anode voltage is applied is termed the "starting time." The experimental results are shown in Fig. 15, where the starting time is plotted as a function of the magnetic field for different voltage amplitudes of an applied square pulse. The time was measured by an oscilloscope across a 200-ohm resistance in series with the tube.

B. Deionization Time

The second factor, deionization time, was found to be the predominant one in frequency limitation. A tube

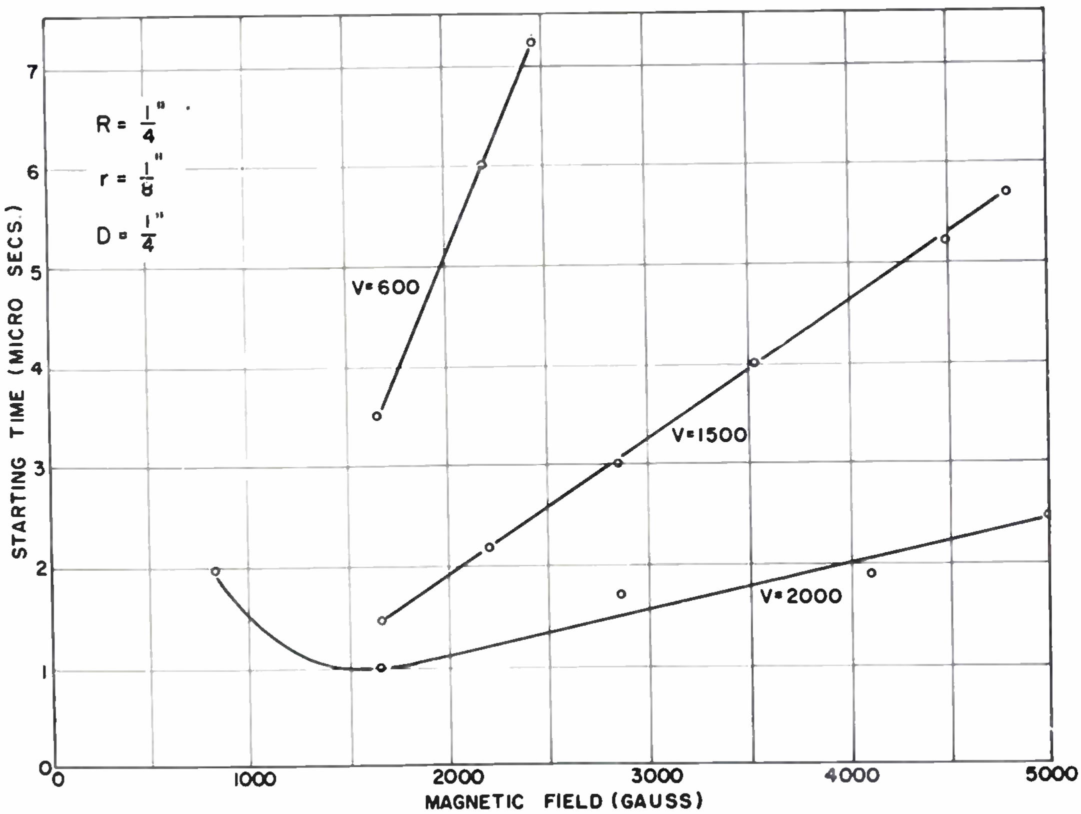

Fig. 15—Starting time.

with one-half-inch R operated satisfactorily at 90 kc with neon gas at 10^{-3}-mm Hg pressure, but gave no dc output with mercury vapor at the same pressure. Calculations of ambipolar diffusion rates showed the positive-ion current would decrease to $1/e$ of the initial value in 0.6 microsecond for neon, but would take 11.6 microseconds with mercury vapor. At 90 kc, this diffusion time for mercury is long compared to the duration (of 6 microseconds) of the positive half cycle, resulting in positive ions remaining in the interelectrode space on the negative half cycle. As the decay time varies with the square of the electrode spacing, calculations showed that for R equal to 3/16 inch the characteristic time should be reduced to 2 microseconds. A tube based on these calculations gave satisfactory operation at 90 kc.

C. Frequency Characteristics

To determine the rectification efficiency and the regulation characteristics as functions of frequency, the tubes were tested at 60 cps, 60 kcps, 120 kcps, and 240 kcps. The results for 60 cps are shown in Fig. 16, together with similar data for the 1B3-GT thermionic vacuum rectifier for comparison. A tube having slightly convex end plates is also included. As the frequency increases, the internal voltage drop, for all tubes, increases

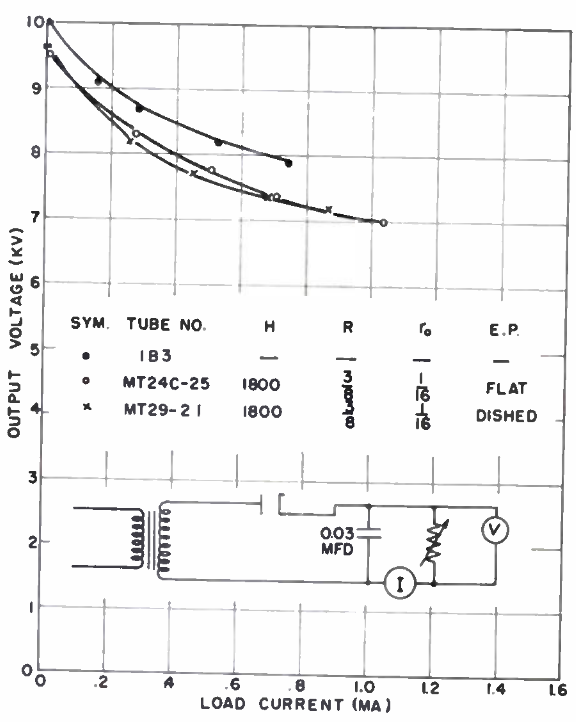

Fig. 16—Single-tube rectifier data at 60 cps.

slightly. The 120 kcps data are shown in Fig. 17. The 60 kcps results are intermediate between these and the 60 cps data, while the 240 kcps results show an output of about 7.5 kv for no current, dropping to 4 kv at one ma. However, the 240 kcps tests showed erratic performance. Some tubes did not conduct at all except with excessive magnetic fields. This frequency appears to lie beyond the limit for satisfactory operation for this particular design of rectifier.

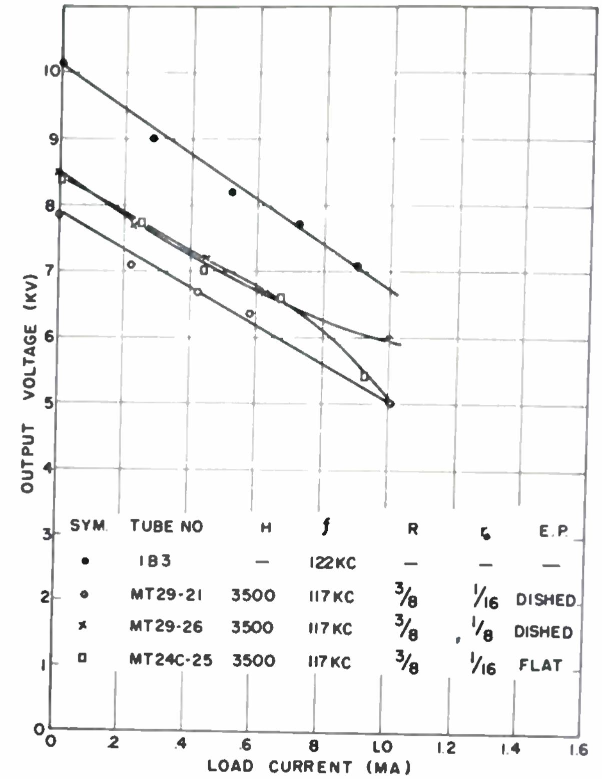

Fig. 17—Single-tube rectifier data at 117 kcps.

The variation of the cutoff characteristic with frequency is shown in Fig. 18 (see following page). It is seen that as frequency increases the voltage required to initiate conduction also increases. For example, at $H=2{,}000$ gauss, the starting voltage is about 500 volts higher at 200 kcps than at 60 cps. This difference is of the same magnitude as the increase in tube drop when this tube is used as a rectifier at these same frequencies.

VI. Temperature Characteristics and Peak Inverse Voltage

The use of saturated mercury vapor in this rectifier imposes both low- and high-temperature limits. When the temperature is reduced below room temperature, conduction requires a higher magnetic field because of a shift of the cutoff curves upward along the magnetic axis. For example, the minimum magnetic field for curve 2 of Fig 11 rises to 1,000 gauss at 0 degrees C, whereas it is 600 gauss at 20 degrees C.

When used in a half-wave rectifier circuit at 60 cps, the output voltage decreases as the temperature falls below 20 degrees C, decreasing by about 25 per cent at 0 degrees C. However, between 20 and 75 degrees C (the highest tried) the output voltage is substantially constant.

As the temperature rises, the inverse breakdown voltage is reduced because of the increase in mercury-vapor

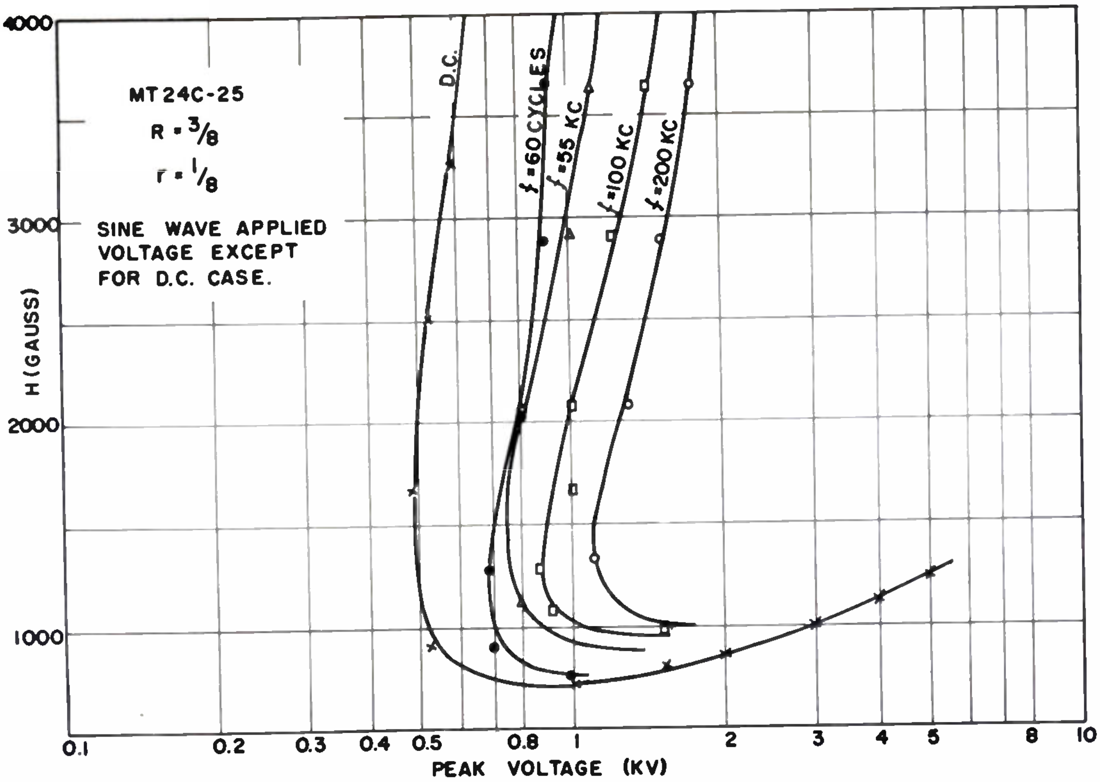

Fig. 18—Variation of cutoff curve with frequency.

pressure. A plot of the peak inverse voltage as a function of temperature (and pressure) is shown in Fig. 19 for two geometries. The larger tube shown in curve 2 breaks down at lower voltages in accordance with Paschen's law.

No inverse breakdown voltage data were taken with dc voltages above 25 kv. However, tubes were operated at 60 cps with inverse breakdown voltages up to about 40 kv.

VII. Life Tests

Extensive life tests have not been made because of lack of sufficient time. However, a life up to 3,000 hours has been obtained with operation at 11-kv output and 0.4 ma of current. At an output of 14 kv and 0.4 ma, a tube currently under test has operated 900 hours with no evidence of deterioration.

In early models of the tube, cathode sputtering severely limited life. The low gas pressure used in this type of tube, coupled with the high voltage, permits high-energy positive ion bombardment of the cathode resulting in rapid cathode sputtering. This caused the walls of the tube to become coated with metal, and led to eventual breakdown of insulation. In tests of a series of materials, both graphite and tantalum were found to have low sputtering rates. Tantalum was eventually used for the cathode center pin since graphite did, after about 1,000 hours, form a carbon desposit on the walls which flaked off and caused flashovers in the tube.

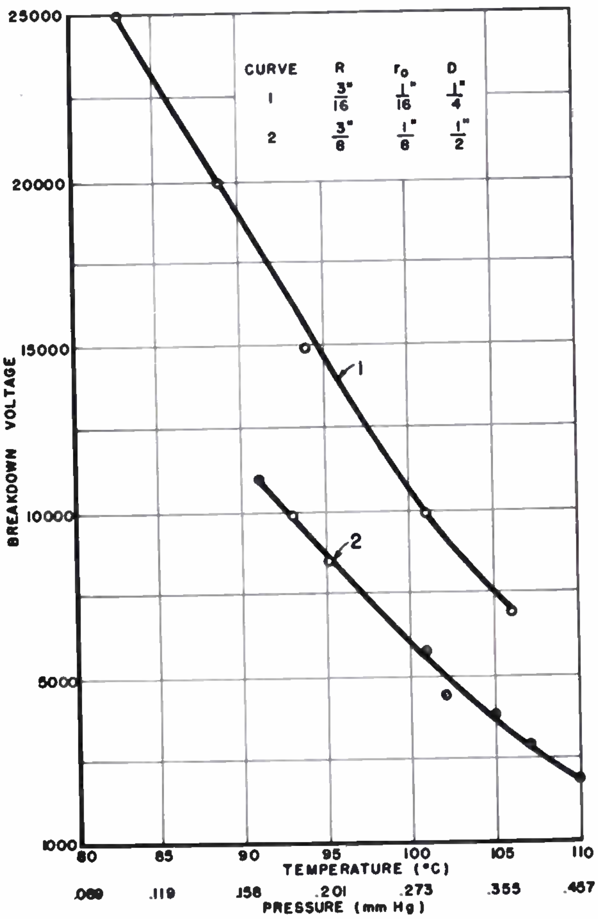

Fig. 19—Inverse breakdown voltage as a function of temperature and mercury vapor pressure.

Gas cleanup also was a serious limitation in early tests. Gases such as neon, argon, helium, and xenon lasted less than an hour or so. These were quickly abandoned in favor of saturated mercury vapor.

VIII. Doubler and Tripler Applications

The advantages of a cold-cathode, high-voltage rectifier such as the present tube are especially evident in the case of voltage multiplier rectifier circuits. Valuable simplification of the circuit results since heater transformer windings with high-voltage insulation are no longer required.

Data obtained with a voltage-doubling rectifier are plotted in Fig. 20. A curve with 1B3-GT tubes is included for comparison. The peak input voltage was 10 kv for both tubes.

Similar data were taken for a tripler circuit. With a peak input voltage of 10 kv, an output of 28 kv was obtained at zero current, dropping to 22 kv at 0.7 ma.

IX. Acknowledgment

The writers are indebted to Messrs. R. F. Snyder and I. H. Sublette, of the RCA Victor Division, Camden, N. J., for much of the experimental data on the use of this tube in rectifier circuits.

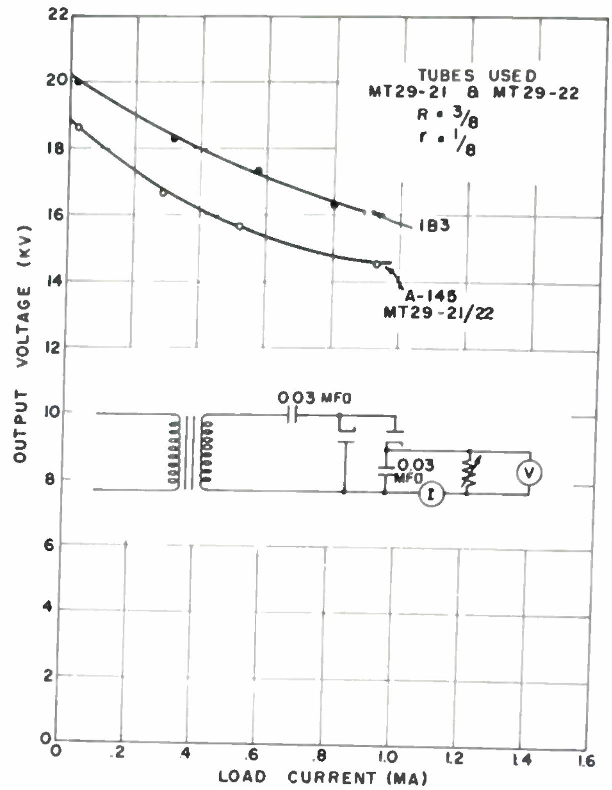

Fig. 20—Voltage doubler rectifier data at 60 cps.

Piezoelectric Transducers for Ultrasonic Delay Lines*

H. N. BEVERIDGE†, ASSOCIATE, IRE, AND W. W. KEITH†

Summary—The over-all frequency response and loss of the transducers of a delay line are derived from Roth's equivalent circuit. The transient response of a delay line is computed from an equivalent transmission-line circuit. The bandwidths and transient responses measured on practical delay lines are given for comparison with theoretical values.

Introduction

ULTRASONIC DELAY LINES are useful for the dynamic storage of wide-band intelligence in applications such as the high-speed memory of a digital computer[1] and the canceller of a moving target indicator.[2] The following discussion involves the theory of the piezoelectric transducer as used in delay lines, or similar ultrasonic devices, where the acoustic loading is much greater than for high-Q crystals in frequency stabilizing circuits or narrow-band filters. In most cases the electromechanical coupling is small, allowing simplifying approximations to be made with respect to the interaction of electrical circuitry on acoustic frequency response. The resulting expressions for the acoustic steady-state and transient responses are independent of all factors except normalized acoustic loading.

Only the longitudinal mode of vibration in thickness is considered in this analysis, but the normalized acoustic frequency and transient responses of a transducer operating in the shear mode is the same as for the longitudinal mode due to identical boundary conditions when shear group velocities and acoustic impedances are substituted in place of the corresponding longitudinal values. Identical transmitting and receiving transducers are assumed. For purposes of analysis, the loading media at the transducer faces are assumed to extend to infinity; hence, reflections in the media are neglected. In practice, the attenuation of the media usually is sufficient to minimize the effect of reflections.

Except where otherwise indicated, the rationalized mks system of units is used.

I. Steady-State Analysis

Steady-state analysis of the piezoelectric transducer is based on an equivalent circuit derived from fundamental physical equations by Dr. W. Roth.[3] His circuit is valid for all conditions of loading, and perfectly general expressions result from its use. The general expres-

* Decimal classification: R265×R117.19. Original manuscript received by the Institute, May 20, 1951; revised manuscript received, March 10, 1952.

† Raytheon Mfg. Co., Newton, Mass.

[1] C. F. West and J. E. DeTurk, "A digital computer for scientific applications," Proc. I.R.E., vol. 36, pp. 1452–1460; December, 1948.

[2] L. N. Ridenour, "Radar System Engineering," McGraw-Hill Book Co., New York, N.Y., pp. 626–679; 1948.

[3] W. Roth, "Piezoelectric transducers," Proc. I.R.E., vol. 37, pp. 750–758; July, 1949.

sions are involved, however, and cannot be made independent of crystal dimensions and external circuit constants unless simplifying approximations are valid.

Transmitting Transducer

Medium 1 of acoustic impedance z_1 is arbitrarily designated as the transmission medium, and medium 2 of acoustic impedance z_2 is designated as the transducer backing material. In this analysis the acoustic impedances are assumed to be resistive and independent of frequency. Assuming a zero impedance source E_3 for electrical drive, the equivalent circuit is solved by straightforward network analysis for the volume velocity Av_1 in mesh 1 using (30) and (34), page 757 of footnote reference 3.

$$Av_1 = \frac{\dfrac{-E_3}{j\omega C_p}}{\dfrac{z}{j\omega C_E A}\left[\dfrac{(\zeta_1+\zeta_2)\cos\delta\pi + j(1+\zeta_1\zeta_2)\sin\delta\pi}{\cos\delta\pi - 1 + j\zeta_2 \sin\delta\pi}\right] + \dfrac{1}{\omega^2 C_p^2}\left[1 + \dfrac{\cos\delta\pi - 1 + j\zeta_1\sin\delta\pi}{\cos\delta\pi - 1 + j\zeta_2\sin\delta\pi}\right]}, \quad (1)$$

where C_p is the piezoelectric capacity, C_E is the electrostatic capacity of the active area A of the transducer, z is the acoustic impedance of the transducer crystal $(=\rho c)$, and where

$$\zeta_1 = \frac{z_1}{z}, \quad (2a)$$

and

$$\zeta_2 = \frac{z_2}{z}. \quad (2b)$$

The normalized frequency δ is given by

$$\delta = \frac{\omega}{\omega_0} = \frac{f}{f_0}, \quad (3)$$

where f_0 is the frequency of fundamental resonance of the transducer. Using the relation that[4]

$$C_p^2 = \frac{\pi A C_E}{K^2 z \omega_0}, \quad (4)$$

it can be shown that, if

$$\frac{\pi(\zeta_1+\zeta_2)}{4K^2} \gg 1.0, \quad (5)$$

the left-hand term of the denominator in (1) is much greater than the right-hand term, and

$$Av_1 \simeq \frac{-E_3 C_E A}{z C_p}\left[\frac{\cos\delta\pi - 1 + j\zeta_2\sin\delta\pi}{(\zeta_1+\zeta_2)\cos\delta\pi + j(1+\zeta_1\zeta_2)\sin\delta\pi}\right]. \quad (6)$$

The electromechanical coupling coefficient K is a dimensionless constant of the transducer crystal. For X-cut quartz, $K \cong 0.1$; hence, (6) is valid for most delay lines.

[4] Equation (4) is derived from (29) and (34), *ibid.*, page 757, using the relation that $z = \rho c$.

At fundamental resonance, $\delta = 1$, and

$$Av_{1_0} = \frac{-2E_3 C_E A}{z C_p(\zeta_1 + \zeta_2)}. \quad (7)$$

Normalizing Av_1 with respect to Av_{1_0},

$$\frac{Av_1}{Av_{1_0}} = \frac{\zeta_1+\zeta_2}{2}\left\{\frac{\cos\delta\pi - 1 + j\zeta_2\sin\delta\pi}{(\zeta_1+\zeta_2)\cos\delta\pi + j(1+\zeta_1\zeta_2)\sin\delta\pi}\right\}, \quad (8)$$

which is the normalized volume-velocity. When the magnitude or phase angle of (8) is plotted against normalized frequency for various loading conditions, the resulting family of universal response curves is independent of crystal constants, crystal dimensions, and circuit constants.

The normalized volume-velocity is a periodic function of frequency, having a magnitude of unity at fundamental resonance and at odd harmonics ($\delta = 1$, 3, 5, and so on), and a magnitude of zero at zero frequency and at even harmonics ($\delta = 0$, 2, 4, etc.). The magnitude response curve is symmetrical about the resonant frequencies. For certain loading conditions, two equal peaks greater than unity are symmetrically located about the resonant frequencies.

The transmitting transducer is frequently driven from the plate of a pentode tube using an inductance L which tunes the transducer capacity C_E and stray capacity C_S to the fundamental resonant frequency and a suitable damping resistance R in parallel with L for the appropriate electrical bandwidth. In practical cases the loading due to the radiation resistance of the transducer may be neglected; hence, the input impedance is the reactance of C_E, and the input voltage E_3 of the transducer is

$$E_3 = g_m e_g Z_L, \quad (9)$$

where Z_L is the impedance of the tuned circuit, g_m is the transconductance, and e_g is the grid voltage. When driven by a tube, the frequency response of the transducer is the product of the acoustic response given by (8) and the response of the tuned circuit of R, L, C_E, and C_S in parallel. Since the frequency response and bandwidth of a parallel resonant circuit are well known, it is convenient to deal with the acoustic response and the response of the circuitry separately.

Since z_1 is resistive, the acoustic power W_1 radiated into the transmission medium is proportional to the square of volume-velocity magnitude. It follows that, for the normalized power,

$$\frac{W_1}{W_{1_0}} = \left|\frac{Av_1}{Av_{1_0}}\right|^2 \quad (10)$$

where Av_1/Av_{1_0} is given by (8) for a zero impedance source.[5] Shown in Fig. 1 is the acoustic power response over the fundamental mode for X-cut quartz crystals and mercury as the transmission medium ($\zeta_1 = 1.29$), with several values of backing impedance. The "acoustic

[5] Equation (10) gives the same results as (15) in the following: H. B. Huntington, A. G. Emslie, and V. W. Hughes, "Ultrasonic delay lines I," *Jour. Frank. Inst.*, vol. 245, page 10; January, 1948.

power response" is a plot of (10) versus normalized frequency. This equation may be solved for the acoustic power bandwidth Δf at a fractional power level $W_1/W_{1_0} = h$.

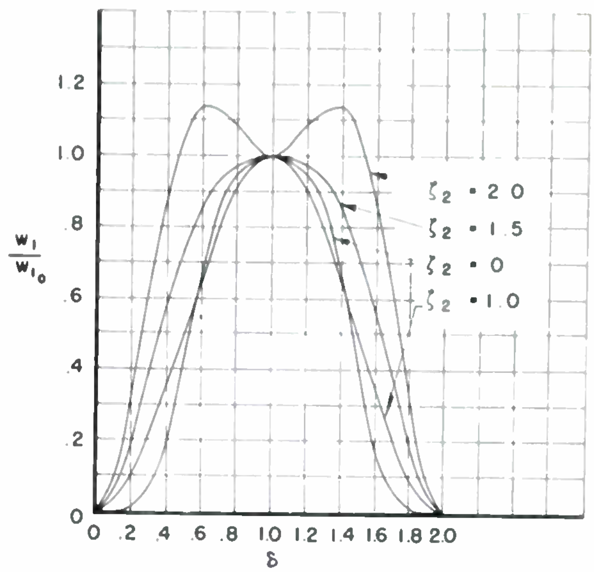

Fig. 1—Power response for mercury ($\zeta_1 = 1.29$).

Of importance in delay line design are the special cases of symmetrical loading ($\zeta_1 = \zeta_2 = \zeta$) and air backing ($\zeta_2 \cong 0$). By virtue of the crystal being a half-wave

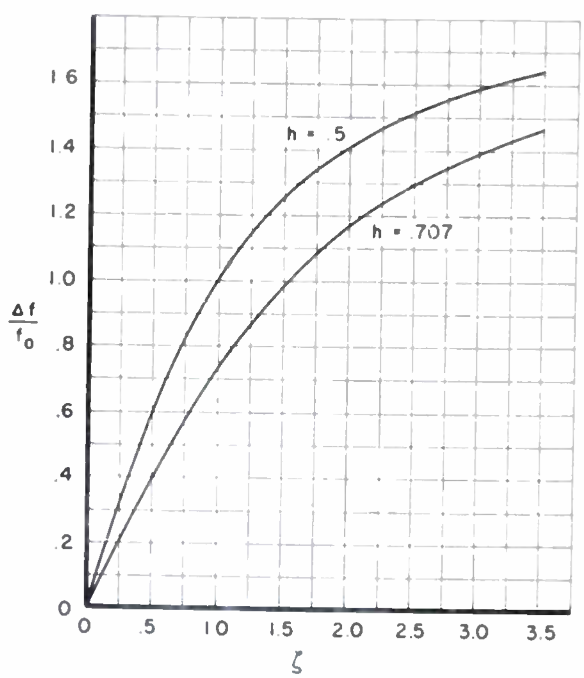

Fig. 2—Fractional bandwidths for symmetrical loading ($\zeta = \zeta_1 = \zeta_2$).

in thickness at resonance, reflection of incident energy at the transducers can be reduced by a backing material having the same acoustic impedance as the transmission medium. When reflections are unimportant, however, lower losses result if the transducers are backed by air. In mercury delay lines a metal plate not in intimate contact with the transducers is used as a backing for mechanical support. The air trapped in surface irregularities of the backing plate destroys the acoustic contact, approximating the conditions for air backing.

Fractional half-power and 0.707-power acoustic bandwidths $\Delta f/f_0$ as functions of loading are shown in Figs. 2 and 3 for symmetrical loading and air backing, respectively.

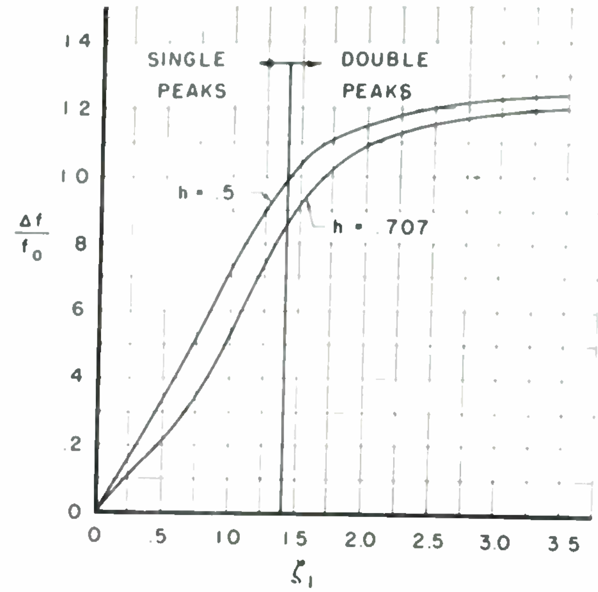

Fig. 3—Fractional bandwidths for air backing ($\zeta_2 \cong 0$).

It should be noted double-peaking does not occur for symmetrical loading. For air backing, the fractional bandwidth between peaks and height of peaks in response as functions of loading are shown in Fig. 4.

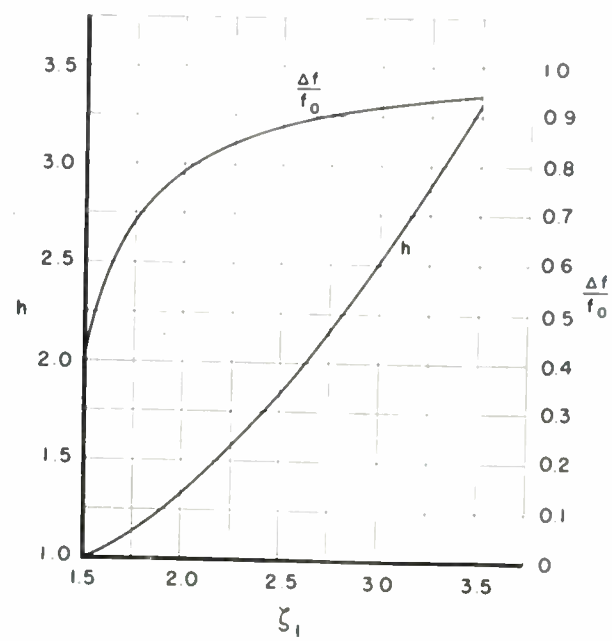

Fig. 4—Bandwidth and height of peaks for air backing ($\zeta_2 \cong 0$).

For X-cut quartz transducers, and mercury as the transmission medium ($\zeta_1 = 1.29$), the fractional bandwidths are shown in Fig. 5 as functions of the backing material impedance.

Over-All Delay Line Response

The over-all frequency response and attenuation of a delay line are functions of the transducers, their circuitry, and propagation effects such as the ultrasonic absorption in the transmission medium and losses due to beam spreading. Propagation effects are neglected n the following analysis because of their wide variation vith the type of transmitting material, with path length, ınd path configuration.[6,7,8] If for a given delay line the ›ropagation losses are known as a function of fre-

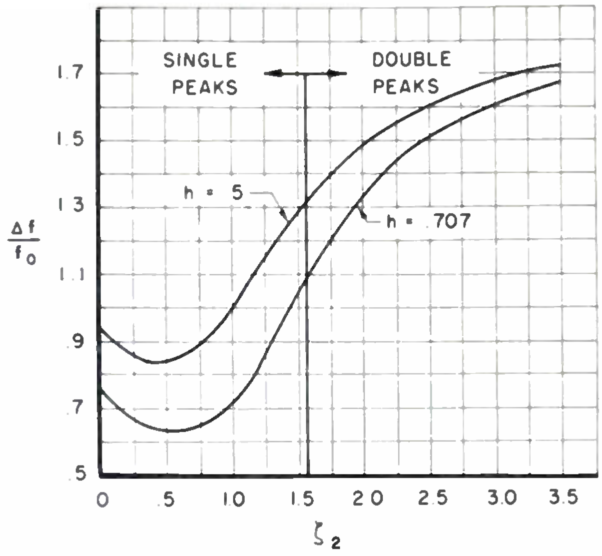

Fig. 5—Fractional bandwidths for mercury ($\zeta_1=1.29$)

quency, the absolute over-all attenuation can be found from the Principle of Superposition.

1. *Frequency Response:* A parallel circuit of L, C_S, and R usually is connected to the electrical terminals of the receiving transducer, as in the case of the transmitting transducer. Assuming identical transmitting and receiving circuit parameters and constant volume velocity incident on the receiving transducer, it can be shown by means of the Reciprocity Theorem that the normalized output voltage of the receiving transducer is identical to the normalized volume velocity radiated by the transmitting transducer.

In consequence of volume velocity on the receiving transducer being proportional to volume velocity radiated by the transmitting transducer, the over-all frequency response of the delay line is the square of the product of (8) and the response of the parallel resonant circuit of R, L, C_S and C_E. The magnitude of the acoustic response is identical to the normalized power given by (10). Therefore, all curves relating to the acoustic power response, bandwidth, or peaks of one transducer may be converted to the over-all frequency response, bandwidth, or peaks of the delay line in voltage, except for the effects of the driving and receiving circuits, by substituting E for W. The frequency response of the transducer's circuitry is well known and can be computed separately.

2. *Transducer Loss:* Because of the low coefficient of electromechanical coupling in piezoelectric transducers, the transducer loss at fundamental resonance accounts for a substantial part of the delay line attenuation. In accordance with the commonly accepted method of measurement, the transducer loss is defined as the ratio of the voltage E_T at the electrical terminals of the transmitting transducer to the voltage E_R at the electrical terminals of the receiving transducer with propagation losses assumed to be negligible. The output voltage of the receiving transducer E_R is found by solving Roth's circuit with the proper connections made for the receiving case. A parallel circuit of L, C_S, and R is connected to the electrical terminals, and a volume-velocity generator Av_R is connected across z_1/A in mesh 1. Solving for E_R at resonance,

$$E_R = -Av_R\frac{2RC_E\zeta_1}{C_p(\zeta_1+\zeta_2)}. \tag{11}$$

Av_R can be found by applying Thevenin's Theorem at the receiving end of the transmission medium. Looking back toward the transmitting transducer (assumed to be at infinity), an acoustic impedance z_1/A in shunt with a volume velocity generator $2\,Av_{1_0}$ is seen, where Av_{1_0} is the volume-velocity radiated by the transmitting transducer and is given by (7), where E_3 is replaced by E_T. Hence,

$$Av_R = \frac{-4E_TC_EA}{zC_p(\zeta_1+\zeta_2)}. \tag{12}$$

Substituting (4) and (12) in (11) and solving for the transucer loss,

$$\frac{E_T}{E_R} = \frac{\pi(\zeta_1+\zeta_2)^2}{8K^2\zeta_1\omega_0C_ER}. \tag{13}$$

In the case of circular, X-cut quartz crystals,

$$C_E = 7.28\,d^2f_0 \text{ micromicrofarads,} \tag{14}$$

where d is the diameter of the active area of the crystal in inches, and f_0 is the fundamental resonant frequency in megacycles.

II. Transient Analysis

In relatively narrow-band applications the over-all transient response of a memory device is chiefly determined by the amplifier used with the delay line. In the high-speed memory of a large scale computer, however, the transient response of the delay line is the ultimate limitation on digital storage capacity and is therefore an important consideration in the logical design. The acoustic transient response of the delay line, neglecting propagation effects and the response of the transducers' circuitry, is developed in the following analysis.

Roth's circuit has served as a basis for the development of the steady-state theory of the piezoelectric transducer as presented above, but, for obvious reasons, it is not suitable for direct application to the transient analysis. The procedure adopted here is to devise an electrical transmission-line circuit having a steady-state response identical to that of the transducer in the approximate case and having a transient response which can be calculated in a straightforward manner. More-

[6] H. B. Huntington, "On ultrasonic propagation through mercury in tubes," *Jour. Acous. Soc. Amer.*, vol. 20, pp. 424–432; July, 1948.
[7] W. Roth, "Scattering of Ultrasonic Radiation in Polycrystalline Metals," Research Laboratory of Electronics, M.I.T., Technical Report No. 52; November 26, 1947.
[8] D. L. Arenberg, "Ultrasonic Solid Delay Lines," *Jour. Acous. Soc. Amer.*, vol. 20, pp. 1–26; January, 1948.

over, the transmission-line circuit presents a clear physical picture of the mechanism involved in both the steady-state and the transient operation of the transducer.

Equivalent Circuit

The equivalent transmission-line circuit, which has a steady-state response identical to the acoustic response of a piezoelectric transducer in the approximate case, is shown in Fig. 6. Heavy lines indicate ideal uniform transmission-line sections of characteristic impedances Z, Z_1, and Z_2, where the central section Z, of length l, is analogous to the transducing crystal and sections Z_1 and Z_2 are analogous to the acoustic loading media at the faces of the transducer. For purposes of analysis, sections Z_1 and Z_2 are assumed to extend to infinity.

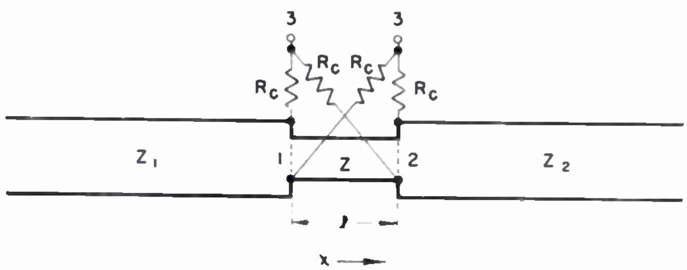

Fig. 6—Equivalent transmission-line circuit.

Terminals 3 are analogous to the electrical terminals of the transducer formed by the conducting coatings on its faces. By analogy with the low electromechanical coupling in a transducer, the coupling resistances R_c are assumed to be much greater than the characteristic impedance Z of the center section. Consequently, the shunting effect of R_c on the transmission-line impedances is neglected in setting up the boundary conditions.

1. *Steady-State Response:* Assuming—for the transmitting case—a sinusoidal constant-voltage generator E_3 connected to terminals 3 of the equivalent circuit, traveling waves are seen to be set up in the transmission lines: waves of voltage $E_1\epsilon^{j\omega x/c_1}$ and current $I_1\epsilon^{j\omega x/c_1}$ propagating to the left in Z_1, waves of voltage $E_2\epsilon^{-j\omega x/c_2}$ and current $I_2\epsilon^{-j\omega x/c_2}$ propagating to the right in Z_2, waves of voltage $E_+\epsilon^{-j\omega x/c}$ and current $I_+\epsilon^{-j\omega x/c}$ propagating to the right in Z, and waves of voltage $E_-\epsilon^{j\omega x/c}$ and current $I_-\epsilon^{j\omega x/c}$ propagating to the left in Z, where the time factor $\epsilon^{j\omega t}$ is implicit in the amplitude coefficients. The velocities of propagation are c, c_1, and c_2 in the corresponding sections.

Since $R_c \gg Z$, the current I_3 flowing in R_c is assumed to be independent of impedances Z_1 and Z_2. Taking boundary 1 as the origin ($x=0$), and applying Kirchoff's laws at the boundaries:

$$I_1 = I_+ + I_- + I_3, \tag{15a}$$

$$I_2\epsilon^{-j\omega l/c_2} = I_+\epsilon^{-j\omega l/c} + I_-\epsilon^{j\omega l/c} + I_3, \tag{15b}$$

$$E_1 = E_+ + E_-, \tag{16a}$$

$$E_2\epsilon^{-j\omega l/c_2} = E_+\epsilon^{-j\omega l/c} + E_-\epsilon^{j\omega l/c}. \tag{16b}$$

In general, for a positively traveling wave,

$$E_+ = I_+Z, \tag{17a}$$

and for a negatively traveling wave,

$$E_- = -I_-Z. \tag{17b}$$

Therefore (16) may be written

$$-I_1Z_1 = Z(I_+ - I_-), \tag{18a}$$

$$I_2\epsilon^{-j\omega l/c_2}Z_2 = Z(I_+\epsilon^{-j\omega l/c} - I_-\epsilon^{j\omega l/c}). \tag{18b}$$

Equations (15) and (18) are simultaneous equations in four unknowns (I_1, I_2, I_+, I_-), which may be solved for I_1 by means of Cramer's Rule

$$I_1 = \frac{I_3Z\{2Z_2 + (Z - Z_2)\epsilon^{-j\omega l/c} - (Z + Z_2)\epsilon^{j\omega l/c}\}}{-\epsilon^{j\omega l/c}(Z^2 + ZZ_1 + Z_1Z_2 + ZZ_2) + \epsilon^{-j\omega l/c}(Z^2 - ZZ_1 + Z_1Z_2 - ZZ_2)}. \tag{19}$$

Using the relations

$$\epsilon^{j\theta} = \cos\theta + j\sin\theta, \tag{20a}$$

$$\epsilon^{-j\theta} = \cos\theta - j\sin\theta; \tag{20b}$$

and rearranging,

$$I_1 = \frac{I_3\zeta_1\left[\cos\left(\frac{\omega l}{c}\right) - 1 + j\zeta_2\sin\left(\frac{\omega l}{c}\right)\right]}{(\zeta_1 + \zeta_2)\cos\left(\frac{\omega l}{c}\right) + j(1 + \zeta_1\zeta_2)\sin\left(\frac{\omega l}{c}\right)}, \tag{21}$$

where

$$\zeta_1 = \frac{Z}{Z_1}, \tag{22a}$$

$$\zeta_2 = \frac{Z}{Z_2}. \tag{22b}$$

At fundamental resonance

$$l = \frac{\lambda_0}{2}, \tag{23}$$

where λ_0 is the wavelength at resonance; hence,

$$\frac{\omega l}{c} = \delta\pi, \tag{24}$$

and

$$I_{1_0} = \frac{2I_3\zeta_1}{\zeta_1 + \zeta_2}. \tag{25}$$

Normalizing I_1 with respect to I_{1_0},

$$\frac{I_1}{I_{1_0}} = \frac{(\zeta_1+\zeta_2)}{2}\left\{\frac{\cos\delta\pi - 1 + j\zeta_2\sin\delta\pi}{(\zeta_1+\zeta_2)\cos\delta\pi + j(1+\zeta_1\zeta_2)\sin\delta\pi}\right\}. \tag{26}$$

Since Z_1 is resistive, the normalized voltage E_1/E_{1_0} is identical to (26). It is clear from the identity of (26) with (8) that the normalized output voltage or current of the transmission-line circuit is identical to the normalized volume velocity radiated by a transmitting transducer or to the normalized output voltage of a receiving transducer for constant received volume velocity. For fundamental resonance at the same frequency, the time delay T_d of electromagnetic energy through Z is identical to the time delay of acoustic energy through the transducer.

Since the steady-state responses are identical, it follows that the transient responses are the same. The transient solution of the transmission-line circuit is readily established by well-known methods.[9]

2. *Impulse Response:* The following analysis assumes an impulse driving source—a generator of a voltage pulse e_3 of zero width occurring at zero time—connected to terminals 3 of the transmission-line circuit. Using the notation that $e = E\,f(T)$, where the function $f(T) = 1$ for time $t = T$, and $f(T) = 0$ for $t > T$ or $t < T$, it follows that $e_3 = E_3\,f(0)$. In the simplest case of symmetrical loading when $\zeta_1 = \zeta_2 = 1$, as shown in Fig. 7, voltage pulses of equal amplitude and opposite polarity are developed at boundaries 1 and 2 at $t = 0$. Arbitrarily assume that the pulse at boundary 1 is positive. A positive pulse (*a*) then propagates to the left of boundary 1 at velocity c_1, and a positive pulse (*b*) propagates to the right of boundary 1 at velocity c. A negative pulse (*c*) simultaneously propagates to the left of boundary 2 at velocity c, and a negative pulse (*d*) propagates to the right of boundary 2 at velocity c_2.

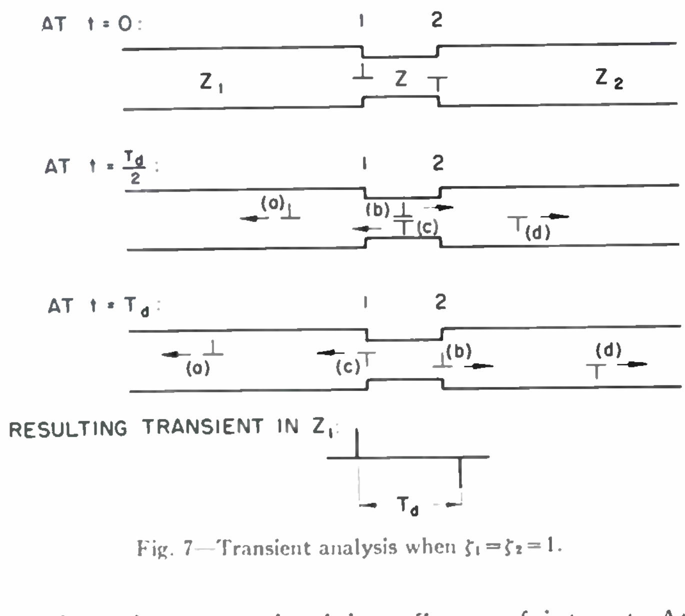

Fig. 7—Transient analysis when $\zeta_1 = \zeta_2 = 1$.

The pulses transmitted into Z_1 are of interest. At $t = T_d$, pulse (*a*) will have reached a position $-x(=c_1T_d)$ in Z_1, pulse (*b*) will have passed into Z_2 with no reflection at boundary 2, and pulse (*c*) will have passed into Z_1 with no reflection at boundary 1. Therefore, the voltage wave propagated by Z_1 is a positive pulse followed by a negative pulse of the same amplitude which is delayed by T_d. In impulse notation the wave at boundary 1 is $E_1\,f(0) - E_2\,f(T_d)$.

[9] See, for example, S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., pp. 22–32; 1944.

In general, the pulses initially developed at the boundaries are unequal due to nonsymmetrical loading, and multiple reflections occur in Z due to mismatches at the boundaries. Change of polarity on reflection also occurs for certain loading conditions.

For a traveling wave in Z incident on boundary 1 or 2,

$$\alpha_1 = \frac{Z_1 - Z}{Z_1 + Z} = \frac{1 - \zeta_1}{1 + \zeta_1}, \tag{27a}$$

and

$$\alpha_2 = \frac{1 - \zeta_2}{1 + \zeta_2}, \tag{27b}$$

where α, the reflection coefficient, is the ratio of reflected voltage to incident voltage. The transmission coefficient β, the ratio of voltage transmitted to voltage in the incident wave, is given by

$$\beta_1 = \frac{2Z_1}{Z + Z_1} = \frac{2}{\zeta_1 + 1}, \tag{28a}$$

and

$$\beta_2 = \frac{2}{\zeta_2 + 1}. \tag{28b}$$

At $t = 0$, the voltages E_1 and E_2 are developed at boundaries 1 and 2 by the current $I\,f(0)$ due to the impulse generator. Hence,

$$E_1 = I\,\frac{ZZ_1}{Z + Z_1} = \frac{IZ\beta_1}{2}, \tag{29a}$$

and

$$E_2 = \frac{-IZZ_2}{Z + Z_2} = \frac{-IZ\beta_2}{2}. \tag{29b}$$

Following the same procedure as in the case of symmetrical loading, but taking into account the losses on transmission and reflection at the boundaries, the general expression for the voltage wave e_1 at boundary 1 may be written:

$$\begin{aligned} e_1 = {} & E_1 f(0) + \beta_1 E_2 f(T_d) + \alpha_2\beta_1 E_1 f(2T_d) \\ & + \alpha_1\alpha_2\beta_1 E_2 f(3T_d) + \alpha_1\alpha_2^2\beta_1 E_1 f(4T_d) \cdots, \end{aligned} \tag{30a}$$

where the terms represent impulses spaced by time T_d.

Evaluating E_2 in (30a) by means of (29b) and normalizing with respect to the initial pulse voltage E_1,

$$\begin{aligned} \frac{e_1}{E_1} = {} & f(0) - \beta_2 f(T_d) + \alpha_2\beta_1 f(2T_d) - \alpha_1\alpha_2\beta_2 f(3T_d) \\ & + \alpha_1\alpha_2^2\beta_1 f(4T_d) \cdots, \end{aligned} \tag{30b}$$

which is the impulse response of one transducer.

Over-All Transient Response for Delay Lines

1. *Theoretical Transient Response:* It has been shown that the transient response of output voltage for a receiving transducer with an impulse of volume velocity incident on interface 1 is identical to (30b). In a delay

line driven by an impulse generator, the wave incident on the receiving transducer is a train of pulses generated by the transmitting transducer, as given by (30b). Each incident pulse produces at the output a train of voltage pulses spaced by T_d and proportional to the amplitude and polarity of the incident pulse. Addition of the pulse trains in the proper time relationship gives the over-all impulse response of the delay line exclusive of circuitry and of propagation effects. For identical transducers, corresponding pulses in the trains are coincident in time due to identical delay times in both transducers. Addition of coincident pulses gives the normalized over-all response

$$
\begin{aligned}
\frac{e}{E} = & \overline{f(0)} - \overline{2\beta_2 f(T_d)} + \overline{(2\alpha_2\beta_1 + \beta_2^2)f(2T_d)} \\
& - \overline{2\alpha_2\beta_2(\alpha_1 + \beta_1)f(3T_d)} \\
& + \overline{\alpha_2(2\alpha_1\alpha_2\beta_1 + 2\alpha_1\beta_2^2 + \alpha_2\beta_1^2)f(4T_d)} \qquad (31) \\
& - \overline{2\alpha_1\alpha_2^2\beta_2(\alpha_1 + 2\beta_1)f(5T_d)} \cdots,
\end{aligned}
$$

where the terms under a bar represent a composite pulse, the pulses being spaced by T_d. Fewest pulses result when $\zeta_2 = 1$, giving

$$
\frac{e}{E} = f(0) - 2f(T_d) + f(2T_d), \qquad (32)
$$

which is independent of ζ_1.[10] Theoretically, if the transducers were backed by a medium which matched the transducer acoustic impedance, optimum transient response would result for any transmission medium between the transducers. Furthermore, the reflection of incident waves at the transducers would be minimized. It should be noted from (13), however, that in this instance the transducer loss is proportional to $(\zeta_1+1)^2/\zeta_1$, which is large when ζ_1 is small.

The response of the delay line to any applied waveform can be found by applying the Principle of Superposition. The method is to assume that the applied wave results from the addition of a large number of impulses of the proper amplitude, polarity, and position in time. Equation (31) gives the response of each impulse, and the sum of all the responses in the proper time and amplitude relations is the response to the applied wave. When the number of impulses is infinite, the superposition integral results.

2. *Measured Transient Responses and Steady-State Bandwidths:* The following measurements of over-all transient response used a 0.02-microsecond input pulse to approximate an impulse. The output of the delay line was viewed on a wide-band video amplifier and synchroscope. Input and output impedances of the delay line were 50 ohms. Propagation effects were minimized by a short path length, all the measurements on mercury utilizing a $\frac{3}{4}$-inch diameter path two inches in length. X-cut quartz crystals were used as transducers in all measurements. The following over-all fractional bandwidths were measured on the same delay lines using rf pulse techniques.

The fundamental resonant frequencies of the transducers were in the range of 6.5 to 10 mc. Unfortunately, cut-off of the video amplifier at 18 mc limited the pass band to the fundamental mode of the transducers, resulting in the 0.02-microsecond input pulse being reshaped as alternations of the fundamental resonant frequency at the output.

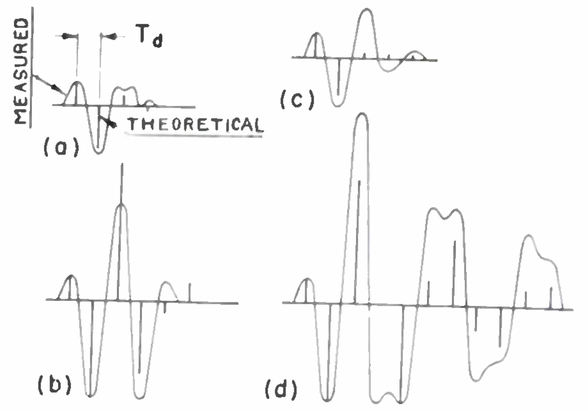

Fig. 8—Theoretical and measured transient responses.

Shown in Fig. 8(a) are the theoretical and measured transient responses for a mercury delay line with mercury backed transducers ($\zeta_1 = \zeta_2 = 1.29$). Higher order pulses of low amplitude are neglected in this curve and in the following curves for theoretical transient response.

The transient response of a mercury delay line with air-backed transducers (steel backing plates) is given in Fig. 8(b) ($\zeta_1 = 1.29$, $\zeta_2 \cong 0$). The measured 3-db bandwidth was 0.62; in comparison, the theoretical bandwidth is 0.76 (Fig. 5).

The transient response of a mercury delay line with lead backings in intimate contact with the transducers is shown in Fig. 8(c) ($\zeta_1 = 1.29$, $\zeta_2 = 1.58$). The measured 3-db bandwidth was 0.93; the theoretical bandwidth is 1.1.

A steel delay line, made from a gauge block $\frac{3}{4}$ inch in diameter and one inch in length, with air-backed transducers, gave the transient response shown in Fig. 8(d) ($\zeta_1 = 2.83$, $\zeta_2 \cong 0$). In both the measured and theoretical responses, higher order pulses of decreasing amplitude follow the eleven pulses shown. The measured bandwidth between peaks was 0.93, and the peaks were 4 db greater than the response at resonance. Theoretically, the bandwidth between peaks is 0.91, and the peaks are 7 db greater than the response at resonance (Fig. 4).

From a comparison of the transient measurements with calculated values it is concluded that the agreement is not exact, but that the theory exhibits the correct trends with sufficient accuracy for most engineering problems. In general, the measured bandwidths are less than the theoretical bandwidths, and the transducer loss measured at resonance is 3 to 6 db greater than the theoretical transducer loss from (13) and (14).

[10] This is in agreement with the steady-state response, which is independent of ζ_1 when $\zeta_2 = 1$; hence, the curve for $\zeta_2 = 1$ in Fig. 1 corresponds to optimum transient response for any ζ_1.

The theory assumes ideal loading conditions which are only approximated in practice. In the case of the mercury delay line with air-backed crystals, for example, the steel backing plates apparently reflect a small acoustic load to the back faces of the crystals, which accounts for the measured bandwidth being less than, and the transducer loss being higher than, the theoretical values as can be seen from Fig. 5 and (13).

Although exact numerical values do not result from the theory as presented above, it should be clear that the theory is fundamentally correct and is of practical importance in the design of ultrasonic delay lines.

ACKNOWLEDGMENT

The authors wish to express their appreciation to Mr. G. B. Parkinson and to Dr. W. M. Hall, who checked the derivations and contributed valuable suggestions for the presentation of the material.

Note on Safe Resonator Current of Piezoelectric Elements*

E. J. POST†

Summary—**An unambiguous quotation of crystal data is proposed combined with some numerical values of much-used quartz cuts. The aim is to prevent a misunderstanding which sometimes exists about maximum electrical loading of piezoelectric elements.**

IN THE APRIL ISSUE of *Electronics* (1951) E. A. Gerber[1] emphasized that the ratio of amplitude of crystal vibration and current in the lead wires depends strongly on the operating conditions of the crystal. To prevent ambiguity in quoting data for maximum crystal current, the limiting data for crystal performance could be given in terms of optimum current in the mechanical branch of the crystal two-terminal.[2] The fictitious (piezo-electric) current in the mechanical branch is directly proportional to the amplitude of vibration, and may be identified with the current in the lead wires in almost any case of a crystal operating in its series resonance.

A dimensional analysis shows that for *uniformly* vibrating, fully plated crystals the piezoelectric current can be written as

$$I = C \cdot \psi;$$

ψ is a function depending upon the geometrical dimensions of the crystal and its mode of vibration only, whereas C is a constant depending on physical and technical quantities characteristic for the crystal cut in question. In Table I particulars about C and ψ are given for computing the optimum current in the mechanical branch of some quartz cuts.

Appropriate use of the table prevents crystal breakage due to electrical overloading. This statement is based on about three years of practical application in the Post Office Laboratory in the Hague in Holland.

TABLE I

FORMULA FOR OPTIMUM CURRENT IN MECHANICAL BRANCH OF RECTANGULAR PLATES

$I_m = C\psi(l, w, t)$ (mA)

l = length w = width t = thickness of resonator

Mode of vibration and cut	$\psi(l, w, t)$ (mm)	C (m.A.mm^{-1})	Remarks
Thickness A_T, B_T, X	wl/t	A_T, $C=0.12$ B_T, $C=0.12$ X, $C=0.4$	
Longitudinal X bar, G_T	w	X cut 5°, $C=0.55$ G_T cut, $C=0.30$	G_T $l<w$
Face shear C_T, D_T	w	C_T, $C=0.25$ D_T, $C=0.15$	$w=l$
Face shear H_T	l	$C=0.25$	$l \gg w$[3]
Flexure Curie strip of X bars	wt/l	$C=0.3$ $C=0.6$	No center electrode center electrode

A direct computation of optimum piezoelectric current based on tensile strength of the material may be found in Cady's "Piezoelectricity."[4] More information about rupture strain of quartz for different orientations is needed to extend the derivation to other crystal cuts. Moreover, one has to take into account the additional stresses due to interfering modes disturbing the uniform vibration pattern in thickness vibrators, and for partially plated crystals one has to take into account the decreased area of the electrode surface. For overtone excitation of thickness vibrators it is worthwhile to point out that some further analysis shows that C is unaffected by the order of the overtone.

No doubt some figures for C are susceptible to correction by those who have more experimental data available. The principle aim of this note is to provide a fair working basis for safe crystal-oscillator performance.

* Decimal classification: R214.211. Original manuscript received by the Institute, June 23, 1951; revised manuscript received, January 9, 1952.

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[1] E. A. Gerber, "Amplitude of vibration in piezoelectric crystals," *Electronics*, vol. 24, p. 143; April, 1951.

[2] A more elaborate form of this note was published in Dutch as an appendix to a paper on oscillator design (Post and Pit, *PTT Bedrijf*, vol. III, p. 39; May, 1950). The material on oscillator design was published in English (PROC. I.R.E., vol. 39, pp. 169–174; February, 1951). Details covered in Gerber's paper have been deleted.

[3] E. J. Post, *Appl. Sci. Res.*, vol. B1, p. 420; 1950.

[4] W. G. Cady, "Piezoelectricity," McGraw-Hill Book Co., Inc., New York, N. Y., p. 324; 1946.

An Experimental Study of Low-Power CW Magnetrons Having Few Segments*

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Summary—**It is shown that, apart from having intrinsically lower efficiency, two- and four-segment magnetrons differ very little in characteristics from multisegment types if the ratio of cathode-to-anode diameter is the same in each case. High efficiency, comparable under some conditions to that of multisegment magnetrons, can be obtained from tubes having few segments when the cathode is relatively much smaller. This results in a number of observable differences between the behavior of the two types, notably a variation of efficiency with diameter in two- and four-segment tubes, which is accompanied by nonconformity to the voltage scaling laws.**

1. Introduction

IT HAS BEEN SHOWN[1] that an approximate value for the theoretical maximum efficiency of a magnetron may be derived from the assumption that all electrons arriving at the anode do so tangentially with an angular velocity equal to that of the rotating rf field, and that the energy of the electrons returning to the cathode may be neglected. Then

$$\eta = 1 - \frac{1}{2}\frac{m}{}\left(\frac{\omega r_a}{n}\right)^2 / eV,$$

where

ω is the rf angular velocity
λ the corresponding wavelength
n the mode number
V the anode voltage
H the magnetic field

and

r_a the anode radius.

By substituting for V the threshold voltage V_T derived by Hartree and Bunemann and introducing the Hull cutoff voltage V_c, this may be written

$$\eta = 1 - \left[\left(\frac{V_c}{V_T}\right)^{1/2} - \left(\frac{V_C}{V_T} - 1\right)^{1/2}\right]^2$$

which, by manipulation and introduction of the reduced[2] values of operating magnetic field (H/H_0) and applied voltage (V/V_0), gives

$$\eta = 1 - \frac{1}{\dfrac{2H}{H_0} - 1}$$

or

$$\eta = 1 - \frac{V_0}{V_T},$$

where

$$V_0 = \left(\frac{r_a}{n\lambda}\right)^2 \cdot \left(2\pi^2 c^2 \frac{m}{e}\right)$$

$$= \left(\frac{r_a}{n\lambda}\right)^2 \cdot (10.1 \times 10^6)$$

and

$$H_0 = \frac{1}{n\lambda\left(1 - \dfrac{r_c^2}{r_a^2}\right)} \cdot \left(\frac{e}{m} \cdot \frac{1}{4\pi c}\right)$$

$$= \frac{1}{n\lambda} \cdot (21.4 \times 10^3)$$

if r_c, the cathode radius, is small compared to r_a.

Comparisons of the predicted and measured values of efficiency of both two- and four-segment magnetrons have been made by Herriger and Hülster,[3] showing that the measured efficiency is never greater than about one half the predicted maximum value. A comparison[4] of available data on multisegment magnetrons of the kind commonly used in radar systems showed that whereas with H/H_0 around 3 the measured efficiency, like that of the types examined by Herriger and Hülster, is about one half of the theoretical maximum value, it rises to within 5 to 10 per cent of that value when H/H_0 is equal to 5.

It will be shown here that efficiencies greater than those found by Herriger and Hülster may be obtained from two- and four-segment magnetrons by the proper choice of geometry and operating conditions, that the difference in efficiency so obtained can be simply and directly related to differences in tube dimensions, and that under certain conditions the efficiency of two- and

* Decimal classification: R355.912.1. Original manuscript received by the Institute, July 20, 1951; revised manuscript received, February 18, 1952.

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[1] W. E. Willshaw, L. Rushforth, A. G. Stainby, R. Latham, A. W. Balls and A. H. King, "The high power pulsed magnetron; development and design for radar applications," *Jour. IEE* (London), vol. 93, no. 5, pp. 991, 992; 1946. See also K. Posthumus, "Oscillations in a split anode magnetron," *Wireless Eng.*, vol. 12, pp. 126–132; March, 1935. Reference to this work is made by Spangenberg in "Vacuum Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 654–660; 1948.

[2] In their original form, as derived by Hartree and Bunemann, the reduced variables do not include the mode number n. Those in common use today, which take n into account, are due to Slater. See, for example, G. B. Collins "Microwave Magnetrons," Radiation Laboratory Series No. 6, McGraw-Hill Book Co., Inc., New York, N. Y., pp. 414–417; 1948.

[3] F. Herriger and F. Hülster, "Die Schwingungen der Magnetfeldrohren," *Hochfreq. und Elektroak.*, Band 49, p. 129; April, 1937.

[4] K. R. Spangenberg, "Vacuum Tubes," McGraw-Hill Publishing Co., New York, N. Y., pp. 665–667; 1948.

four-segment magnetrons is comparable if not almost equal to that of multisegment types. High efficiency can be obtained from tubes having few segments only if the cathode is relatively much smaller than that in multisegment magnetrons. The experimental work to be described has related this difference to differences in the characteristics of the two types, and in particular to a considerable variation of efficiency with diameter in two- and four-segment magnetrons, which is shown to be accompanied by nonconformity to the "voltage-scaling" laws. It has also shown that when the relative size of the cathode is greater than that giving maximum efficiency in magnetrons having few segments, and is the same for these and multisegment tubes, there is little if any fundamental difference between the characteristics of the two types. They differ significantly only in that the magnetrons with fewer segments have intrinsically lower efficiency.

2. Experimental Results

Two-Segment Magnetrons

Fig. 1 shows the relation between efficiency, anode current, field, and frequency. With H and f fixed, η and I_p are inversely proportional over most of the range of current in which oscillation is maintained.

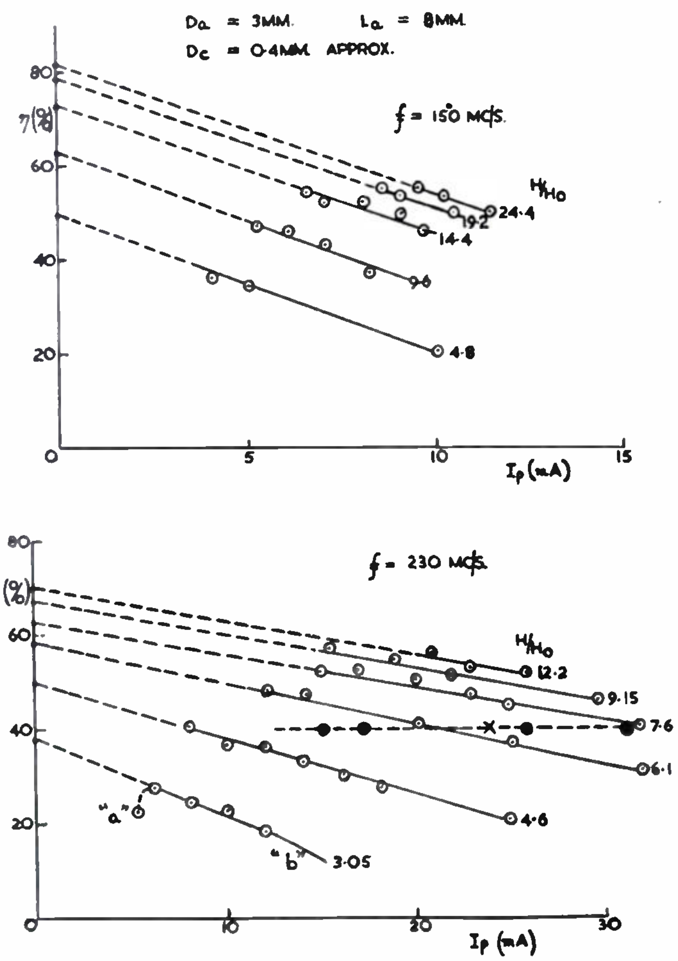

Fig. 1—Efficiency of a two-segment magnetron as a function of anode current and magnetic field.

Near the lower limit of oscillation, there is a characteristic "tail" or rise in efficiency with current as illustrated at "a." This was found in nearly all the tubes examined, but to varying degree, so that it may pass unnoticed, particularly when r_a is small. Large tubes, on the other hand, may give a rise in efficiency with current over a large part of the operating range, and intermediate sizes show the tail as a region of constant efficiency at the low current end of the mode. At the high current limit there may be either an abrupt cessation of oscillation or a falling off in efficiency as illustrated at "b." In both cases, approach to the upper mode boundary is characterized by no increase or, alternatively, a fall in current as the anode voltage is increased, even though the primary emission available from the cathode is many times the oscillating current at any part of the mode.

If the loading is set to give maximum efficiency at low values of H/H_0, and then H/H_0 is increased, there is an abrupt change in the η/I_p characteristic at some particular value of H/H_0, as illustrated by the broken curve "X" of Fig. 1. The efficiency no longer shows inverse proportionality to anode current, but appears to be sensibly independent of it. A slight reduction in loading restores the inverse relationship, the value of η_{min} being then roughly the same as the efficiency previously obtained over the whole range of operating current.

Changes in the relative dimensions of the gaps and the interaction space displace the lower and upper mode boundaries in the same direction, and affect the incidence of tail. A tube with narrow gaps ($<r_a/5$) will oscillate at lower current and show little backlash at the lower mode boundary and more tail than a similar one with relatively large gaps (about $r_a/2$), which will jump into oscillation at a higher minimum current, have considerable backlash, and little if any tail. At the upper mode boundary, tubes with narrow gaps drop abruptly out of oscillation at lower current and therefore give less maximum power output than those having wide gaps. The efficiency and operating current in the central linear part of the curve are not greatly affected by gap width.

At any fixed frequency, as H is increased above a critical value for a particular tube, the mode boundaries move very close together, the average current in the mode remains constant or falls, and the measured values of efficiency show little or no increase. When H exceeds another critical value slightly higher than the first, some tubes show a rapid decrease in operating current and measured efficiency with increasing H, and in some extreme cases cannot be made to oscillate at all. The fall off may be minimized or possibly eliminated by the presence of a minute trace of gas, or by an increase in temperature of the cathode, even though the emission available at the lower temperature is several times the oscillating current. It is particularly sensitive to changes in cathode radius, even though this is small compared to the diameter of the interaction space. Control by the cathode is a function of size only, and not of any increase in available emission that may be a result of an increase in radius, which raises the critical value of H, and will, in most cases, eliminate the falling-off effect or the refusal to oscillate above some critical field. In one particular experiment, by a change in radius from roughly $r_a/10$ to $r_a/5$, normal operation was obtained from a tube which had previously failed to oscillate at all except at very low field. Similar, but not so marked effects may be obtained by off-centering a small cathode.

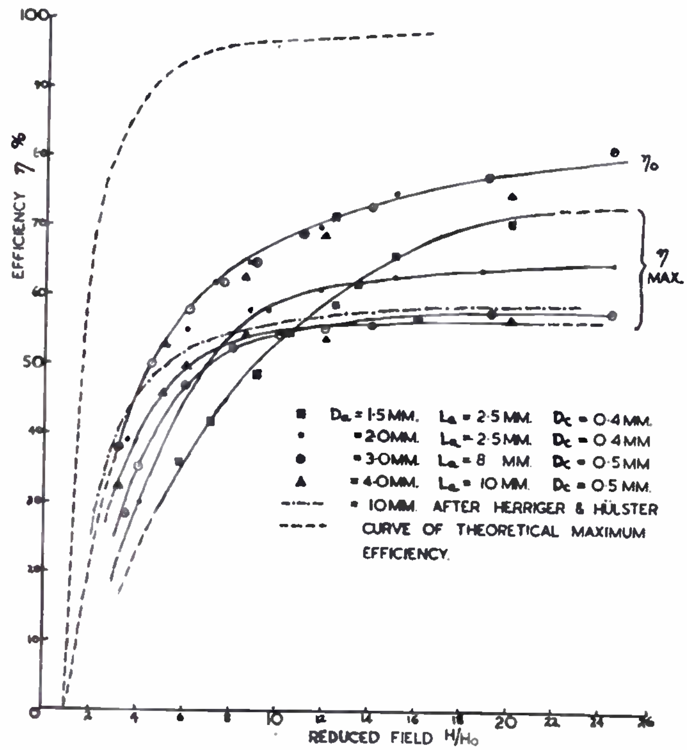

Fig. 2—Efficiency of a number of two-segment magnetrons as a function of reduced magnetic field, showing effect of changing anode diameter.

Fig. 2 shows a family of curves of efficiency against a reduced magnetic field for tubes of different anode diameter, together with a curve taken from the Herriger and Hülster paper: η_{max} is the measured maximum efficiency; η_0 the efficiency at very low current, obtained by extrapolating curves like those of Fig. 1 to $I_p = 0$. The former, η_{max}, shows little or no increase above some critical value of H/H_0 that depends upon anode diameter, but η_0 shows a steady but diminishing increase with H/H_0 over the whole range of measurement. For fixed values of H/H_0, there is a small but definite increase of η_{max} with frequency; but for convenience in comparison, a single curve is drawn to represent the relation between η_{max} and H/H_0. η_0 was found to be entirely independent of frequency and anode diameter.

When H/H_0 is less than 5, one obtains the empirical relationship

$$\eta_{max} \doteqdot 0.42 D_a^{1/2} \cdot \eta_0 \not> \eta_0,$$

where D_a is in mm, and it is assumed that the leveling off of η_{max} above the critical value of H/H_0 has not taken place. The value of H/H_0 at which leveling off in η_{max} occurs may also be related to the anode diameter by the approximate empirical formula

$$(H/H_0)_l \doteqdot 30/D_a.$$

Both formulas give good agreement with the results of our experiments, but do not fit those obtained by Herriger and Hülster.

Four-Segment Magnetrons

When curves of efficiency against current of four-segment tubes were first examined, it appeared that apart from a fall in efficiency near the mode boundaries, the efficiency was sensibly constant over the whole of the operating range. Closer examination of curves taken on a number of tubes and measurements made with loading slightly lighter than that found to give maximum efficiency near the low-current-mode boundary showed, however, that the curves for four-segment tubes are not fundamentally different from those for two-segment types. The curves consist, in general, of an extended region of constant efficiency near the low-current-mode boundary, corresponding to the "tail" in the two-segment characteristic, and a region of linearly decreasing efficiency near the upper limit. As the loading is increased, the curve is flattened out and η appears to be almost independent of I_p. A similar effect can be produced by heavy loading of two-segment types. It is characteristic of the four-segment tubes that however they are loaded, the measured values of η_{max} and η_{min} differ less than those of two-segment types. The effect of changes in loading and gap dimensions is similar but not so pronounced as in two-segment types. The fall-off effects in high magnetic fields are affected by changes in cathode diameter in a similar way.

Fig. 3 shows a set of three families of curves of efficiency against a reduced magnetic field for magnetrons of three different sizes. Above a critical frequency for each size of tube, which decreases as the anode diameter increases, the curves of η_{max} closely resemble those of two-segment magnetrons of similar size, except that efficiency is always higher at the same values of H/H_0 and the leveling off in efficiency occurs at higher values of η_{max} and H/H_0 for the four-segment types. As the frequency is reduced, the curves of η_{max} versus H/H_0 show a broad peak at values of H/H_0 which increase as the anode diameter decreases. The peak becomes more pronounced until a critical frequency is reached, when no further changes are observed.

In Fig. 4, curves of η_{max} taken from Fig. 3 and the curve of η_0 are shown for comparison with the curve of theoretical maximum efficiency, an averaged curve for multisegment magnetrons, and a curve given by Herriger and Hülster for a magnetron having $D_a = 10$ mm. As in the case of two-segment magnetrons, the larger tubes are more efficient when H/H_0 is low but show a limiting value of η_{max} lower than that of the smaller tubes, and η_0 is independent of diameter and frequency.

If η_{max} is read from the upper branches and H/H_0 is less than 5, efficiency is given by the approximate empirical expression

$$\eta_{max} \doteqdot 0.4 D_a^{1/2} \cdot \eta_0 \not> \eta_0,$$

where D_a is in mm. This is almost the same formula as that obtained for two-segment tubes. The efficiency in the lower branches of the curves of η_{max} is independent of diameter and is approximately equal to $2\eta_0/3$. The values of H/H_0, after which no great increase in efficiency is obtained in the lower family of curves, are related by the approximate formula

$$(H/H_0)_l \doteqdot 40/D_a,$$

and the position of the peaks in the upper branches by

$$(H/H_0)_p \doteqdot 30/D_a,$$

which is the same formula as that obtained for the leveling-off point in two-segment types.

Herriger and Hülster, attribute the trough in their η versus H/H_0 curve to an unfavorable spatial relationship between the RF gaps and the loops in the electron paths from cathode to anode. No sign of such a trough was found during our measurements of the efficiency of the 3-, 4-, and 6-mm tubes. The 6-mm tube was made to give a characteristic somewhat similar to the Herriger and Hülster curve by deliberately undercoupling the load when the curve "Z" of Fig. 3 was obtained. The tube then generated a large amount of second harmonic in the region of the depression in η_{max}. This could be eliminated, and the normal characteristic obtained, by an increase in loading too small to affect the measured values of efficiency at low values of H/H_0. This generation of harmonics was found in all the tubes examined if the loading was not properly adjusted, being more pronounced in the larger types. It therefore seems possible that the trough in the Herriger and Hülster curve might be explained by an effect of this kind.

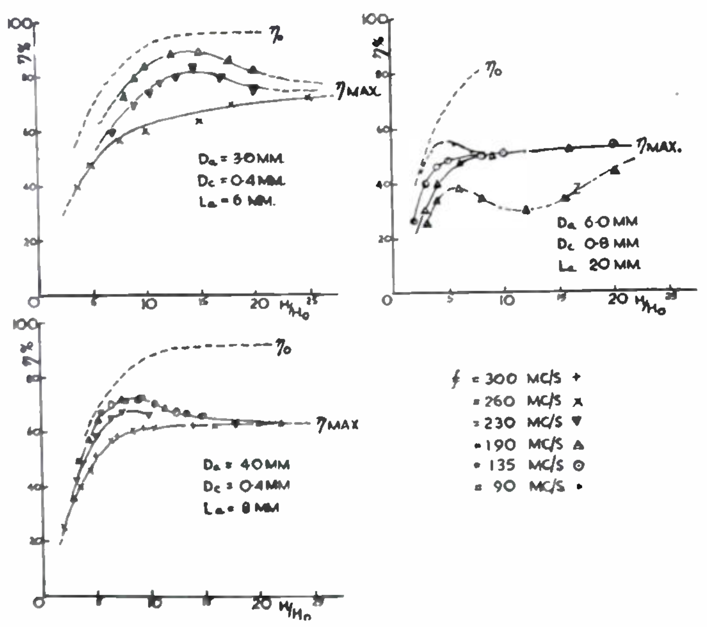

Fig. 3—Efficiency of a number of four-segment magnetrons as a function of reduced magnetic field, showing effect of changing anode diameter and frequency.

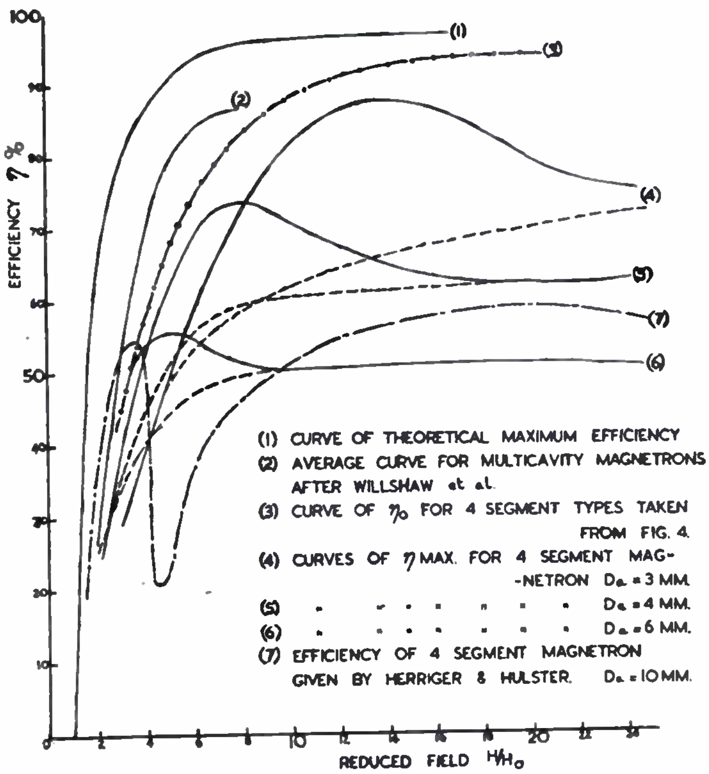

Fig. 4—Efficiency of four-segment magnetrons as a function of reduced magnetic field compared with the theoretical maximum efficiency and that of multisegment types.

The Relation between Anode Diameter and Efficiency

Having seen that magnetrons having few segments and relatively small cathodes show considerable variation of efficiency with anode diameter, one asks whether this is associated with a corresponding variation in some other electrical characteristic, for it seems improbable that efficiency should be affected by changes of dimension alone. It may be argued that if I_p', the oscillating current per unit length of tube, increases with anode diameter less rapidly than the applied voltage V_p, conditions are more favorable to electron bunching and spoke formation in the larger tubes; so their efficiency should be higher. If the ratio V_p/I_p' varies with anode diameter, one cannot apply to magnetrons having few segments the voltage scaling laws applicable to multisegment types, whose efficiency is generally assumed to be independent of anode diameter.

For two-segment magnetrons, direct comparison measurements of the minimum and maximum operating current were made at selected fixed frequencies in the range 150 to 300 mc with the magnetic field chosen to make H/H_0 vary in steps from 2 to 10. Interpretation of the results was complicated by the effect of gap dimensions, waste current, and minor irregularities in construction; but it was concluded that:

(a) For low values of H/H_0, where η_{max} is roughly proportional to $D_a^{1/2}$, V_p is more nearly proportional to $D_a^{3/2}$ than to D_a^2. For higher values, where the levelling off in efficiency takes place, V_p is approximately proportional to D_a^2.
(b) I_m', the minimum oscillating current per unit length of tube, is roughly the same for all tubes at a given value of H/H_0.
(c) The maximum value of oscillating current per unit length is roughly proportional to the diameter.

Thus there appears to be at least an empirical connection between η_{max} and the ratio V_p/I_m'; for when H/H_0 is less than 5, and η_{max} is proportional to $D_a^{1/2}$, so also is the corresponding value of V_p/I_m'.

For four-segment magnetrons, direct-comparison measurements were limited to values of H/H_0 between 3 and 8. Interpretation of data was complicated in the same way as for two-segment types, and also by the ability of the four-segment tubes to operate anywhere in the region bounded by the upper and lower curves of Fig. 4. Even so, there is evidence to show that:

(a) V_p is proportional to D_a^2 to within the limits of measurement.
(b) When H/H_0 is less than 5, and the tubes are operating on the lower branches of the efficiency curves, I_m' is roughly proportional to D_a^2, and so to V_p. Under these conditions, η_{max} varies very little with anode diameter.
(c) When H/H_0 is greater than 5, I_m' increases more rapidly than D_a^2 and V_p. Under these conditions, the larger tubes have lower efficiency.
(d) If comparisons are made between tubes operating on the upper branches of the curves, or between one on the upper and one on the lower, the relation of I_m' to D_a, and so to V_p, is modified in a way that is compatible with the initial assumption that efficiency is in some way proportional to V_p/I_p'.

The measured characteristics of four-segment magnetrons are therefore intermediate between those of two-segment and multisegment types. Operating on the lower branches of the curves, their efficiency is sensibly independent of anode diameter, and they conform to the voltage scaling laws that are applied to multisegment tubes. On the upper branches, they show the lack of proportionality of V_p and I_m' to D_a^2 and the variation of efficiency with diameter that is characteristic of two-segment magnetrons. In this connection, the agreement between the empirical formulas for $(H/H_0)_l$ for two-segment tubes and $(H/H_0)_p$ for four-segment magnetrons may be more than coincidence.

The Effect of Changes in Cathode-to-Anode Ratio

The two types have one great difference in geometry if high efficiency is obtained from each—the relative sizes of cathode and anode. In multisegment tubes, D_c is roughly equal to or greater than $D_a/2$, whereas the cathodes in our tubes were relatively much smaller, D_c being never more than $D_a/5$. One therefore asks whether this could account for the difference in characteristics.

Fig. 5 shows a family of curves for a 6-mm tube with various sizes of cathode. Similar results were obtained from a 3-mm tube, and from the two sets of observations it was concluded that when D_c is roughly equal to or greater than $D_a/4$ efficiency is not significantly dependent on anode diameter. Under these conditions there is no peak in the efficiency curves and the tubes conform to the scaling laws. They differ from multisegment magnetrons only in having intrinsically lower efficiency.

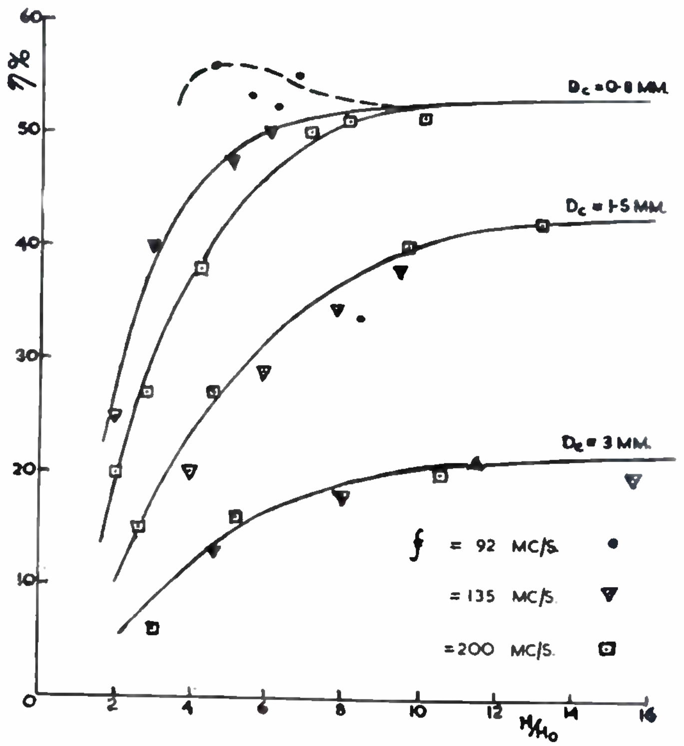

Fig. 5—Efficiency of a four-segment magnetron as a function of reduced magnetic field, showing effect of increasing cathode diameter.

The disappearance of the peaks in η_{max} as the cathode diameter is increased suggests that the upper branches of Fig. 4 may correspond to the peak and trough in the Herriger and Hülster curve for their 10-mm tube, particularly if they used a filamentary cathode and so obtained a cathode-to-anode ratio very much lower than the $\frac{1}{8}$ of the 3-, 4-, and 6-mm types used in the experiments described above.

3. Summary of Experimental Results

It has been shown that efficiency comparable with that of multisegment magnetrons can be obtained from two- and four-segment types by suitable choice of geometry and operating conditions. For high efficiency, the cathodes are relatively much smaller in the tubes having few segments, which show a variation of efficiency with anode diameter.

This variation is accompanied by a corresponding variation in the ratio of applied voltage to the current per unit length of tube, so that tubes of this kind do not conform to the voltage scaling laws that are commonly applied to multisegment types. In other respects, notably the restriction of the range of oscillating current as the magnetic field is increased and the rise in efficiency with current near the low-current boundary, the behavior of two- and four-segment magnetrons closely resembles that of multisegment types, as does that of the four-segment tubes in the apparently small variation of efficiency with operating current.

Finally, it has been shown that the differences in the measured characteristics of multisegment magnetrons and of those having few segments can be related to the difference in the relative size of the cathode that is necessary if both types are to give their maximum possible efficiency. When the cathode-to-anode ratio is comparable for the two types, tubes having few segments have intrinsically lower efficiency, but their characteristics show no fundamental dissimilarity to those of multisegment magnetrons.

4. Comment

It has been suggested that a relatively small cathode gives high efficiency because electrons that might otherwise dissipate their energy in back-bombardment of the cathode can pass through the center of the interaction space and give up their energy to the rf fields during a subsequent traverse to the anode. However, experiments made during this study have shown that six-segment magnetrons with small cathodes either fail to oscillate at all, or will oscillate only with low values of reduced magnetic field and then give very low efficiency. The same effect was also found in four-segment tubes, though not so marked. In both cases it could be eliminated by an increase in cathode diameter, or by equivalent off-centering of a small cathode.

An alternative explanation of the role of the cathode in determining magnetron characteristics assumes that too small a cathode will prevent build-up of a stable electron spoke system, because this would result in regions near the cathode having a space-charge density exceeding that which the fields therein could maintain in equilibrium. Thus for easy starting and efficient operation in the normal magnetron mode, a tube of given size requires a particular minimum size of cathode. As the number of segments is increased, penetration of the rf fields into the interaction space will decrease, and so the minimum size of cathode will increase. By reason of the greater bunching and possibly higher operating current density, one would also expect an increase with increasing field and applied voltage, or alternatively, a drop out of oscillation above some critical field dependent upon cathode-to-anode ratio. Off-centering a small cathode will move it into a region of higher fields, and so permit build-up of oscillation, even though efficiency would probably be lower than that obtained with a larger axially mounted cathode. The results of our experiments agree with all these conclusions.

If the cathode-to-anode ratio is greater than the permissible minimum, space-charge defocusing and collection of electrons before they give up their energy to the rf fields will be more pronounced in magnetrons having few segments, and so their efficiency should be lower than that of multisegment types.[5] This effect will become more marked as the cathode-to-anode ratio is increased. If the reduced magnetic field and applied voltage are increased, the higher operating current density and greater degree of bunching will accentuate these effects. One would therefore expect the difference in efficiency of the two types to increase with both D_c/D_a and H/H_0. Experimentally this is so.

If the cathode is too small for operation in the normal magnetron mode, the tubes can oscillate only in the "minimum voltage" regime[6] described by Willshaw and Robertshaw, in which the fields near the cathode have no part in the mechanism of oscillation. Using a block with a large number of segments, they have found that with the low cathode-to-anode ratio required for efficient operation in the minimum voltage regime, the normal magnetron mode of operation is suppressed. An increase in cathode diameter permits the tube to operate either in the minimum voltage regime or in the normal mode with low values of reduced field. A further increase results in suppression of the minimum voltage mode and extension of the normal magnetron mode to higher values of reduced magnetic field. This effect of changing cathode-to-anode ratio is much the same as that which has been found in our experiments with two-, four-, and six-segment magnetrons. Operation in the minimum voltage regime is limited to values of H/H_0 around 1.5. Under these conditions the power output from the magnetrons used in our experiments would be too small to be measured by our apparatus. However, some of the four-segment tubes, but none of the two-segment ones, gave some indication of operation in the minimum voltage regime. Curves of efficiency against a reduced magnetic field, such as those of Figs. 3 and 4, should theoretically pass through zero efficiency at $H/H_0=1$; and in the case of two-segment magnetrons, the experimental curves do likewise, apart from a slight falling away near the origin of the curves, which can be attributed to an imperfect cutoff characteristic. Some four-segment tubes, however, gave curves of efficiency which fell away from these curves and would, if extrapolated, have passed through zero efficiency at H/H_0 be-

[5] Slater's rule cannot be applied to magnetrons having four segments or less, for it gives $r_c=0$. Experimentally, the optimum cathode-to-anode ratio for two- and four-segment tubes is between 1:10 and 1:8. For six-segment tubes the rule gives $r_c/r_a=1/5$. Experimentally, the optimum value is found to be somewhat higher than this.

[6] W. E. Willshaw and R. G. Robertshaw, "The behaviour of multiple circuit magnetrons in the neighbourhood of the critical anode voltage," *Proc. Phys. Soc.*, vol. 63, p. 41; January, 1950.

tween 1.5 and 2. At the same time, measured efficiencies of a few per cent were obtained at these values of reduced magnetic field. In some extreme cases, when very small cathodes were used, operation was limited to values of H/H_0 below 2, when only by careful adjustment of field and applied voltage could a measurable efficiency be obtained.

5. Conclusion

Experiment has shown that there are no fundamental differences between the characteristics of multisegment magnetrons and those of tubes having few segments. Observed differences in their behavior have been related to the difference in the ratio of cathode-to-anode diameter which is necessary if maximum efficiency is to be obtained from each type.

It appears that for all magnetrons the cathode-to-anode ratio must exceed a certain minimum if oscillation is to be in the normal mode. It must not exceed a certain maximum if high efficiency is required. Both minimum and maximum increase with the number of segments.

The effect of variations in cathode-to-anode ratio has been explained in terms of the maximum permissible space-charge density near the cathode and of space-charge defocusing of the rotating electron cloud. But this explanation is purely qualitative and deals only in very general terms with the phenomena involved. It must be concluded that the exact role of cathode in determining magnetron characteristics is still a matter for speculation.

6. Acknowledgment

This paper is published by permission of the Admiralty. The experimental work which it describes was carried out at the Services Electronics Research Laboratory. The author wishes to record his appreciation of the help of F. A. Howe who constructed most of the experimental tubes and assisted in the measurement of their characteristics.

Appendix

Details of the Experimental Method

The construction of some typical tubes used in the experiments here reviewed is illustrated in Fig. 6. The interaction space is enclosed by a comparatively massive copper anode which permitted operation of the tubes without overheating or other adverse effects at high values of reduced field. In nearly all tubes the length of the anode was roughly twice the diameter of the interaction space, and in no case was it less than one and a quarter times the diameter. The width of the gaps between the segments in both two- and four-segment types was approximately one fifth of the internal diameter of the anodes. Earlier experiments had indicated that the size and position of cathode end-discs had little effect on efficiency, and therefore no end-discs were incorporated, except those formed by the flanges of the eyelets into which the ends of the cathodes were anchored. The available primary emission was usually several times the oscillating current, even when operating with the maximum permissible power input.

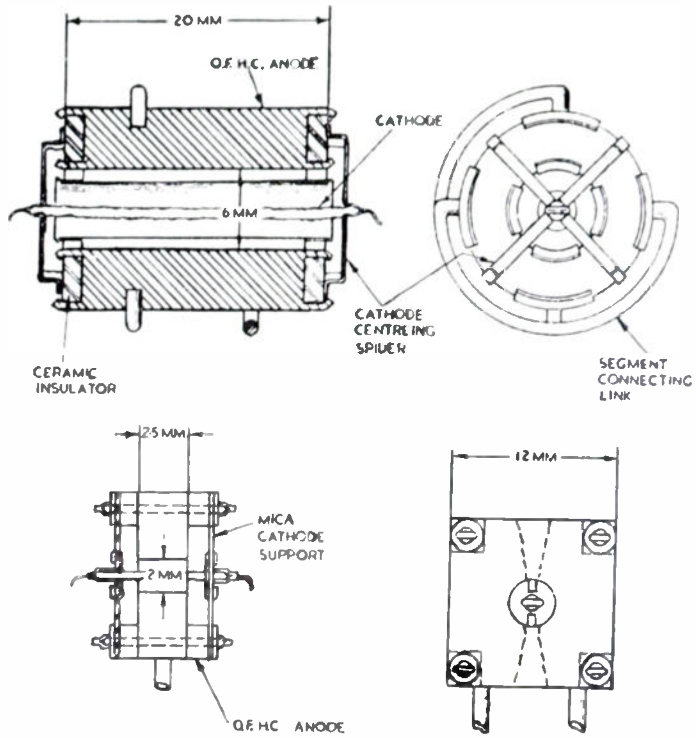

Fig. 6—Construction of two typical experimental tubes.

To avoid as far as possible any freak results due to some undiscovered imperfection or peculiarity of any one tube, measurements, with one or two exceptions, were made on two or more samples of each type. At the beginning of each set of measurements, the tube being tested was checked for cutoff characteristic and aligned in the field to give maximum efficiency at some convenient field strength. In general, the alignment found to give highest efficiency also gave the best cutoff characteristic. A few tubes, however, showed a small but definite increase in efficiency when given a few degrees displacement from the optimum alignment for cutoff. Power output was measured by visual comparison of the brilliance of the rf load lamp with that of a similar lamp placed next to it and operated from a variable dc supply. In general, positioning of the load lamp with respect to the shorting bar was not critically dependent on magnetic field, a minor readjustment being required only when the field was set to values much higher than that for which the optimum position was selected. The load and comparison lamps were selected to have the same resistance when cold and to have similarly shaped filaments. Measurements of power output by three different observers gave general agreement to within 5 per cent of the measured power, as did substitution of one load lamp by another of a different kind, thus indicating that the accuracy of measurement was high enough for reliable and self-consistent results to be obtained. This was confirmed by the close agreement between measurements repeated at considerable time intervals. Measurement of the loaded and unloaded Q's of the rf circuit indicated that the circuit efficiency was of the order 95 per cent or greater, so that the efficiency as measured was taken as the electronic efficiency of the tube being tested. No correction was made for anode input power dissipated in back bombardment of the cathode, which, over most of the range of measurement, appeared to be an insignificant fraction of the total input.

Heater-Voltage Compensation for Alternating-Current Amplifiers*

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Summary—By the use of a direct-current feedback network which minimizes variations in plate current, the output of a radio-frequency amplifier can be made to be independent of heater-voltage fluctuations. Under- or over-compensation may be obtained by choice of circuit components to meet individual requirements. The design procedure for this network is indicated, and the experimental results for a compensated 30-mc broad-band intermediate-frequency amplifier are shown.

Introduction

HEATER-VOLTAGE variations cause changes in the characteristics of vacuum tubes, with a resultant deterioration of gain stability. Methods of reducing and eliminating this troublesome effect have been devised for electrometer circuits[1,2] and direct-current amplifiers,[3,4,5] but little has been done to compensate for these changes in alternating-current amplifiers, especially at high frequencies. In high-gain applications where gain stability is of any consequence some method of controlling heater voltage or compensating for changes in heater voltage must be used. Since the change in output of a high-gain amplifier may be several times the change in heater voltage, methods of obtaining sufficient heater-voltage control become increasingly difficult as greater gain stability is required. Seeger[6] has shown that large cathode and screen-grid resistors greatly improve the stability of tube types 6AK5 and 6AG5. In this paper a feedback network (Fig. 1) for stabilizing tube characteristics and completely compensating for heater-voltage changes is described, and the experimental results for a 30-mc compensated amplifier and a conventional 30-mc amplifier are shown for purposes of comparison.

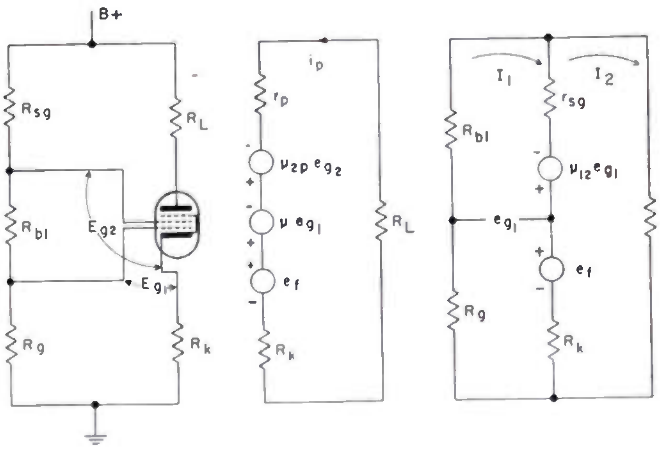

Fig. 1—Basic circuit for heater-voltage compensation.

General

For a tube operated under space-charge conditions, variations of the cathode temperature vary the average initial electron velocity, resulting in a change of plate current. Since transconductance is approximately proportional to the plate current,[7] a change in heater voltage results in a change in transconductance. Conversely—any method of preventing plate current from changing will stabilize the transconductance. The circuit shown in Fig. 1 operates on the above principle. Any tendency for plate and screen-grid current to change is counteracted by a change of bias voltage applied to the grid by the feedback network R_{sg}, R_{b1}, and R_g.

* Decimal classification: R363.1. Original manuscript received by the Institute, June 27, 1951; revised manuscript received, February 25, 1952.

† Radio and Electrical Eng. Div., National Research Council, Ottawa, Canada.

[1] F. E. O'Meara, "Circuits and tubes for ultra-sensitive electrometers, *Rev. Sci. Inst.*, vol. 22, pp. 106–108; February, 1951.

[2] J. J. Dowling and C. O'Ceallaigh, "The balancing of valve circuits," *Phil. Mag.*, vol. 20, pp. 532–542; September, 1935.

[3] A. G. Bousquet, "Random emission compensation," *Gen. Rad. Exper.*; May, 1944.

[4] G. E. Valley, Jr., and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Radiation Laboratories Series No. 18, pp. 458–467; 1948.

[5] J. M. Sowerby, "Reducing drift in d-c amplifiers," *Wireless World*, vol. LVI, pp. 293–295; August, 1950; and pp. 350–352; October, 1950.

[6] C. L. Seeger, "Notes on the Design of Extra-terrestrial Radio Noise Receivers," URSI Convention, Washington, D.C.; April, 1950.

[7] For some of the factors affecting this relationship, see: (a) J. Deketh, "Fundamentals of Radio Valve Techniques," Book I, N. V. Philips Gloeilampenfabrieken, Eindhoven, pp. 431–444; 1949. (b) W. Raudorf, "Change of mutual conductance with frequency," *Wireless Eng.*, vol. 26, pp. 331–337; October, 1949.

Theory and Design Formulas

For purposes of solution it is postulated that the cathode potential variations (due to heater-voltage changes) can be represented in the cathode lead as a generator, e_f, of zero internal impedance. It is assumed that the effect of the suppressor-grid is negligible and that the change in plate current due to heater voltage variations will have negligible effect on the screen-grid current. We may then set up the equivalent plate and screen-grid circuits, as shown in Fig. 1. From the equivalent plate circuit, we have

$$i_p(R_1 + r_p + R_k) = e_f - \mu e_{g1} - \mu_{2p} e_{g2}, \qquad (1)$$

where e_{g1}, e_{g2} are functions of e_f, and μ_{2p} is the screen-grid to plate-amplification factor.

The right-hand side of this equation must be zero in order that plate current be independent of e_f. Thus

$$e_f = \mu e_{g1} + \mu_{2p} e_{g2}. \qquad (2)$$

We may solve for e_{g1}, e_{g2} from the equivalent screen-grid circuit by taking loop currents I_1 and I_2,

$$\begin{aligned} I_1(R_g + R_{b1} + r_{sg} + R_k) - I_2(R_k + r_{sg}) &= e_f - \mu_{12} e_{g1} \\ - I_1(R_k + r_{sg}) + I_2(R_k + r_{sg} + R_{sg}) &= - e_f + \mu_{12} e_{g1} \\ I_1(R_k + R_g) - I_2 R_k &= e_f + e_{g1}. \end{aligned} \qquad (3)$$

Solving,

$$e_{g1} = - e_f \frac{1 - \dfrac{\alpha}{\beta}}{1 + \mu_{12} \dfrac{\alpha}{\beta}} \qquad (4)$$

and,

$$e_{g2} = e_f \frac{R_{sg}(R_g + R_{b1})}{\beta} \frac{(1 + \mu_{12})}{1 + \mu_{12} \dfrac{\alpha}{\beta}}, \qquad (5)$$

where

$$\alpha = R_{sg}(R_k + R_g) + R_k(R_g + R_{b1})$$

$$\beta = R_{sg}(R_{b1} + R_g + R_k + r_{sg}) + (R_g + R_{b1})(R_k + r_{sg}).$$

The conditions for $i_p = 0$ may be obtained by substituting (4) and (5) into (2);

$$1 + \mu_{12} \frac{\alpha}{\beta} = \mu_{2p} \frac{R_{sg}(R_g + R_{b1})}{\beta} (1 + \mu_{12}) - \mu\left(1 - \frac{\alpha}{\beta}\right). \qquad (6)$$

Equation (6) may be expanded and, in most practical cases, simplified into

$$\begin{aligned} & R_{b1}[r_{sg}(1 + \mu) - R_{sg}(\mu_{2p} - 1)] \\ & - R_g[R_{sg}(\mu + \mu_{2p} - \mu_{12} - 1) - r_{sg}(1 + \mu)] \\ & + r_{sg} R_{sg}(1 + \mu) = 0. \end{aligned} \qquad (7)$$

Since the amplification factors and the screen-grid resistance r_{sg} become fixed when the operating potentials are chosen, (7) becomes a function of R_{b1}, R_g, and R_{sg}. If R_{sg} is determined, values of R_{b1} and R_g that satisfy (7) may be chosen.

The following procedure may be followed if it is desirable to employ a common voltage source for the plate and screen-grid circuits:

(a) The grid bias E_{g1}, the screen-cathode voltage E_{g2}, and the plate-cathode voltage E_p are chosen. The characteristics of the tube operated under these conditions are determined. For the average 6AK5, with $E_{g1} = -2$v, $E_{g2} = 110$v, and $E_p = 180$v, the amplification factors μ, μ_{12}, and μ_{2p} were found to be 2,300, 26.0, and 88.8, respectively. The plate and screen-grid currents I_p and I_{sg} were 7.6 and 2.5 ma, and the internal screen-grid resistance was approximately 19 k.

(b) Neglecting the bleeder current, R_{sg} is calculated from the approximate formula

$$R_{sg} \approx \frac{I_p R_L + E_p - E_{g2}}{I_{sg}}, \qquad (8)$$

and is approximately 39 k for the 6AK5 operated under the above conditions.

(c) R_g and R_{b1} are chosen from (7). Substituting the above values into (7), we have

$$R_{b1} - 1.20 R_g + 0.04 = 0, \qquad (9)$$

which is plotted in Fig. 2.

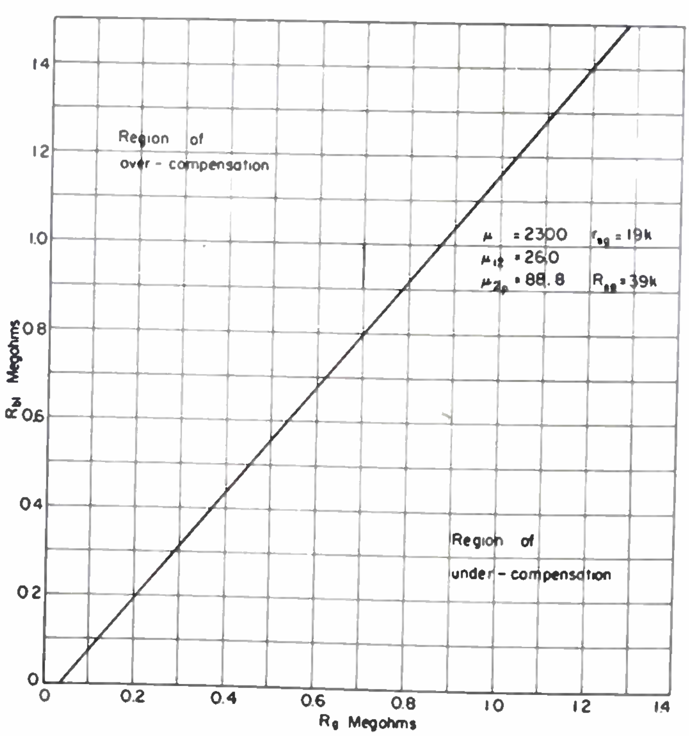

Fig. 2—Plot of $R_{b1} - 1.20\ R_g + 0.04 = 0$.

One megohm and 820,000 ohms were selected for R_{b1} and R_g in this experimental circuit. It can be seen that this is in the region of over-compensation, but it was desirable to use standard values as far as possible. If

R_g and R_{b1} are made smaller, bleeder current must be taken into consideration in step (2). It may be necessary to determine the correct values of resistors by a series of successive approximations.

(d) The cathode resistance that will provide the correct grid bias is given by

$$R_k = \frac{E_{g2}R_g}{I_kR_{b1}} - \frac{E_{g1}}{I_k}\,\frac{R_{b1}+R_g}{R_{b1}}, \tag{10}$$

where I_k is the total cathode current. For the above conditions R_k was 9,600 ohms.

Experimental Results

Fig. 3 shows the average change in transconductance with heater voltage for the 6AK5 tube. Component values as developed in the example above were used in the compensating circuit of Fig. 1.

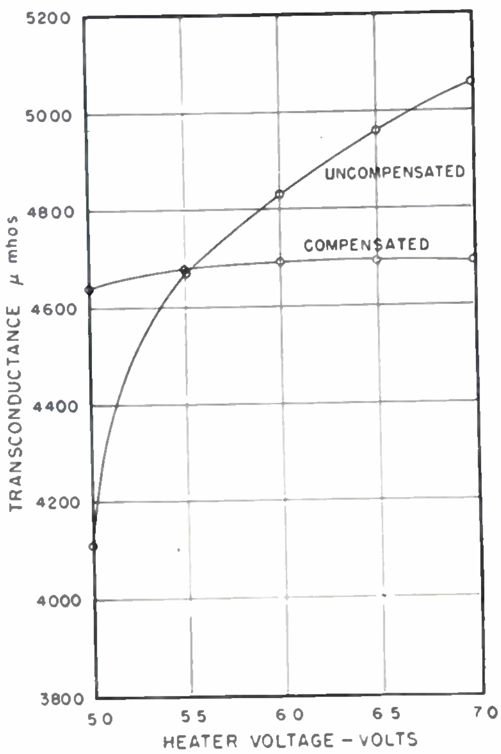

Fig. 3—Average transconductance versus heater-voltage characteristics for the 6AK5.

Characteristics of individual tubes varied as much as ±15 per cent from the average values, resulting in some tubes showing either under- or over-compensation when tested in the circuit of Fig. 1. Approximately half of the tubes tested showed excellent compensation, while about 10 per cent had such extreme changes in transconductance with heater voltage that they were unacceptable for highly stable amplifiers. In every case compensation decreased the change in transconductance with heater-voltage changes.

It is possible to obtain under- or over-compensation by a suitable choice of the circuit resistors. In a multistage amplifier, over-compensation may be applied to a single stage to counteract the changes in the remaining stages. If complete compensation of a single stage is required, the circuit components may be adjusted to give compensation for any tube; however, in practice it may be easier to select a tube which nearly conforms to the average amplification factors.

Where several stages are compensated, as in a multistage amplifier, the over-compensation that may occur in one stage may be counteracted by the under-compensation of one of the remaining stages. As a consequence of this probable distribution, the over-all heater-voltage compensation may be several times better than that of one stage; and there will be less stringent requirements on the uniformity of tube characteristics. The values of the resistors in the compensating network may be altered, or the supply voltage may be changed slightly to achieve the best compensation.

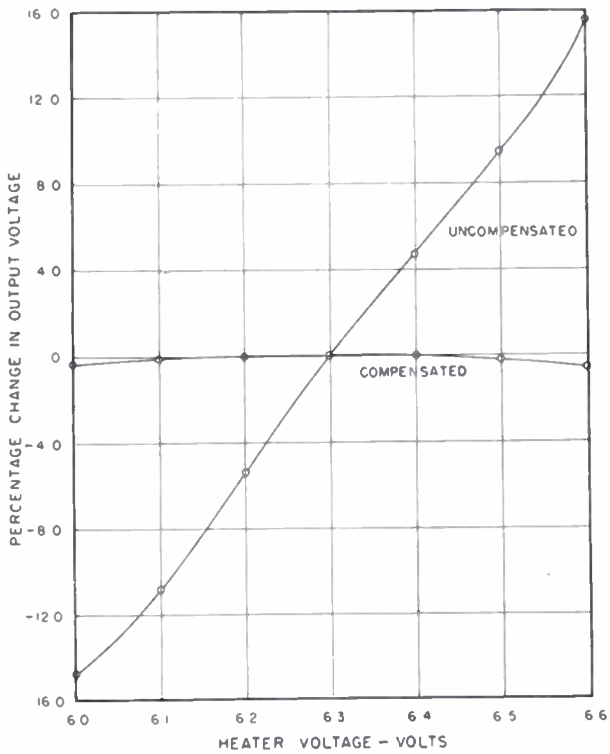

Fig. 4—Percentage change in output voltage versus heater voltage for two 30-mc intermediate-frequency amplifiers, identical in gain, bandwidth, and noise figure.

Fig. 4 is a comparison between two 30-mc intermediate-frequency amplifiers of the feedback-pair type. They had a bandwidth of 10 mc, a gain of 125 db, and nearly equal noise figures. The compensated amplifier employed the dc network of Fig. 1 in each of the ten stages. For a heater-voltage change of ±1 per cent from 6.3 volts, the output of the uncompensated amplifier changed by ±3 per cent, while the output of the compensated amplifier changed by only −0.05 per cent.

Acknowledgment

The author wishes to express his appreciation of the assistance given him by W. J. Medd in checking the formulas and for his many valuable suggestions.

Design of Optimum Buried-Conductor RF Ground System*

FRANK R. ABBOTT†

Summary—From the design equations developed, a radial-conductor ground system may be constructed which will provide maximum power radiated per dollar of over-all cost. The fundamental equations are particularly applicable to any installation operating in the frequency range below 1,000 kc. A radial-conductor ground system based on the design parameters here presented will assure low-frequency antenna efficiencies of 50 per cent or better as compared to efficiencies of 25 per cent or less usually encountered in similar antennas without such a ground system.

A technique for achieving optimum design of radial-conductor ground systems was developed. Curves were prepared which facilitate the determination of optimum spacing and length of the ground radials. A typical computation is included.

Introduction

It is an accepted fact that transmitting antennas operating in the frequency range below 1,000 kc are in most cases rather inefficient radiators. The greatest single factor leading to this inefficiency is usually the ground system. With few exceptions[1] (see list of references), rules for ground-system design are empirical and based on results from existing installations.

The problem of obtaining optimum design in a radial-conductor ground system lies in attaining maximum power radiated per dollar of over-all cost. The design of such a system is analogous to determining optimum conductor size by Kelvin's Law.[2] The method offered here is more involved, but provides a precise and definitive procedure for the design of an optimum ground system for any antenna installation, provided the soil conductivity is accurately known. Both calculations and measurements indicate that the use of such properly designed radial-conductor ground systems should lead to antenna-radiation efficiencies of 50 per cent or more compared to the 25 per cent or less now tolerated.

In the following text, the step-by-step analytical procedures are explained, and a typical problem is solved in detail. Appendix I contains a derivation of the admittance of a single conductor in the system, a factor required in the analysis.

Theory of Design

Optimum design of the ground system, when the design of the antenna and the transmitter is fixed, is accomplished when the total annual expense of such a ground system is at a minimum. The annual expense includes amortization cost of copper and its installation, plus the cost of the power dissipated in the ground. Optimum design, therefore, requires investigation of proper spacing of conductors in any area over which the total ground-system current density remains reasonably independent of the ground conductivity.

Copper costs are determined by wire diameter and conductor spacing. Wire diameter is dictated by practical considerations. Assume a wire of radius a, costing, when placed a safe distance under the surface, K_1 dollars per meter-year. Then the yearly cost per square meter is K_1/d, where d is the space between wires measured in meters.

The second factor in the annual expense is the cost of the power dissipated in the ground. This cost is determined by adding the amortization cost of the station, exclusive of the ground system, to the annual operating costs of the station, dividing the sum by the mean antenna input power in watts and multiplying by the watts dissipated per square meter, w, with a conductor spacing d. This cost term obviously increases with increased spacing of conductors. The total annual cost per meter, K_t, to be minimized by varying d then becomes

$$K_t = \frac{K_1}{d} + K_2 w, \tag{1}$$

in which

$$K_2 = \frac{C_a + C_b}{W}, \tag{2}$$

where C_a is the annual amortization cost of the station exclusive of the ground system, C_b is the annual operating cost of the station, and W is the average input power to the antenna in watts.

The power dissipated per square meter of ground surface is given by the real part of $|J^2|Z$, in which J is the total current flowing across a square meter of area and Z is the impedance of a square meter. Since the current divides between the embedded conductors and the surrounding soil, it is easier to work with admittances. Thus the loss, w, is expressed as the real part of $|J^2|/Y$, in which Y is the sum of Y_s, the characteristic admittance of the soil, and Y_g, the admittance of the grid of wires on the surface of the meter square, which will be almost purely susceptive.

* Decimal classification: R126×R320.3. Original manuscript received by the Institute, September 10, 1951; revised manuscript received, February 25, 1952. This report covers work to October 30, 1950.

† Development Division, U. S. Navy Electronics Laboratory, San Diego, Calif.

[1] H. E. Girhing and G. H. Brown, "General considerations of tower antennas for broadcast use," Proc. I.R.E., vol. 23, pp. 311–356; April, 1935 and G. H. Brown, R. F. Lewis, and J. Epstein, "Ground systems as a factor in antenna efficiency," Proc. I.R.E., vol. 25, pp. 753–787; June, 1937.

[2] "Standard Handbook for Electrical Engineers," McGraw-Hill Book Co., Inc., New York, N. Y., 6th ed., p. 1488; 1933.

The characteristic admittance of the soil, Y_s, is given by[3]

$$Y_s = \sqrt{\frac{\epsilon - \frac{j\sigma}{\omega}}{\mu}}.$$

In the notation employed,
μ is the permeability of the soil,
σ is the conductivity of the soil,
ϵ is the dielectric constant of the soil,
ω is 2π times the frequency, f, and
j is equal to $\sqrt{-1}$.

In the case of a low frequency installation, any reasonable soil conductivity makes $(\sigma/\omega) \gg \epsilon$ and the admittance becomes

$$Y_s \approx \sqrt{\frac{-j\sigma}{\omega\mu}} = \frac{1-j}{\sqrt{2}}\sqrt{\frac{\sigma}{\omega\mu}} = (1-j)\sqrt{\frac{\sigma}{4\pi f\mu}} \text{ mhos.}$$

The grid susceptance to uniformly distributed excitation will be designated Y_g, and is a function of the spacing, d, and radius, a, of the conductors.

It is shown analytically in Appendix I, and has been confirmed experimentally,[4] that for $d \ll$ one wavelength, λ, the admittance of soil covered by a grid assumes a simple form, given by

$$Y = Y_s + Y_g.$$

It is further shown in Appendix I that

$$Y_g = -j\frac{1}{f\mu d \log_e \frac{d}{2\pi a}} \text{ mhos.}$$

Thus the total admittance of the soil and grid becomes

$$Y = Y_s + Y_g \approx (1-j)\sqrt{\frac{\sigma}{4\pi f\mu}} - \frac{j}{f\mu d \log_e \frac{d}{2\pi a}} \text{ mhos.}$$

Recalling that the power dissipated per square meter, w watts, is equal to the real part of $|J^2|/Y$, we obtain

$$w = \frac{|J^2|\sqrt{\frac{\sigma}{4\pi f\mu}}}{\left[\frac{\sigma}{4\pi f\mu} + \left(\sqrt{\frac{\sigma}{4\pi f\mu}} + \frac{1}{f\mu d \log_e \frac{d}{2\pi a}}\right)^2\right]} \text{ watts/meter}^2. \tag{3}$$

This analytical expression relating watts lost per square meter and the grid spacing, d, permits expression of annual cost of power lost, wK_2, and finally the total annual cost associated with one square meter of ground surface, in terms of the grid spacing, d.

In cases where the frequency is high and/or the soil conductivity is low, the dielectric constant, ϵ, is no longer unimportant compared to σ/ω. In these cases, the admittance of the soil is given by

$$Y_s = \sqrt{\frac{\epsilon}{\mu} - j\frac{\sigma}{\omega\mu}},$$

which can be expressed as

$$Y_s = U + jV.$$

The power dissipated into the ground, w, then becomes the real part of

$$\frac{|J^2|}{U + jV - j\frac{1}{f\mu d \log_e \frac{d}{2\pi a}}}, \text{ or}$$

$$w = \frac{|J^2|\,U}{U^2 + \left(V - \frac{1}{f\mu d \log_e \frac{d}{2\pi a}}\right)^2}. \tag{4}$$

Determination of Optimum Number of Radial Conductors

In the region where the ground system is effective, the final quantity in the denominator of (3) is dominant, and the expression for power dissipated into the ground reduces to

$$w \approx |J^2|\sqrt{\frac{\sigma}{4\pi f\mu}}\left(f\mu d \log_e \frac{d}{2\pi a}\right)^2$$
$$\approx |J^2|\, d^2 \sqrt{\frac{\sigma f^3\mu^3}{4\pi}} \log_e^2 \frac{d}{2\pi a}. \tag{5}$$

Because $\log_e (d/2\pi a)$ varies slowly with changing values of d, the yearly cost of power lost in the ground may be expressed as

$$wK_2 \approx C_1 d^2,$$

in which

$$C_1 \approx \left(|J^2|\sqrt{\frac{\sigma f^3\mu^3}{4\pi}} \log_e^2 \frac{d}{2\pi a}\right) K_2. \tag{6}$$

The total cost, K_t, then becomes

$$K_t = C_1 d^2 + K_1/d \text{ dollars per year.} \tag{7}$$

Differentiating (7) with respect to d, and setting the derivative equal to zero to obtain a minimum, leads to

$$2C_1 d - \frac{K_1}{d^2} = 0, \quad \text{or} \quad C_1 d^2 = \frac{K_1}{2d}. \tag{8}$$

[3] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand and Co., Inc., New York, N. Y., p. 243; 1943.

[4] F. R. Abbott, "Standard Radial Ground Systems for LF and MF Monopole Transmitting Antennas," Navy Electronics Laboratory Report 219; February 9, 1951.

Substituting the expression for C_1 from (6) gives

$$\left(|J^2|\sqrt{\frac{\sigma f^3\mu^3}{4\pi}}\log_e{}^2\frac{d}{2\pi a}\right)K_2 d^2 = \frac{K_1}{2d}, \tag{9}$$

so that

$$d^3\log_e{}^2\frac{d}{2\pi a} = \frac{K_1}{2K_2|J^2|\sqrt{\dfrac{\sigma f^3\mu^3}{4\pi}}}. \tag{10}$$

Let

$$\alpha = \frac{K_1}{K_2} = \frac{\text{yearly cost of one meter of buried conductor}}{\text{cost of one watt year of rf power}}.$$

By assuming nonferromagnetic ground, the evaluation of

$$2\sqrt{\frac{\mu^3}{4\pi}} = 2\sqrt{\frac{(4\pi\times 10^{-7})^3}{4\pi}} = 7.9\times 10^{-10}$$

reduces (10) to

$$d^3\log_{10}{}^2\frac{d}{2\pi a} = \frac{2.38\times 10^8\times a}{|J^2|\sqrt{\sigma f^3}} = F_1, \tag{11}$$

where F_1, a function of wire diameter and conductor spacing, becomes the optimum-spacing parameter.

Just as (3) leads to (11) defining the optimum-spacing parameter where the earth conductivity is relatively good, (4) leads to a value of F_1 for the case where both terms in the complex admittance of the soil are important. In this case,

$$F_1 \approx d^3\log_{10}{}^2\frac{d}{2\pi a} = \frac{0.094a}{|J^2|\mu^2 f^2 U}. \tag{12}$$

If the dielectric character of the soil becomes dominant,

$$F_1 = \frac{0.094a}{|J^2|f^2\sqrt{\mu^3\epsilon}}. \tag{13}$$

In mks units, the ground-surface current is numerically equal to the field of magnetic intensity, H, just above the surface. If the antenna system is energized, H may be established by a field-strength meter, and the spacing at any point may be determined by evaluating F_1.

For simple vertical radiators, the surface current density, J, can be calculated as a function of the radial distance from the base of the antenna.[2]

$$\frac{\lambda|J|}{I_0} = \frac{1}{2\pi\dfrac{r}{\lambda}\sin\beta h}\left(e^{-j\beta\rho} - e^{-j\beta r}\cos\beta h\right), \tag{14}$$

where

β is the propagation constant, $2\pi/\lambda$,
ρ is the distance from the top of the antenna, and
r is the distance from the base of the antenna.

Fig. 1 presents computed[5] values of $(\lambda|J|)/I_0$ versus the distance, r/λ, enabling the rapid evaluation of surface current density, $|J|$, as a function of antenna base current, I_0 and antenna height, h/λ.

Specification of the most appropriate values for current density, J, ground conductivity, σ, frequency, f, and the annual-cost ratio, a, enables evaluation of the right-hand side of (11), designated by F_1.

[5] Sine-wave current distribution is assumed.

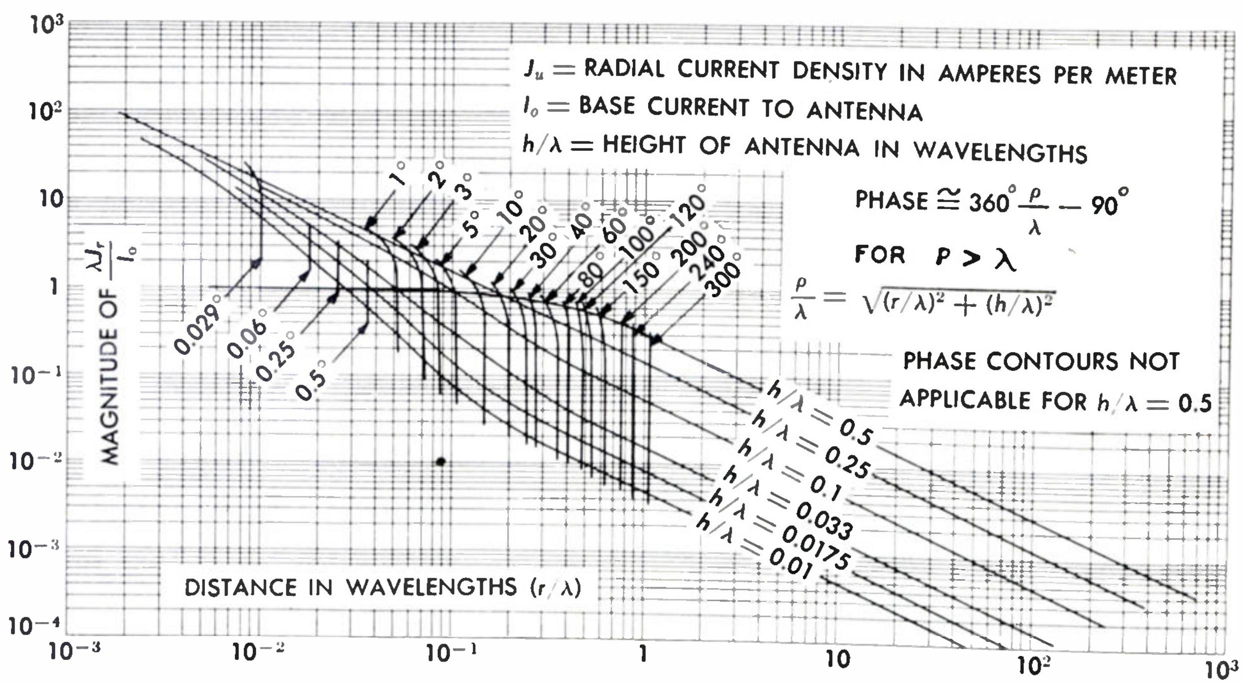

Fig. 1—Phase and density of radial ground current.

Curve I of Fig. 2 gives optimum spacing, d, as a function of the parameter F_1, using No. 8 awg wire (diameter equals 0.00325 meter) as conductor size.

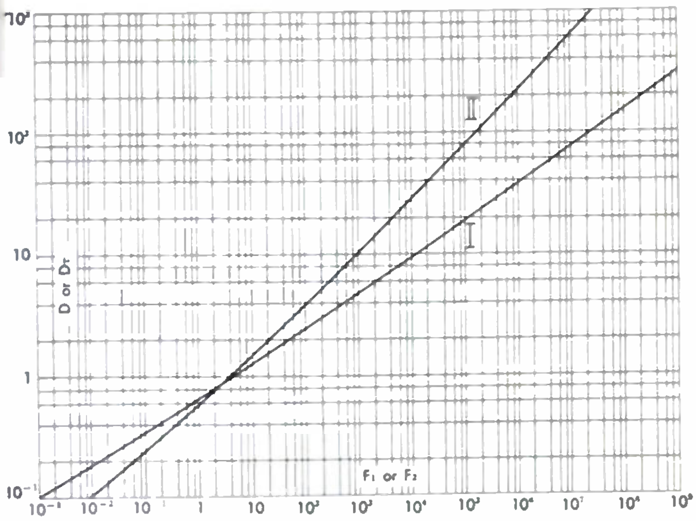

Fig. 2—Optimum and terminal spacing of buried conductors.

Having determined from Fig. 2 the appropriate conductor spacing at several distances, r, the desirable number of radials at any distance is given by

$$N = \frac{2\pi r}{d} = \text{optimum number of radial conductors.} \quad (15)$$

Usually the number remains substantially constant and permits selection of a fixed number for the system.

The application of infinite parallel-wire grid theory to the foregoing problem leads to valid results only when the optimum number of radial conductors is large enough that the angle between adjacent conductors is small, and the electrical length of the conductors in the soil exceeds one wavelength. The latter condition can be violated in the case of a ground system for a low-power, very low-frequency installation if soil conductivity is extremely poor. Such cases are unusual, and generally require special consideration.

Determination of Optimum Length of Radial Conductors

Consider the second term in the denominator of (3),

$$\sqrt{\frac{\sigma}{4\pi f\mu}} + \frac{1}{f\mu d \log_e \dfrac{d}{2\pi a}}.$$

As previously stated, the final quantity is dominant in the region where the radials are essential. However, prescribing the total number of radials automatically determines a radial distance past which an increase in the spacing, d, makes the first term become the larger. Since the optimum spacing increases very rapidly beyond this point, it becomes unimportant to continue the radials farther.

The proper length for the N radials is then obtained when

$$f\mu d_t \log_e \frac{d_t}{2\pi a} \approx \sqrt{\frac{4\pi f\mu}{\sigma}}, \quad (16)$$

where d_t is the terminal spacing.

Rewriting (16) and changing the logarithm base gives

$$d_t^2 \log_{10}^2 \frac{d_t}{2\pi a} \approx \frac{2.36}{f\mu\sigma} = F_2, \quad (17)$$

in which F_2 becomes the length parameter.

Where no ferromagnetic material is present,

$$\mu = 4\pi \times 10^{-7},$$

and

$$d_t^2 \log_{10}^2 \frac{d_t}{2\pi a} = \frac{1.88 \times 10^6}{f\sigma} = F_2. \quad (18)$$

Curve II of Fig. 2, which is a plot of d_t versus $2.36/f\mu\sigma$, defines the terminal spacing for a conductor of No. 8 awg wire. Since $d = 2\pi r/N$, the value of r at $d = d_t$ gives the length of the radial conductors. Thus,

$$r_t = N \frac{d_t}{2\pi}, \quad (19)$$

which is the recommended length of the ground system.

For the case where earth conductivity is relatively good ($\sigma \gg \omega\epsilon$), the approximate equality, $Y_e \approx Y_g$, leads to a value of the recommended length parameter given by

$$F_2 = d_t^2 \log_{10}^2 \frac{d_t}{2\pi a} = \frac{0.188}{f^2\mu^2} \left| \sqrt{\frac{\omega\mu}{\omega\epsilon - j\sigma}} \right|; \quad (20)$$

and again, if the conductivity may be neglected ($\sigma \ll \omega\epsilon$), the length parameter reduces to

$$F_2 = \frac{0.188}{f^2\sqrt{\mu^3\epsilon}}. \quad (21)$$

Power Loss in an Optimum Radial Ground System

The integrated power loss, P, in a radial ground system as derived from (5) can be expressed by

$$P \approx \int_0^{2\pi} \int_{r_1}^{l} \left\{ 20\sqrt{\sigma f} \frac{I_0^2}{N^2} \left[\frac{2\pi r}{\lambda} \left(\frac{I_r}{I_0} \right) \right]^2 \log_{10} \frac{r}{Na} \right\} dr\, d\phi, \quad (22)$$

in which r_1 is the effective radius of the transmission line exciting the antenna. Curves of the quantity in the brackets have been prepared for various antenna heights, and are available.

A simpler approximation for the power absorbed in the soil, from the antenna to the extremity of an optimum ground system, is easily derivable from (8), which states that, with an optimum system, the annual cost of power dissipated in the ground is one-half the annual cost of the buried copper. Since the annual cost for buried copper is K_1Nr_t, the annual cost for power dissipated is

$$PK_2 = \frac{K_1Nr_t}{2} \text{ dollars,}$$

or

$$P = \frac{\alpha Nr_t}{2} \text{ watts.} \qquad (23)$$

Application of Theory to a Specific Installation

Assume a broadcast station operating at 600 kc ($\lambda = 500$ meters) with a power of one kw. The antenna is a monopole 125 meters, or one-quarter wavelength high, whose radiation resistance is 40 ohms; the ground conductivity is 0.002 mho per meter; and the base current, I_0, is $\sqrt{1{,}000/40}$, or 5 amperes. The cost of one watt year of rf power, K_2, is one dollar, and the amortization cost of one meter of buried No. 8 copper wire is 0.01 dollar. The cost ratio, α, is then 0.01. Evaluating the optimum-length parameter gives

$$F_2 = \frac{1.88 \times 10^6}{f\sigma} = \frac{1.88 \times 10^6}{6 \times 10^5 \times 2 \times 10^{-3}} = 1.566 \times 10^3.$$

From curve II of Fig. 2, the terminal spacing, d_t, is found to be 12.5 meters. From curve I of Fig. 2, the spacing parameter, F_1, is then found to be 1.86×10^4. The expression

$$F_1 = \frac{2.38 \times 10^8 \times a}{|J^2| \sqrt{f^3\sigma}}$$

can now be solved for $|J^2|$.

$$|J^2| = \frac{2.38 \times 10^8 \times 10^{-2}}{1.86 \times 10^4 \times \sqrt{0.002 \times (6 \times 10^5)^3}} = 6.15 \times 10^{-6};$$

$$|J| = 2.48 \times 10^{-3} \text{ amperes per meter,}$$

and

$$\frac{|J|\lambda}{I_0} = 0.248. \qquad (24)$$

Entering Fig. 1 with this value shows that r/λ is 0.64, from which r is determined to be 0.64×500 or 320 meters. This immediately gives the optimum number of radials as

$$N = \frac{2\pi \times 320}{12.5} = 161.$$

It is apparent from the analysis, in which the number of radials, N, is treated as a continuous dependent variable function of current density, that no specific number can be a precise optimum number over any great radial distance. Thus, the optimum number at several radii must be calculated, and, from inspection of the values, engineering judgment used to specify a number reasonably close to the optimum at all radial distances.

In order to illustrate this process, the optimum number at approximately two-thirds the terminal distance, or 200 meters, is calculated below.

At $r = 200$, $r/\lambda = 0.4$. From Fig. 1, this gives a value for $|J|\lambda/I_0$ of 0.396, from which $|J|$ is determined to be 3.96×10^{-3}, and $|J^2| = 1.57 \times 10^{-5}$.

Now by substituting values in (11), the spacing parameter, F_1, is evaluated.

$$F_1 = \frac{2.38 \times 10^8 \times 2 \times 10^{-2}}{1.57 \times 10^{-5} \times 2.08 \times 10^7} = 7.29 \times 10^3.$$

Curve I of Fig. 2 gives a value for optimum spacing, d, of 9.4 meters, from which

$$N = \frac{2\pi \times 200}{9.4} = 134 \text{ radials.}$$

The optimum number of radials may be computed for several other distances, and the results set down in tabular form:

r	50	100	150	200	250	300	320
N	70	102	119	134	145	157	161

From this table, 145 is selected as being a convenient and reasonable number of radials for the system. The 12.5-meter terminal spacing now defines the terminal length as 290 meters.

The power dissipated in the derived system as given by (23) becomes

$$P = \frac{aNr_t}{2},$$

where $a = 0.01$, $N = 145$, and $r_t = 290$ meters, from which P is found to be 210 watts.

Appendix I

Calculation of the Inductance of a Wire Grid[6]

Assume an infinite grid of perfectly conducting wires. Let E equal the rms longitudinal impressed electric field at the surface of a conductor.

Let d equal the spacing between wires (assumed $\ll\lambda$), a equal the radius of the wires (assumed $\ll d$), and λ equal the wavelength of radiation.

[6] W. Wessel, "Passage of electric waves through wire gratings" (in German), *Hochfreq. und Elektroak.*, vol. 54, pp. 62–69; August, 1939. An alternate derivation is given by G. G. MacFarlane, "Surface impedance of an infinite parallel-wire grid at oblique angles of incidence," *Jour. IEE* (London), vol. 93, pt. IIIA, pp. 1523–1527; October, 1947.

On the surface of each wire a field, E, is produced by the current in the wires.

Let J_z be the surface current density on a conductor. Then I equals the total current in any wire (equal to $2\pi a J_z$). Furthermore, let μ equal the permeability of the medium, ϵ equal the dielectric constant of the medium, ω equal 2π times the frequency, f, γ equal the propagation constant of the medium $=\sqrt{j\omega\mu(\sigma+j\omega\epsilon)}$, A equal the magnetic vector potential due to a total current, I, per wire, and j equal $\sqrt{-1}$.

The inductance per unit length of a wire is defined as

$$L = \frac{E}{\omega I} = \frac{\text{longitudinal field}}{2\pi(\text{frequency})(\text{current per wire})}.$$

In mks units, $E = j\omega\mu A$, and

$$A = \sum_{n=-\infty}^{n=\infty} \int_{-\infty}^{\infty}\int_0^{2\pi} J_z \frac{e^{-\gamma R}}{4\pi R}\, dz\,a\,d\phi,$$

the summation being carried over all the wires.

We are concerned only with the value of E on the surface of a conductor. The value of R for the wire considered becomes, as shown in Fig. 3,

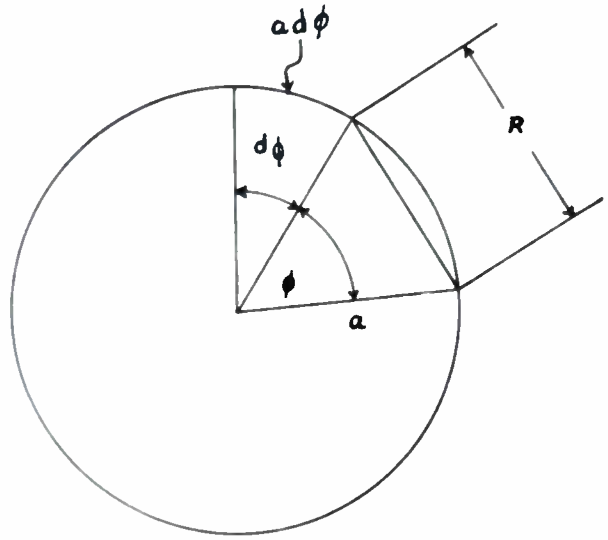

Fig. 3—Co-ordinate system for wire.

$$R_1 \approx \sqrt{z^2 + (2a \sin 1/2\phi)^2}$$

and

$$I = \int_0^{2\pi} J_z a\,d\phi.$$

For all other wires,

$$R_n = \sqrt{z^2 + n^2d^2},$$

and the current,

$$I = \int_0^{2\pi} J_z a\,d\phi,$$

is considered to flow along the axis of the conductor. For simplicity, let $\psi = \phi/2$; then we have for E,

$$E = \frac{j\omega\mu}{4\pi}\left[\int_{-\infty}^{\infty}\int_0^{\pi} \frac{I e^{-\gamma(z^2+4a^2\sin^2\psi)^{1/2}}}{\pi(z^2+4a^2\sin^2\psi)^{1/2}}\,d\psi\,dz + 2\sum_{n=1}^{n=\infty}\int_{-\infty}^{\infty}\frac{I e^{-\gamma(z^2+n^2d^2)^{1/2}}}{(z^2+n^2d^2)^{1/2}}\,dz\right]. \quad (25)$$

Consider the expression

$$F = \int_{-\infty}^{\infty} \frac{e^{-\gamma(z^2+\eta^2)^{1/2}}}{(z^2+\eta^2)^{1/2}}\,dz.$$

Let $\sqrt{(z^2/\eta^2)+1} = \cosh t$.

Then $z/\eta = \sinh t$, and $dz = \eta \cosh t\,dt$. For $z=\infty$, $t=\infty$, and the integral becomes

$$F = \int_{-\infty}^{\infty} e^{-\gamma\eta\cosh t}\,dt.$$

As shown by Watson,[7] this is equal to the Hankel function, $j\pi H_0^{(1)}(j\gamma\eta)$. Thus we have

$$E = \frac{j\omega\mu}{4\pi} I\left[\int_0^{\pi} jH_0^{(1)}(2j\gamma a \sin\psi)\,d\psi + j2\pi\sum_{n=1}^{\infty} H_0^{(1)}(j\gamma nd)\right]. \quad (26)$$

To evaluate the first term of (26), we recall that $H_0^{(1)}(z) = J_0(z)+jN_0(z)$, where J_0 is a Bessel function of zero order and N_0 is a Weber function of zero order. Since we are concerned with cases where $|j\gamma a| \ll 1$, we may write, $J_0(z)\approx 1$, and $N_0(z)\approx(2/\pi)(\log_e z + C - \log_e 2)$, where C is Euler's constant, 0.5772. Let $C = \log_e k$. Then

$$N_0(z) \approx \frac{2}{\pi}\log_e \frac{kz}{2}.$$

We have for the first term in (26),

$$\begin{aligned} j\int_0^{\pi} H_0^{(1)}(2j\gamma a\sin\psi)\,d\psi &= j\int_0^{\pi}\left[1 + \frac{2j}{\pi}\log_e (j\gamma ka \sin\psi\right]d\psi \\ &= j[\pi + 2j\log_e j\gamma ka - 2j\log_e 2] \\ &= j\pi - 2\log_e \frac{j\gamma ka}{2}. \end{aligned} \quad (27)$$

To evaluate the second term in (26), we again express $H_0^{(1)}$ in parts as $H_0^{(1)}(z) = J_0(z)+jN_0(z)$. It has been shown by von Ignatowsky[8,9] that for $d<\lambda$

$$\sum_{n=1}^{\infty} J_0(jn\gamma d) = \frac{-j}{\gamma d} - \frac{1}{2}$$

[7] G. N. Watson, "Bessel Functions," Cambridge University Press, Cambridge, England, 2nd. ed., p. 180, eq. (10); 1944.

[8] W. Von Ignatowsky, "Reihen mit Zylinderfunktionen nach dem Vielfachen des Argumentes," *Archiv der Math. und Phys.*, series 3, vol. 23, pp. 193–220; 1914.

[9] W. Von Ignatowsky, "Zur Theorie der Gitter," *Ann. der Phys.*, series 4, vol. 44, pp. 369–436; 1914.

and

$$\sum_{n=1}^{\infty} N_0(jn\gamma d) = \frac{-1}{\pi} \log_e \frac{j\gamma kd}{4\pi}$$

so that

$$2j\pi \sum_{n=1}^{\infty} H_0^{(1)}(jn\gamma d) = 2j\pi\left[\left(\frac{-j}{\gamma d} - \frac{1}{2}\right) - \frac{j}{\pi} \log_e \frac{j\gamma kd}{4\pi}\right]. \quad (28)$$

Substituting (27) and (28) in (26), we obtain

$$E = \frac{\omega\mu i}{2}\left(\frac{j}{\gamma d} + j\frac{2}{2\pi} \log_e \frac{d}{2\pi a}\right),$$

or

$$E = J\left(-\frac{1}{2}\eta_0 + jf\mu d \log_e \frac{d}{2\pi a}\right), \quad (29)$$

in which $J = I/d$ is the current in a strip of unit width, $\eta_0 = -j\omega\mu/\gamma$ is the characteristic impedance of the media, and $f = \omega/2\pi$ is the excitation frequency. The first term, $-(I_{\eta 0}/2d)$, leads to the energy absorbed by the space on either side of the grid when the excitation field, E, is applied to the grid alone. The admittance, Y, of a unit area of grid and surrounding media in parallel becomes the sum of their individual admittances. Thus

$$Y = \frac{1}{\eta} - j\frac{1}{f\mu d \log_e \dfrac{d}{2\pi a}} = Y_s - j\frac{1}{f\mu d \log_e \dfrac{d}{2\pi a}}.$$

in which Y_s is the soil admittance.

Appendix II

Glossary of Symbols[10]

- a Radius, in meters, of wire used in grid.
- C_a Annual amortization cost of the station, exclusive of ground system.
- C_b Annual operating cost of the station.
- C_1 Parameter defined by (6).
- d Spacing between buried conductors in meters.
- d_t Terminal spacing of conductors in meters.
- F_1 Optimum spacing parameter defined by (11).
- F_2 Length parameter defined by (17).
- f Frequency of incident rf energy in cps.
- H Field of magnetic intensity in ampere-turns per meter.
- h Height of monopole antenna, or equivalent height of top-loaded monopole, in meters.
- I_0 Current at base of antenna in amperes.
- J Ground-system current density in amperes per meter.
- j $\sqrt{-1}$.
- K_t Total annual cost for one meter of ground system.
- K_1 Annual cost of one meter of buried copper.
- K_2 Cost of amortization and operation in dollars per watt year.
- N Optimum number of radials.
- P Integrated power loss in the ground system.
- r Radial distance from base of antenna in meters.
- r_t Optimum length of radials in meters.
- U The real part of the admittance of the soil.
- V The imaginary part of the admittance of the soil.
- W Mean input power to the antenna in watts.
- w Power dissipated per square meter in the ground in watts.
- Y Admittance of one square meter of ground surface in mhos.
- Y_g Admittance of buried grid of wires in mhos.
- Y_s Characteristic admittance of the soil in mhos.
- Z Characteristic impedance of the soil in ohms.
- α Ratio of annual cost of one meter of buried conductor to cost of one watt year of rf power.
- β Propagation constant $= 2\pi/\lambda$.
- ϵ Dielectric constant of the soil in farads per meter.
- λ Wavelength of exciting rf energy in meters.
- μ Permeability of the soil in henrys per meter. Equal to $4\pi \times 10^{-7}$ for nonferromagnetic ground.
- ρ Distance from top of antenna in meters.
- σ Conductivity of the soil in mhos per meter.
- ω $2\pi f$.

[10] Meter-kilogram-second units are used.

Acknowledgment

Valuable suggestions for the development of this theory were made by S. A. Schelkunoff, L. J. Chu, J. L. Finch, J. Bolljahn, and several members of the NEL consulting staff. Waveguide measurements to check the analysis were performed by E. Devaney, J. Wade, and N. Balli of the NEL antenna-range staff.

The Absorption Gain and Back-Scattering Cross Section of the Cylindrical Antenna*

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Summary—The method of Hallén as modified by King and Middleton is applied to the case of the receiving dipole antenna whose axis is parallel to the electric vector of a far-zone field. First-order formulas are given for gain and broadside back-scattering cross section. Measured values of broadside gain and of broadside back-scattering cross section for both unloaded and matched-loaded dipoles are compared with the theory.

I. Introduction

Up TO the present time, the study of the dipole antenna has consisted primarily of the theoretical and experimental determination of the current distribution and the input impedance. The remainder of the antenna properties consisting of absorption or transmitting gain, scattering gain, effective length, and absorption and scattering cross sections have received less attention.

In this paper, formulas are developed, using the method of Hallén[1] as modified by King and Middleton,[2,3] for the power gain and back-scattering cross section of the cylindrical dipole antenna. The theoretical expressions obtained for the above quantities are compared with measured values of absorption gain and of back-scattering cross section for both shorted and matched-loaded dipole receiving antennas. Both the experimental results and the detailed calculations from the theory are limited to the values in the plane perpendicular to the axis of the dipole at its center. Significant differences exist between theory and experiment. However, the discrepancies may be logically ascribed to specific limitations in both the theoretical and experimental techniques.

II. The General Formulas

It can be shown by the use of the reciprocity theorem and the bilateral quality of mutual impedance between two antennas that the absorption gain of a receiving antenna is equal to its transmitting gain.[4,5] In other words the radiation and reception patterns of a given antenna are equal.

Using the notation of Schelkunoff,[6] the power gain of a linear antenna is given by

$$G = 4\pi\Phi/W, \tag{1}$$

where for an antenna of length $2h$ located along the z axis, with its center at the origin of the usual spherical co-ordinate system,

$$\Phi = \frac{\zeta}{8\lambda^2}|N_\theta|^2, \qquad \zeta = 120\pi, \tag{2}$$

$$N_\theta = -\sin\theta \int_{-h}^{h} I(z)e^{j\beta z\cos\theta}dz, \tag{3}$$

and W, the total power radiated, is

$$W = 2\pi\int_0^\pi \Phi\sin\theta d\theta, \tag{4}$$

or

$$W = \tfrac{1}{2}|I_0|^2R_a. \tag{5}$$

R_a is the radiation resistance, and is defined as the real part of the ratio of the terminal voltage, V_0, to the terminal current, I_0. Substitution of (5) in (1) gives

$$G_t = \frac{8\pi\Phi_v}{R_a|I_0|^2}. \tag{6}$$

If P is the power scattered back toward the source per unit solid angle, then the back-scattering cross section, σ, is defined as

$$\sigma = 4\pi P/S. \tag{7}$$

The incident power density is

$$S = |E|^2/2\zeta, \tag{8}$$

where $\zeta = \sqrt{\mu/\epsilon} = 120\pi$ ohms and E is the electric field. This allows (7) to be expressed in terms of the incident and scattered fields, E_i and E_s, and the range R;

$$\sigma = 4\pi R^2|E_s|^2/|E_i|^2. \tag{9}$$

Since the field radiated by a linear antenna at a great distance R is given by

$$E = \frac{-j\beta\zeta}{4\pi R}N_\theta e^{-j\beta R}, \tag{10}$$

where $\beta = 2\pi/\lambda$ and the time factor is omitted, the back-scattering cross section may be written as

* Decimal classification: R326.611. Original manuscript received by the Institute, July 20, 1951; revised manuscript received, February 28, 1952.

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[1] E. Hallén, *Nova Acta Upsala*, Royal Soc. Sciences, vol. 11, pp. 1-14; November, 1938.

[2] R. King and D. Middleton, "The cylindrical antenna: a comparison of theories," *Quart. Appl. Math.*, vol. 3, pp. 302-335; January 1946.

[3] D. Middleton and R. King, "The thin cylindrical antenna: a comparison of theories," *Jour. Appl. Phys.*, vol. 17, pp. 273-284; April, 1946.

[4] S. Silver, "Microwave Antenna Theory and Design," McGraw-Hill Book Co., Inc., New York, N. Y., chap. 2; 1949.

[5] J. Aharoni, "Antennae," Oxford University Press, London, p. 205; 1946.

[6] S. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., p. 335 ff; 1943.

$$\sigma = 8\pi\zeta\Phi_E/|E_i|^2. \quad (11)$$

The symbol Φ_E signifies that the current distribution along a receiving antenna is to be used in the formula for Φ. The symbol Φ_v in (6) denotes that the transmitting current distribution is involved. In general, these two distributions are not the same even for matched loads.

Insofar as the load is concerned, a receiving antenna acts like a generator, and the equivalent circuit of Fig. 1 applies. The equivalent generator impedance is the

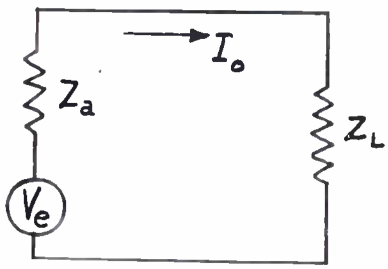

Fig. 1—Equivalent circuit of receiving antenna.

input impedance of the antenna, Z_a, when it is driven as a transmitter.[4,7] The voltage, V_e, of the generator is proportional to the incident field. The factor of proportionality, d, is called the *effective length* of the antenna.

$$V_e = dE_i. \quad (12)$$

The effective length of a dipole is

$$d = -N_\theta/I_0, \quad (13)$$

where the currents involved are those on the same antenna when driven as a transmitter.[8]

It has been shown[9] that the total terminal current may be written as

$$I_0 = I_{0E} - I_{0v}, \quad (14)$$

where I_{0E} is the current due to the external field at the center of an unloaded antenna, and I_{0v} is the terminal current in a driven antenna with the voltage V_L maintained at its terminals. The voltage V_L is given in terms of the external field by

$$V_L = I_{0E}Z_aZ_L/(Z_a + Z_L). \quad (15)$$

By the principle of superposition, (14) is true not only at the center of the antenna but also at any point z on the antenna.[10] Thus the total current distribution on a receiving antenna is

$$I_r(z) = I_E(z) - I_v(z). \quad (16)$$

It must be emphasized that the circuit of Fig. 1 is valid only for absorption of power by the load. *The power dissipated in the internal impedance of the equivalent generator has no direct relation to the energy reradiated by the antenna and in general cannot be used in discussing the scattering cross section.*[4] The current distribution of (16) must be used in calculating the scattering properties of a receiving antenna.

[7] C. Bouwkamp, "Hallén's theory for a straight perfectly conducting wire, used as a transmitting or receiving aerial," *Physica*, vol. 9, p. 609; July, 1942.

[8] E. Jordan, "Electromagnetic Waves and Radiating Systems," Prentice-Hall, Inc., New York, N. Y., p. 335; 1950.

[9] R. King, H. Mimno, and A. Wing, "Transmission Lines, Antennas, and Wave Guides," McGraw-Hill Book Co., Inc., New York, N. Y., p. 163; 1945.

[10] R. King and C. Harrison, "The receiving antenna," Proc. I.R.E., vol. 32, pp. 18–34; January, 1944.

III. The Current Distributions

The discussion in the previous section indicates that the determination of any of the properties of an antenna requires a knowledge of the current distribution along the antenna. In the classical theory of antennas, the assumption is made that the distribution of current is sinusoidal. The gains of dipole antennas usually given in standard texts are subject to this restriction. The sinusoidal distribution is correct for an indefinitely thin, perfectly conducting linear radiator.[11] In recent years various attempts have been made to solve the current-distribution problem for antennas of finite cross section. The method of Hallén forms the basis for the present calculations. Hallén's integral equation for the transmitting distribution of current in the absence of an external field is

$$\int_{-h}^{h} I(s)r^{-1}e^{-j\beta r}ds = \frac{-j4\pi}{\zeta}\left[C_1 \cos \beta z + \frac{1}{2} V_0 \sin \beta|z|\right], \quad (17)$$

where

$$r = \sqrt{(z-s)^2 + a^2}; \quad (18)$$

z is a point on the dipole axis, $-h \leqq z \leqq h$, s is a point on the surface of the dipole, $-h \leqq s \leqq h$, and a is the dipole radius. V_0 is the potential of a generator of negligible thickness at $z = 0$. The constant C_1 is determined from the condition that

$$I(\pm h) = 0. \quad (19)$$

The integral equation involving the current in an unloaded receiving antenna which is immersed in a field, E_i, parallel to the dipole axis, is

$$\int_{-h}^{h} I(s)r^{-1}e^{-j\beta r}ds = \frac{-j4\pi}{\zeta}[C_1 \cos \beta z + E_i/\beta]. \quad (20)$$

The general solution of this equation is the superposition of two solutions,

$$I(z) = I(-z) \quad (21a)$$

and

$$I(z) = -I(-z). \quad (21b)$$

The first or symmetrical solution gives a current which is unidirectional through the center so that a potential difference is established across the load. The anti-

[11] S. Schelkunoff, *op. cit.*, p. 142.

symmetrical solution (21b), on the other hand, calls for zero current at the center. If one is interested in the voltage across the load, only the symmetrical solution need be considered. But in the determination of scattered fields where the current at all points is of interest, both solutions must be included. However, for the special case in which the receiving antenna is placed parallel to the incident field, the antisymmetrical current vanishes, and the total current is given by the symmetrical solution alone.[10] Furthermore, in the measurements to be described, the antenna is placed over an image plane. By the principle of images, only the symmetrical current distribution can exist.

The first-order solution of (17), given by King and Middleton, is[2]

$$I_v(z) = j2\pi V_0 f_1(z)(\zeta\psi_t H_2)^{-1}, \tag{22}$$

where

$$f_v(z) = f_v'(z) + jf_v''(z) = b_1 \cos\beta z - b_2 \sin\beta|z| - C(z)\sin\beta h + S(z)\cos\beta h, \tag{23}$$

$$b_1 = [2\psi_t + E(h)]\sin\beta h - S(h), \tag{24a}$$

$$b_2 = [2\psi_t + E(h)]\cos\beta h - C(h), \tag{24b}$$

$$H_2 = H_2' + jH_2'' = [\psi_t + E(h)]\cos\beta h - C(h), \tag{25}$$

$$C(z) = \int_{-h}^{h} r^{-1}\cos\beta s\, e^{-j\beta r}ds, \tag{26}$$

$$S(z) = \int_{-h}^{h} r^{-1}\sin\beta|s|\, e^{-j\beta r}ds, \tag{27}$$

$$E(z) = \int_{-h}^{h} r^{-1}e^{-j\beta r}ds. \tag{28}$$

The functions of (26), (27), and (28) are solved in footnote reference 2. These functions depend upon the thickness of the dipole. It is customary to designate this thickness parameter by Ω, where

$$\Omega = 2\log_e\frac{2h}{a}\cdot \tag{29}$$

The expansion parameter, ψ_t, of King and Middleton is defined as

$$\psi_t = \begin{cases} |C(0)\sin\beta h - S(0)\cos\beta h|\,(\sin\beta h)^{-1}, & \beta h \leqq \pi/2 \\ \left|C\left(h-\dfrac{\lambda}{4}\right)\sin\beta h - S\left(h-\dfrac{\lambda}{4}\right)\cos\beta h\right|, & \beta h \geqq \pi/2. \end{cases} \tag{30}$$

The first-order solution for the current is considered here because higher-order solutions involve functions which cannot be evaluated by ordinary methods.[7]

Similarly, the first-order solution of (20) for the current on an unloaded receiving antenna is

$$I_E(z) = \frac{j4\pi E_i f_E(z)}{\beta\zeta\psi_r H_1}, \tag{31}$$

where

$$f_E(z) = f_E'(z) + jf_E''(z) = 2\psi_r\cos\beta z + E(z)\cos\beta h - C(z) + C(h) - \cos\beta h[2\psi_r + E(h)], \tag{32}$$

$$H_1 = H_1' + jH_1'' = \psi_r\cos\beta h + E(h)\cos\beta h - C(h), \tag{33}$$

and where $I_E(z)$ is used to denote the current in an unloaded receiving antenna.

By arguments analogous to those used by King and Middleton, the expansion parameter ψ_r is found to be

$$\psi_r = \begin{cases} |C(0) - E(0)\cos\beta h|\,(1-\cos\beta h)^{-1}, & \beta \leqq \pi \\ \frac{1}{2}|C(0) + E(0)|, & \beta \geqq \pi. \end{cases} \tag{34}$$

For $\beta h \ll 1$, $\psi_r = \Omega - 1$.

IV. Calculation of Broadside Absorption or Transmitting Gain

The computation of gain first requires the calculation of Φ_v. From (2) and (22) the value of Φ_v for the broadside case ($\theta = 90$ degrees) is

$$\Phi_v = \frac{\pi^2|V_0|^2|g_v(h)|^2}{2\lambda^2\zeta\psi_t^2|H_2|^2}, \tag{35}$$

where

$$g_v(h) = g_v'(h) + jg_v''(h) = \int_{-h}^{h} f_v(z)dz. \tag{36}$$

Substitution of (35) into (6) gives the broadside gain as

$$G_t = \frac{\pi|Z_a|^2|\beta g_v(h)|^2}{\zeta\psi_t^2 R_a|H_2|^2}\cdot \tag{37}$$

The function $\beta g_v(h)$ is found to be[12]

$$\begin{aligned}\beta g_v(h) &= 4\psi_t - 2\Omega + 2\,\mathrm{Cin}\,2\beta h + 2j\,\mathrm{Si}\,2\beta h \\ &\quad + \cos\beta h(2\Omega - 4\psi_t + 2\,\mathrm{Cin}\,2\beta h - \mathrm{Cin}\,4\beta h \\ &\quad - 4\,\mathrm{Cin}\,\beta h + 2j\,\mathrm{Si}\,2\beta h - j\,\mathrm{Si}\,4\beta h \\ &\quad - j4\,\mathrm{Si}\,\beta h) + \sin\beta h(\mathrm{Si}\,4\beta h - j\,\mathrm{Cin}\,4\beta h).\end{aligned} \tag{38}$$

For the first-order theory,

$$|Z_a|^2/R_a = (2\pi)^{-1}|H_2|^2\zeta\psi_t[H_2''f_v'(0) - H_2'f_v''(0)]^{-1}. \tag{39}$$

Substituting this in (36) gives the first-order gain as

$$G_t = \tfrac{1}{2}\psi_t^{-1}|\beta g_v(h)|^2[H_2''f_v'(0) - H_2'f_v''(0)]^{-1}. \tag{40}$$

Equation (40) is plotted in Fig. 2.

If the antenna is very thin, $\Omega \rightarrow \infty$, then (37) reduces to

$$G_t = 4(1-\cos\beta h)^2[\sin 2\beta h(\mathrm{Si}\,4\beta h - 2\,\mathrm{Si}\,2\beta h) - \cos 2\beta h\,\mathrm{Cin}\,4\beta h + 4\cos^2\beta h\,\mathrm{Cin}\,2\beta h]^{-1}. \tag{41}$$

These values of G_t are the same as those obtained by assuming a sinusoidal current distribution, and are the

[12] $\mathrm{Cin}\,(x) = \int_0^x \frac{1-\cos t}{t}dt;\ Si(x) = \int_0^x \frac{\sin t}{t}dt.$

values generally given in standard texts for center-fed dipoles.[13] They are, of course, valid only for the indefinitely thin dipole.

Note that the curves of Fig. 2 do not approach $G_t = 1.5$ as $\beta h \to 0$. The relation for G_t in the limit of very short βh becomes independent of βh, but does remain a function of Ω. This is a questionable result and may be ascribed to the particular choice of ψ_t.

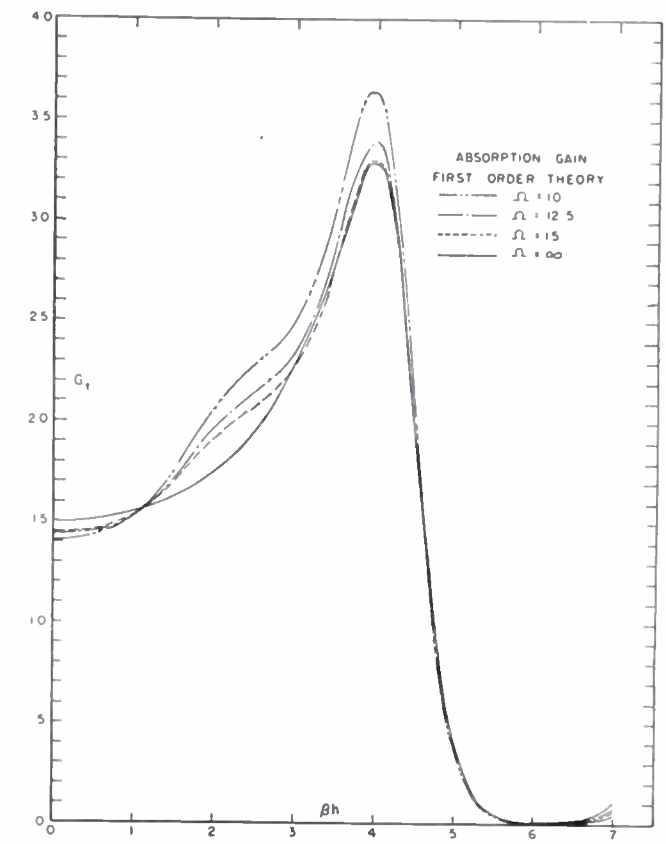

Fig. 2—Broadside absorption gain. First-order theory.

V. Calculation of the Broadside Back-Scattering Cross Section

The back-scattering cross section was given by (11). The evaluation of Φ_E in this relation requires the current distribution given by (16). From (31) and (22), the current distribution is given by

$$I_r(z) = \frac{j2\pi}{\zeta}\left[\frac{2E_i}{\beta\psi_r H_1} f_E(z) - \frac{V_L}{\psi_t H_2} f_v(z)\right]. \quad (42)$$

Using (15), this becomes

$$I_r(z) = \frac{j4\pi E_i}{\beta\zeta\psi_r H_1}\left[f_E(z) - \frac{j2\pi}{\zeta\psi_t H_2} \cdot \frac{Z_a Z_L}{Z_a + Z_L} f_E(0) f_v(z)\right]. \quad (43)$$

[13] R. King, H. Mimno, and A. Wing, *op. cit.*, pp. 186–187.

Thus, from (2) with (3), for the broadside case,

$$\Phi_E = |E_i|^2 (2\zeta\psi_r^2 |H_1|^2)^{-1} |g_E(h) - BZ_e f_E(0) g_v(h)|^2, \quad (44)$$

where

$$g_E(h) = \int_{-h}^{h} f_E(z)dz, \quad (45)$$

$$B = j2\pi(\zeta\psi_t H_2)^{-1} = 2\pi(H_2'' + jH_2')(\zeta\psi_t |H_2|^2)^{-1} \quad (46)$$

$$Z_e = R_e + jX_e = Z_L Z_a/(Z_L + Z_a). \quad (47)$$

From (11) and (44),

$$\sigma/\lambda^2 = |\beta g_E(h) - BZ_e f_E(0)\beta g_v(h)|^2 (\pi\psi_r^2 |H_1|^2)^{-1}. \quad (48)$$

It is evident from (48) that the back-scattering cross section depends upon the antenna-load impedance. For the unloaded receiving antenna, $Z_e = Z_L = 0$, and for this case,

$$\sigma/\lambda^2 = |\beta g_E(h)|^2 \pi^{-1} \psi_r^{-2} |H_1|^{-2}, \; Z_L = 0. \quad (49)$$

For the match-loaded case ($Z_L = Z_a^*$),

$$Z_e = R_e = |Z_a|^2/2R_a. \quad (50)$$

Using the result of (39) for Z_e in (48), the first-order formula becomes

$$\frac{\sigma}{\lambda^2} = \left|\beta g_E(h) - \frac{(H_2'' + jH_2')f_E(0)\beta g_v(h)}{2[H_2'' f_v'(0) - H_2' f_v''(0)]}\right|^2 \cdot [\pi\psi_r^2 |H_1|^2]^{-1}, \; Z_L = Z_a^*. \quad (51)$$

The function $\beta g_E(h)$ is found to be

$$\begin{aligned}\beta g_E(h) = {} & \sin\beta h(4\psi_r - 2\Omega + \beta h \operatorname{Si} 4\beta h - 4\log 2 \\ & + 2\operatorname{Cin} 2\beta h + \operatorname{Cin} 4\beta h - j\beta h \operatorname{Cin} 4\beta h \\ & + j\operatorname{Si} 4\beta h + j2\operatorname{Si} 2\beta h) + \cos\beta h[2\beta h(\Omega - 2\psi_r \\ & + 2\log 2) - \beta h \operatorname{Cin} 4\beta h - 2\beta h \operatorname{Cin} 2\beta h \\ & - 2\sin 2\beta h + \operatorname{Si} 4\beta h - j\beta \operatorname{Si} 4\beta h \\ & - j2\beta h \operatorname{Si} 2\beta h - j2\cos 2\beta h + j2 \\ & - j\operatorname{Cin} 4\beta h]. \end{aligned} \quad (52)$$

The broadside back-scattering cross section for $Z_L = 0$ and $Z_L = Z_a^*$ is plotted in Figs. 3 and 4 for several values of Ω.

The curves for σ/λ^2 with matched load do not approach the value $9/16\pi = 0.179$ as $\beta h \to 0$. The same difficulty with the expansion parameters occurs here as in the case of absorption gain.

VI. Experimental Measurements

Measurements were made of relative absorption gain as a function of βh by mounting a dipole on an image plane and measuring the relative power absorbed in a matched load. The frequency used was 3,000 mc and a block diagram of the experimental equipment is shown

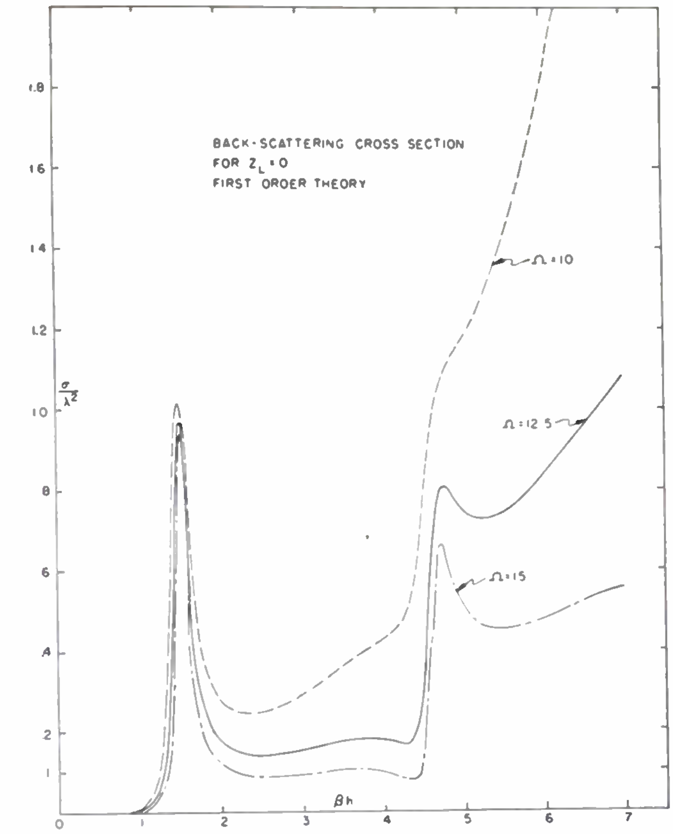

Fig. 3—Broadside back-scattering cross section of unloaded dipole. First-order theory.

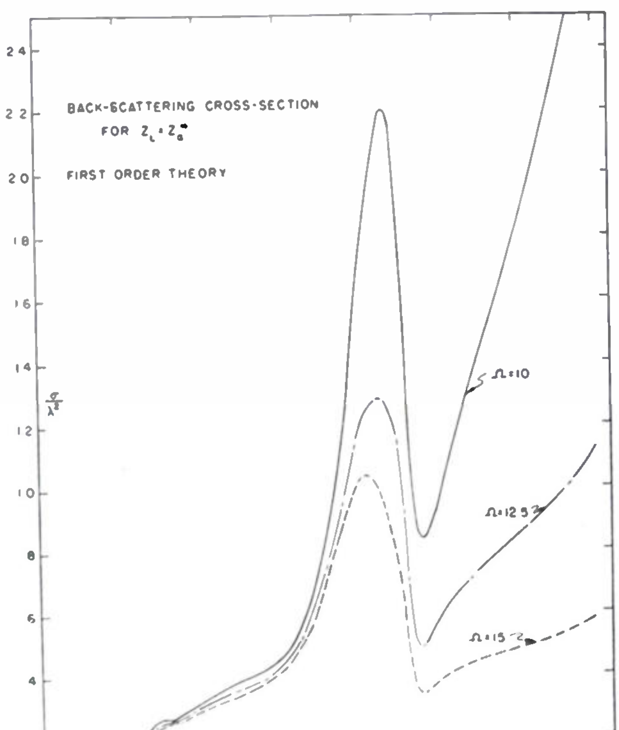

Fig. 4—Broadside back-scattering cross section of dipole with matched load. First-order theory.

in Fig. 5. The image plane was eight by twelve feet in size and was made from 3/16-inch aluminum sheet.

The location of the dipole on the image plane was determined after making measurements of the standing

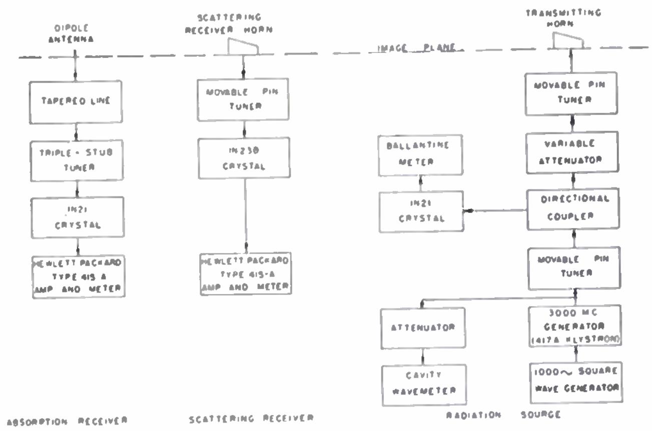

Fig. 5—Block diagram of measuring system.

waves on the plane. These measurements were made with a surface probe. Fig. 6 shows the result of one run down the center line of the plane in the region from 130 to 230 cm from the radiating horn aperture. The region around 180 cm was further explored with two different probe lengths and found to be reasonably uniform in amplitude and phase. The dipole was accordingly mounted at 183.5 cm from the radiating aperture D ($D^2/\lambda = 102$ cm).

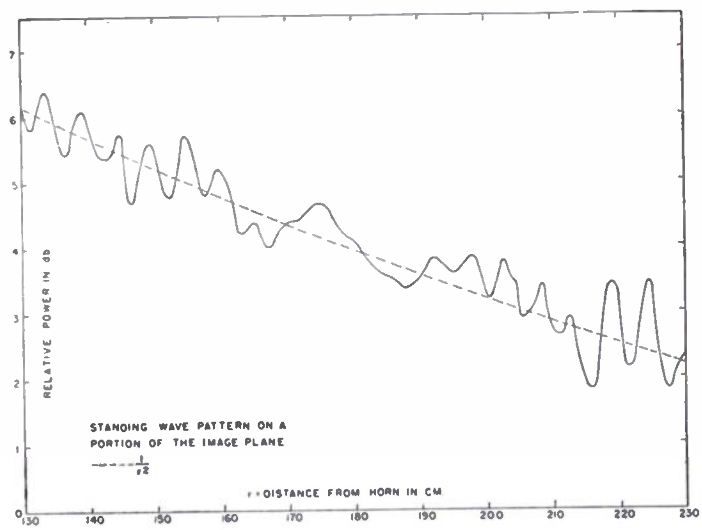

Fig. 6—Standing-wave pattern along center line of the image plane.

The scattering receiver horn was adjusted in location so that the power received in the absence of the scattering dipole was minimized, and yet the power received from a very short scatterer was easily readable. In the absence of the dipole to be measured the power in the scattering receiver was more than 10 db below that received from the shortest dipole length measured. It was also confirmed that the presence or absence of the scattering horn did not perceptibly alter the power absorbed by the dipole.

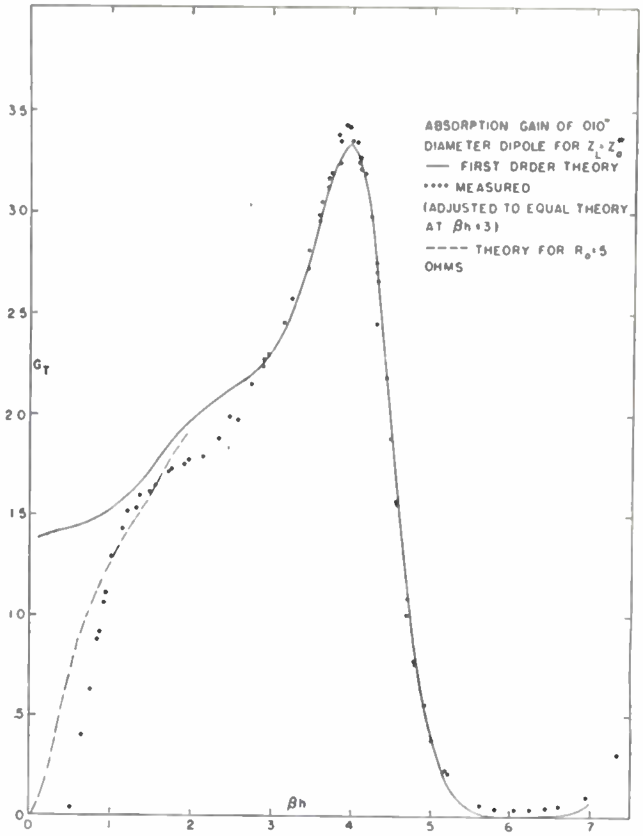

Fig. 7—Broadside absorption gain of 0.010-inch diameter dipole.

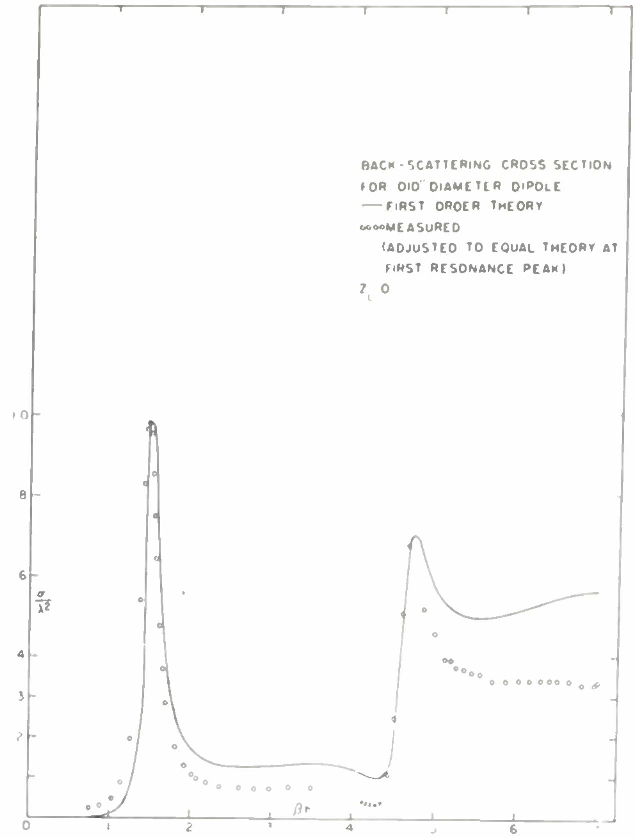

Fig. 9—Broadside back-scattering cross section of unloaded 0.010-inch diameter dipole.

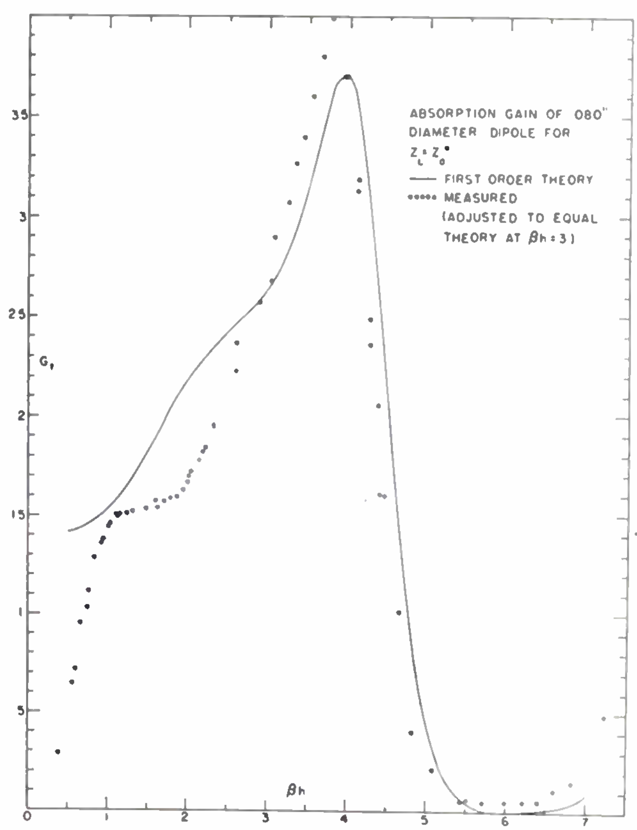

Fig. 8—Broadside absorption gain of 0.080-inch diameter dipole.

The antenna was matched with a triple-stub tuner capable of developing a vswr as high as 100 at 3,000 mc. As the antenna becomes short, its impedance becomes more reactive with a decreasing resistive component. For short antenna lengths, a match cannot be obtained because of the high vswr required of the tuner. The combination of mismatch and ohmic losses in the system causes the experimental data for G_t and σ/λ^2 to tend rapidly toward zero as length becomes less than about $\beta h=1$.

Two different dipole diameters were measured and theoretical curves were computed for each size. These curves were required since Ω is not constant as βh is varied at constant diameter. The experimental results are shown in Figs. 7 and 8. In each case the experimental data were normalized to equal the theory at $\beta h=3$. The theoretical gain for short lengths due to a loss resistance of 5 ohms is shown as the dashed line in Fig. 7. From both sets of experimental results there appears to be rather conclusive evidence that the "bulge" in the theoretical gain curves below $\beta h=3$ does not exist in actuality. One would, of course, expect closer agreement between theory and experiment for the smaller diameter wire, and this is evident from the two figures.

The back-scattering cross section as measured for the two diameters for both zero (shorted) and matched loads

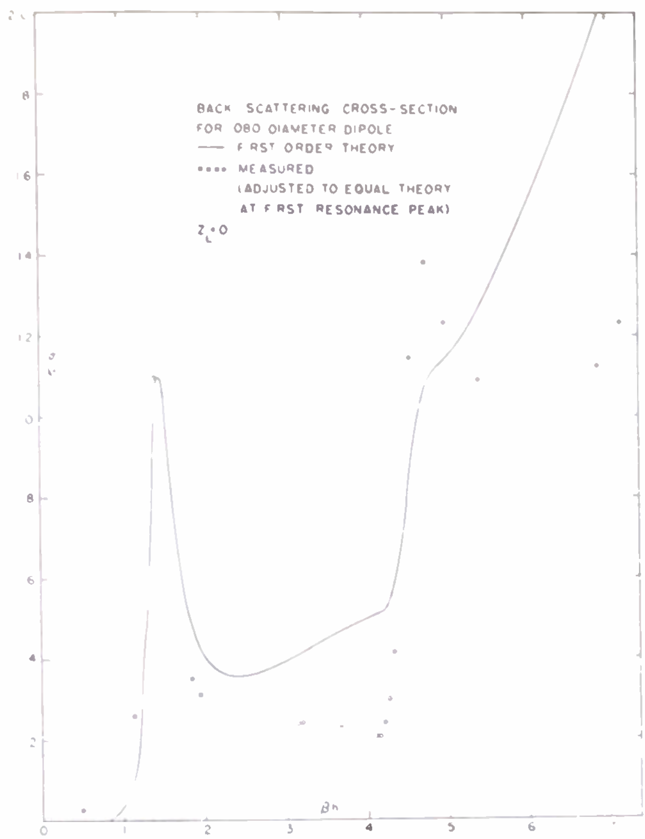

Fig. 10—Broadside back-scattering cross section of unloaded 0.080-inch diameter dipole.

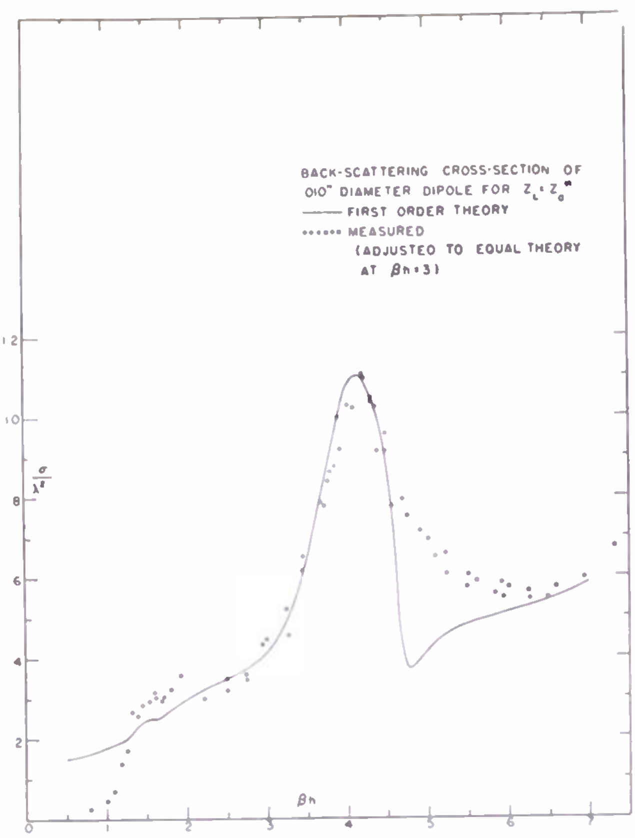

Fig. 11—Broadside back-scattering cross section of 0.010-inch diameter dipole with matched load.

is compared with theory in Figs. 9 to 12. The data indicate that this is an antenna parameter which is extremely sensitive to prediction by present theory.

VII. Conclusion

Various difficulties have become apparent in the King-Middleton type of first-order solution of the Hallén integral equation. These are particularly apparent in the scattering problem. In order to illustrate the sensitivity of the back-scattering cross section to dipole radius, Fig. 13 compares the first-order theoretical results using the methods of previous authors. In the figure are presented the results using ψ_r; the ψ_t of King and Middleton; the expansion parameter of Gray;[14] the simple expansion parameter, Ω, of Hallén; and also the results of Van Vleck, et al.[15] These are plotted for $\Omega = 10$, and may be compared at the longer lengths with the experimental results of Fig. 10 since, for the 0.080-inch diameter dipole, the value of Ω varies from 10 at $\beta h = 3\pi/2$ to 10.78 at $\beta h = 7$. It is interesting to note that the results of Van Vleck compare most favorably

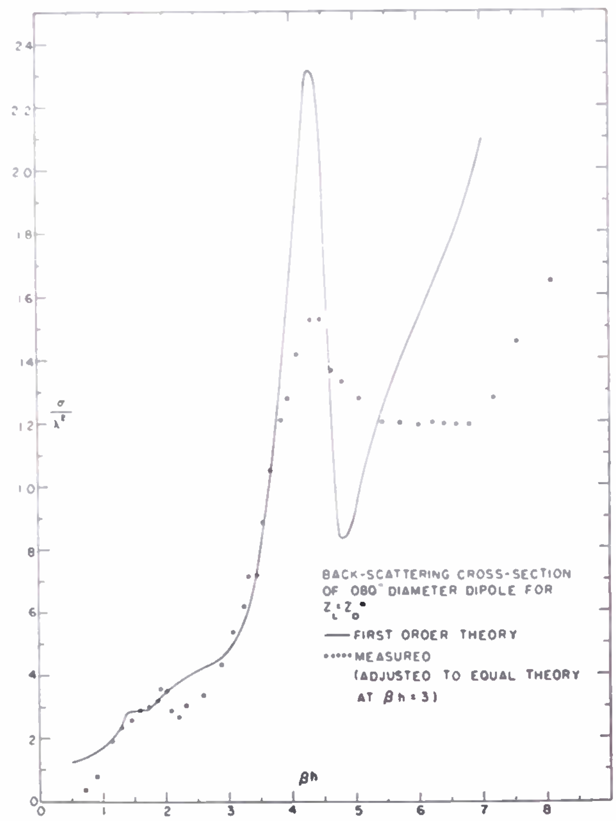

Fig. 12—Broadside back-scattering cross section of 0.080-inch diameter dipole with matched load.

[14] M. Gray, "A modification of Hallén's solution of the antenna problem," *Jour. Appl. Phys.*, vol. 15, pp. 1–65; January, 1944.

[15] J. Van Vleck, F. Bloch, and M. Hamermesh, "Theory of radar reflection from wires or thin metallic strips," *Jour. Appl. Phys.*, vol. 18, p. 274; March, 1947.

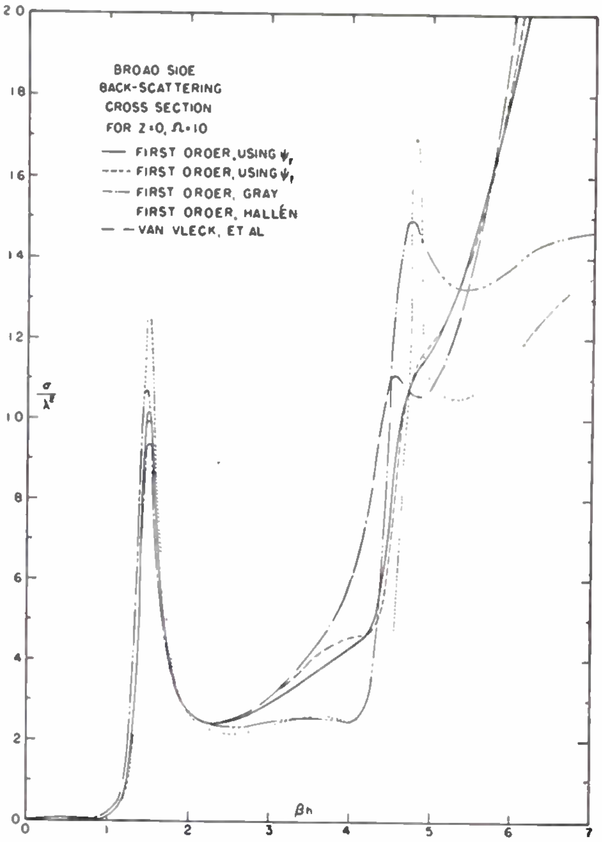

Fig. 13—Broadside back-scattering cross section of unloaded dipole. A comparison of several solutions for $\Omega=10$.

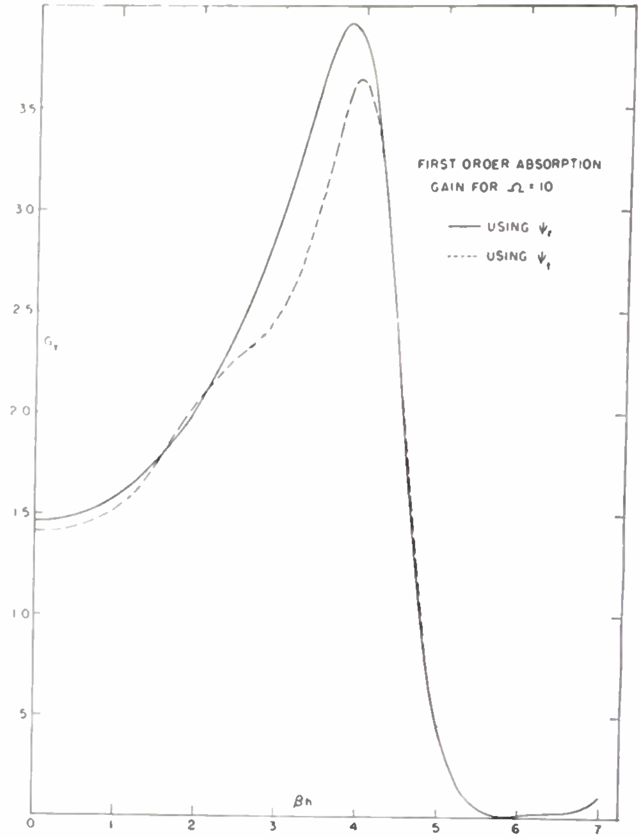

Fig. 14—Comparison of first-order solutions for absorption gain using ψ_r and ψ_t with $\Omega=10$.

with these experimental values. Fig. 14 compares the first-order gain using the ψ_t of King and Middleton with that obtained when ψ_r is used as the expansion parameter. In the latter case, the bulge near $\beta h=2$ is not obtained.

Thus it is seen that the differences between theory and experiment are drastically influenced by the choice of the expansion parameter in a first-order theory. A discussion of some of the difficulties is presented elsewhere.[16]

[16] S. H. Dike, "Difficulties with present solutions of the Hallén integral equation," *Quart. Appl. Math.* (to be published).

Correction

H. C. Hurley, S. R. Anderson, and H. F. Keary, co-authors of the paper, "The Civil Aeronautics Administration VHF Omnirange," which appeared on pages 1506–1520 of the December, 1951 issue of the Proceedings of the I.R.E., have requested that the editors publish the following corrections:

Interchange $E_1{}^1$ and $E_2{}^1$ in Fig. 13 (a), page 1513.

Interchange E_{VAR} and E_{REF} in the first and third lines under the section entitled "Phase Comparison Methods," page 1513.

Single- and Multi-Iris Resonant Structures*

IRVING REINGOLD†, JOHN L. CARTER†, AND KENTON GAROFF†, ASSOCIATE, IRE

Summary—An experimental investigation has been made of the characteristics of single- and multi-iris rectangular structures of various dimensions in a thin diaphragm which is placed transversely across a rectangular, standard *S*-band waveguide. The irises (or windows) are parallel to the "a" dimension of the waveguide, and varied in position along the "b" dimension of the waveguide. Measured values of resonant frequency, bandpass and *Q*, and the correlation between single- and double-iris structures are presented as a function of the position of the iris in the diaphragm. Data are also presented for other parameters of interest, such as the impedance. It is shown that the *Q* of a single iris of fixed dimensions in the center of a thin diaphragm can be appreciably increased by moving it closer to the wide waveguide wall. In addition, a low-*Q* multi-iris structure can be realized in which the *Q* of each iris is comparatively high.

I. Introduction

It is a well-known fact that various sized and shaped openings in a transverse diaphragm in a waveguide may be chosen so that they are resonant at a particular frequency, with practically all of the energy incident on the diaphragm being transmitted. However, comparatively little information is available on the effect of varying the position of a rectangular window in a thin diaphragm along the "b" dimension of a waveguide. As far as the authors of this paper can ascertain, no information is available concerning the characteristics of multi-iris structures or the relationship between single- and multi-iris structures in a thin diaphragm. It is the objective of this paper to present experimental data concerning the characteristics of single- and multi-iris structures in a thin diaphragm, their behavior as a function of position in the diaphragm, and the correlation between the single- and double-iris structures.

II. Test Procedure and Measurements

Data are presented for diaphragms with rectangular windows of three separate dimensions: 2.111 inches ×0.075 inch×0.050 inch, 2.113 inches×0.100 inch ×0.050 inch, and 2.119 inches×0.150 inch×0.050 inch. Since it was desired to have a common starting point for all of the data presented, each case of a single iris in the center of the diaphragm was designed for the same resonant frequency.

The diaphragm was placed in the waveguide perpendicular to the direction of propagation. Measurements for the single-window cases were made for windows which varied in position along the "b" dimension of the waveguide. Measurements for the double-window cases were made for windows which were equidistant from the center of the diaphragm and varied in position along the "b" dimension of the waveguide. The larger dimension of the rectangular window is parallel to the "a" dimension of the waveguide in all cases. Spacing from the center line of the diaphragm, which is also the center line of the waveguide, to the center line of the window is the variable against which the data are presented. (See Fig. 1.)

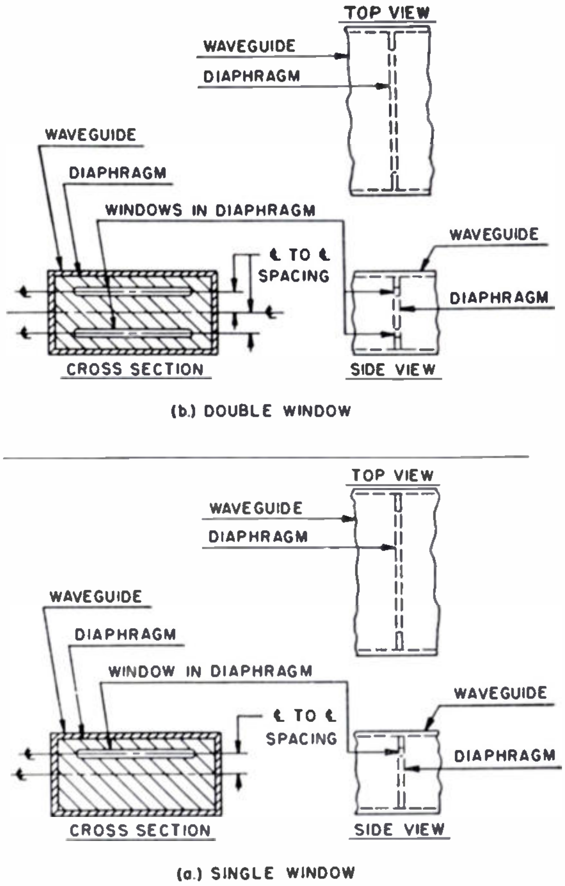

Fig. 1—(a) Single-iris diaphragm. (b) Double-iris diaphragm.

Measurements were made in a conventional manner, using a klystron signal source, calibrated attenuators, a waveguide slotted line, and a matched load.

The band pass, Δf, is arbitrarily chosen as the difference in frequency between the voltage standing-wave ratio values of 1.4 on either side of the resonant frequency, f_0. Q is calculated from the relationship,

$$Q = \left(\frac{r-1}{2\sqrt{r}}\right)\left(\frac{f_0}{\Delta f}\right), \tag{1}$$

* Decimal classification: R118.1. Original manuscript received by the Institute, July 17, 1951; revised manuscript received, February 15, 1952.

† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

where r is the voltage standing-wave ratio. Equation (1) is derived from the relationships for the parameters of a resonant element such as an iris in a thin diaphragm, as given in volume 14 of the Radiation Laboratory Series.[1]

The insertion loss of the various structures measured was spot checked throughout the tests. Impedance data as a function of frequency are presented for one single-iris and one double-iris structure. The characteristics of one triple-iris structure were explored, and its resonant frequency, band pass, and Q determined. The behavior of two double-iris structures separated by $\lambda g/4$ in the waveguide, and of three double-iris structures, each separated by $\lambda g/4$, was also determined. Proper precautions were taken to minimize the inherent sources of error in the series of measurements performed.

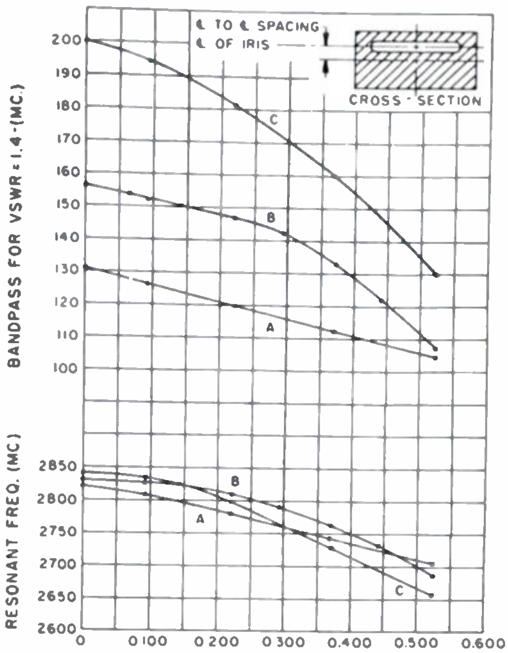

Fig. 2—Band pass and resonant frequency as a function of position of a single iris in a thin diaphragm. Curve A, iris dimensions = 2.111 inches × 0.075 inch × 0.050 inch. Curve B, iris dimensions = 2.113 inches × 0.100 inch × 0.050 inch. Curve C, iris dimensions = 2.119 inches × 0.150 inch × 0.050 inch.

III. Results

The curves shown in Fig. 2 indicate the relationships between the resonant frequency and band pass as a function of the position in the diaphragm of a single iris. As expected, the band pass increases as the iris width is increased. It will be noted that in each case the resonant frequency decreases as the iris spacing from the center of the guide increases. Note, also, that for any one particular iris the band pass decreases with increasing iris spacing from the center of the guide. This increase, which is linear for the narrowest iris, departs from linearity as the iris width is increased, the point of departure moving closer to the center of the waveguide with increasing iris width. This phenomenon is also observed in Fig. 3, where Q is shown as a function of the position of the iris in the diaphragm.

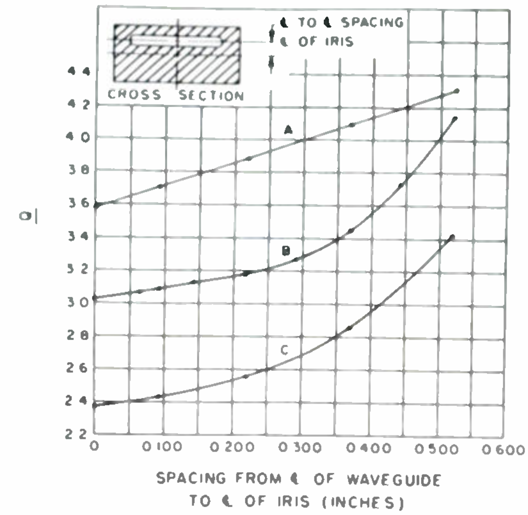

Fig. 3—Q as a function of position of a single iris in a thin diaphragm. Curve A, iris dimensions = 2.111 inches × 0.075 inch × 0.050 inch. Curve B, iris dimensions = 2.113 inches × 0.100 inch × 0.050 inch. Curve C, iris dimensions = 2.119 inches × 0.150 inch × 0.050 inch

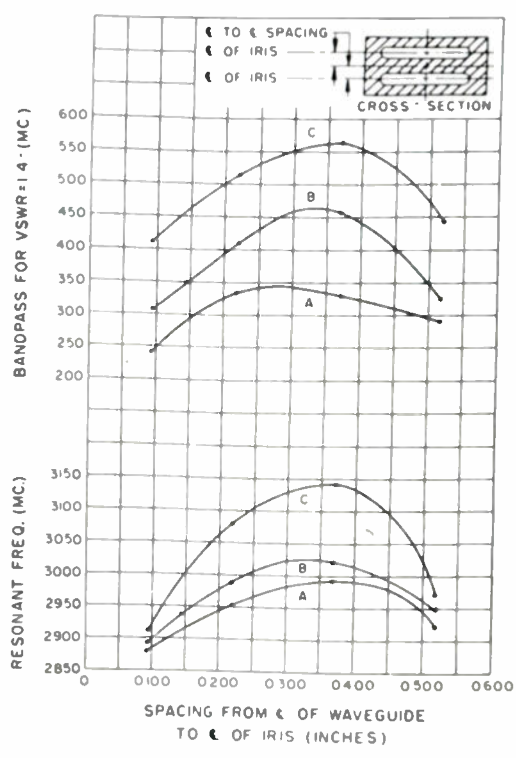

Fig. 4—Band pass and resonant frequency as a function of position of a double iris in a thin diaphragm. Curve A, iris dimensions = 2.111 inches × 0.075 inch × 0.050 inch. Curve B, iris dimensions = 2.113 inches × 0.100 inch × 0.050 inch. Curve C, iris dimensions = 2.119 inches × 0.150 inch × 0.050 inch.

The curves shown in Fig. 4 indicate the relationship between the resonant frequency and band pass as a function of position in the diaphragm of the double

[1] Smullin and Montgomery, "Microwave Duplexers," Rad. Lab Series, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 14, p. 73; 1948.

iris. In each case, the resonant frequency increases to a maximum value at some intermediate value of the iris spacing from the center of the guide and then drops off as this spacing is increased further. The band pass also rises to a peak value for an intermediate value of the iris spacing from the center of the guide and then decreases with a further increase in the spacing. Note, also, that the resonant frequency is greater than for the corresponding single-iris cases. Curves for Q as a function of the position in the diaphragm of the double iris are shown in Fig. 5. As a consequence of the behavior of the resonant frequency and band pass, Q reaches a minimum value at an intermediate value of the spacing from the center of the waveguide to the center of the iris, the minimum moving closer to the center of the guide as the iris width is increased. Note, also, that the total deviation in Q decreases as the iris width is increased.

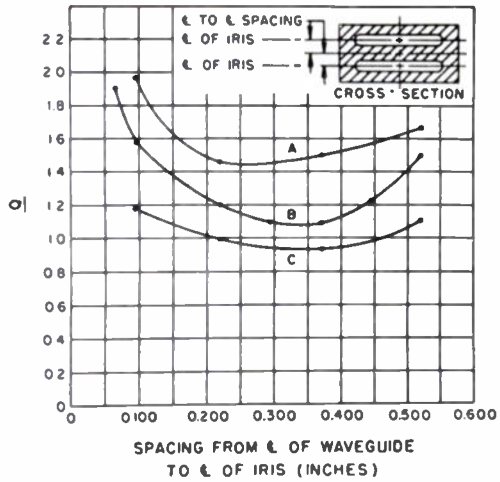

Fig. 5—Q as a function of position of a double iris in a thin diaphragm. Curve A, iris dimensions = 2.111 inches × 0.075 inch × 0.050 inch. Curve B, iris dimensions = 2.113 inches × 0.100 inch × 0.050 inch. Curve C, iris dimensions = 2.119 inches × 0.150 inch × 0.050 inch.

Fig. 6 indicates the correlation between Q for the single-iris cases and Q for the corresponding double-iris cases as a function of the iris position in the diaphragm. The Q-ratio curves for each iris dimension vary slightly for small spacings between the center of the guide and the center of the iris, but these curves diverge as this spacing is increased.

A typical plot of frequency versus vswr for one single-iris and the corresponding double-iris diaphragm is shown in Fig. 7. These curves are characteristic of the frequency versus vswr relationship for a shunt resonant element, that is, a resultant curve where the vswr rises smoothly on either side of resonance as the frequency deviation from resonance is increased.

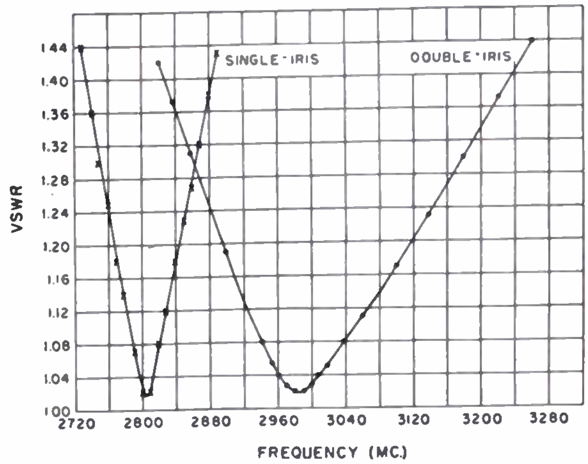

Fig. 7—Typical frequency versus vswr curves for single- and double-iris structures. Iris dimensions = 2.113 inches × 0.100 inch × 0.050 inch. Spacing from ℄ of waveguide to ℄ of iris = 0.220 inch.

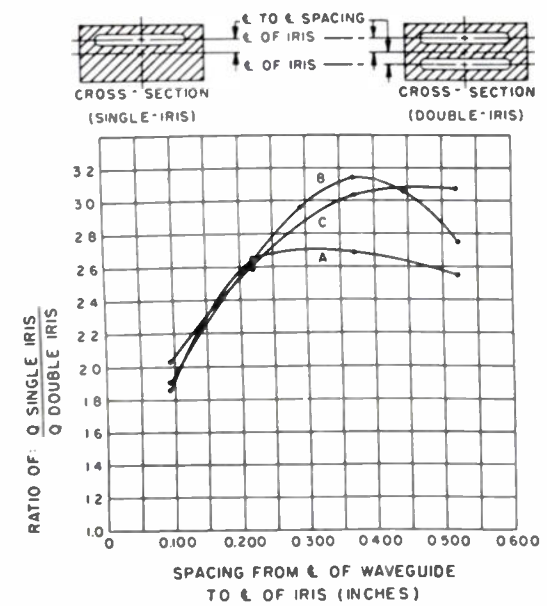

Fig. 6—Ratio of $\frac{Q \text{ single-iris}}{Q \text{ double-iris}}$ as a function of position of the irises in a thin diaphragm. Curve A, iris dimensions = 2.111 inches × 0.075 inch × 0.050 inch. Curve B, iris dimensions = 2.113 inches = 0.100 inch × 0.050 inch. Curve C, iris dimensions = 2.119 inches × 0.150 inch × 0.050 inch.

Impedance data for the single-iris and double-iris diaphragms discussed above are shown on an expanded-scale Smith chart in Fig. 8. Here again the resultant curves are characteristic of that of a shunt resonant element; that is, a reactive component of opposite sign exists on either side of resonance and diminishes steadily as the resonant frequency is approached until, at resonance, the impedance is purely resistive.

The band pass for two double-iris structures and three double-iris structures in which the spacing between elements is $\lambda g/4$, is characteristic of the behavior of multi-element structures separated by approximately a quarter guide wavelength; that is, the band pass increases as the number of elements increases.[2] The curve for the two-element case has one minimum, and the curve for the three-element case has two minima.

[2] *Ibid.*, p. 73.

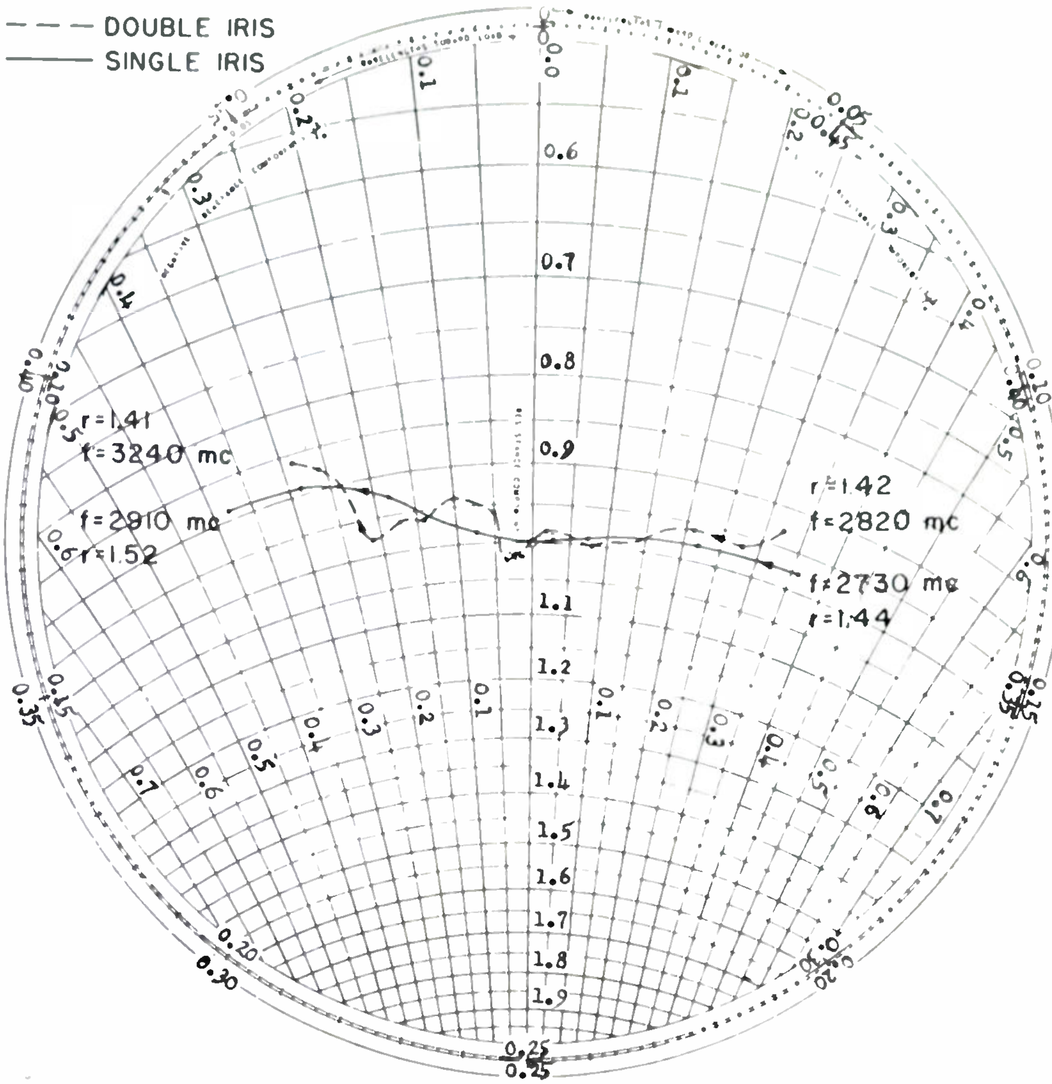

Fig. 8—Expanded Smith chart showing impedance as a function of frequency for the resonant structures of Fig. 7.

The insertion loss of the various resonant structures was negligible in all cases, being in the order of 0.1–0.2 db.

An examination of the data for the single- and double-iris diaphragms would lead one to predict that triple-iris structures would have larger band-pass characteristics and consequently a lower Q than single- or double-iris structures. The characteristics of the one triple-iris diaphragm investigated, which had one iris in the center of the diaphragm and the other two spaced symmetrically above and below the center of the diaphragm, are as follows: resonant frequency = 2,920 mc; band pass = 1,050 mc; and $Q = 0.46$. The values obtained, therefore, verify the predicted increase in band pass and decrease in Q.

IV. Discussion of the Results

An important factor involved in the behavior of the resonant iris is the proximity effect of the waveguide wall as the irises are moved further away from the center of the waveguide. Since the data are presented as a function of the spacing from the center of the iris to the center of the waveguide, this effect will first become apparent for the wider iris. This will explain, at least qualitatively, the departure from linearity of the band pass and Q curves for the wider single irises, and the fact that this point of departure is first evident in the diaphragms with the widest iris. Similarly, the slope of the resonant-frequency curve for the single-iris diaphragm is steepest for the widest iris, an indication again that the proximity effect of the waveguide wall becomes an important factor in the behavior of the iris as the iris is moved further away from the center of the waveguide.

In the case of the double-iris diaphragm several effects are present. The two irises appear to act like a single iris of twice the width for the closest spacing between the irises, since for each size iris the ratio of the Q of the single iris to the Q of the corresponding double iris is very close to two. As the spacing is increased, a coupling effect is introduced which results in lowering the Q to some minimum value. The proximity effect of the waveguide wall appears to be the predominant factor as the iris separation is increased. The portion of the Q curves over which this effect is apparent is quite similar in appearance to the curves for the single-iris diaphragms. It will be noted that the ratio of the Q curves are similar to each other for each size iris when the irises are close to the center of the waveguide. The ratio changes rapidly, however, as the spacing between the irises increases and the irises begin to approach the waveguide wall.

The magnitude of the effect of using multiresonant elements is best illustrated by the triple iris. The Q of the two off-center irises individually is 3.25, while their combined Q is equal to 1.02. The center iris alone has a Q of 2.4. The combination of all three irises into a triple iris results in a Q of 0.46. It should be noted that the band pass (vswr = 1.4) associated with this Q is approximately 80 per cent of the recommended band pass of the standard S-band waveguide.[3]

[3] Moreno, "Microwave Transmission Design Data," McGraw-Hill Book Co., Inc., New York, N. Y., p. 128; 1948.

V. Conclusions

The investigation discussed in this paper has shown that the Q of an iris of fixed dimensions in the center of a thin diaphragm placed transversely in a waveguide can be appreciably increased by moving it closer to the wider waveguide wall. In addition, a low-Q multi-iris structure can be realized in which the Q of each iris is comparatively high. The resonant structures described can be used as filter elements of desired characteristics by judiciously positioning and dimensioning the irises.

Experimental evidence has been presented which shows that the double-iris diaphragm behaves in a manner similar to a single-resonant structure. Q for both the single- and multi-iris structures varies smoothly as a function of the position of the iris in the diaphragm. Therefore, it should be possible to describe this phenomenon by the evaluation of the proper equivalent circuit. To date, however, efforts in this direction have not been successful.

On the Excitation of Surface Waves*

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Summary—**It is shown that surface waves are actually contained in complete solutions of excitation problems of cylindrical surface waveguides. The total field can be divided into surface waves and a supplementary field; these field components satisfy certain orthogonality relations between each other. The amplitude of a surface wave which is excited by a dipole or any other power source can be determined without solving the entire excitation problem.**

Introduction

Surface waves are waves which are guided by and along an interface of two different media without losing energy by radiation. The physical reality of such waves (particularly the Uller-Zenneck wave[1] and Sommerfeld's single wire wave[2]) has been questioned frequently. The fact that such waves are solutions of Maxwell's equations is not, *a priori*, a proof for their existence. Such solutions may have the meaning of asymptotic solutions and describe the field infinitely remote from the power source where the field intensity approaches zero.

The two-wire, or Lecher wave has been commonly accepted as real since it can be easily established by experiments. Actually it is much more complicated from the theoretical point of view than, for example, Sommerfeld's wave. It only exists if both conductors have infinite conductivity or if they are of identical structure.[3] If a two-wire line section is excited by a concentrated power source acting at one end the measured field is in very good agreement with the field of a Lecher wave. If a similar experiment is made with a single wire, for instance by exciting it in the middle, the result is quite different. The field shows no indication of Sommerfeld's wire wave. From such an experiment one might conclude that the single-conductor wave is nonexistent. The explanation, however, is that the guided wave is overshadowed by the supplementary field, a field component which is required in order to preserve the continuity of the total field in the vicinity of the power source and at the ends of the wire.

Although surface waves other than the two-wire wave have been established by experiments,[4] still missing is a general theoretical proof of their existence. (For the single conductor wave this proof has been recently given by T. E. Roberts[5] who treated the excitation problem of a wire with modified surface.) It is the object of this pa-

* Decimal classification: R118. Original manuscript received by the Institute, July 19, 1951; revised manuscript received, March 13, 1952. Presented, 1951 IRE National Convention.

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[1] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, p. 584; 1941.

[2] *Ibid.*, pp. 527–531.

[3] Shown by A. Meyerhoff, Signal Corps Engineering Laboratory Technical Memorandum M-1376, the two-wire line should actually be considered as system of two coupled single wires.

[4] C. H. Chandler, "An Investigation of Dielectric Rod as Waveguide," *Jour. Appl. Phys.*, vol. 20, p. 1188; December, 1949. G. Goubau, "Surface Waves and Their Application to Transmission Lines," *Ibid.*, vol. 21, p. 1119; November, 1950; "Single Conductor Surface Wave Transmission Line," Proc. IRE, vol. 39, p. 619; June 1951.

[5] T. E. Roberts, Jr., "Currents Induced on an Infinitely Long Wire by a Slice Generator," and, "Properties of A Single-Wire Line," Technical Report Nos. 129 and 134, Cruft Laboratory, Harvard University, Cambridge, Mass.; May 2, 1951 and November 1, 1952.

per to present such proof and to show that the supplementary fields and the surface waves are mathematically separable by means of orthogonality relations which are also valid in the case of closed waveguides. The amplitude of a surface wave which is excited by a given power source can be calculated without solving the entire excitation problem. From the analysis it will become obvious that there is no fundamental difference between the excitation problem of an ordinary two-wire line and that of any other open waveguide.

Relations Between Surface Waves

Assuming there are two or more possible solutions of Maxwell's equations describing surface waves, as for instance in the case of dielectric rods, there exist the following relations between the field components of any two of these waves:

$$\int_F (E_{S\gamma} \times H_{S\mu}) n df = \int_F (E_{S\mu} \times H_{S\gamma}) n df = 0, \tag{1}$$

where $E_{S\gamma} H_{S\gamma}$ and $E_{S\mu} H_{S\mu}$ are the fields of these waves and n is a unit vector normal to the surface of integration. The integration is performed over the entire area of a plane perpendicular to the guide. The relations are well-known for closed waveguides, and they are valid also for open waveguides.[6] They can be easily derived by means of the reciprocity theorem.

Relations Between Surface Waves and Supplementary Fields for Lossless Waveguides

We consider first a lossless surface waveguide of unlimited length excited by an electric dipole as shown in Fig. 1. Provided one or more surface waves are excited, they represent the asymptotic solution of the problem

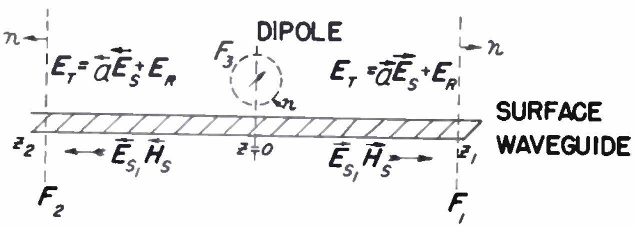

Fig. 1—Excitation of a surface waveguide by a dipole.

since their amplitudes remain constant along the line, while the amplitude of the radiating field decreases. The total field E_T, H_T can be written in the form:

$$z > 0;\ E_T = \sum_\gamma \vec{a}_\gamma \vec{E}_{S\gamma} + E_R \quad H_T = \sum_\gamma \vec{a}_\gamma \vec{H}_{S\gamma} + H_R$$

$$z < 0;\ E_T = \sum_\gamma \overleftarrow{a}_\gamma \overleftarrow{E}_{S\gamma} + E_R \quad H_T = \sum_\gamma \overleftarrow{a}_\gamma \overleftarrow{H}_{S\gamma} + H_R. \tag{2}$$

[6] R. Adler, "Waves on inhomogeneous cylindrical structures," Proc. IRE, vol. 40, pp. 339–348; March, 1952.

$\vec{E}_{S\gamma}$, $\vec{H}_{S\gamma}$ and $\overleftarrow{E}_{S\gamma}$, $\overleftarrow{H}_{S\gamma}$ are the field components of surface waves of unit amplitude traveling in the positive and negative z-direction $\vec{a}_\gamma$ and $\overleftarrow{a}_\gamma$, the corresponding amplitude factors. E_R, H_R is the supplementary field. The summation is performed over all the surface waves.

The total field and the unit surface waves are solutions of Maxwell's equations. Therefore they satisfy the reciprocity relations

$$\oint_F (E_T \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times H_T) n df = 0. \tag{3}$$

n is the unit vector normal to the surface of integration. The integral shall be taken over the surface of the space between the two cross planes F_1 and F_2 with the exception of a small sphere F_3 inclosing the dipole. The integral over the cylindrical part of the surface, which is located at infinity, vanishes since the field of the surface wave decreases in the radial direction exponentially or with $1/r^2$ (in the case of a two-wire line).

If E_T and H_T are expressed in terms of the surface waves and the radiating field according to (2) and if the orthogonality relations (1) are observed, the reciprocity integrals over F_1 and F_2 become

$$\begin{aligned}\int_{F_1} (E_T \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times H_T) n df &= \vec{a}_\mu \int_{F_1} (\vec{E}_{S\mu} \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times \vec{H}_{S\mu}) n df \\ &\quad + \int_{F_1} (E_R \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times H_R) n df\end{aligned} \tag{4a}$$

$$\begin{aligned}\int_{F_2} (E_T \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times H_T)\, n df &= -\overleftarrow{a}_\mu \int_{F_2} (\overleftarrow{E}_{S\mu} \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times \overleftarrow{H}_{S\mu}) n df \\ &\quad - \int_{F_2} (E_R \times \overleftrightarrow{H}_{S\mu} - \overleftrightarrow{E}_{S\mu} \times H_R) n df.\end{aligned} \tag{4b}$$

The transverse field components of the unit surface waves can be written in the form:

$$\begin{aligned}\vec{E}_{S\mu} \cdot t &= E_\mu e^{-jhz} \cdot t & \vec{H}_{S\mu} \cdot t &= H_\mu e^{-jhz} \cdot t \\ \overleftarrow{E}_{S\mu} \cdot t &= E_\mu e^{+jhz} \cdot t & \overleftarrow{H}_{S\mu} \cdot t &= -H_\mu e^{+jhz} \cdot t\end{aligned} \tag{5}$$

with E_μ, H_μ being transverse vectors. t is a unit vector in any transverse direction. The subscript μ on the propagation constant h is omitted. Thus (4a) and (4b) with the upper arrows become

$$\int_{F_1} (\ ) n df = -2\vec{a}_\mu \int_{F_1} (E_\mu \times H_\mu) n df - \int_{F_1} (E_R \times H_\mu + E_\mu \times H_R) e^{jhz_1} n df \tag{6a}$$

$$\int_{F_2} (\ )ndf = \int_{F_2} (E_R \times H_\mu + E_\mu \times H_R) e^{jhz_2} ndf, \tag{6b}$$

and with the lower arrows:

$$\int_{F_1} (\ )ndf = \int_{F_1} (E_R \times H_\mu - E_\mu \times H_R) e^{-jhz_1} ndf \tag{7a}$$

$$\int_{F} (\ )ndf = -2\overset{\leftarrow}{a}_\mu \int_{F_2} (E_\mu \times H_\mu) ndf$$

$$- \int_{F_2} (E_R \times H_\mu - E_\mu \times H_R) e^{-jhz} ndf. \tag{7b}$$

If the position of one of the cross planes, say F_1 is changed, the reciprocity integrals over F_2 and F_3 are unchanged. Therefore we obtain from (6a):

$$e^{jhz_1} \int_{F_1} (E_R \times H_\mu + E_\mu \times H_R) ndf = \text{constant}, \tag{8a}$$

and from (7a):

$$e^{-jhz_1} \int_{F_1} (E_R \times H_\mu - E_\mu \times H_R) ndf = \text{constant}. \tag{8b}$$

Since we consider a lossless guide, h is real. If the position of F_1 approaches $z_1 = +\infty$, E_R, H_R approach zero. Therefore the constants in (8a) and (8b) are zero. The same pertains to the integrals over F_2. Thus,

$$\int_F (E_R \times H_\mu) ndf = \int_F (E_\mu \times H_R) ndf = 0 \tag{9}$$

for any cross plane on either side of the dipole.

The reciprocity integral which is taken over the sphere surrounding the dipole can be evaluated easily:

$$\int_{F_3} (E_T \times \overset{\leftrightarrows}{H}_{S\mu} - \overset{\leftrightarrows}{E}_{S\mu} \times H_T) ndf = \begin{cases} -j\omega P \overset{\leftrightarrows}{E}_{S\mu} & \text{(electric dipole)} \\ j\omega M \overset{\leftrightarrows}{H}_{S\mu} & \text{(magnetic dipole)} \end{cases} \tag{10}$$

Where P is the moment of an electric dipole and M the moment of a magnetic dipole, $\overset{\leftrightarrows}{E}_{S\mu}$ and $\overset{\leftrightarrows}{H}_{S\mu}$ are the field components of the unit waves in the direction of the dipoles.

Thus, all parts of the reciprocity integral (3) are evaluated. With (6), (7), (9) and (10) substituted in (3) we obtain the amplitudes a_μ of the surface waves excited in both directions.

$$\overset{\rightleftarrows}{a}_\mu = -\frac{j\omega P \overset{\leftrightarrows}{E}_{S\mu}}{2N} \quad \text{or} \quad \overset{\rightleftarrows}{a}_\mu = +\frac{j\omega M \overset{\leftrightarrows}{H}_{S\mu}}{2N} \tag{11}$$

with

$$N = \int_F (E_\mu \times H_\mu) ndf. \tag{12}$$

N is the power of a unit surface wave if E_μ and H_μ are defined as real quantities.

The relations (11) do not contain the radiating field. Therefore the amplitudes of the surface waves excited by a dipole or any given power source can be determined without solving the entire excitation problem. Furthermore, (11) shows that surface waves are actually excited by an electric or magnetic dipole provided the unit waves have the proper field component in the direction of the dipole.

Waveguide With Losses

If the losses are taken into account, the preceding considerations must be revised since h is now complex and e^{jhz} and e^{-jhz} approach infinite values at $z = +\infty$ and $z = -\infty$ respectively. Thus, we cannot conclude that the constant in (8a) is zero. Instead of (9) we obtain then

$$e^{jhz_1} \int_{F_1} (E_R \times H_\mu) ndf = e^{jhz_1} \int_{F_1} (E_\mu \times H_R) ndf = \text{constant} \tag{13a}$$

and correspondingly,

$$e^{-jhz_2} \int_{F_2} (E_R \times H_\mu) ndf = e^{-jhz_2} \int_{F_2} (E_\mu \times H_R) ndf = \text{constant}. \tag{13b}$$

If we consider waveguides for which the conductivity is not essential for the existence of the surface wave, for instance in the case of a two-wire line or a single-wire line having a dielectric coat, the field distribution close to the dipole will be practically the same whether the line has losses or not—provided the dipole is not so far removed from the line that at its location the field distribution of the surface waves is noticeably modified. Therefore the amplitudes of the excited surface waves are essentially the same as in the lossless case, and given by (11). From this it follows that the orthogonality relations (9) between the radiating field and the surface waves remain valid.

The situation is different in the case of Sommerfeld's wire wave, since this wave exists only by reason of the finite conductivity, and the field structure depends to a high degree on the conductivity. If we consider the total field, E_T, H_T and subtract from this field surface waves going in both directions, each of which has an amplitude as given by equation (11), then the remainder of the field satisfies the orthogonality relations with the surface waves [equation (9)]. Therefore it is reasonable to consider the amplitudes of the surface waves given by equation (11) as the true amplitudes and the remainder as the supplementary field. This does not mean that these amplitudes of the surface waves actually could be measured since we know from experience they are overshadowed by the radiating field. However, if we consider a power source which would excite Sommerfeld's wave only, i.e., a layer of properly distributed dipoles, and if we then calculate the amplitude of this wave, using equation (11) for each individual dipole, we obtain the correct result.

It should be mentioned that in the case of lossy waveguides, particularly in the case of Sommerfeld's wave, the total energy delivered by the power source is *not* the sum of the power propagated by the surface waves and the power radiated by the supplementary field, since both fields are—if dissipation is present—not orthogonal with regard to power considerations.

Examples

A lossless single conductor guide with a dielectric coating is excited by a concentrated voltage source (Fig. 2). The voltage source is represented by a magnetic current ring around the wire producing a voltage

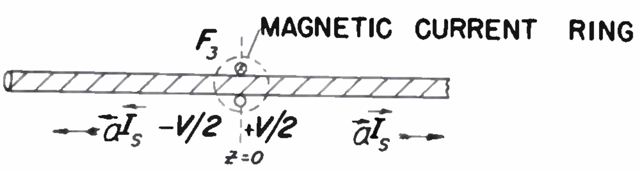

Fig. 2—Excitation of a single conductor guide by a concentrated voltage source.

step from $-V/2$ at the left side of the ring to $+V/2$ on the right side. In this case there is only one surface wave existing, provided the dielectric layer is small compared with the wavelength. According to the procedure outlined before we exclude the power source by a snugly fitting surface F_3 and evaluate the reciprocity integral for this surface. If we express the unit surface waves in terms of their currents $\vec{I}_S$ and $\overleftarrow{I}_S$ in the wire we obtain

$$\int_{F_3} (E_T \times \overleftrightarrow{H}_S - \overleftrightarrow{E}_S \times H_T) n\, df = -V \overleftrightarrow{I}_S. \qquad (14)$$

We define the unit waves so that at $z=0$, $\vec{I}_S = \overleftarrow{I}_S = I_S$. The integral in (14) replaces the integral which before was taken over the sphere surrounding the dipole. Thus we obtain with (11) the amplitudes of the surface waves excited by the voltage source:

$$\vec{a}\, I_S = \overleftarrow{a}\, I_S = \frac{V}{2Z_S} \quad \text{with} \quad Z_S = \frac{N}{|I_S|^2} \qquad (15)$$

where the amplitudes are expressed in terms of the currents of the unit waves at $z=0$. N is the power of a unit wave; Z_S can be called the characteristic impedance of the guide.

This result may appear obvious if the two halves of the line, the one to the left and the one to the right of the power source are considered as two transmission lines which are connected in series with the voltage source. But we must remember that the definition of the characteristic impedance in (15) is arbitrary and purposely chosen so that this simple result is obtained. Usually the impedance is defined as the ratio of the transverse electric and magnetic field components. Using this definition the result is different. It shall be noticed that the current determined by equation (15) is not the total current delivered by the voltage source. The total input impedance of the line consists of two impedances acting in parallel: one ($2Z_S$) representing the surface waves and the other one representing the supplementary field. The latter one is unknown. It contains a positive resistive component due to the radiation loss. In order to have little radiation loss a power source of large extension is required.

As another example a two-wire line is considered in Fig. 3. First let the line be excited by a voltage source acting on one wire only (Fig. 3a). The voltage source is

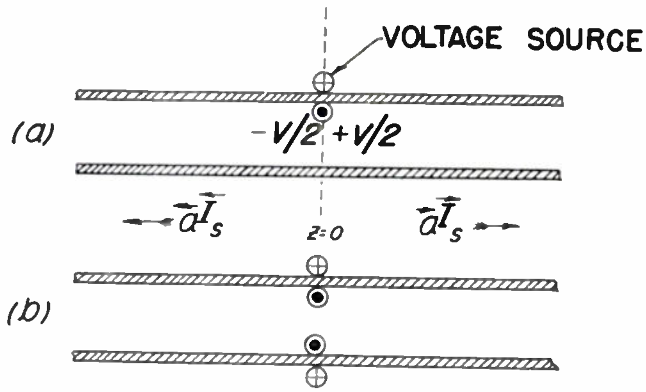

Fig. 3—Excitation of a two-wire line.

represented again by a magnetic current. If we assume no losses, the only type of existing surface wave is the ordinary two-wire or Lecher wave. The result is the same as in the preceding example. Besides the Lecher wave, a considerable radiating field is excited. The characteristic impedance defined by N/I_S^2 is here identical with the impedance defined by the ratio of the transverse field components. If the losses are considered or if the wires have a dielectric coating, the problem becomes more complicated since in addition to the Lecher wave a wave like that on a single conductor is possible, and this wave also is excited. Moreover, there are waves of the Hondros type[7] present, but these are of no practical consequence.

If the power source is split into two halves which act on both wires but in opposite directions (Fig. 3b), the amplitude of the Lecher wave is the same as before. However, since the supplementary fields for both power sources have the same structure but opposite direction, they almost cancel. Therefore only a small radiation will remain due to the displacement of the two fields. This compensation of the radiating field is the reason why a two-wire line can be excited efficiently with a concentrated power source provided the excitation is symmetrical. All other surface waveguides requires launching devices of comparatively large extension in order to avoid excessive radiation loss.

[7] D. Hondros, *Ann. der Phys.*, vol. 30, p. 905; 1909.

Correspondence

Errors in a Microwave Rotary Phase Shifter*

An interesting microwave phase shifter, making use of two fixed quarter-wave plates with a rotatable half-wave plate between them, was described in a paper by Fox (see Fig. 1).[1] Such a device is useful in the laboratory since the phase shift in degrees equals twice the mechanical angle of rotation of the half-wave plate in degrees. The deviation from linearity of phase shift versus rotation depends on the deviation from correct positioning of the two fixed plates and the errors in phase shift of the individual plates. Some time ago M. A. Meyer[2] showed, in an unpublished calculation, that small deviations in the design of the phase-shifter components could be tolerated since the deviation from linearity depended only to the second order on the variations in the dielectric plates. This fact was thought of sufficient interest to warrant being brought to the attention of PROCEEDINGS readers.

The proof of the above assertion may be demonstrated in several ways. Perhaps the most systematic are the use of complex vectors to represent the magnitude and phase of the electric field and the use of a different x-y co-ordinate system in each section with the x axis always parallel to the particular plate, as shown in the figure. The field in each section may be resolved into linear x and y components and a relationship between output and input established. This was essentially the method used by Meyer.

A qualitative physical proof may be based on a consideration of circularly polarized waves. When operating perfectly, the first quarter-wave plate converts vertical linear to right-hand circular polarization. The half-wave plate converts this to left hand, whose phase is determined by the angle of rotation. Finally, the second quarter-wave plate converts left-hand circular back to vertical linear polarization. If the plates or their alignments are not perfect, a small amount of circularly polarized energy of wrong sense will be generated. Thus, the first-order effect of errors will be to produce a cross-polarized component in the output which, if absorbed by a horizontal resistance card, will not affect linearity of phase shift.

Consider the second quarter-wave plate first. Incident upon it is mainly a left-hand circular, and also a small right-hand component due to imperfections. The left hand is converted mainly into linear vertical (whose phase is certainly determined by the phase of the left-hand circular), plus a small amount of horizontal which is absorbed in a resistance card. The small right-hand component is converted mainly to horizontal, which is also absorbed, and a fractional amount of vertical. This fractional amount of a small quantity is of second order. Thus, at the second quarter-wave plate, only the left-hand-circular component contributes, to the first order, to the vertically polarized output signal.

A similar argument may be carried through for the first plate. A large amount of right-hand plus a small left-hand circular component is produced by this plate. The large amount of right hand is largely converted to left hand in passing through the half-wave plate; the fraction of this signal that is converted to right hand is discarded, as seen in the preceding paragraph. The small left-hand component from the first plate is converted through half-wave plate mainly to right hand, which also does not contribute, in first order, to final output.

It remains merely to be shown, therefore, that the phase of the left-hand circular output of the half-wave plate with respect to a fixed-phase right-hand circular input is a linear function of the angle of rotation. From symmetry considerations the incident right-hand circularly polarized wave looks the same to the half-wave plate for any angle of rotation and thus the output will be the same when referred to a set of axes moving with the plate. The transformation of the reference phase from the moving axes to a fixed set of axes introduces exactly the linear dependence of phase on rotation. (Thus, if the quarter-wave plates are perfect, an imperfect half-wave plate will not affect the linearity of the phase shifter at all, but will only cause a loss in power because of the setting up of a cross-polarized component.)

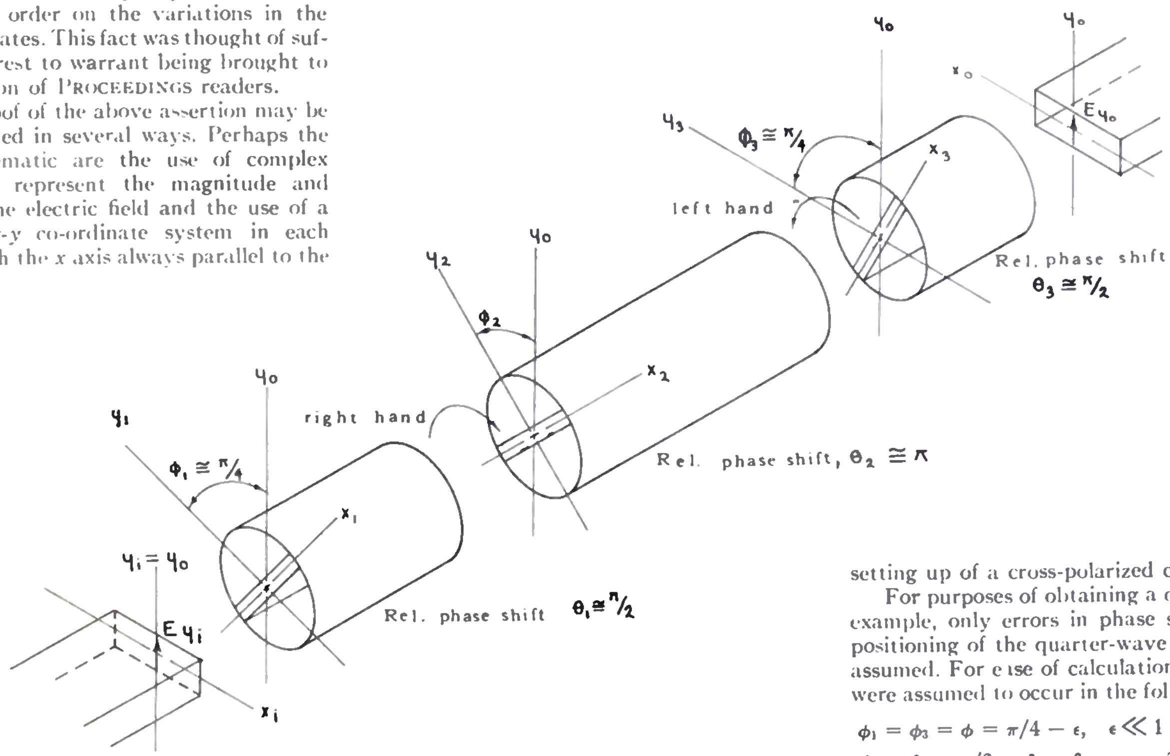

Fig. 1—Rotary phase shifter, schematic.

For purposes of obtaining a quantitative example, only errors in phase shift and in positioning of the quarter-wave plates were assumed. For ease of calculation, the errors were assumed to occur in the following way:

$$\phi_1 = \phi_3 = \phi = \pi/4 - \epsilon, \quad \epsilon \ll 1$$

$$\theta_1 = \theta_3 = \pi/2 - \delta, \quad \theta_2 = \pi - 2\delta, \quad \delta \ll 1.$$

Then, letting $\psi = \phi_2 - \phi_3 = -(\phi_1 - \phi_2)$, an expression for the output voltage divided by the input voltage is

$$\frac{E_{y0}}{E_{yi}} = \cos 2\epsilon \cos \delta \, e^{-2j\delta} e^{-2j\psi} + e^{-2j\delta}\big[\cos 2\psi(1 - \cos 2\epsilon \cos \delta) - 2\cos^2\psi \sin^2\delta + j\cos^2\psi \sin 2\delta \sin 2\epsilon\big].$$

Note the phase of the first term varies linearly with ψ and the second term contains only expressions of second order in ϵ and δ.

A numerical calculation for $\delta = 5°$, $\epsilon = 1°$ shows that the maximum deviation from linearity is less than 0.75 degrees.

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* Received by the Institute, October 11, 1951.

[1] A. Gardner Fox, "An adjustable wave-guide phase changer," PROC. I.R.E., vol. 35, pp. 1489–1498; December, 1947.

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Mr. Callick has been with the Services Electronics Research Laboratory from 1949 to the present.

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John L. Carter was born April 6, 1920 in Bramwell, W. Va. He received the B.S. degree in mathematics and physics from West Virginia State College in 1947, and did graduate work in physics at New York University from 1947 to 1949.

John L. Carter

Mr. Carter joined the Thermionics Branch of the Signal Corps Engineering Laboratories, Fort Monmouth, N. J. in 1949, where he has been engaged in the development of radar duplexing systems and associated microwave gas-discharge devices.

J. H. Coleman (A'47–A'51) was born in Danville, Va., on August 21, 1925. He received the B.E.E. degree from the University of Virginia in 1946 on the Navy V-12 program. He attended Princeton University as a half-time graduate student in the physics department from 1946 to 1950.

J. H. Coleman

During this time Mr. Coleman was a research engineer for the RCA Laboratories Division in Princeton, N. J., working on klystrons, radioactive voltage sources, and magnetic gas-discharge tubes.

In 1950 Mr. Coleman, as research director, organized the Radiation Research Corporation in West Palm Beach. Projects included radiation survey meters, electrometer vacuum tubes, and applications of radioisotopes.

Mr. Coleman is a member of the American Physical Society and a lieutenant in the Naval Reserve.

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Huntington W. Curtis was born January 31, 1921, in Bailey Island, Maine. He received the B.S. degree from the College of William and Mary in 1942, with majors in chemistry and in physics, the M.S. degree from the University of New Hampshire in 1948, and the Ph.D. degree from the University of Iowa in 1950, the latter degrees in electrical engineering.

H. W. Curtis

During the war years Dr. Curtis served as an officer in the Army Signal Corps; this duty included two years on the teaching staff of the United States Military Academy as instructor in electrical engineering. After finishing at the University of Iowa in 1950, he became assistant professor in the Electronics Division, E.E. Department, of the University of Virginia. Recalled to active duty in June, 1951, he is at present assistant professor in the Department of Electricity of the United States Military Academy. Dr. Curtis is a member of Phi Beta Kappa, Tau Beta Pi, and Sigma Xi.

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Sheldon H. Dike (S'40–M'41–SM'48) was born in Atlantic City, N. J., on October 23, 1916. He attended Colgate University, and received the B.S. degree in electrical engineering from the University of New

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Mexico in 1941. He received the Ph.D. degree from Johns Hopkins University in 1951.

During 1941–42 Dr. Dike was a physicist at the Department of Terrestrial Magnetism of the Carnegie Institution, where he was engaged in research leading to the first practical proximity fuze for projectiles. He has several patents on proximity fuzes and related devices, and in recognition of his work was given the Naval Ordnance Development Award.

S. H. Dike

Dr. Dike was a research engineer at the University of New Mexico in 1942–43, and a research associate at the University of Michigan, 1943–44. He was appointed to the staff of the Los Alamos Scientific Laboratory in 1944. As a civilian consultant with the atomic-bomb group overseas, it was his responsibility to supervise the loading into aircraft of the two bombs that were dropped on Japan.

In 1946 Dr. Dike joined the Glenn L. Martin Company as head of electronics research and development for guided missiles. He was associated with the Radiation Laboratory of Johns Hopkins University from 1947 to 1951, where he was engaged in proximity-fuze research. At present he is with the Research Department of the Sandia Corporation, Albuquerque, New Mexico.

Dr. Dike's activities include membership in the American Institute of Electrical Engineers, the American Physical Society, and Sigma Xi.

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Kenton Garoff (A'44) was born in Bridgeport, Conn. in June 1918. He was graduated from Brooklyn College in 1940 with a B.A. degree in chemistry and physics. Since then he has done graduate work in physics and mathematics at Rutgers University and Brooklyn Polytechnic Institute.

Kenton Garoff

Mr. Garoff joined the Thermionics Branch of the Signal Corps at Fort Monmouth, N. J. in 1942 where he worked on sweep circuits, IF amplifiers and uhf receivers. In 1943, he joined the group working on microwave gas-switching TR and ATR tubes, and has been in charge of this group since 1944. He is currently engaged in research and development of radar duplexers and microwave gas-discharge devices.

George H. Gordon was born in Kansas City, Mo., on July 18, 1917. He received the B.S. degree in chemical engineering from the University of Kansas in 1939.

G. H. Gordon

From 1939 to 1947 Mr. Gordon was on the staff of the Kodak Research Laboratories, where he worked on color photographic processes. During World War II he was in the U. S. Naval Reserve as an electronics officer. Since 1947 he has been in the New York office of the Motion Picture Film Department of the Eastman Kodak Company.

Mr. Gordon is a member of the Society of Motion Picture and Television Engineers, Tau Beta Pi, and the Video Utilization Committee of the IRE.

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Georg Goubau (A'49) was born in Munich, Germany, on November 29, 1906. He received the Dipl. Phys. degree in 1930, and the Dr. Ing. degree in 1931, both from the Munich Technical University. From 1931 to 1939 he was employed in research and teaching in the physics department of the same university, under Professor Zenneck. During this time he was principally concerned with ionospheric investigations. He established the first German Ionospheric Research Station (Herzogstand/Kochel), and was in charge of the research work carried on at this station.

Georg Goubau

In 1939 Dr. Goubau was appointed professor and director of the department of applied physics at the Friedrich-Schiller University, in Jena, Germany. Before he arrived in this country, he was the senior author of the volumes on electronics of the FIAT Review of German Science, published by the Military Government for Germany. Dr. Goubau is now a consultant at the Signal Corps Engineering Laboratories, in Fort Monmouth, N. J.

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Leonard R. Kahn (S'46–M'51) was born on June 16, 1926, in New York, N. Y. He received the B.E.E. degree, cum laude, from the Polytechnic Institute of Brooklyn in 1951.

Mr. Kahn attended Syracuse University under the Army Specialized Training Program from 1943–1944. From 1944–1946, he served in the USA Signal Corps. From 1947–1950, Mr. Kahn was employed by RCA Communications, and during the last year of his employment was on part-time loan to RCA Laboratories. In 1950 he joined the Crosby Laboratories at Mineola, N. Y., and is now a project engineer specializing in communication - circuits analysis and design.

Leonard R. Kahn

Mr. Kahn is a member of the IRE Subcommittee on Single-Sideband Transmitters. He is an associate member of the American Institute of Electrical Engineers and a member of Tau Beta Pi and Eta Kappa Nu.

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W. W. Keith was born in Magnolia, Ark., on February 16, 1923. He has completed courses in radio engineering from the Capitol Radio Engineering Institute. From 1940 to 1942 he did radio servicing.

W. W. Keith

In 1942 he joined the Raytheon Mfg. Co., doing system and type testing of radar equipment. In 1946 he was transferred to the receiver section, where he has developed moving-target-indicators and memories for digital computers, and has developed and designed mercury delay lines.

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Arthur E. Kerwien (A'29–M'40–SM'43) was born in Fort Lee, N. J., on April 16, 1904. After receiving a B.S. degree in Physics at Union College in 1926, he joined Bell Telephone Laboratories. In 1947 he received an M.S. degree from Stevens Institute of Technology.

A. E. Kerwien

Throughout his twenty-five years' service with Bell Laboratories, he has been concerned with both research and development of radio transmitters. He was instrumental in the development of radiotelephony, including multi-channel radiotelephone. He also worked on problems of negative feedback, and is currently engaged in the development of single-sideband, multi-channel radiotelephone transmitters for overseas use.

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D. D. King (M'46) was born on August 7, 1919, in Rochester, New York. He received the A.B. degree in engineering sciences from Harvard College in 1942 and the Ph.D. degree in physics from Harvard University in 1946. He was a teaching fellow in physics and communication engineering in 1943, serving as a staff member of the pre-radar Officer's Training School at Cruft Laboratory, Harvard University. During 1945 he was a research associate at Cruft Laboratory. In 1946 he was appointed research fellow in electronics and in 1947 assistant professor of applied physics in Harvard University.

D. D. King

In 1948 Dr. King was appointed associate professor of physics in the Institute for Co-operative Research of the Johns Hopkins University, and in 1950 assistant director of the Radiation Laboratory.

Dr. King is a member of Sigma Xi and the American Physical Society.

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L. M. Klenk (M'41–SM'43), a native of Kearny, N. J., was born on February 19, 1907. He was graduated from Newark College of Engineering with a B.S. degree in 1929. That year he joined Bell Telephone Laboratories, where, until World War II his time was devoted to the design and installation of short-wave radio transmitters used for overseas communication and high-frequency short-hop communication equipment. During the war he designed radar and counter-measure equipment for the government.

L. M. Klenk

Since 1945, Mr. Klenk has been engaged in the mechanical design of an overseas radio transmitter.

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Nean Lund (A'43) was born in Madison, Wis., on May 13 1912. He was graduated in 1935 from the University of Wisconsin, where he received a B.S. degree in E.E. After a year of post-graduate study at the University, and a brief period as a radio consultant, he joined the staff of Bell Telephone Laboratories.

Nean Lund

He first worked on the development and testing of single-sideband radio receivers. From 1941-43, he was concerned with circuit development of loran, radar and other equipment used by the military. In recent years he has worked on the measurement of interchannel modulation noise from multiplex single sideband transmitters, and on the measurement of extra-band radiation from multiplex single sideband radio transmitters.

Mr. Lund is a member of the American Institute of Electrical Engineers, Phi Eta Sigma, and Eta Kappa Nu.

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E. G. Linder (A'40–SM'49) was born in 1902 in Waltham, Mass. He received the B.A. degree from the University of Iowa in 1925, and the M.S. degree in 1927.

E. G. Linder

Dr. Linder was an instructor in physics at the State University of Iowa from 1925 to 1927; an instructor at California Institute of Technology from 1927 to 1928; and a Detroit Edison Fellow at Cornell University from 1928 to 1932, where he received the Ph.D. degree in 1931.

Dr. Linder was employed in the Research Division of the RCA Victor Company from 1932 to 1935 and the RCA Manufacturing Company, RCA Victor Division, from 1935 to 1942. At present he is with the RCA Laboratories at Princeton, N. J.

Dr. Linder is a member of the American Physical Society.

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John L. Moll (A'51) was born on December 21, 1921, in Wauseon, Ohio. He received the B. Sc. degree in engineering physics in December, 1943 from the Ohio State University.

John L. Moll

Mr. Moll worked at the Radio Corporation of America during the years 1944 and 1945. He was employed as a graduate assistant in the mathematics department at Ohio State University from October 1945 until June 1948. Since 1948 he has been an instructor in the electrical engineering department, as well as a research associate for the Ohio State University Research Foundation.

Mr. Moll is a member of Sigma Xi and the American Mathematical Society.

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A. J. Munn (A'41) was born in Belfast, Ireland, on March 9, 1908. As a boy of 10, he came to the United States and at 15 went to work at Bell Telephone Laboratories. In 1936 he received a B.S. degree in electrical engineering from Cooper Union and in 1941 a B.E.E. degree from New York University.

A. J. Munn

Mr. Munn first learned instrument and tool making at the Laboratories, and for a few years conducted research on telephone voice recording. From the early thirties up to the present, however, he has been concerned principally with the electromechanical design of radiotelephone receivers and transmitters, except during World War II, when he had special military assignments.

Mr. Munn is a member of Eta Kappa Nu.

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Joseph Nedelka was born in Bohemia on November 13, 1905; he came to the United States six months later with his family. In 1937 he received an E.E. degree from Brooklyn Polytechnic Institute.

J. Nedelka

Mr. Nedelka joined Western Electric Company's Engineering Department, now known as Bell Telephone Laboratories, in 1922. While in the drafting department, he took a technical assistant's course provided by the company. He was engaged in systems engineering and drafting until 1940 when he worked on the design of radar equipment. After World War II, he engaged in the design of overseas radiotelephone work, and is currently occupied with the design of television terminal and coaxial carrier equipment.

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Evert J. Post was born in Rotterdam, Netherlands, on October 20, 1914. He received the degree of Physical Engineer from the Delft Institute of Technology.

Evert J. Post

In 1946 Mr. Post joined the Central Laboratory of the Netherlands Postal and Telecommunications Services at The Hague where he has been primarily concerned with the development of quartz crystals and oscillators.

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Irving Reingold was born in Newark, N. J., on November 13, 1921. He received the degree of B.S. in mechanical engineering in 1942, and the professional engineers' degree in 1949, both from the Newark College of Engineering. He has also done graduate work in electronics at the Brooklyn Polytechnic Institute.

IRVING REINGOLD

From 1943 until 1945 he was an electronics manufacturing engineer in the Electronic-Tube Section of the Westinghouse Electric Company, Bloomfield, N. J.

From 1945 until 1951, Mr. Reingold was a project engineer in the Radar Section of the Watson Laboratories, Air Materiel Command, Red Bank, N. J., where he was engaged in the development and application of microwave radar-systems transmission-line and antenna components.

He joined the Thermionics Branch of the Signal Corps Engineering Laboratories at Fort Monmouth, N. J., in 1951, as a project engineer engaged in the development of microwave duplexers. Mr. Reingold is a licensed professional engineer in the State of New Jersey.

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Charles F. P. Rose (A'22–M'40–SM'43) was born in Montclair, New Jersey, September 19, 1901. He entered the Radio Research Department of the Western Electric Company in 1920. He was graduated as a Student Assistant in 1924, and subsequently attended Columbia University Extension School. Since 1925, he has served as a member of the technical staff of the Bell Telephone Laboratories.

C. F. P. ROSE

Mr. Rose has been engaged in designing, developing, installing, and testing short-wave transoceanic radio telephone transmitters for the Bell System in New Jersey, Argentina, and California. During the war he was engaged in developing special electronic equipment used by the U. S. government. In 1944, he was transferred from the Radio Research Department to the Transmission Development Department where he is continuing circuit development work for radio systems.

Edwin A. Speakman (A'43–M'44–SM'48) received his B.S. degree in physics in 1931 from Haverford College, Haverford, Pa., where he was a scholarship student. From 1931 to 1934, he was an instructor in physics at Haverford College and as part of his graduate study, invented a photoelectric timing system, the first using a light beam for timing races automatically and to accuracy of 0.01 sec.

E. A. SPEAKMAN

From 1934 to 1939 Mr. Speakman was a radio engineer in the research laboratories of the Philco Corporation, where he invented the telescopic rod antenna which is used on radio-equipped automobiles.

After serving as a physicist with the Curtis Publishing Company for a brief period in 1939, Mr. Speakman was appointed to the Naval Research Laboratory, where he held the positions of assistant superintendent of the Radio Division and head of the Countermeasures Branch, from 1940 to 1949, receiving the Navy's Meritorious Civilian Service Award in 1946 for his research on naval radar equipment. He also served as a civilian scientific expert on several special Naval missions to the Mediterranean and Pacific areas during the war.

Mr. Speakman joined the staff of the Research and Development Board in May, 1949 as executive director of the Committee on Electronics. He is a member of the American Physical Society and former member of the U. S. Civil Service Board of Examiners for Scientific and Technical personnel.

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Ludovicus R.M. Vos de Wael was born in Westervoort, the Netherlands, on July 21, 1905. He took his degree of electrotechnical engineer at the Technical University at Delft in 1930, and was appointed an engineer of the Netherlands Postal and Telecommunications Services in the same year.

LUDOVICUS R.M. VOS DE WAEL

For some years he was engineer in charge of the transmitting station at Kootwijk, and was also engaged there in development work on the first short wave ssb equipment for the radio telephone channel with Bandung.

After World War II he was, for a short time, in charge of the receiver station, and in 1947 he joined the Radio Laboratory at the Hague.

Mr. Vos de Wael was appointed chief engineer in 1948 and is now engaged in research work on frequency standards, auxiliary measuring equipment, and receiver design.

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Robert W. Wilmarth (S'47–A'51–M'52) was born on May 18, 1922, in Yonkers, N. Y. He received the B.E.E. degree in 1949, and the M.S. degree in 1950 from Ohio State University. From 1940 to 1945, Mr. Wilmarth was engaged in the construction of electron tube at the Bell Telephone Laboratories, Inc. During the years of 1945 to 1950, he was employed on a part time basis as a research associate at the Electron Tube Laboratory at the Ohio State University.

R. W. WILMARTH

Since April 1951, Mr. Wilmarth has been a senior engineer in the vacuum tube group at the Federal Telecommunications Laboratories, Inc. He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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Laurence G. Young (A'37, SM'48) is a native of Pittsburgh, Penn., where he was born January 15, 1897. He was graduated from Carnegie Institute of Technology in 1919 with a B.S. degree in E.E.

L. G. YOUNG

The following year he joined Bell Telephone Laboratories and for six years worked on coil design. In 1926 he transferred to a radio development group, where at first he was engaged in antenna design, and later, high-power radio transmitter design. During World War II, Mr. Young worked on high-power amplifiers and, also, radar and other equipment. At present he is studying short-wave radio transmitters used in transoceanic service.

Institute News and Radio Notes

Nominations—1953

At its May 7, 1953, meeting, the IRE Board of Directors received the recommendations of the Nominations Committee and the reports of the Regional Committees for officers and directors for 1953. They are:

President, 1953: J. W. McRae

Vice-President, 1953: S. R. Kantebet

Director-at-Large, 1953 1955 (*two to be elected*): S. L. Bailey, K. C. Black, B. E. Shackelford, L. C. Van Atta

Regional Directors, 1953–1954 (*one to be elected in each Region*):

Region 2.—F. M. Bailey, R. D. Chipp, J. R. Ragazzini

Region 4.—S. J. Begun, H. R. Hegbar, C. A. Priest, B. R. Teare, Jr.

Region 6.—M. C. Scott, Jr., A. W. Straiton

Region 8.—J. F. Henderson, A. B. Oxley

According to Article VI, Section 1, of the IRE Constitution, nominations by petition for any of the above offices may be made by letter to the Board of Directors, setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance, a letter of petition must reach the executive office before twelve o'clock noon on August 14, 1952, and shall be signed by at least 100 voting members qualified to vote for the office of the candidate nominated.

Calendar of COMING EVENTS

AIEE-IRE Telemetering Conference, Los Angeles, Calif., August 26–27

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29

Cedar Rapids IRE Technical Conference, Roosevelt Hotel, Cedar Rapids, Iowa, September 19–20

National Electronics Conference, Sherman Hotel, Chicago, Ill., September 29–October 1

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

IRE-AIEE Computers Conference, New York, N. Y., December 10–12

IRE-AIEE Meeting on High-Frequency Measurements, Washington, D. C., January 14–16

Radar Weather Conference, McGill University, Montreal, Canada, September 15–17

Southwestern IRE Conference and Electronics Show, Plaza Hotel, San Antonio, Tex., February 5–7

1953 IRE National Conference on Airborne Electronics, Dayton, Ohio, May 11–13

1953 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 23–26

Radio Propagation Disturbance Warnings

On July 1, 1952, the National Bureau of Standards started broadcasting new short wave radio disturbance forecasts via the NBS standard frequency broadcasting station WWV. The new service replaces the radio disturbance warning notices that have been transmitted by WWV since 1946. The broadcasts tell users of radio transmission paths over the North Atlantic the condition of the ionosphere at the time of the announcement and communication conditions to be expected for the next 12 hours.

The NBS radio disturbance forecasts, prepared four times daily, are transmitted in Morse code twice each hour—$19\frac{1}{2}$ and $49\frac{1}{2}$ minutes past the hour—on WWV standard frequencies of 2.5, 5, 10, 15, 20, and 25 mc, as was done prior to July 1. The notices heretofore have included a letter indicating present radio reception conditions. As before, the letters used are "N," "U," and "W," signifying that radio propagation conditions are normal, unsettled, or disturbed, respectively. The new notices also contain a digit indicating the expected quality of future reception. The digit is the forecast of expected quality of transmitting condition on the NBS-CRPL scale of 1 (impossible) to 9 (excellent).

RADIO FORECAST SCALE

Digit (Forecast)	Propagation Condition	Letter (Current)
1	Impossible	W
2	Very Poor	W
3	Poor	W
4	Fair to Poor	W
5	Fair	U
6	Fair to Good	N
7	Good	N
8	Very Good	N
9	Excellent	N

In addition, WWV renders the following services on standard radio frequencies of 2.5, 5, 10, 15, 20, 25, 30, and 35 megacycles; time announcements at 5-minute intervals by voice and International Morse Code; standard time intervals of 1 second, and 1, 4, and 5 minutes; and standard audio frequencies of 440 cycles (the standard musical pitch A above middle C) and 600 cycles.

The audio frequencies are interrupted at precisely one minute before the hour and are resumed precisely on the hour and each five minutes thereafter. Code announcements are in Universal Time using the 24-hour system beginning with 0000 at midnight; voice announcements are in Eastern Standard Time. The audio frequencies are transmitted alternately: the 600-cycle tone starts precisely on the hour and every 10 minutes thereafter, continuing for 4 minutes; the 440-cycle tone starts precisely five minutes after the hour and every 10 minutes thereafter, continuing for 4 minutes. Each carrier frequency is modulated by a seconds pulse which is heard as a faint tick; the pulse at the beginning of the last second of each minute is omitted.

Cedar Rapids Section to Hold Conference

The IRE Section of Cedar Rapids will sponsor a two day communications conference, September 19–20, 1952, at the Roosevelt Hotel, Cedar Rapids, Iowa.

The conference will feature outstanding speakers in the field of communications, exhibits, plant inspections, and a banquet. Among the speakers assured for the conference and their topics are: L. V. Berkner, president, Associated Universities, Inc., banquet speaker; I. S. Coggeshall, general traffic manager, Western Union Telegraph Company, speaking on information theory, M. G. Crosby, president, Crosby Laboratories, Inc., speaking on diversity reception, G. Q. Herrick, chief, State Department Broadcast Plans and Development, speaking on Voice of America operations; and, L. M. Craft, vice-president, Collins Radio Company, speaking on equipment design trends.

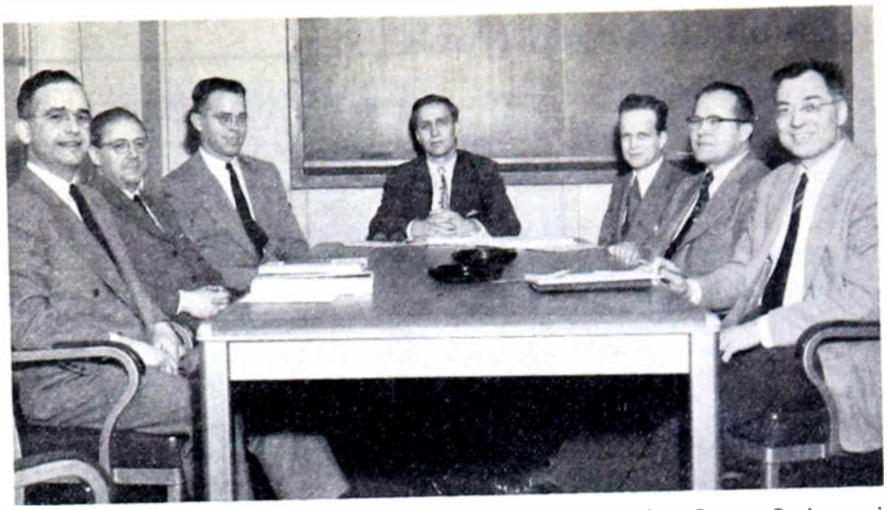
Technical Committee Notes

The **Standards** Committee met on April 10, M. W. Baldwin, Jr., was acting Chairman in the absence of A. G. Jensen. A. G. Clavier reported on the status of graphical symbols for semiconductor devices. The matter had been discussed at a meeting of the Symbols Committee at which C. D. Mitchell, Chairman of the Task Group on Graphical Symbols for Semiconductor Devices, reported on the group's activities. The work of the Task Group has been delayed for some time to prevent conflict between the IRE and ASA drafts of proposed Standards. Most of the proposed standards coming from the Task Group have been adopted by the ASA Subcommittee and incorporated in the recent ASA draft. The Symbols Committee feeling that the work still was incomplete asked Mr. Mitchell to resume the work of his task group. A meeting was planned, and the previous work will be revised and extended by the task group with a view of offering a list of symbols for the Standards Committee approval. Mr. Clavier then proceeded to the status of symbols for computers stating that W. B. Callaway, Chairman of the Task Group on Symbols for Functional Operation of Control, Computing and Switching Equipment, had reported on the work of his Group at the last meeting of the Symbols Committee. Reviewing the purpose of functional symbols as differentiated from standard graphical symbols, Mr. Callaway told the Symbols Committee that while the latter show how devices, such as relays, are connected, they do not show the operation of the devices. Mr. Callaway displayed two sets of drawings utilizing functional symbols and expressed the opinion that since there was an active demand for both systems, it might be advisable to standardize on both. The Symbols Committee agreed that for the present it would be advisable to standardize only the basic symbols used in the two systems and use these systems as the "building blocks" for a future complete standard. Mr. Callaway has agreed to revive the task group and to try to get the telephone and computer people together in this work. E. S. Seeley has offered his resignation from the ASA Committee Z24, and it was proposed that W. M. Goodale be appointed IRE Representative and H. Olson serve as his alternate. These appointments are to be sent to the Executive Committee for approval. The Committee reviewed the scopes of the technical committees. The next item on the agenda concerned the ASA (Z57) definition for "Flutter." W. M. Goodale pointed out at the March 20th meeting that the definition, previously approved by the Standards Committee, was at variance with the Receiver definition approved at the February 14 meeting. H. E. Roys, IRE Representative on ASA Z57, reviewed the work leading to approval of the ASA definition which was originally proposed and written by SMPTE. It was agreed that Mr. Roys should be asked to inform ASA that the Standards Committee will not withdraw its approval of the ASA definition for "Flutter," and that it has also approved the Receivers definition. The committee then turned its attention to a further consideration of the Proposed Standards on Receivers: Definitions of Terms.

Under the Chairmanship of A. G. Fox, the **Antennas and Waveguides** Committee met on April 15. Detailed consideration of the proposed Standard on Wave Propagation, 51 IRE 24.1 PS1 was given by the Committee. Comments and criticisms received from members of the Committee were reviewed for the benefit of H. O. Peterson representative of the Wave Propagation Committee. The Committee agreed to revise the two definitions of Q given in 51 IRE 2. PS1. A discussion was given to Cymomotive Force but the Committee was unable to favor the use of such a term until more data on the proposal could be obtained.

The **Circuits** Committee convened on May 2, under the Chairmanship of C. H. Page. The revised feedback definitions were discussed, and it was decided that the basic term of the feedback set is "feedback loop" Two members of the committee were asked to submit definitions incorporating their concepts of feedback. The use of "port" as a general term for input or output access to a network was discussed. R. M. Foster, W. A. Lynch, and H. J. Carlin were asked to make a study of the advantages and ramifications of the use of this term and to write an essay on the subject to be circulated to all technical committee chairmen for their reaction. The list of active circuit definitions of Subcommittee 4.7 is finished, except for the feedback terms, and is to take the grand tour in the fall.

On April 4, the **Electron Devices** Committee met under the Chairmanship of A. L. Samuel. Chairman G. A. Espersen reported that his Magnetron Subcommittee was writing definitions and methods of test for Q, circuit efficiency, pulling figure, etc. They had proposed to collaborate with the klystron and traveling-wave task groups on common methods of test but had difficulty in arranging this. Chairman T. J. Henry of the Small-Signal High-Vacuum Tubes Subcommittee reported that their only activity was the task groups on klystron and traveling-wave tubes. E. M. Boone, reporting for the combined task group on uhf definitions, submitted some general definitions



Executive Committee, IRE Syracuse Section, D. C. Pinkerton, Chairman of the Syracuse Section, presides at a monthly meeting of the local section steering body. (left to right) W. H. Hall, A. D. Haedecke, L. R. Fink, D. C. Pinkerton, J. W. Downie, R. A. Galbraith, and Samuel Seeley.

proposed by this group. Chairman H. L. Thorson of the Gas Tubes Subcommittee reported no present activity but suggested that the development of the plasmatron might require future activity. C. E. Fay pointed out that T-R tubes had originally been assigned to the POHVT Subcommittee but since most of these tubes contain gas, they might be more appropriately in the domain of the Gas Tube Subcommittee. The Electron Devices Committee agreed to this transfer and suggested that a task group of the Gas Tube Subcommittee be formed to deal with T-R tubes. The proposed Standards on Noise Measurements which were reviewed and revised were approved, and the secretary was instructed to transmit this material and that of the Noise Definitions, which were approved on January 18, 1952, to the Standards Committee.

The **Sound Recording and Reproducing** Committee met on April 23, under the Chairmanship of H. E. Roys. Mr. Roys announced the new Chairman for 1952–1953, A. W. Friend, and the new Vice Chairman Lincoln Thompson. Dr. Friend announced that A. P. G. Peterson had accepted temporary Chairmanship of the Subcommittee on Magnetic Recording. Lincoln Thompson will continue as Chairman of the Mechanical Recording Subcommittee and R. M. Fraser as Chairman of the Optical Recording Subcommittee. The Chairman announced that the work of the Task Force on Definitions of Magnetic Recording in Computer Work would cease since a subcommittee of this nature was being set up by the Electronic Computers Committee. Likewise, the Task Group on Symbols, 19.2.1, would be discontinued. The Committee believes that the tentative Standard on Methods of Measurement of Noise in Sound Recording and Reproducing Systems, prepared by the Subcommittee on Magnetic Recording, is now ready for submission to the Standards Committee. Lincoln Thompson reported that his Subcommittee on Mechanical Recording was preparing a draft on record calibrations and may have something to submit to the Committee by the end of the year. CCIR is preparing standards on disks and magnetic recording for the international exchange of program material. One of the big problems is the standardization of recording reproducing frequency response characteristics for which a method of measurement is needed. The Chairman pointed out that the Magnetic and Mechanical Recording Subcommittees could greatly benefit the U. S. position by working on this problem.

Under the Chairmanship of W. J. Poch the **Video Techniques** Committee met on April 15. Stephen Doba, Jr. presented a report on measurement of nonlinear characteristics involved in a television system. It was agreed that this report should be circulated to Committee members and also to Subcommittee members. A. J. Baracket reviewed the current activities of his Subcommittee on Video Systems. Activities of the Subcommittee on Video Transmission were presented by Stephen Doba, Jr. and J. L. Jones. R. L. Garman outlined the results of a recent meeting of his Subcommittee on Video Utilization. A revised tutorial paper prepared by L. D. Grignon is ready for re-submission, with approval of the Video Techniques Committee. Mr. Schade also has been asked to prepare a paper in this series with special emphasis on the cathode-ray tube aspects on the recording problem. The Committee reviewed definitions submitted by the Receivers Committee and approved by the Standards Committee.

Appointments to the **Joint Technical Advisory** Committee for the two-year period, July 1, 1952–June 30, 1954, were recently approved. Beginning July 1, 1952, the JTAC will consist of the following members: Ralph Bown, *Chairman;* A. V. Loughren, *Vice Chairmen;* L. V. Berkner; D. G. Fink; J. V. L. Hogan; D. D. Israel; I. J. Kaar; and P. F. Siling. L. G. Cumming will serve as non-member Secretary.

Radar Conference Slated

The Third Annual Radar Weather Conference will be held, September 15–17, 1952, at McGill University, Montreal, Canada. The conference will provide symposia on radar meteorology and related studies, and informal discussions of recent advances.

Those interested in participating should communicate with the Committee for 1952 Radar Weather Conference, McGill University, Montreal 2, Canada.

Professional Group News

Professional Groups Committee

D. D. Israel has been appointed Chairman of a Professional Groups Committee Subcommittee to evaluate an experimental plan for the year 1952, for the administration and financing of Professional Group chapters. This Subcommittee will report on September 1, 1952 and January 1, 1953, on Professional Group-Section relationship.

A plan is being formulated to offer technical libraries in municipalities and in technical schools a minimum number of Professional Groups' TRANSACTIONS at a flat rate.

Audio

F. X. Byrnes and M. E. Brady are the organizers of a Chapter of the Audio Group in San Diego, Calif.

The Audio Group is planning to publish the papers on Audio given at the 4th Southwestern IRE Conference and Radio Engineering show, May, 1952, in Houston, Tex. The PGA-8 TRANSACTIONS will be mailed to members of the Group in July.

Antennas and Propagation

The Proceedings of the URSI-IRE Spring meeting April 21–24, 1952, Washington, D. C., jointly sponsored by the Antennas and Propagation Group, will be published in the Group's TRANSACTIONS.

A planning committee recently met in Washington, D. C., to discuss the joint Meeting on Radio-Meteorology to be held in 1953, at Austin, Tex. The Group on Antennas and Propagation will be a sponsor to the meeting.

Circuit Theory

Officers of the Circuit Theory Group have been elected to serve for the term July 1, 1952 to July 1, 1953. They are: *Chairman*, R. L. Dietzold; *Vice-Chairman*, C. H. Page; West Coast Chairman, D. L. Trautman; *Secretary-Treasurer*, W. H. Huggins; Symposium Committee, H. J. Carlin Chairman, W. H. Huggins, C. H. Page; Sections Committee Chairman, J. J. Gershon; Papers Committee, W. N. Tuttle, Chairman, J. L. Bower, C. H. Page, L. Zadeh; New Members of the Administrative Committee, W. H. Huggins, J. L. Barnes, H. J. Carlin.

The Group in conjunction with its Los Angeles Chapter will present a session on broadband amplifiers and a session on general advances in network theory at the IRE Western Convention. The Circuit Theory Group has voted to assess its members $2.00 to enable them to publish the papers of these sessions in the TRANSACTIONS of the Group.

Plans are also being made by the Circuit Theory Group to jointly sponsor the next symposium on network synthesis, with the Polytechnic Institute of Brooklyn, Microwave Research Institute. It is tentatively planned to cover varying parameter and nonlinear circuit theory at the symposium.

Industrial Electronics

The Group on Industrial Electronics will publish in its first TRANSACTIONS the papers of the recent symposium on "Electronics and Machines," held in Chicago, Ill.

Electronic Computers

Proceedings of the Electronic Computer Symposium "Engineering Tomorrow's Computers," held at the University of California, Los Angeles, Calif., will be published and distributed to members of the Electronic Computers Group. Other persons desiring a copy should send $2.00 to George Gourrich, 12830 Parkyns Street, Los Angeles 49, Calif. The Proceedings will include the papers and transcriptions of the two panel discussions on reliability and germanium bodies.

The Group is planning its sessions for the joint IRE-AIEE Computers Conference, December, 1952, New York, N. Y. Papers of the Conference will be published in the Group's first TRANSACTIONS.

W. Buchholtz of the IBM Engineering Laboratory, Poughkeepsie, N. Y., is the new Chairman of the Publications Committee.

Airborne Electronics

The papers from the Airborne Electronics Conference, May 12–14, 1952, in Dayton, Ohio, will be published in the PGAE-4 TRANSACTIONS of the Group.

A local chapter of the Airborne Electronics Group has been established in Baltimore, Md. Plans are being made to establish chapters in Syracuse, Philadelphia, Washington, D. C., Kansas City, Dallas, and other cities where there is sufficient interest.

Communications

The recently formed Group on Communications held a successful Symposium on, "Radio Telegraph Transmission Systems," at Brentwood, N. Y., June 21, 1952.

Plans for a membership drive for the Group are now under way.

Engineering Management

The Group on Engineering Management is planning a symposium in conjunction with the 6th Annual Conference on Administration of Research at the Georgia Institute of Technology, September 8–10, 1952, Atlanta, Ga. The Group is planning a second Symposium to be held jointly with the National Electronics Conference, October, 1952, in Chicago, Ill.

Electron Devices

Papers from the technical sessions on Electron Devices, given at the Fall Meeting of IRE-RTMA, October, 1952, in Syracuse, N. Y., and the 1952 IRE Western Convention in August, will be published in the first TRANSACTIONS of this Group.

Vehicular Communications

Papers for the Fall meeting of the Vehicular Communications Group should be submitted to the Chairman of the Papers Committee, A. A. MacDonald, Westinghouse Electric Corp., 2519 Wilkens Ave., Baltimore, Md. The time and place of the meeting will be announced.

PGVC-2 TRANSACTIONS will be mailed to members of the Group in July.

Quality Control

The Group on Quality Control has appointed the following committees: Papers and Speakers Procurement Committee, Papers Review Committee, Publication and Publicity Committee, Professional Relations Committee, Meeting and Symposium Committee, Membership Committee, and Finance Committee.

The Group plans to publish in its first TRANSACTIONS the papers on Quality Control from the 1952 IRE National Convention technical sessions and the Toronto Fall Meeting.

Information Theory

A symposium on, "Applications of Communication Theory," is being planned for the Fall of 1952 for the Information Theory Group.

The Administrative Committee of the Information Theory Group has been elected for the next three-year term. They are: M. J. DiToro, Federal Communications Labs., Nutley, N. J.; Meyer Leifer, Sylvania Electric Products Inc., Bayside, N. Y.; and W. D. White, Airborne Instruments Laboratory, Mineola, N. Y.

This group is planning their first publication of TRANSACTIONS in the near future.

Microwave Electronics

J. G. McCann of California has been elected to the Administrative Committee and Harald Schutz has been appointed Chairman of the Membership Committee of the recently formed Microwave Electronics Group.

New plans for the Group include consideration of a Group chapter in Dayton, Ohio, and a one day symposium in New York City, during the middle of October, or November, 1952. A. C. Beck has been appointed chairman of the symposium. The Group is also considering the publication of a NEWSLETTER and a TRANSACTIONS later this year.

Instrumentation

Over 1,000 persons attended the Symposium on Quality Electronic Components, jointly sponsored by the Group on Instrumentation, RTMA, and AIEE, May 5–7, 1952, in Washington, D. C. Proceedings of this meeting will be published by the middle of July and can be obtained through the Washington, D. C. office of RTMA.

IRE Western Convention

Eighteen technical sessions at the IRE Western Convention and Electronics Show will be sponsored by the following Professional Groups: Audio, Airborne Electronics, Antennas and Propagation, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Electronic Computers, Electron Devices, Information Theory, Instrumentation, Radio Telemetry and Remote Control.

Over 1,100 Attend Electronic Components Sessions

More than 1,100 engineers and technical authorities participated in the second Symposium on Progress in Quality Electronic Components, sponsored by IRE, AIEE, and RTMA, held May 5–7, 1952, in Washington, D. C. This attendance surpassed all previous records. More than 40 technical papers were presented by leading authorities from government, industry, and research laboratories.

The meeting was opened with a welcoming address by J. G. Reid, Jr., chairman of the Joint Symposium Committee. Glen McDaniel, president of RTMA, delivered the keynote address of the initial session, "Electronics Today." Among other outstanding speakers and subjects presented were: J. A. Milling, chairman of the Electronics Production Board, on "Electronics in the Defense Production Program"; Captain Rawson Bennett, USN, Bureau of Ships, former director of the Electronics Production Resources Agency on "Some Factors in Today's Electronics Production"; Lieutenant Colonel C. B. Lindstrand, USAF, Electronics Production Resources Agency, on the "European Electronic Components Industry"; and, W. A. G. Dummer, Telecommunications Research Establishment, London, Eng., on "Electronic Components in Great Britain."

Copies of the addresses given during the initial and only general session have been mailed to RTMA member-companies. Copies of the complete proceedings, including technical papers, may be obtained from RTMA headquarters for $5.00 each, when the printing is completed in July.

IRE Convention Symposium

Symposium: TV Station Construction and Theater Conversion, sponsored by the Broadcast Transmission Systems Group during the IRE National Convention in March. (left to right) J. G. Leitch, WCAU, Philadelphia; Carlos Dodd, WFAA-TV, Dallas, Tex., Newland Smith, WOR-TV; A. B. Chamberlain, CBS; A. A. Walsh, NBC; Clure Owen, ABC; and (standing rear) R. F. Guy, NBC.

New IRE Section Formed in Hamilton, Ontario

The IRE Board of Directors approved establishment of the Hamilton, Ontario, Canada, IRE Section of Region 8 at its May meeting. Formerly a Subsection of Toronto, Hamilton brings the total number of IRE Sections to sixty-three.

Also, some changes were made in the Subsection group. A Wichita Subsection of the Kansas City Section (Region 5) was established, and the Urbana Subsection of the Chicago Section (Region 5) was dissolved. The present total of Subsections is fourteen.

Addition of a Student Branch, operated jointly with the AIEE at Brown University, brought the Student Branch total to one hundred and eleven.

A list of IRE Sections and Subsections, together with respective Chairmen and Secretaries, appears in alternate issues of Proceedings of the I.R.E. This issue contains a new amended list on pages 881 and 882.

IRE Professional Group on Audio

Chairman's Final Report, 1951–1952

Benjamin B. Bauer, Chairman

The Professional Group on Audio has been extremely active and has initiated a number of procedures of value. The Final Report of its Chairman for 1951–1952 is here presented in the thought that the membership will be interested in the progress of this Group and that other Groups may find material of interest to them in this report —*The Editor.*

The progress of the IRE Professional Group on Audio during the past year has been threefold: (1) Increase in publications and services to members; (2) Promotion of individual participation through active organization of local chapters; (3) Increase in the active membership. Equally important, though less well publicized, has been the latent energy expended by the officers and by the Administrative Committee to establish Group policies and procedures in harmony with the IRE objectives; to select and appoint capable committees to perform in accordance with these procedures; and to provide for continued development and growth of the PGA.

Among the first actions evident to the members was the issue of improved Newsletters, beginning with the July, 1951 issue, dressed in a distinctly "IRE" cover page. Among the features in these Newsletters have been the technical editorials contributed especially for the Professional Group on Audio by foremost scientists in audio and acoustics. These editorials have received the unqualified approval of PGA members. Requests for republication of some of them have already been granted by the IRE to foreign publications.

A large share of the credit for the success of PGA publications goes to our Regional Editors. Jordan J. Baruch, our Eastern Editor and Chairman of the PGA Publications Review Committee has contributed a number of papers and book reviews. Dan Martin, Midwestern Editor, and Vincent Salmon, Western Editor, have contributed immeasurably to sound publication policy, and each has written excellent technical editorials.

Our total volume of publications has fallen somewhat short of the more optimistic expectations, initially because of lack of funds, and subsequently because of difficulties in procuring suitable material. Nevertheless, including the March issue of Transactions of the IRE-PGA (which has recently been merged with the Newsletter), PGA members will have received during the year approximately 160 pages of material, containing eight technical editorials and nine technical papers. With the announcement of the new PGA republication procedure in the current issue of Transactions, it is expected that the amount of available material will increase substantially during the coming year.

This new republication procedure is among the most important results of intensive and cordial co-operative effort between the Editorial Committee of the PGA and the Editor of the Institute and its PROCEEDINGS. Briefly speaking, this procedure will permit the PGA to make available for republication in non-IRE technical journals all articles not judged to contain subjects of compelling interest to a wide segment of IRE membership; many of these papers contain important contributions along specialized lines of audio technology. Mechanism has been set up to provide liaison between the authors of these papers and the publishers to facilitate arrangements for republication in technical and trade journals. While we have not had adequate experience with this procedure, initial reactions of all the parties concerned have been favorable.

It is evident that the keystone committees to implement this republication procedure are the Papers Procurement Committee under the chairmanship of Leo L. Beranek, and the Papers Study Committee, under the Chairmanship of Hugh S. Knowles since January, 1952. As a result of the efforts of Dr. Beranek's committee, some of the papers presented at the IRE Spring Convention were transmitted to members within two weeks after presentation, and a repetition of this performance should become routine matter in the future. Hugh Knowles brings to the Papers Study Committee the degree of expertness, impartiality, and judgment which is the "sine-qua-non" for this committee to make its recommendations valuable to the authors and to the PROCEEDINGS OF THE I.R.E.

Another PGA sponsored activity has been the presentation of recorded papers. Andrew B. Jacobsen has been the leading exponent of this idea, and the PGA is happy to have espoused this activity. A recorded paper procured by Mr. Jacobsen at the 1951 Fall West Coast IRE Meeting has already been presented at the Seattle, Milwaukee, and Chicago PGA Chapters, and in each case it was extremely successful. Plans were made to record the technical audio sessions at the 1952 IRE National Convention.

During the past year, program committees of the Group have sponsored the audio sessions at three national conventions, as follows: 1951 Chicago Electronics Conference, Chairman, Benjamin B. Bauer; 1951 Radio Fall Meeting in Canada, Chairman, Frank Slaymaker; 1952 IRE National Convention in New York, Chairman, W. S. Bachman.

PGA Chapters have been active in sponsoring programs on audio among IRE Sections, the details of which have been publicized in past issues of the NEWSLETTER.

Another manifest activity has been a concerted drive to sponsor the formation of PGA Chapters, carried on by the Chapters Committee under the Chairmanship of Stan Almas. Intense correspondence has been carried on with the sixty-odd IRE Sections and Subsections in an effort to promote the formation of chapters. At the present writing, the PGA has six known active chapters in Boston, Mass.; Chicago, Ill.; Cincinnati, Ohio; Detroit, Mich.; Milwaukee, Wis.; and Seattle, Wash. Several other Sections have expressed interest in chapter formation.

During the past year we have acquired a substantial active membership. Before the billing of assessments, PGA listed approximately 1,000 names. Approximately 400 of this number became active members at the time the assessment was levied, and an additional 400 members joined between July, 1951 and January, 1952, giving us a total active membership of approximately 800. We believe that our membership will continue to grow at about the same rate for some time to come, and that some of the presently inactive members may become active as a result of the continued efforts of the PGA Membership Committee, under the Chairmanship of A. M. Wiggins. Recently, approval was obtained to permit the enrollment of Student Members, and we trust that many of them will take this opportunity.

During the past year, a major task of the Administrative Committee was to establish a sound financial policy. Last year, IRE Headquarters contributed $1,000 in matching funds to help pay direct costs of Group operation, and we have been advised that this year $500 will be similarly contributed. Next year, the matching funds may be withdrawn altogether. The necessity of raising funds by means other than increased assessments seemed to be imminent, and several means of raising additional revenue were studied and submitted for the approval of the Professional Groups Committee. From this work has emerged the present Institutional Listing scheme, in which six firms are already participating. Promotion of Institutional Listings is administered by the PGA Ways and Means Committee, under the chairmanship of John Hilliard.

To mention some additional tasks satisfactorily completed: the Administrative Committee has developed the policy for accepting "Situations Wanted" ads at a nominal cost to members; several methods were studied for prorating assessments of members joining after the billing date, with the final adoption of a system proposed by the IRE Executive Secretary G. W. Bailey; assent of Headquarters was obtained to act as bursar for the Group, thus relieving the Secretary-Treasurer of a major banking chore; provisions were made for nominal aid to Sections wishing to form PGA Chapters; action has been initiated to increase the Administrative Committee from six to nine; several written procedures for the transaction of the business of the Group have been prepared.

This report would not be complete without recognition and appreciation of the effort of many friends and associates. We wish to mention especially W. R. G. Baker, George W. Bailey, L. G. Cumming, E. K. Gannett, A. N. Goldsmith, and Miss Emily Sirjane for their co-operation and counsel. Last but not least, our sincerest thanks are due to our Secretary, Miss Henelee Goldman of Shure Brothers, Inc., for her efficiency in handling the incredible amounts of composition and correspondence arising from our PGA activities.

IRE People

Arnold A. Cohen (A'42–M'47–SM'51) has been appointed director of telecomputing systems development of Engineering Research Associates, Inc., St. Paul, Minn. In his new capacity, he will direct the development of special purpose digital computers, and the application of digital techniques to control, instrumentation, and communications problems.

Dr. Cohen was born in 1914, in Minnesota, and received the B.E.E. degree in 1935, and the M.S. and Ph.D. degrees in physics, in 1938 and 1947, respectively, from the University of Minnesota.

Dr. Cohen was associated with RCA, as an engineer in gas tube and phototube development, from 1942–1946, at which time he joined the Engineering Research Associates, Inc. With this company, he formerly worked as a project engineer in high-speed electronic computing, magnetic recording, and ultrasonics.

Dr. Cohen is a member of the Association for Computing Machinery, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

Lewis E. Pett (M'52) has been appointed western district manager for the television transmitter division of the Allen B. Du Mont Laboratories, Inc. Mr. Pett has been with the Du Mont Laboratories since 1944.

Mr. Pett received the B.S. degree in electrical engineering at Columbia University in 1924, and studied at Harvard for two years. He has been associated with the radio and TV industries since 1926.

Charles B. Raybuck (A'43) has been appointed chief engineer of Melpar, Inc., Alexandria, Va.

C. B. RAYBUCK

Mr. Raybuck was born in Chicago, Ill., in 1917, and is a graduate of the Armour Institute of Chicago and the University of Maryland, where he specialized in microwave theory. From 1935–1941, he was with the Belmont Radio Corporation working on design of radio test equipment and special projects. Before joining Melpar, Inc., Mr. Raybuck was chief engineer of the Maryland Electronics Company.

Ivan S. Coggeshall (A'26–M'29–F'42) has been promoted to director of planning of Western Union Telegraph Company's International communications.

I. S. Coggeshall

Dr. Coggeshall was born in Newport, R. I., in 1896, and attended Worcester Polytechnic Institute, where he later received an honorary doctor of engineering degree. He was an early wireless amateur and a telegraph operator with Postal Telegraph-Cable Company from 1912–1917, when he joined the Western Union Telegraph Company as an engineering apprentice, at Boston, Mass. Transferring to Western Union's New York headquarters in 1920, Dr. Coggeshall has graduated to various executive positions since that time. In 1946, he was made general traffic manager of the submarine cable system.

Dr. Coggeshall was on active duty with the United States Navy during both World Wars. In World War II, he was the representative for Western Union on the cable committee of the Board of War Communications, and had active duty assignments under the Director of Naval Communications in Washington, D. C.

The Junior Past President of the Institute, Dr. Coggeshall has been active on numerous IRE Committees. He also has been closely associated in the communication division of the American Institute of Electrical Engineers and is an AIEE fellow.

❖

William G. Tuller (S'37–A'40–M'45–SM'48) has been elected vice president in charge of engineering of Melpar, Inc., Alexandria, Va. He was formerly the director of engineering of that company.

W. G. Tuller

Dr. Tuller was born on September 8, 1918, in Rutherford N. J. From 1939–1944, he was a staff member of the electrical engineering department of the Massachusetts Institute of Technology, attached to the ultra-high-frequency laboratory, and later became a member of the radiation laboratory. From 1941–1947, he worked for the Raytheon Manufacturing Company as development engineer and consultant. He received the Sc.D. degree from MIT in 1947, and has been with Melpar, Inc. since that year.

Dr. Tuller is a member of the IRE Standards Committee, and Chairman of the IRE Technical Committee on Information Theory and Modulation Systems. He is chairman of the subpanel on packaged subassemblies of the Research and Development Board for the Armed Forces, and a member of the Research and Development Board Panel on Components.

Dr. Tuller has been a contributor to the Proceedings and is a member of Sigma Xi.

Sir Robert Watson-Watt (SM'46–F'47) has accepted a directorship on the Board of Canadian Aviation Electronics Ltd., and has become a member of their executive committee. Also, recently, he was named top consultant on radar problems to Canada's Defense Research Board.

Sir Robert Watson-Watt

Sir Robert was presented £50,000, this year, by Great Britain's Royal Commission on Awards to Inventors for his contribution to the development of radar. He is the recipient of many other honors and awards and has held many notable positions with the British Government in the capacity of scientific advisor and consultant.

❖

Arthur C. Weid (M'44), formerly chief engineer of Melpar, Inc., has been named executive assistant to the executive vice president of Melpar, Inc., Alexandria, Va.

A. C. Weid

Mr. Weid was born in New York, N. Y., in 1916, and received the B.S. degree in electrical engineering from Alabama Polytechnic Institute, and the M.Sc. degree in physics from New York University. He was an instructor in physics in 1938–1939, and an instructor in mathematics in 1939–1942, at the Alabama Polytechnic Institute. During World War II, Mr. Weid was a project engineer with the Division of War Research of Columbia University. He also was a project engineer with the Airborne Instruments Laboratory, specializing on uhf measurements and antenna patterns, before he joined Melpar in 1947.

❖

Ira Kamen (M'48) has been appointed vice president of the manufacturing corporation division of the General Bronze Corporation. He will be responsible for organizing the industrial and government sales operations of that company.

Ira Kamen

Mr Kamen was born in New York City and studied at the New York University College of Engineering and the College of the City of New York. He began his career as a development engineer with Dictograph Products. In 1941 he was a project engineer with the Kurman Electric Corporation, and then served as a supervisor professional radio engineer with the United States Navy Electronics Office. He also has been associated with the Intra-Video Corporation as a general manager, and with RCA directing the sales and engineering of RCA, antenaplex systems.

Mr. Kamen is the author of three books on television and technical articles.

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Edmund G. Shower (A'38–SM'46) has been appointed head of the transistor division of the National Union Radio Corporation. Prior to his recent assignment, he served many years with the Bell Telephone Laboratories, where he was in charge of transistor production.

E. G. Shower

Mr. Shower was born in 1903, in Baltimore, Md., and received the B.E. degree in electrical engineering from Johns Hopkins University, in 1925. Following this, he did graduate work in physics at Columbia.

On leave of absence from Bell Telephone Laboratories, Mr. Shower served with the United States Navy, from 1943–1946, as a member of the electronics division of the Bureau of Ships. He was placed in charge of co-ordination of electron tube design for the naval establishments with the Army, Air Force, Marines, and allied agencies.

Mr. Shower is the author of a number of papers in scientific publications and holds four patents on electronic tubes and circuits.

❖

Fred O. Grimwood (A'46–SM'49) vice-president and sales manager of the Gates Radio Company, died recently in Quincy, Ill. He was 41 years of age.

Mr. Grimwood, who was a native of Indiana, studied engineering management at Evansville College, Ind., and began his professional career as chief engineer of radio station WFIW in Hopkinsville, Ky. Later, he became a consulting radio engineer in Evansville, Ind., and Washington, D. C. He was the former owner of station KLCN in Blytheville, Ark., and station WTOM, in Bloomington, Ind. Mr. Grimwood was the chief engineer of the Gates Radio Company from 1942–1948, when he became vice-president of the company.

Mr. Grimwood had supervised the manufacture and design of much radio equipment for foreign governments and designed much equipment for the United States Signal Corps. He was licensed to practice before the Federal Communications Commission, and was one of the first engineers in the United States to conduct what is known as field measurement of radio broadcasting stations, now recognized as the standard measurement of coverage in the broadcasting industry.

Mr. Grimwood was a registered professional engineer in Illinois, and a member of the National Association of Radio and Television Broadcasters.

Books

The Oxide-Coated Cathode by G. Herrmann and S. Wagener

Published (1951) by Chapman & Hall Ltd., 37 Essex St., London W.C. 2, Eng. 298 pages+4-page appendix+2-page bibliography+7-page index. 154 figures. 5½×8½. Price, 42s.

G. Herrmann is an electrical engineer, Berlin, Germany, and S. Wagener is a member of the Post Office Research Station, Dollis Hill, London, Eng.

This is an English edition of a work first published in 1944 in Germany. It is not simply a translation of the original book; revisions have been made and material added that is based on literature available to about the end of 1949.

The chapter headings are as follows: (1) The thermal emission of electrons from metals; (2) Methods of measuring the work function of metals; (3) Phenomena in ionic solids; (4) The mechanism of the emission from an activated oxide coating in equilibrium; (5) Oxide coatings of different composition; and, (6) Variations in the equilibrium of the oxide coating.

It will be noticed that the title is a bit misleading since only the last half of the book is devoted to the oxide-coated cathode. The authors stated that inclusion of the first three chapters was needed to provide a better understanding of the material presented in the last half of the book. This is true, at least, in regard to chapter 3.

In the first two chapters, thermionic emission from metals is discussed in the conventional manner, beginning with the Fermi-Dirac energy distribution of the conduction electrons. Sections are included on the space-charge-limited current, current in a retarding field, the effects of absorbed impurities, measurement of fundamental quantities.

Chapter 3 is especially interesting as it provides most of the material essential to an understanding of conduction in, and thermionic emission from, ionic solids and semiconductors. To a degree, this chapter reminds one of Mott and Gurney's book, emphasis here being on those parts which contribute particularly to clarification of the energy-band picture. (A requirement if one is to understand the remaining material.)

While the treatment employed cannot be criticized for lack of accuracy, a greater service might have been performed by the authors if they had developed the band picture as it is commonly done now. I refer, of course, to the scheme introduced some years ago by Slater to show the development of energy bands and their overlapping as related to interatomic distance.

In chapter 3, the equation for thermionic emission from the usual model of the excess semiconductor is derived. The derivation follows the general method of Blewett based on the concepts discussed in the previous chapter. The terminology and symbols employed are not those most commonly used in this country. This is naturally no fault of the authors, but the annoyance is often great enough to make one hope that efforts at standardization will eventually be extended beyond present aims and geographical boundaries.

A rather short chapter 5 discusses the effects of variations in electrical results when the chemical composition of the coating is varied, including the addition of materials other than the alkaline earth oxides. The experimental results of numerous workers are reviewed.

The final chapter presents a good discussion of the activation process and the influence of the base-metal composition on activation and the formation of interface layers. Important effects such as the surface evaporation of barium, poisoning, and diffusion are covered in sufficient detail for most readers.

In a general over-all appraisal of the book, one is apt at first to be rather critical of its omissions. For example, pore conduction is not mentioned and the marked effect of a small amount of silicon in the base material, as compared with larger amounts of other reducing materials, was not pointed out. On the other hand the rate at which mysteries, in regard to the coated cathode, are being cleared up is so rapid that two or three years are enough to produce significant changes in the state of this branch of electronic science, and no book on such a lively topic can be up to date when published. In conclusion, this book warrants careful reading by all concerned with electron emission, either from a theoretical or an engineering point of view.

George D. O'Neill
Sylvania Electric Products Inc.
Bayside, N. Y.

Materials Technology for Electron Tubes by Walter H. Kohl

Published (1951) by the Reinhold Publishing Corporation, 330 West 42 St., New York, N. Y. 474 pages+19-page index+xv pages. 217 figures. 6×9. $10.00.

Walter H. Kohl is the consultant to the director of research, Collins Radio Company, Cedar Rapids, Iowa.

This book, which includes many useful tables, gives much information on standard materials used as envelopes, structural numbers, and seals in sealed-off electron vacuum tubes.

Approximately a third of the book is devoted to six chapters on various glasses, their compositions and properties; while another six chapters discusses the common metals used in these glasses: w, mo, cu, ni, and c. The information and tables are carried out to a greater extent than is normally needed for the common problems faced in the design and fabrication of electron tubes however, much enlightening and applicable information is also presented on soldering and brazing, ceramics and mica, and ceramics to metal seals. Perhaps for the first time, such materials of electron vacuum-tube construction are fully covered in a book whose principal concern is with these materials.

The chapters on high-vacuum techniques and thermionic emission contain many practical aspects on such problems as the speed of pumps and the fabrication of cathodes. Another two chapters which are concerned with electron structure of matter and the phase rule are interesting to read, but could have been excluded without detracting from the value or quality of the book. The chapter on phase rule concludes with phase diagrams of a glass system. The chapter on electrons, atoms, crystals, and solids attempts to cover too much basic quantum physics and not enough chemistry. Nevertheless, it is hoped that these chapters will stimulate the reader to investigate the references listed which introduce these subjects in a more fundamental way.

Each chapter is reviewed by at least one person who is experienced in that particular field, thus bringing us up to date on the best current practices. Well written, this handbook gives the technical bases for certain procedures followed, enabling one to acquire some physical intuition and understanding for doing certain things in a certain way. For example, in introducing strain analysis of glass, a discussion is first presented on the fundamental of polarization of light by birefringent crystal, such as identifying the ordinary and extraordinary rays, the optical axis, etc.

This book will be a profitable addition in every laboratory with interests in fabrication, design, or construction of vacuum tubes.

E. R. Piore
Office of Naval Research
Washington 25, D. C.

Fundamentals of Radio Communications by Abraham Sheingold

Published (1951) by the D. Van Nostrand Company, Inc., 250 Fourth Avenue, New York 3, N. Y. 431 pages+11-page index+v pages. 333 figures. 6×9.

Abraham Sheingold is an associate professor of electronics United States Naval Postgraduate School.

This is the first of a new series of text and reference books in communications engineering, which is being edited by H. J. Reich, Dunham Laboratory, Yale School of Engineering. The aim of the text is to provide a basic understanding of principles and techniques employed in currently used systems. Therefore, elementary concepts and theory are treated in detail, while brief and qualitative description of systems occupy less than 15 per cent of the 430 pages of the text.

The material is presented at an intermediate level, and differential equations are entirely avoided. The complex number concept, its vector representation and the necessary elements of vector analysis, are discussed. It will be found that algebra and trigonometry are prerequisite, as are an elementary knowledge of electricity and magnetism. Quantitative circuit theory is adequately presented with the usual problems of radio circuit analysis included for the student. The newer statistical approach to communication theory is appropriately omitted.

Since this is the first edition, one can expect to find an occasional error, such as the interchange of R_1 and R_2 (on page 203); however, this book must not be criticized on that account. The concepts of the book are presented with a rare combination of lucidity and economy of words which should win for it a high standing among textbooks in its field.

R. M. Page
Naval Research Laboratory
Washington, D. C.

Sections*

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The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February 1951, through January 1952, may be obtained for 2s8d. 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

016:534 1487

References to Contemporary Papers on Acoustics—R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 724–730; November, 1951.) Continuation of 293 of March.

534.121.2.001.362 1488

Transformer Analogs of Diaphragms—B. B. Bauer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 680–683; November, 1951.) The action of a diaphragm is shown to be analogous to that of a system of ideal transformers, each corresponding to a particular area of the diaphragm. Equivalents for various types of diaphragm are described.

534.152 1489

Demonstration of Standing Waves in the Free Acoustic Field, and a Simple Receiver for Short Acoustic Waves—R. W. Pohl. (*Naturwiss.*, vol. 38, pp. 486–490; November, 1951.) Two methods of recording a standing-wave pattern are described. A shadow picture is obtained of (*a*) the disturbance of a liquid surface (water or gasoline) over which the sound wave is projected, or (*b*) the turbidity in a soap film arranged obliquely in the sound field. Application of the first method in acoustic measurements is illustrated.

534.21-14 1490

Normal Mode Propagation in Three-Layered Liquid Half-Space by Ray Theory—C. B. Officer, Jr. (*Geophys.*, vol. 16, pp. 207–212; April, 1951.) The integral expression for the field due to a point source is derived by summation of multiple reflections of waves from a plane-wave source. Its physical significance is discussed.

534.21-14+621.3].001.362 1491

The Formal Connection between the Corpuscular Theory of Sound Propagation in Liquids and Problems of Electrical Engineering, particularly the Theory of Quadripoles—K. Altenburg. (*Frequenz*, vol. 5, pp. 285–289; October, 1951.)

534.213.4 1492

Propagation of Sound in a Duct with Constrictions—U. Ingård and D. Pridmore-Brown. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 689–694; November, 1951.) Iris partitions are uniformly distributed along the duct. Attenuation is determined as a function of frequency for ducts with hard and with absorptive walls, charts being given for hard-wall ducts. By proper choice of wall absorption and iris dimensions and spacing, a wide attenuation band can be obtained. Measured values of attenuation are in good agreement with theory.

534.231:534.26 1493

On the Relation between the Sound Fields Radiated and Diffracted by Plane Obstacles—F. M. Wiener. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 697–700; November, 1951.) The analogy between acoustic diffraction and radiation problems is discussed; a very simple relation exists between a plane rigid scatterer exposed to a perpendicularly incident plane wave and a plane piston radiator. The radiated field is proportional to the scattered field. A similar relation exists between the radiation impedance and the force per unit pressure exerted on the scatterer.

534.232:538.652 1494

Magnetostriction Transducer Measurements—H. J. Round. (*Wireless Eng.*, vol. 29, pp. 101–105; April, 1952.) Description of a simplified method, based on equivalent circuits, for determining the constants of a magnetostriction oscillator.

534.24:534.213.4 1495

An Analysis of the Effect of the Discontinuity in a Bifurcated Circular Guide upon Plane Longitudinal Waves—L. L. Bailin. (*Bur. Stand. Jour. Res.*, vol. 47, pp. 315–335; October, 1951.) A theoretical treatment of the scattering of sound by a semi-infinite tube of small diameter inserted axially into a larger tube of infinite length. The theory is rigorous and explicit provided the waves incident at the discontinuity are restricted to the lowest mode of propagation. The problem is a particular case of the general problem of the effects of obstacles on the propagation of acoustic or em waves in guides.

534.321.9.047:621.431.75 1496

Ultrasonic Spectra emitted by Aircraft Propulsion Mechanisms, and their Physiological Effects—P. Grognot. (*Ann. Télécommun.*, vol. 6, pp. 341–344; November, 1951.)

534.373-13 1497

The Absorption of Sonic and Ultrasonic Waves in Gases—M. Dubois. (*Jour. Phys. Radium*, vol. 12, pp. 876–884; November, 1951.) Theoretical and experimental investigation of the general problem of the attenuation of a plane wave due to the physical properties of the medium and to variations of pressure, temperature, and molecular energy caused by the passage of the wave. 80 references.

534.422:534.321.9.043/.047 1498

A New Improved Type of Ultrasonic Siren—L. Pimonow. (*Ann. Télécommun.*, vol. 6, pp. 337–341; November, 1951.) An earlier design (1822 of 1951) has been improved by truncating the rotor, increasing its speed, and introducing a reflecting surface.

534.522 1499

Theory of Optical Method of Sound Analysis—J. Picht. (*Ann. Phys.* (Lpz.), vol. 5, pp. 117–132 and 1951, vol. 9, pp. 381–400; 1949.) Development of the theory and derivation of formulas applicable to the method described by Schouten (1549 of 1939, 1062 and 1452 of 1940), and modification of the method for combinations of acoustic frequencies with arbitrary phase relations.

534.75 1500

Compression Properties of the Ear—H. Mol. (*Tijdschr. ned. Radiogenoot*, vol. 16, pp. 277–291; November, 1951.) Analysis of the compressive action of the ossicle chain. The large dynamic range of the human ear is ascribed to this.

534.79:534.839 1501

On the Measurement of the Loudness of White Noise—I. Pollack. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 654–657; November, 1951.) A scale of loudness for white noise was obtained by independent subjective methods which show consistency among themselves. It is suggested that this scale forms a better measure for complex sounds than a pure-tone scale.

534.79:534.839 1502

On the Threshold and Loudness of Repeated Bursts of Noise—I. Pollack. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 646–650; November, 1951.) Full paper. Summary noted in 543 of 1951.

534.79:534.839 1503

Sensitivity to Differences in Intensity between Repeated Bursts of Noise—I. Pollack. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 650–653; November, 1951.)

534.833.4-13/-14 1504

Absorption of Sound in Fluids—J. J. Markham, R. T. Beyer, and R. B. Lindsay. (*Rev. Mod. Phys.*, vol. 23, pp. 353–411; October, 1951.) A critical review and unified presentation of the fundamental theories of absorption of sound in gases and liquids, together with an outline of experimental techniques. A summary of the reliable experimental results (mainly at frequencies between 1 mcps and 100 mcps) is included and these are interpreted in terms of the theoretical concepts. 225 references.

534.844.1:621.396.615.11 1505

Equipment for Acoustic Measurements: Part 3—Acoustic Pulse Measurements—Mayo, Beadle, and Wharton. (See 1681.)

534.846.4 1506
Speech Reinforcement in St. Paul's Cathedral—P. H. Parkin and J. H. Taylor. (*Wireless World*, vol. 58, pp. 54–57 and 109–111; February and March, 1952.) Discussion and illustrated description of the system recommended, showing results of acoustic tests. The dome area is served by two 11-ft. vertical arrays of loudspeakers mounted beside the pulpit and the lectern; they comprise eleven 10-in. and nine $3\frac{1}{2}$-in. loudspeakers covering the restricted range of 250–4,000 cps with crossover at 1,000 cps. Six 6-ft. loudspeaker arrays mounted on piers in the nave are fed through a time-delay mechanism consisting of a turntable carrying an 11-in. disk of plastic magnetic material with recording, playback, and erasing heads, an ultrasonic erasing head being provided to guard against failure of the magnetic erasing equipment.

534.85:681.84 1507
New Developments in the Gramophone, World—L. Alons. (*Philips Tech. Rev.*, vol. 13, pp. 134–144; November, 1951.) Properties required of record disks and reproducing apparatus for long playing are discussed; a description is given of Philips apparatus which gives a playing time of $22\frac{1}{2}$ minutes for a microgroove 12-in. record and which also takes ordinary records.

534.861.1 1508
Orchestral Studio Design. Recent Modifications to the B.B.C. Maida Vale Studio—T. Somerville and H. R. Humphreys. (*Wireless World*, vol. 58, pp. 128–131; April, 1952.) The acoustic properties of this studio in its original state were characterized by excessive reverberation time at low frequencies and extreme deadness at high frequencies. Architectural modifications which have led to very large improvements are described; these include the application to the walls of box-type "membrane absorbers" resonant at various low frequencies, and the rebuilding of the orchestra platform.

621.395.61/.62:621.395.625.3 1509
Investigation of Transducers by Repeated and Retrograde Re-recording—W. Meyer-Eppler. (*Fernmeldetech. Z.*, vol. 4, pp. 507–512; November, 1951.) Analysis showing how slight linear distortion occurring in an electro-acoustic transducing system due to transit-time effects may be determined by repeated re-recording. By reversing the direction of motion of the magnetic tape, phase distortion can be totally compensated. Application of the technique in acoustic tests of rooms is outlined.

621.395.623.7 1510
Direct Radiator Loudspeaker Enclosures—H. F. Olson. (*Audio Eng.*, vol. 35, pp. 34–38, 64; November, 1951.) A comprehensive analysis of the effects of various shapes of cabinet shows that for optimum performance the cabinet front must have no sharp edges.

621.395.625.3 1511
Magnetic Sound-Recording—F. Duchâteau (*HF* (Brussels), no. 11, pp. 303–312; 1951.) Discussion of the main technical difficulties involved and of the means devised to overcome them.

621.396.645.029.3 1512
An Ultra-Linear Amplifier—D. Hafler and H. I. Keroes. (*Audio Eng.*, vol. 35, pp. 15–17; November, 1951.) The screen-grid of a tetrode is energized with dc from a low-impedance source through a special winding on an Acrosound TO-300 output transformer, in which the effects of the anode and screen-grid currents are combined. The screen-grid load impedance must be about 18.5 per cent of the anode load impedance. A circuit incorporating this output stage is shown diagrammatically, with component values. The power output of over 20 w is undistorted within 1 db from 20 to 20,000 cps, intermodulation being less than 2 per cent.

ANTENNAS AND TRANSMISSION LINES

621.392.26 1513
The Susceptance of a Thin Iris in Circular Wave Guide with the TM_{01} Mode Incident—K. L. Dunning and R. G. Fellers. (*Jour. Appl. Phys.*, vol. 22, pp. 1316–1320; November, 1951.) "Certain waveguide boundary value problems can be formulated in terms of lumped-constant circuits and distributed-constant transmission lines. Equivalent circuit voltages and currents can be introduced as measures of the transverse electric and magnetic fields. Making use of these concepts and Schwinger's integral equation method, the susceptance of a thin circular iris in circular cylindrical waveguide with the TM_{01} mode incident is discussed and calculated. Results are compared with experimental data."

621.392.26.012.3 1514
$TM_{1,1}$ Waves in Rectangular Waveguides—(*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, p. 32; September, 1951.) Nomogram for determination of cut-off frequency from dimensions of guide.

621.392.43.012.3 1515
Matching-Stub Calculators—S. Yamasita. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 32, 31; August, 1951.) A nomogram for determining the position and length of matching stubs on Lecher-wire lines.

621.396.67 1516
General Theory of Symmetric Biconical Antennas—S. A. Schelkunoff. (*Jour. Appl. Phys.*, vol. 22, pp. 1330–1332; November, 1951.) The input admittance of a biconical antenna of arbitrary angle is expressed as the limit of a certain sequence of functions. The first term of this sequence approaches the exact expressions for input admittance as the cone angle approaches either zero or 90°; hence it probably constitutes a good first approximation for all angles.

621.396.67:621.397.6 1517
Antennas for U.H.F.—E. C. Johnson and J. D. Callaghan. (*FM-TV*, vol. 11, pp. 16–18, 56; November, 1951.) Illustrated descriptions of different types of antenna for television reception, showing their field patterns and gain characteristics.

621.396.677 1518
The Necessary Number of Elements in a Directional Ring Aerial—H. L. Knudsen. (*Jour. Appl., Phys.*, vol. 22, pp. 1299–1306; November, 1951.) Discussion based on the theories of Page (1862 of 1948 and 308 of 1949) and Stenzel (1929 Abstracts, p 450). An expression is derived for the characteristic of an array with a finite number of elements. This expression shows that no uniform improvement of the approximation of the array characteristic to the ideal is obtained by increasing the number of elements, but in this respect odd numbers of elements are better than even numbers.

621.396.677.029.6 1519
Measurement of Radiation of U.S.W. Aerials—J. Delcambe. (*HF* (Brussels), nos. 11 and 12, pp. 293–302 and 327–335; 1951.) If the antenna aperture field distribution is very nearly plane and equiphase, application of the Kottler formulas previously noted [2109 of 1951 (Divoire and Delcambe)] gives numerical values for the directive properties of waveguide, horn, and microwave-lens antennas with an accuracy sufficient to meet practical requirements. Gain can be estimated to within 7 per cent. The measurements methods applied and precautions taken are detailed. An appendix describes a microwave bolometer-type wattmeter developed for the purpose.

621.396.677.5†:621.318.424 1520
Ferromagnetic Loop Antennas—W. J. Polydoroff. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 11–13, 24; November, 1951.) From results on loops with ferrite cores it is concluded that, in their design, (a) a balance must be struck between an acceptable value of Q and maximum effective permeability, a Q of 125–150 being considered most suitable, (b) cylindrical cores of length/diameter ratio >10 give greater effective height; (c) the winding should cover 80 per cent of the core length; (d) the wire, if insulated by vinylite or double cotton covering, may be wound directly on the core; (e) the core should be in the shape of hollow tubing. Various applications where reduction in antenna size is important are suggested. See also 2485 of 1946 (Burgess).

621.396.677.6.029.6:621.396.931/.933].2 1521
Rotating H-Type Adcock Direction-Finders for Metre and Decimetre Wavelengths—H. G. Hopkins and F. Horner. (*Proc. IEE* (London), Part III, vol. 99, pp. 96–97; March, 1952.) Long summary of 598 of April.

CIRCUIT AND CIRCUIT ELEMENTS

621–52:621.396.611.3 1522
Applications of Electrical Methods of Differentiation to Control Problems—Rateau. (See 1688.)

621.3.015.7 1523
A Pulse Mixing Unit—R. R. Rathbone and R. L. Best. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 10, 11, 27; September, 1951.) The unit described accepts pulses from up to eight external lines, mixes them at the input, and delivers them with a delay of 0.08 μs as a single output train. Positive pulses of random form with amplitudes from 6 to 60 v are converted to pulses of half-sine-wave form, of duration 0.1 μs, with amplitudes varying by not more than 5 per cent.

621.3.015.7 1524
The Pulse Standardizer—R. R. Rathbone (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 6–7, 31; November, 1951.) Description of equipment which accepts pulses of random amplitudes (>6v), with repetition rate up to $3+10^6$/sec, and converts them to a set of pulses of uniform amplitude (up to 37v) and with the same recurrence frequency. Pulses with repetition rates up to $5+10^6$/sec can be accepted if a 10 per cent reduction of the maximum output-pulse voltage can be tolerated.

621.3.015.7:621.396.6 1525
Pulse Circuits for the Millimicrosecond Range—F. H. Wells. (*Jour. Brit. I.R.E.*, vol. 11, pp. 491–503; November, 1951.) The circuits described include pulse-shaping circuits, pulse generators, amplifiers, scalers, and recording oscilloscopes. Applications to high-speed coincidence measurements, millimicrosecond time-interval measurements, and fast counting are described.

621.3.015.7:621.396.619.16 1526
Conversion of Rectangular Pulses of Given Width and Variable Height into Rectangular Pulses of Given Height and Variable Width—W. Vogt. (*Funk u. Ton*, vol. 5, pp. 578–584; November, 1951.) In the circuit described, AM pulses charge a capacitor in the positively biased grid circuit of a switching tube which is conductive during the capacitor discharge. The discharge time is a function of the charging voltage. The nonlinearity of this relation is studied. Methods of improving the linearity are indicated.

621.314.222.012.3 1527
Charts for the Calculation of Mains Transformers—G. Pavel. (*Funk u. Ton*, vol. 5, pp. 561–577; November, 1951.) Design parameters are discussed. The charts relate core size,

number of turns, wire diameter, and resistance of primary and secondary windings of mains transformers for powers up to 150w. Special charts are provided for E- and M-shaped laminated cores of standard materials.

621.314.3†:621.314.5 1528
Magnetic Modulators—E. P. Felsh, V. E. Legg and F. G. Merrill. (*Electronics*, vol. 25, pp. 113–117; February, 1952.)

621.316.8 1529
The Problem of a Non-ohmic Resistor in Series with an Impedance—E. B. Moullin. (*Proc. IEE*, Part I, vol. 98, pp. 344–346; November, 1951.) Discussion on 2657 of 1951.

621.316.8.029.5 1530
Resistors at Radio Frequency—T. J. F. Pavlasek and F. S. Howes. (*Wireless Eng.*, vol. 29, pp. 31–36; February, 1952.) The rf resistance and distributed capacitance of resistors of the metallized-filament type enclosed in insulating sleeves were investigated at frequencies from 0.5 to 40 mc. Curves of R_{rf}/R_{dc} $2/R_{dc}$ and ϕ are plotted against $f \cdot R_{dc}$. R_{rf} and R_{dc} being respectively the rf and dc resistances, f the frequency and ϕ the phase angle. A comparison is made with the theoretical values obtained by considering the resistor as a transmission line having distributed resistance and capacitance.

621.318.572:621.385.2 1531
R. F. Bursts actuate Gas-Tube Switch—Geisler. (See 1786.)

621.318.572+681.142]:621.385.5.032.216 1532
The Single-Pulse Dekatron—Acton. (See 1787.)

621.385.2:546.289]+621.314.7 1533
Germauium Crystal Valves—Bettridge. (See 1786.)

621.392.012:517.63 1534
Block-Diagram Network Transformation—T. D. Graybeal. (*Elec. Eng.* (N.Y.), vol. 70, pp. 985–990; November, 1951.) Essentially full text of 1951 AIEE Pacific General Meeting paper. A convenient method particularly applicable to the analysis and synthesis of servo systems. The Laplace transform equations of the system are expressed in the form of a block diagram. By methods similar to the star-delta transformation, complicated systems may be reduced to one of a few simple forms.

621.392.5 1535
A Network Theorem—E. E. Zepler. (*Wireless Eng.*, vol. 29, pp. 44–45; February, 1952.) To find the effect of connecting an impedance Z across two points of a network, the impedance may be replaced by one in parallel with the generator. A simple formula is given from which the value of this parallel impedance can be found.

621.392.5 1536
Minimum Phase Networks—J. A. Tanner. (*Electronic Eng.*, vol. 23, pp. 418–423; November, 1951.) A "minimum phase network" is a feedback network having a minimum value of phase lag at every frequency while satisfying a specified gain/frequency characteristic. The properties of such networks are summarized in relation to their use in servo systems.

621.392.5:621.396.621.53 1537
The Parallel-T Network as a Linear Mixer—J. S. Nisbet. (*Electronic Eng.*, vol. 23, pp. 432–433; November, 1951.) Two oscillations are mixed by applying them across opposite sides of a parallel-T network and taking the output from one of the mid-shunt arms. The circuit operation is analyzed. Coupling between the two oscillators is avoided over a small frequency band.

621.392.5:681.142 1538
Linear Networks with Time-Varying Lumped Parameters—J. Brodin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1168–1170; November 12, 1951.) The equations of a system are expressed by means of an integration operator. The method is applied to the calculation of the pass band of an electroanalogue multiplier with a time-varying network.

621.392.5.029.3:621.3.012.3 1539
The Prediction of Audio-Frequency Response: No. 1—Circuits with Single Reactance Element—N. H. Crowhurst. (*Electronic Eng.*, vol. 23, pp. 440–443; November, 1951.) The first of a series of data sheets; applicability is not restricted to the af range.

621.392.5.029.3:621.3.012.3 1540
The Prediction of Audio-Frequency Response: No. 2—Circuits with Two Reactance Elements—N. H. Crowhurst. (*Electronic Eng.*, vols. 23/24, pp. 483–489, 33–38 and 82–86; December, 1951—February, 1952.) Formulas and charts are presented applicable to circuits reducible to a combination of series inductance and shunt capacitance, or series capacitance and shunt inductance, together with resistances.

621.392.52 1541
Tchebyshev Filters and Amplifier Networks—V. Belevitch. (*Wireless Eng.*, vol. 29, pp. 106–110; April, 1952.) Darlington's method of filter synthesis (1361 of 1940) is applied to the design of simple low-pass filters composed of alternate series coils and shunt capacitors. Resistive termination of both ends is considered and also the case where one end is on open circuit. When filters of this type are used as amplifier input or output networks, the prescribed value of the terminal shunt capacitance imposes a physical limitation on the gain-bandwidth product of the stage; this limitation has been studied by Bode (583 of 1946), but additional precision results from the present analysis. More general filters with one open-circuit termination are mentioned and a new method of design is outlined. The results can be extended to other than low-pass filters by means of frequency transformations.

621.392.52 1542
RC Networks as Filters—H. Pieplow. (*Arch. tech. Messen*, nos. 189 and 191, pp. T117–T118 and T141–T142; October and December, 1951.) Common types are analyzed and tabulated according to their filter properties. Means of improving the filtering action are considered, and it is found possible, by the addition of amplitude-stabilizing devices, to reduce distortion at the output of an RC oscillator to about 0.1–0.2 per cent. Negative feedback in conjunction with specially designed networks enables amplifiers to be constructed with almost any desired filter properties and stability of operation.

621.392.52 1543
Ladder Filters without Attenuation Fluctuations in the Pass Band (Power-Law Filters)—G. Bosse. (*Frequenz*, vol. 5, pp. 279–284; October, 1951.) An analytical treatment deriving design formulas for such filters, the over-all group delay of which is expressed by a power series. Attenuations, phase delays and transfer functions are shown for symmetrical low-pass filters with 1, 2, and 3 T-type units.

621.392.52 1544
Composite Ladder Filters. Second-Order Image Impedances—R. O. Rowlands. (*Wireless Eng.*, vol. 29, pp. 50–55; February, 1952.) Design formulas are derived for filters comprising up to three halfsections and having equal or inverse impedance functions of the second order. Their performance is equal to that of the conventional m-derived type, but fewer components are required. Other filters are described having a second-order impedance function at only one pair of terminals.

621.392.52 1545
The Numerical Calculation of Filter Circuits with Generalized Parameters, using Modern Theory with Special Attention to Cauer's work—V. Fetzer. (*Arch. elect. Übertragung*, vol. 5, pp. 499–508; November, 1951.) Cauer applied the operating-parameter theory to the calculation of Tchebycheff-type filters; in the present paper this theory is applied to filters having infinite-attenuation points anywhere in the attenuation band while conforming to Tchebycheff type within the pass band. Explicit formulas are derived for the coefficients of the characteristic function, which is developed as a polynomial; the zero-attenuation points are found by solving the corresponding higher-order equation. The relations between Cauer's Q functions and the characteristic function are shown for both symmetrical and "antimetrical" filters. Darlington's formulas (1361 of 1940) for the basic low-pass circuit are used to calculate the circuit elements from the no-load impedance.

621.392.52:518.4 1546
New Graphical Methods for Analysis and Design—W. Saraga and L. Fosgate. (*Wireless Eng.*, vol. 29, pp. 68–79; March, 1952.) Methods are developed for transforming a given performance characteristic into a straight line or a set of straight lines. The method is demonstrated by applying it to the analysis and design of image-parameter and insertion-parameter filters.

621.392.6 1547
Synthesis of Passive Electrical Networks with n Pairs of Terminals and Prescribed Scattering Matrix [matrice de répartition]—V. Belevitch. (*Ann. Télécommun.*, vol. 6, pp. 302–312; November, 1951. Full paper referred to in 2128 of 1951. The term "matrice de repartition" is used to denote the matrix répresenting the distribution of power between given terminating impedances.

621.395.661.1 1548
Repeater Coil with Divided Secondary Winding and Single Stray-Resonance Peak—O. Illner. (*Frequenz*, vol. 5, pp. 265–272; October, 1951.) From an equivalent circuit the conditions are determined under which only one resonance peak occurs in the response characteristic. In this case the pass band can be about 40 per cent wider than that obtainable when the response characteristic has two peaks. See also 1088 of 1951 (Schmitt and Schrag).

621.396.6–181.4 1549
Miniaturization—Crux of Contemporary Product Design—W. H. Hannahs and B. S. Ellefsen. (*Elec. Mfg.* (N.Y.) vol. 45, pp. 86–91. 200; June, 1950.) Review of techniques applied to circuit components and subassemblies.

621.396.611.4 1550
Some Results from the Theory of Coupled Electromagnetic Cavity Resonators—E. Ledinegg and P. Urban. (*Acta Phys. austriaca*, vol. 4, pp. 180–196; December, 1950.) Based on the theory developed previously (1115 of 1951), the coupling frequencies are calculated for some systems of particular interest, e.g., cylindrical cavities coupled at the flat ends by windows or coaxial-line sections. Comparison is made with experimental results.

621.396.615 1551
Blocking-Oscillator Amplitude Control—(*Electronic Eng.*, vol. 23, p. 439; November, 1951.) Circuits are described in which the amplitude of the oscillator output is adjusted by varying the bias applied to the suppressor grid.

621.396.615 1552
Distortion in Beat-Frequency Sources—C. G. Mayo. (*Wireless Eng.*, vol. 29, pp. 92–94; April, 1952.) At higher beat frequencies than that at which the two oscillators of a beat-frequency source lock, there is distortion of the beat-frequency wave form. Analysis shows that at any frequency $\omega/2\pi$ the second-harmonic distortion is $\omega_0/2\omega$, where $\omega_0/2\pi$ is the

highest beat frequency at which locking occurs.

621.396.615.077.2/.3 1553

An Amplidyne Phase Shift Oscillator—J. C. West. (*Jour. Sci. Instr.*, vol. 28, pp. 336–339; November, 1951.) An oscillator is described capable of developing a peak current of 3a in a 17-Ω load over the frequency range 0.06–18 cps. The amplidyne with its associated electronic amplifier has a local feedback loop to reduce the effects of saturation and hysteresis in the magnetic circuit. Application is to the determination of the frequency response of servomechanisms.

621.396.615.17 1554

Three-Valve Pulse Generator with Fixed Repetition Rate—F. A. Benson and G. V. G. Lusher. (*Wireless Eng.*, vol. 29, pp. 90–91; April, 1952.) A development of the 2-valve generator previously described (1244 of June), producing short positive pulses of amplitude about 50 v, using a square-wave input from either a multivibrator or a clipping circuit.

621.396.615.17:621.314.7 1555

Transistors as Multivibrators—I. Queen. (*Radio-Electronics*, vol. 22, pp. 92, 94; September, 1951.) A multivibrator using two transistors, and a flip-flop circuit using a single transistor, are described. Both are triggered by a differentiated square-wave voltage.

621.396.615.17.012.3 1556

A Nomogram for Multivibrator Design—W. R. Luckett. (*Electronic Eng.*, vol. 23, p. 448; November, 1951.)

621.396.615.18 1557

Decade Multivibrator Design. Method of Stabilizing Frequency Division from a Crystal Drive—J. E. Attew. (*Wireless World*, vol. 58, pp. 114–116; March, 1952.) Spasmodic momentary jumping to the next lower ratio of division often occurs in a crystal-driver decade multivibrator, even though synchronized for even division. This is prevented by applying a pulse of opposite polarity just before the true synchronizing pulse. A complete circuit diagram of a stable frequency division system embodying this principle is shown. Setting-up procedure is outlined.

621.396.645 1558

A New Push-Pull Amplifier Circuit—A. P. G. Peterson. (*Gen. Rad. Exper.*, vol. 26, pp. 1–7; October, 1951.) Power-amplifier applications of the circuit described in 1250 of June are considered.

621.396.645 1559

Cathode-Coupled Amplifier—J. A. Lyddiard. (*Wireless Eng.*, vol. 29, pp. 63–67; March, 1952.) Analysis is presented in which the cathode-coupled amplifier is treated as a cathode follower driving an earthed-grid voltage amplifier. The method is simple, and leads to a straightforward design procedure.

621.396.645:621.3.015.3 1560

Overvoltage Effect in R. F. Power Amplifiers—E. Rizzoni. (*Alta Frequenza*, vol. 20, pp. 200–209; October, 1951.) The effect of overvoltage at the terminals of the anode oscillatory circuit of a RF power amplifier, due to detuning of the anode circuit, is discussed, and a method of calculating it as a function of load admittance and amount of detuning is described. Calculations and graphs are presented for the FIVRE beam tetrode Type 4-C500 and the RCA triode Type 893 A-R.

621.396.645:621.315.612.4 1561

Dielectric Amplifier Fundamentals—A. M. Vincent. (*Electronics*, vol. 24, pp. 84–88; December, 1951.) The input voltage is applied across a capacitor with ferro-electric dielectric, whose reactance is thereby varied; the capacitor is in circuit with an ac power source and load, and amplified power variations appear across the latter. The impedance of the circuit is relatively high. Practical forms of the amplifier are described, and its operation is compared with that of magnetic and tube amplifiers; at present, the frequency range appears to have an upper limit at about 10 mc. Numerous applications are indicated.

621.396.645:621.392.52 1562

Broad-Banding by Stagger Tuning—R. C. Wittenberg. (*Electronics*, vol. 25, pp. 118–121; February, 1952.) Simplified methods are presented for design calculations of multicircuit stagger-tuned IF amplifiers to have either Butterworth or Tchebycheff type of response. Tables and charts are given and their use illustrated. Circuits with bandwidths up to twice the center frequency are designed using noncritical values of components.

621.396.645.37 1563

Highly-Selective Amplification at Low Frequencies—F. H. Hyde. (*Wireless Eng.*, vol. 29, pp. 85–90; April, 1952.) Analysis is presented of the unbalanced twin-T RC filter with series arms R and C and shunt arms $aR/2$ and 2C. The locus of the transmission vector, for $0.5<a<1.0$ is approximately circular and the phase angle varies continuously from 0 to 2π radians. A single-stage amplifier with series feedback through a balanced twin-T filter ($a=1$) has a Q value of $A/4$ when A, the loaded stage gain without feedback, is large. Unbalancing of the filter by making $a<1$, gives improved selectivity, and a working Q of 20 can be readily obtained. Experimental results are given for filters with resonance frequencies of about 0.25, 0.5, and 360 cps.

621.396.645.37 1564

Dual Circuit of a Feedback Amplifier—D. A. Bell. (*Wireless Eng.*, vol. 29, pp. 40–43; February, 1952.) The dual circuit of a single-stage amplifier is first obtained. The duals of (*a*) a single-stage amplifier with a voltage-feedback branch, (*b*) a combination of two amplifiers with local feedback round each and additional feedback over both stages, are then derived.

621.396.645.37.012.8 1565

Equivalent Circuits to Simplify Feedback Design—R. S. Burwen. (*Audio Eng.*, vol. 35, pp. 11–12, 45; October, 1951.) Equivalent circuits are used in analysis of the general feedback amplifier. Voltage and current feedback are considered separately. The effect of a small amount of feedback on power amplifiers is studied; a 16-Ω loudspeaker winding in the cathode circuit is sufficient to provide 6 db feedback in a single-pentode output stage. Application of equivalent circuits in the design of a preamplifier circuit with a prescribed response curve is described.

621.396.645.371.011.1 1566

Expressions for the Reduction of Distortion and Output Impedance in Terms of db of Feedback—W. J. Kessler and S. E. Smith. (*Audio Eng.*, vol. 35, p. 13; October, 1951.) The feedback factor is eliminated from the formulas usually applied, in order to obtain formulas expressed in terms of parameters easily measured.

621.3.015.7 1567

Pulse Techniques. [Book Review]—S. Moskowitz and J. Racker. Publishers: Prentice-Hall, New York, 1951, 300 pp., $5.00. (*Jour. Frank. Inst.*, vol. 252, p. 203; August, 1951.) "The book is apparently intended as a text for a brief undergraduate course in the pulse aspects of electronics."

621.392 1568

Circuits in Electrical Engineering. [Book Review]—C. L. Vail. Publishers: Prentice-Hall, New York, 1950, 560 pp., $5.75. (*Jour. Appl. Phys.*, vol. 22, p. 1391; November, 1951.) "Primarily intended as a textbook [of circuit analysis] and not as a reference work."

621.396.6+621.385 1569

Introduction to Electronic Circuits. [Book Review]—Feinberg. (See 1792.)

GENERAL PHYSICS

519.27:517.433 1570

Two Classes of "Observation Operators"—R. Vallée. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1350–1351; November 26, 1951.) Two types of linear operators are described, with a dual correspondence between the space-time variables on the one hand and the space-periodicity variables on the other. Special cases of these operators are frequently met with in all experimental fields, and they play an important part in information theory.

534.014.2 1571

Predominantly Subharmonic Oscillations—C. A. Ludeke. (*Jour. Appl. Phys.*, vol. 22, pp. 1321–1326; November, 1951.) Theory of the demultiplication of the frequency applied to a nonlinear system. Subharmonic resonances of amplitude greater than the applied fundamental are particularly considered. Experiments with electromechanical apparatus confirmed the theory. Transitions between different subharmonics are discussed.

534.21+538.56 1572

Wiener-Hopf Techniques and Mixed Boundary-Value Problems—S. N. Karp. (*Commun. pure appl. Math.*, vol. 3, pp. 411–426; December, 1950.) The parallelism between the method of separation of variables and the Green's function integral-equation method is shown to hold as in problems of more classical type. This relation leads to a characterization (from the standpoint of co-ordinate systems) of those problems in which the Wiener-Hopf type of problem (in an extended sense) arises. Certain heuristic advantages of the separation-of-variables procedure are also pointed out. Special applications considered include the diffraction of a plane wave by a staggered array of semi-infinite planes, and the es charge distribution on a cone, including the special case of a disk.

535.13:538.3 1573

Is there an Aether?—P. A. M. Dirac. (*Nature*, (London, vol. 168, pp. 906–907; November 24, 1951.) The difficulty of reconciling the concept of the aether with the principle of relativitiy is removed by applying quantum mechanics; the existence of an aether is implicit in the new theory of electrodynamics (1574 below).

537.122 1574

A New Classical Theory of Electrons—P. A. M. Dirac. (*Proc. Roy. Soc. A.*, vol. 209, pp. 291–296; November 7, 1951.) In the theory of the electromagnetic field without charges, the potentials are not fixed by the field, but are subject to gauge transformations; thus more variables are involved than are physically needed. It is possible by destroying the gauge transformations to make the superfluous variables acquire a physical significance and describe electric charges. This leads to a simplified classical theory of electrons which appears to be more suitable than the usual one as a basis for a passage to the quantum theory.

537.226:539.11 1575

Note on the Interaction of an Electron and a Lattice Oscillator—E. P. Gross. (*Phys. Rev.*, vol. 84, pp. 818–823; November 15, 1951.)

537.311.1 1576

Application of Collective Treatment of Electron and Ion Vibrations to Theories of Conductivity and Superconductivity—D. Bohm and T. Staver. (*Phys. Rev.*, vol. 84, pp. 836–837; November 15, 1951.)

537.311.33 1577

The Effect of the Mean Free Path of Electrons on the Electrical Properties of Non-metals—R. W. Wright. (*Proc. Phys. Soc.*, vol. 64, pp. 984–999; November 1, 1951.) The conductivity, thermoelectric power, Hall coefficient, fractional change of conductivity in a magnetic field and the Nernst, Ettinghausen, and Righi-Leduc coefficients are calculated on the Lorentz-Sommerfeld theory, using the most appropriate mean-free-path function for the non-metal concerned. The theoretical variations of the electrical properties with temperature so obtained agree well with experimental results.

537.531+539.18]:535.43 1578

Multiple Scattering of Waves—M. Lax. (*Rev. Mod. Phys.*, vol. 23, pp. 287–310; October, 1951.) Coherent and incoherent scattering of light, X-rays, neutrons, and photon waves are fully discussed. Using the self-consistent field method the scattered fields are derived, including the effects of anisotropic scattering, scattering of quantized waves, creation and absorption of particles. Doppler shifts, and randomly, partially, or completely ordered systems of scatterers. The connection between collisions with a multiparticle system and multiple scattering is considered.

538.114 1579

Application of the Bethe-Weiss Method to the Theory of Antiferromagnetism—Yin-Yuan Li. (*Phys. Rev.*, vol. 84, pp. 721–730; November 15, 1951.)

1580

538.311:621.318.423:513.647.1:621.385.029.6

Properties of the Electromagnetic Field of Helices—É. Roubine. (*Compt Rend. Acad. Sci.* (Paris), vol. 233, pp. 1174–1176; November 12, 1951.) Field distributions corresponding to the theory for the thin-wire helix given in 2978 of 1951 are compared with those corresponding to the theory of the continuous-cylinder guide; the two agree under certain stated conditions, which are satisfied for traveling-wave tubes with narrow beams, but not for wide-beam tubes.

538.522 1581

The Proximity Effect and Coefficient of Mutual Induction at High Frequency for a Wire and Part of a Very Thick Plate, both being Conducting and Parallel—A. Colombani. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1267–1269; November 19, 1951.) Assuming the wire radius is smaller than the depth of penetration of the current into the wire, the mutual inductance is given by $M=\mu^{\frac{1}{2}}/d(\pi\omega\gamma')^{\frac{1}{2}}$, where d is the distance of the wire from the plane, and γ' the conductivity of the plate. This result is extended to the case in which the wire radius is not negligible compared with d.

538.56:535.42 1582

On Systems of Linear Equations in the Theory of Guided Waves—W. Magnus and F. Oberhettinger. (*Commun. Pure Appl. Math.*, vol. 3, pp. 393–410; December, 1950.) An investigation of the diffraction of an em wave by a plane strip between two parallel planes, or in a rectangular waveguide, assuming that essentially only one type of wave exists. For an incoming wave of the type $\exp(i\alpha x)\cos \beta y$, α and β being real, the diffracted wave components can be expanded in a Fourier series, whose coefficients are uniquely determined by the condition of the finiteness of the total energy in any finite part of the space; they are given by an infinite set of linear equations. The special case of a diffracting strip half the width of the waveguide is treated in detail: in this case the linear equations can be dealt with by successive approximation, and the first steps can be carried out explicitly.

538.56:535.42 1583

On the Theory of Electromagnetic-Wave Diffraction by an Aperture in an Infinite Plane Conducting Screen—H. Levine and J. Schwinger. (*Commun. Pure Appl. Math.*, vol. 3, pp. 355–391; December, 1950.) A procedure for solving this problem exactly has been described by Meixner (94 of 1951), but approximations are required which are suitable for computation and accurate over a frequency range. This paper, a sequel to previous ones concerned with diffraction in a scalar field (83 and 1897 of 1950), described variational principles for obtaining some of the desired information. A formal description is given of the fields and boundary conditions involved. Expressions for the field vectors in any region are derived in terms of the tangential components of the electric or magnetic field components on its boundary, with the aid of tensor Green's functions. These expressions are first found for the regions on each side of the screen by using integrals involving the tangential electric field over the aperture. From the equality of the tangential magnetic fields for each region in the aperture, an integral equation for the tangential electric aperture field is obtained. From this, a stationary property of the radiation field at large distances from the aperture is derived. Similarly, variational principles are obtained by consideration of the tangential magnetic field over screen and aperture, and of the current over the screen. The connection between the principles and the plane wave "transmission crosssection," which is a measure of the ratio of energy passing through the aperture per second to that transported per unit area of the incident wave, is demonstrated. Numerical results are given for the cross-section of a circular aperture with normally incident plane wave, and are compared with those from the Kirchhoff and Rayleigh approximations.

538.566 1584

The Analytical Expression of Huyghens' Principle for Electromagnetic Waves—A. da Silveira. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1269–1272; November 19, 1951.)

538.566 1585

The Magnetic Dipole over the Horizontally Stratified Earth—J. R. Wait. (*Canad. Jour. Phys.*, vol. 29, pp. 577–592; November, 1951.) Analysis of the radiation characteristics of a vertical magnetic dipole above a two-layer or three-layer earth, with particular reference to the effects produced within the earth. Transient effects are considered in some cases.

538.566 1586

Transient Electromagnetic Propagation in a Conducting Medium—J. R. Wait. (*Geophys.*, vol. 16, pp. 213–221; April, 1951.) Expressions are developed by Laplace transformation for the electric fields due to different types of step-function current source in an infinite conducting medium. Sources considered are the electric dipole, the magnetic dipole, and step-function currents in insulated wires of finite and infinite length.

538.566:538.63 1587

The Influence of Magnetic Fields upon the Propagation of Electromagnetic Waves in Artificial Dielectrics—E. R. Wicher. (*Jour. Appl. Phys.*, vol. 22, pp. 1327–1329; November 1951.) "An effect corresponding to the Faraday effect in natural dielectrics is predicted for a class of artificial dielectrics because of the existence of a Hall effect in the metallic components of the structure. A formula for Verdet's constant as a function of element polarizability and Hall coefficient is obtained. The resonance shift to be expected in a cavity resonator, filled with an artificial dielectric, and subjected to a strong magnetic field, is calculated."

538.691 1588

The Effect of a Magnetic Field on Electrons in a Periodic Potential—J. M. Luttinger. (*Phys. Rev.*, vol. 84, pp. 814–817; November 15, 1951.) A theorem due to Wannier for treating the motion of electrons in a perturbed periodic field is generalized to include the effect of a slowly varying magnetic field. The problem reduces to that of solving a Schrödinger equation.

546.212:536.421.4 1589

Experimental Investigation of Icing Phenomena—D. Melcher. (*Z. angew. Math. Phys.*, vol. 2, pp. 421–443; November 15, 1951.) Detailed study of the formation of ice (*a*) artificially in a wind tunnel, (*b*) in the open air, taking account of the influence of an electric field.

501:530.12 1590

Mathematics of Relativity. [Book Review]—G. Y. Rainich. Publishers: Chapman and Hall, London, Eng., 1950, 174 pp., 28s. (*Beama Jour.*, vol. 58, pp. 375, 377; November 1951.) "This book is an admirable survey of relativity theory and it can well be recommended on that account."

537.1+538.1 1591

A History of the Theories of Aether and Electricity. [Book Review]—E. Whittaker. Publishers: Nelson, London, Eng., 1951, 434 pp., 32s, 6d. (*Jour. Frank. Inst.*, vol. 252, p. 441; November, 1951.) To be completed in two volumes; this first volume deals with classical theories.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:551.510.535 1592

The Wave-Frequency Dependence of the Duration of Radar-Type Echoes from Meteor Trails—V. C. Pineo and T. N. Gautier. (*Science*, vol. 114, pp. 460–462; November 2, 1951.) Simultaneous measurements on 27.2 and 41.0 mc recorded at the National Bureau of Standards between November 1, 1948 and October 1, 1949 support Lovell's conclusion (3402 of 1948) that the duration of radar echoes from meteor trails is approximately proportional to the square of the wavelength. An indication is given of the methods used.

523.72:621.396.822 1593

On Bailey's Theory of Amplified Circularly Polarized Waves in an Ionized Medium—R. Q. Twiss. (*Phys. Rev.*, vol. 84, pp. 448–457; November 1, 1951.) Detailed critical analysis of Bailey's theory (1909 of 1950). The growing waves, which Bailey interprets as amplified waves, can only be excited by reflection. It is contended that Bailey's theory can explain neither the excess RF radiation from sunspots nor that from discharge tubes. Power amplification is, however, possible in a drifting ionized medium under certain ideal conditions, which are discussed. See also 624 of 1951.

538.12:521.15 1594

A Fundamental Theory of the Magnetism of Massive Rotating Bodies—G. Luchak. (*Canad. Jour. Phys.*, vol. 29, pp. 470–479; November, 1951.) A theory based on a relativistically covariant generalization of Maxwell's equations to include gravitational fields.

550.372+550.382 1595

An Electromagnetic Interpretation Problem in Geophysics—L. B. Slichter. (*Geophys.*, vol. 16, pp. 431–449; July, 1951.) A flat earth in which permeability μ, conductivity σ, and permittivity ϵ vary only with depth, is considered subjected to an alternating field produced by a vertical magnetic dipole above the surface. Expressions for the variation of μ, σ, and ϵ are obtained in the form of Taylor's series, the coefficients of which may be determined by measurement of the magnetic field intensity H at the surface. The horizontal and vertical components of H above the surface are shown to be mutually dependent, and formulas independent of the electrical characteristics of the ground are derived which connect the two.

550.381 1596

Measurements of the Variation with Depth of the Main Geomagnetic Field—S. K. Runcorn, A. C. Benson, A. F. Moore, and D. H. Griffiths. (*Phil. Trans.*, vol. 244, pp. 113–151; November 27, 1951.) The main geomagnetic field is attributable either to a source seated at the core, or to a fundamental property of rotating matter corresponding to a source distributed throughout the earth. Measurements made in five mines in northern England provide evidence in favor of the core theory.

550.384.4 1597
The Equatorial Electrojet as Detected from the Abnormal Electric Current Distribution above Huancayo, Peru, and Elsewhere—S. Chapman. (*Arch. Met. Geophys. Bioklimatol, A*, vol. 4, pp. 368–390; 1951. In English.) Abnormally large daily variations of the horizontal component of magnetic force observed at Huancayo indicate the daily rise and decline of a concentrated eastward electric current, termed "equatorial electrojet," above the station. Similar effects have been observed at stations in Africa and India. The influence on the phenomenon of position with respect to geographic and magnetic equators is examined, and observations required to determine the height, intensity, width, and return current flow are discussed.

551.5+550.37+550.38 1598
General Assembly of the International Union of Geodesy and Geophysics, Brussels, 1951—H. W. L. Absalom. (*Met. Mag.*, vol. 80, pp. 326–330; November, 1951.) Brief report of the proceedings; recent work in the fields of meteorology and terrestrial magnetism and electricity was reported and discussed.

551.510.5 1599
Abrupt Seasonal Changes in Tropopause Level and Stratosphere Temperature at Habbaniya—D. Dewar. (*Met. Mag.*, vol. 80, pp. 323–326; November, 1951.)

551.510.53:551.557 1600
Evidence for a Stratospheric Circulation in Vertical Meridional Planes between Polar and Equatorial Regions in Winter—L. S. Clarkson. (*Met. Mag.*, vol. 80, pp. 309–318; November, 1951.)

551.510.535 1601
Some Characteristics of the Ionosphere E Region—K. Rawer and E. Argence. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1208–1210; November 12, 1951.) Recent observational data indicate either that the dissociation of O_2 takes place at a height greater than that suggested by Penndorf (2224 of 1949) or that the ionization process does not involve the dissociation of O_2.

551.510.535 1602
The Half-Year Period in the Ionization of the F_2 Layer—O. Burkard. (*Arch. Met. Geophys. Bioklimatol, A*, vol. 4, pp. 391–402; 1951. In German.) The amplitude and phase of the observed half-yearly variations of ionization depend on the geographical location of the observation stations. A straightforward explanation of the effect is based on the assumption that the intensity of the solar ultraviolet radiation depends on latitude and is least at the solar equator. This theory also explains the half-yearly variation of apparent height of the layer.

551.510.535:550.386 1603
Geomagnetic Bays and their Relation to Ionospheric Currents—H. Wiese. (*Z. Met.*, vol. 5, pp. 341–347; November, 1951.) Daily and yearly variations of the frequency of occurrence of bays as shown by the geomagnetic records obtained at Niemegk during the period 1937–1944 are investigated. A tendency to repetition at 27-day and 24-hour intervals is indicated. Correlation with ionospheric air currents is observed; air currents in the ionosphere are related to those in lower atmospheric layers, at least in winter.

551.510.535:551.510.4 1604
Ozone Measurements during Sudden Ionospheric Disturbances—S. Fritz. (*Arch. Met. Geophys. Bioklimatol, A*, vol. 4, pp. 343–350; 1951. In English.) The measurements were made in order to study the effect on total atmospheric ozone of the enhanced solar emission of ultraviolet radiation associated with the flares causing the ionospheric disturbances. On the assumption that the ratio of the extraterrestrial intensity of sunlight at 3,110 Å to that at 3,300 Å (the observation wavelengths) is unaffected, the observations indicate that the variation of ozone content due to the disturbances is small or nil, as would be expected from theoretical considerations.

551.510.535:621.396.11 1605
Ionosphere Review: 1951. Greatly Reduced Rate of Decrease in Sunspot Activity and M.U.Fs—T. W. Bennington. (*Wireless World*, vol. 58, pp. 121–122; March, 1952.) Shows the monthly mean sunspot numbers and F_2-layer noon and midnight critical frequencies from the last sunspot maximum to 1951, and 12-month running averages of the same three quantities since the last sunspot minimum. Solar activity is likely to decrease slowly during 1952. An estimate is made of sw propagation conditions during 1952.

551.510.535:621.396.72:621.3.087.47 1606
Ionospheric Souuding Stations—(*U.R.S.I. Inform. Bull.*, no. 72, pp. 20–23; November, December, 1951.) Stations in Austria and those under the Bureau Ionosphérique Français and the Service de Prévision Ionosphérique Militaire are listed, with operating data.

551.515.4 1607
The Electrical and Meteorological Conditions Inside Thunderclouds—J. Kuettner. (*Jour. Met.*, vol. 7, pp. 322–332; October, 1950.)

551.594.12 1608
Height Variations in the Concentration of Ions near the Ground during Quiet Summer Nights at Uppsala—H. Norinder and R. Siksna. (*Tellus*, vol. 3, pp. 234–239; November, 1951.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 1609
Origins of Radar. Background to the Awards of the Royal Commission—(*Wireless World*, vol. 58, pp. 95–99; March, 1952.) An account of the development of radar in Britain from 1935 to the war years, based on evidence given before the Royal Commission on Awards to Inventors.

621.396.9:526.9 1610
The Effect of Meteorological Conditions on the Measurement of Long Distances by Electronics—C. I. Aslakson and O. O. Fickeissen. (*Trans. Amer. Geophys. Union*, vol. 31, pp. 816–826; December, 1950.) Surveying using shoran (frequency range 220–350 mc) requires data on meteorological conditions between the two receiving stations. From these the changes in atmospheric refractive index with height are computed and the necessary corrections to velocity of wave propagation and path length determined. Correction methods are described, with numerical examples.

621.396.9:526.9 1611
Accuracy in Electromagnetic Distance Measurement—J. Moline. (*Radio franç*, no. 11, pp. 1–5; November, 1951.) Principles of pulse, FM, and phase-displacement methods of distance measurement by means of em waves reflected from the distant point are outlined, including that combining pulse and phase measurement. In phase-displacement systems accuracy to within $\lambda/100$ is attainable for frequencies up to 30 mc.

621.396.9:551.5 1612
Abnormal Displacement of some Echoes from Rain—R. Lhermitte. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1210–1212; November 12, 1951.)

621.396.9:551.578.4 1613
Some Quantitative Measurements of Three-Centimeter Radar Echoes from Falling Snow—R. C. Langille and R. S. Thain. (*Canad. Jour. Phys.*, vol. 29, pp. 482–490; November, 1951.) The variation of back-scatter intensity with rate of snowfall was observed during four storms. Analysis of the size distribution of snowflakes, which appears to be important in calculating radar echo intensity, was only carried out for one storm. Observed echo intensities are in fair agreement with values calculated from Ryde's formula (2062 of 1948).

621.396.91:551.594.22 1614
Lightning Detection by Radar—M. G. H. Ligda. (*Bull. Amer. Met. Soc.*, vol. 31, pp. 279–283; October, 1950.) Description of the method used, showing records obtained.

621-526:621.396.93 1615
Analysis and Construction of a Position-Fixing Servomechanism—Klein. (See 1739.)

621.396.932 1616
Radio Aids to Marine Navigation: The Seaman's Requirements—F. J. Wylie. (*Jour. Brit. I.R.E.*, vol. 11, pp. 478–490; November, 1951.)

621.396.933 1617
Aircraft Navigational Aids—(*Engineer* (London), vol. 192, pp. 702–703; November 30, 1951.) An account of the facilities provided by (*a*) the Marconi vhf df system for instantaneous visual indication of bearings and position fixing by the ground station; (*b*) the Mullard "telescribe" equipment in experimental use at London airport, by which written messages, maps, etc. are instantaneously reproduced on the cathode-ray screen of a distant receiving unit; and (*c*) the Decca "flight log."

621.396.933 1618
A Simplified Multiple-Track Range Airborne Equipment—R. S. Styles. (*Proc. IEE*, Part III, vol. 99, pp. 88–92; March, 1952.) Description of Australian M.T.R. light-weight equipment suitable for installation in small aircraft. In association with the appropriate ground-station installation it provides pilots with accurate azimuthal track guidance for a range of 100 miles, flying at 5,000 feet, and/or a localizer path, for use during final approach to a runway, accurate to within ±22 yards at the normal touch-down point.

621.396.933:629.13.053 1619
Automatic Track-Plotting Instrument for Aircraft—(*Engineering* (London), vol. 172, p. 589; November 9, 195?.) Description of the operation of the Decca "flight log" in use with the Mark VII receiver, which displays aircraft position and ground track directly on a chart. See also 2192 of 1951.

621.396.933.2.001.4 1620
ILS Field Test Set—Ellis. (See 1685.)

629.13.05:538.74 1621
Stroboscopic Earth-Inductor Compass—S. A. Schwartz. (*Elec. Eng.* (N. Y.), vol. 70, pp. 1001–1003; November, 1951.) A compass is described in which a coil carrying a compass card is rotated in the earth's magnetic field by a small air turbine. The sinusoidal voltage induced in the coil is amplified and squared, and the leading edge of the square wave is used to trigger a stroboscopic source of light. When observed by this light the compass card appears to be stationary and indicates a direction relative to magnetic north.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.787.7 1622
Sensitive Differential Manometer—J. M. Los and J. A. Morrison. (*Rev. Sci. Instr.*, vol.

22, pp. 805–809; November, 1951.) The mercury surfaces in the manometer arms serve as the moving plates of two separate parallel-plate capacitors which form part of the oscillatory circuits of two similar 3.6-mc oscillators. The capacitance change, due to a differential pressure change, varies the beat frequency of the oscillators, which is measured by reference to an af signal generator. An accuracy to within 0.1–0.2μ is obtained for pressure changes between 0 and 0.02 cm Hg. For larger pressure changes (up to 0.25 cm Hg) the accuracy is within 0.1 per cent.

535.37 1623

Recent Developments in Luminescent Materials—S. T. Henderson. (*Research* (London), vol. 4, pp. 492–497; November, 1951.) The subject is considered under the headings (a) research on known materials to discover the mechanism of luminescence, (*b*) investigations to improve utility and extend applications of known materials, and (*c*) discovery of new materials. 105 references.

537.228.1:546.431.824.31 1624

Electrical, Particularly Piezoelectric Properties of Barium Titanate—J. H. van Santen and G. H. Jonker. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 259–274; November, 1951. Discussion, pp. 275–276.) These properties are discussed in relation to crystal structure. The piezoelectricity of pre-polarized $BaTiO_3$ can be considered as a combination of a linear electrostriction and a piezoelectric effect. Possible applications of $BaTiO_3$ ceramics are noted.

537.311.33+535.37 1625

New Views on Oxidic Semi-Conductors and Zinc-Sulphide Phosphors—F. J. W. Verwey and F. A. Kroger. (*Philips Tech. Rev.*, vol. 13, pp. 90–95; October, 1951.) The mechanism of conduction in nonstoichiometric oxidic semiconductors is similar to that in other semiconductors in which desired changes of valency are produced by the admixture of suitable impurities. Members of the latter group are not characterized by lack of thermal stability (due to vacant lattice sites) as are nonstoichiometric compounds. Similar considerations are applicable to ZnS phosphors, in which the fluorescence centers consist of activator ions surrounded by sulphur ions. The accuracy of this hypothesis is confirmed by the fact that the role of the halogen ions in the formation of Zn phosphors may be taken over by trivalent cations.

537.311.33:517.944 1626

A Note on the Partial Differential Equations describing Steady Current Flow in Intrinsic Semiconductors—R. C. Prim. (*Jour. Appl. Phys.*, vol. 22, pp. 1388–1389; November, 1951.)

537.311.33:537.211 1627

An Effect of Light on Semiconductors: Variation of the Contact Potential Difference—W. Veith and G. Wlérick. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1097–1101; November 5, 1951.) Measurements were made of the contact potential with and without illumination for a CdS layer in vacuum, using a retarding-potential method. The results support a formula obtained by a classical calculation [see e.g., 4444 of 1940 (Mott and Gurney)]. Variation of the contact potential with temperature is also noted.

537.311.33:546.482.21 1628

Measurements of the Electrical Conductivity of CdS Crystals Irradiated by Medium Energy Electron Beams—H. Benda. (*Ann. Phys.* (Lpz.), vol. 9, pp. 413–422; November 15, 1951.)

537.312.8:546.87 1629

The Anomalous Magnetoresistance of Bismuth at Low Temperatures—P. B. Alers and R. T. Webber. (*Phys. Rev.*, vol. 84, pp. 863–864; November 15, 1951.) Results are reported of measurements made on Bi crystal rods 2 mm in diameter and 2–3 cm long, using transverse fields of strength 60 kilogauss and over.

537.533.9:537.226 1630

Direct Demonstration of the Conductivity of a Thin Dielectric under Electron Bombardment—C. Dufour. (*Jour. Phys. Radium*, vol. 12, pp. 887–888; November, 1951.) Measurements were made of the current through a composite target comprising an evaporated layer of ZnS between evaporated layers of Al across which a variable voltage was applied.

538.221 1631

Systematic Relations between Hysteresis, Creep of Nonlinearity Products, and the Richter After-Effect—R. Feldtkeller, H. Wilde and G. Hoffmann. (*Z. angew. Phys.*, vol. 3, pp. 401–409; November, 1951.) Measurements made on three similarly treated specimens of Si/Fe-alloy stampings are reported; systematic differences between them are discussed. The measurement equipment is described.

538.221 1632

Ferromagnetic Resonance and the Internal Field in Ferromagnetic Materials—J. R. MacDonald. (*Proc. Phys. Soc.*, vol. 64, pp. 968–983; November 1, 1951.) "A classical treatment of the domain energy terms of a homogeneous ferromagnetic solid leads to a formula for the internal field contributions from these terms. With this result, modifications in the resonance condition of ferromagnetic resonance arising from self energy, exchange energy, magnetocrystalline anisotropy and applied or intrinsic stress are obtained and are applied to various crystalline anisotropy and stress conditions of interest in ferromagnetic resonance experiments. Finally, the bearing of the results on the anomalous *g*-values obtained in resonance experiments is considered."

538.221 1633

The Magnetization Process in Ferrites—H. P. J. Wijn and J. J. Went. (*Physica*, vol. 17, pp. 976–992; November/December, 1951.) "The initial magnetization curve of ferrites has been measured as a function of frequency up to 2 mc. It has been found that the magnetization of sintered ceramic ferrites with a high permeability is brought about by at least two processes, one of which, in the frequency range covered, is independent of frequency and determines the initial permeability. The other process has a relaxation frequency of about 200 kc and is responsible for the irreversible processes during magnetization. From measurements on samples of sintered ferrites fired at different temperatures it has been concluded that the frequency-dependent magnetization is caused by irreversible Bloch-wall displacements, while the initial permeability is caused by a reversible rotation of the magnetization in Weiss domains in the direction of the external magnetic field (in contrast to what is believed to be the case in cast ferromagnetic metals). A discussion shows that neither eddy current effects nor any inertia effects so far known are responsible for the relaxation frequency of the Bloch wall at about 200 kc."

539.23:537.311.31 1634

Variation, as a Function of Temperature and Applied E.M.F., of the Electrical Resistance of Very Thin Metal Films Deposited on Diamond, Amber and Plexiglass—N. Mostovetch and T. Duhautois. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1265–1267; November 19, 1951.) The results do not differ appreciably from those previously obtained (1701 and 2535 of 1950); they suggest that the semiconductivity of very thin films is not an impurity-type semiconductivity in which the support plays an essential part.

546.289+546.815.221]:537.311.33 1635

A Study of Rectification Effects at Surfaces of Germanium and Lead Sulphide—C. A. Hogarth and J. W. Granville. (*Proc. Phys. Soc.*, vol. 64, pp. 992–998; November 1, 1951.) Chemical etching or thermal treatment in vacuo can remove the amorphous surface layer produced by polishing. Heat treatments up to 900° C can improve the rectification characteristics considerably, and may be preferable to etching. Surface modifications resulting from the various treatments were examined by electron diffraction.

546.289:539.164.9 1636

Electron-Hole Production in Germanium by Alpha-Particles—K. G. McKay. (*Phys. Rev.*, vol. 84, pp. 829–832; November 15, 1951.) The number of electron-hole pairs produced in Ge by alpha-particle bombardment was determined by collecting the internally produced carriers across a reverse-biased $n-p$ junction. There was no evidence of trapping of carriers in the barrier region. The energy lost by a bombarding particle per electron-hole pair produced is 3.0 ± 0.4 ev. The difference between this and the energy gap is attributed to losses to the lattice from the internal carriers.

546.289:539.185.9 1637

Evidence for Production of Hole Traps in Germanium by Fast Neutron Bombardment—J. W. Cleland, J. H. Crawford, Jr., K. Lark-Horovitz, J. C. Pigg and F. W. Young, Jr. (*Phys. Rev.*, vol. 84, pp. 861–862; November 15, 1951.)

546.289:621.314.7:535.215 1638

The *n-p-n* Junction as a Model for Secondary Photoconductivity—K. G. McKay. (*Phys. Rev.*, vol. 84, pp. 833–835; November 15, 1951.) Experiments are discussed in which a Ge *n-p-n* junction is subjected to bombardment by alpha particles, producing excess-hole currents in the *p* region. Since the secondary currents observed in photoconductive insulators have similar characteristics, study of the *n-p-n* junction is expected to lead to better understanding of secondary photoconductivity.

547.476.3:537.228.1 1639

Rochelle-Salt Specimens moulded from Crystal Plates under High Pressure—F. Blaha. (*Acta Phys. austriaca*, vol. 4, pp. 272–277; December, 1950.) Disks cut from Rochelle-salt crystals were moulded into pastilles by application of pressure up to 17,000 kg/cm² in the direction of the α axis. Conductivity and permittivity measurements over a range of temperatures are given.

549.211.091.3 1640

The Effect of Inhomogeneities on the Electrical Properties of Diamond—A. J. Ahearn. (*Phys. Rev.*, vol. 84, pp. 798–802; November 15, 1951.)

549.514.51 1641

A New Crystal Cut for Quartz with Zero Temperature Coefficient—E. J. Post. (*Appl. Sci. Res.*, vol. B1, pp. 420–428; 1950.) See 115 of 1950.

621.314.634 1642

Reversible Changes in the Boundary Layer of Selenium Rectifiers—A. Hoffmann, F. Rose, E. Waldkötter and E. Nitsche. (*Z. Naturf.*, vol. 5a, pp. 465–467; August, 1950.) Two observed phenomena are discussed: (*a*) an increase of working resistance on changing over from application of forward voltage only to application of alternating voltage; (*b*) a decrease of the capacitance of the boundary layer with increase of the operating bias voltage, for the same instantaneous total voltage. Both effects can be explained by assuming a migration of impurity centers due to increased mean field strength at the boundary.

621.314.7 1643

Transistors: Part 2—Physics and Construction of the Transistor—J. Malsch. (*Arch. elekt. Übertragung*, vol. 5, pp. 425–433 and 467–573; September and October, 1951.) Addendum, *ibid.*, vol. 6, pp. 73–79; February, 1952. A survey paper. Part 1: 2726 of 1951.

621.315.616:547-128† 1644

Silicone Rubber emerges as a Dielectric Material—J. F. Dexter. (*Elec. Mfg.* (N. Y.), vol. 45, pp. 100–103, 204; June, 1950.) Results of tests on the effects of aging and temperature on the physical and electrical properties of silicone rubber are shown. Samples with brittle points at −90°C are serviceable at 250°C and withstand temperatures up to about 175°C indefinitely. They show exceptional resistance to heat, cold, moisture, oxidation, corona discharge and fatigue. Points of production technique are noted.

621.396.622.63 1645

The Temperature Dependence of the Static Characteristics of Crystal Rectifiers, and its Theoretical Significance—K. Seiler. (*Z. Naturf.*, vol. 5a, pp. 393–397; July, 1950.) An experimental investigation was made of rectifiers composed of a layer of Si with traces of Al, in combination with a Mo contact point; the temperature range covered was −80°C to +95°C. I/V characteristics are plotted; the significance of the results for determining the concentration of impurity centers is discussed.

669.715:537.311.31 1646

Effect of Alloying Elements on the Electrical Resistivity of Aluminum Alloys—A. T. Robinson and J. E. Dorn. (*Jour. Metals*, vol. 3, pp. 457–460; June, 1951.) "The electrical resistivities of aluminum alloys containing Cu, Ge, Zn, Ag, Cd, and Mg were found to increase linearly with the atomic percentage of the solute atoms. Application of Linde's rule to these data suggests that each aluminum atom contributes 2.5 electrons to the metallic bond."

778.3:[621.317.755+621.397.621.2 1647

Photography of Oscillograms and Television Images—H. Aberdam. (*Toute la Radio*, nos. 160 and 161, pp. 339–342 and 365–368; November and December, 1951.) An examination of the technique, with particular reference to the photographic emulsion required, the spectral brightness of the scanning spot, and the fluorescence of the screen.

621.315.59 1648

Semiconducting Materials. [Book Review]—H. K. Henisch (Ed.). Publishers: Academic Press, New York, N. Y., 1951, 281 pp., $6.80. (*Electronics*, vol. 25, pp. 336–338; February, 1952.) Proceedings of conference held at the University of Reading in July, 1950; contains the full text of the 28 papers presented.

MATHEMATICS

681.142 1649

A General Purpose Differential Analyser: Part I—Description of Machine—G. L. Ashdown and K. L. Selig. (*Elliott Jour.*, vol. 1, pp. 44–48; September, 1951.) This analyzer is designed for robustness and for rapidity of problem setting rather than for high accuracy. The integrator is of ball-and-disk type. Mechanically independent units are connected by servo-links with electrical connections through a central cross-connection panel.

681.142 1650

The Use of the EDSAC for Mathematical Computation—M. V. Wilkes. (*Appl. Sci. Res.*, vol. B1, pp. 429–438; 1950.) A simple explanation of the constituent elements of any program.

681.142 1651

On the Background of Pulse-Coded Computors—T. J. Rey. (*Electronic Eng.*, vol. 24, pp. 28–32 and 66–69; January and February, 1952.)

681.142 1652

An Electronic Multiplier—M. J. Somerville. (*Electronic Eng.*, vol. 24, pp. 78–80; February, 1952.) Description of a multiplier circuit with negligible time lag for use in an analogue computer. The frequency of a carrier is modulated in proportion to one of the multiplicands, while its amplitude is modulated in proportion to the other. The resulting signal is fed to a phase discriminator, whose output is proportional to the required product. The use of the technique is illustrated in solving Airey's equation.

51:62 1653

Advanced Engineering Mathematics. [Book Review]—C. R. Wylie, Jr. Publishers: McGraw-Hill, New York, N. Y., 1951, 640 pp., $7.50. (*Electronics*, vol. 25, p. 330; February, 1952.) ". . . strongly recommended as text book or reference for advanced students in electrical engineering."

519.2 1654

Statistische Methoden für Naturwissenschafter, Mediziner und Ingenieure (Statistical Methods for Scientists, Physicians and Engineers). [Book Review]—A. Linder. Publishers: Birkhauser, Basle, Switzerland, 2nd enlarged ed., 238 pp., 31.20 Swiss francs. 1951. (*Z. angew. Math. Phys.*, vol. 2, pp. 494–495; November 15, 1951.) ". . . will have a wide circulation and give excellent service in both theory and practice."

681.142 1655

The Preparation of Programs for an Electronic Digital Computer. [Book Review]—M. V. Wilkes, D. J. Wheeler and S. Gill. Publishers: Addison-Wesley Press, Cambridge Mass., 1951, 167 pp., $5.00. (*Jour. Frank. Inst.*, vol. 252, pp. 445–446; November, 1951.) "Although the system of subroutines discussed and given in this book might not be applicable to all machines, it can serve as a pattern which will greatly facilitate the development of such a system for a particular machine."

681.142 1656

Synthesis of Electronic Computing and Control Circuits. [Book Review]—Staff of the Computation Laboratory, Harvard University. Publishers: Harvard University Press, Cambridge, Mass., 1951, 278 pp., $8.00. (*Electronics*, vol. 25, pp. 400, 406; March, 1952.) "The book deals entirely with digital computing circuits, considering no analog devices. The control circuits mentioned in the title are of the type in which all of the relevant information is handled in digital form, rather than of the type associated with servomechanisms."

MEASUREMENTS AND TEST GEAR

538.71 1657

Some Developments in Electronic Magnetometers—A. W. Brewer, J. Squires and H. McG. Ross. (*Elliott Jour.*, vol. 1, pp. 38–43; September, 1951.) Developments discussed include (*a*) airborne equipment suitable for geophysical surveys and capable of measuring variations as small as 1 gamma in the total geomagnetic field, (*b*) equipment for absolute measurements.

621.3.018.41(083.74) 1658

Standard-Frequency Transmissions—(*Wireless Eng.*, vol. 29, p. 82; March, 1952.) Actual values for the frequencies of the standard-frequency transmissions from Rugby (1027 of May) and Droitwich (718 of April), as determined at the National Physical Laboratory, are to be reported regularly in *Wireless Engineer*. The first report, presented here, gives values for February 1952.

621.3.018.41(083.74)+529.786]:538.569.4 1659

"Atomic" Clocks and Frequency Stabilization on Microwave Spectral Lines—C. H. Townes. (*Jour. Appl. Phys.*, vol. 22, pp. 1365–1372; November, 1951.) "Application of the various types of radiofrequency spectral lines to accurate frequency stabilization and time standards is surveyed. Pertinent characteristics of microwave gas absorption lines and the various types of errors in frequency stabilization due to the nature of these absorption lines or to fundamental thermal noise are discussed in detail. It is shown that time standards synchronized with microwave absorption in ammonia or resonances in molecular or atomic beams have limits of accuracy of the order of 1 part in 10^{12} for a short time, and still smaller limiting fractional errors over longer periods of time."

621.317.18.083.4 1660

Balance Approach in A.C. Measurement Circuits: Part 1—Theoretical Bases—H. Poleck. (*Arch. Tech. Messen*, pp. T115–T116; October, 1951.) An ideal balancing operation is characterized by unambiguous indication of the direction in which the balancing elements must move, and by a single movement of the balancing elements to reach zero. Operation of null indicators with and without phase dependence is analyzed.

621.317.3:621.385.032.216 1661

A Method of Measuring the Interface Resistance and Capacitance of Oxide Cathodes—C. C. Eaglesfield and P. E. Douglas. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 318–320; November, 1951.) The interface impedance, consisting of a resistance and a capacitance in parallel, causes frequency-dependent feedback. Another frequency-dependent network is added to make the gain independent of frequency, in which case the interface components are equal to the measurement components. The apparatus and its operation are described briefly.

621.317.328.029.62 1662

V.H.F. Microvoltmeter and Field-Strength Measurement Set—P. Lygrisse. (*Électronique* (Paris), pp. 29–31; November, 1951.) Circuit diagram and description of an instrument for field-strength measurements at levels between 5μV/m and 0.1 V/m in the frequency range 75–195 mc.

621.317.335.3.029.64 1663

Balance Methods for the Measurement of Permittivity in the Microwave Region—T. J. Buchanan. (*Proc. IEE*, Part III, vol. 99, pp. 61–66; March, 1952.) From measurements of propagation constant the permittivity of water and aqueous solutions were calculated for wavelengths of 3.2 cm and 1.26 cm.

621.317.335.3.029.64 1664

A New Method for Measurement of the Dielectric Constant of Low-Conductivity Fluids in the Centimetre Waveband—E. Ledinegg, P. Urban and F. Reder. (*Acta Phys. austriaca*, vol. 4, pp. 9–17; July, 1950.) A method using a cylindrical cavity resonator is described; operation is at fixed frequency and at fixed cylinder length. The determination is made by measuring the volume of fluid introduced into the cavity to make it resonate at the same frequency as when empty. Accuracy to within a few parts per thousand is possible.

621.317.336 1665

Impedance Measurement at High Frequency using Bridged-T and Parallel-T Elements—K. Lamberts. (*Arch. Tech. Messen*, no. 189, pp. T108–T109; October, 1951.)

621.317.374 1666

Determination of Loss Angle of Materials with High Dielectric Constant—E. Ledinegg and P. Urban. (*Acta Phys. austriaca*, vol. 4, pp. 197–212; December, 1950.) The method is based on the introduction into a cylindrical cavity resonator of a layer of the dielectric material of thickness such that the resonance frequency is the same as for the empty cavity. See also 1664 above.

621.317.411.029.62/.63 1667

On the Determination of the Complex Permeability of Ferromagnetic Conductors at High Frequencies—A. Wieberdink. (*Appl. Sci. Res.*, vol. B1, pp. 439–452; 1950.) The complex propagation constant for em waves in a concentric Lecher system, the outer conductor being copper, while the inner conductor is a wire of the material under investigation, is measured. From this constant the complex permeability of the wire can be calculated. Results of measurements on a Ni-Fe wire show a decrease of permeability with increasing frequency and a resonance phenomenon at a frequency of 320 mc.

621.317.7:621.396.615.17 1668

The Multivibrator as Test Apparatus—O. Limann. (*Funk u. Ton*, vol. 5, pp. 585–599; November, 1951.) Description of the circuit and its operation and of many applications in square-wave testing.

621.317.723:621.385.5 1669

A Stabilized Mains-Supplied Valve Electrometer Circuit—G. Bonfiglioli and G. Montalenti. (*Alta Frequenza*, vol. 20, pp. 210–213; October, 1951.)

621.317.725 1670

Linear Diode Voltmeter—R. E. Burgess. (*Wireless Eng.*, vol. 29, p. 80; March, 1952.) Correction to paper abstracted in 736 of April.

621.317.725:621.314.671 1671

Valve Voltmeter. The Rectifier Section.—M. G. Scroggie. (*Wireless World*, vol. 58, pp. 89–94; March, 1952.) Detailed discussion of the design of a double-diode rectifier unit suitable for use in the measurement of alternating voltages with the dc instrument described in 1041 of May.

621.317.725:621.317.32:621.396.822 1672

Slideback and Infinite-Impedance Voltmeters—R. E. Burgess. (*Wireless Eng.*, vol. 29, pp. 59–62; March, 1952.) Extension of analysis given previously for diode and anode-bend voltmeters (189 of February). In the slideback voltmeter the increase of mean current through a diode or triode on application of an input voltage is counterbalanced by additional negative bias; this type of voltmeter gives indications lower than the peak voltage of a cw signal and has a square-law response to noise. The infinite-impedance voltmeter has the same rectification characteristics as a simple diode voltmeter, but takes no power from the source. Curves are given for the rectification characteristics of the different types of voltmeter for cw signals and fluctuation noise applied separately; formulas are derived for the response to any arbitrary mixture of signal and noise.

621.317.73 1673

A Microwave Swept-Frequency Impedance Meter—E. A. N. Whitehead. (*Elliott Jour.*, vol. 1, pp. 57–58; September, 1951.) Description of an instrument based on the use of directional couplers, for the rapid testing of waveguide components. A klystron oscillator has its frequency swept by a mechanical drive, the waveband of 3.0 to 3.4 cm being covered once in two seconds. The amplitude and phase of the voltage reflection coefficient of the component on test are displaced on a cro.

621.317.733.011.21:621.392.26 1674

A Reflectionless Wave-Guide Termination—R. E. Grantham. (*Rev. Sci. Instr.*, vol. 22, pp. 828–834; November, 1951.) Developed as a reference standard for microwave impedance bridges, the termination, also applicable to coaxial transmission lines, consists of a section of waveguide with a movable dissipative load, and is preceded by a tuner which can be adjusted to cancel in magnitude and phase the small reflection coefficient of the load. Reflection coefficients of 0.001 were obtained with X-band waveguide terminations.

621.317.733.011.21:621.396.611.21 1675

The Design and Use of an Admittance Bridge for Piezoelectric Crystals—J. F. W. Bell. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 324–327; November, 1951.) A radio-frequency bridge for the rapid measurement of resistance and Q-factor of piezoelectric crystals is described. The limitation of the accuracy of measurement due to frequency fluctuations of the generator used and to variations in the stray capacitance of the variable resistance arm of the bridge is discussed. Examples of the use of the bridge at 250 kc are given.

621.317.737 1676

A Q-Meter based on Free Damped Oscillations—K. Franz and S. F. Pinasco. (*Rev. telegr. Electronica* (Buenos Aires), vol. 40, pp. 731–733; November, 1951.) The meter is designed for determining accurately the Q value of the tank circuits of power oscillators, with Q values <20. The circuit under test is introduced in the anode lead of a type-6SH7 pentode with pulsed input. The arrival of a pulse charges the tank-circuit capacitor, which discharges by free oscillation of the circuit; the voltage changes across the capacitor are applied to the grids of a double triode (the two sections connected in parallel) which is cathode coupled to a cro. The Q of the tank circuit is given by $Q=n\Pi/\log(B_0/B_n)$, where B_0 is the initial amplitude and B_n that of the nth wave in the damped wave train displayed on the cro. Typical oscillograms corresponding to Q values ranging from 63 to <0.5 are shown.

621.317.755:621.385.012 1677

Electron-Tube Curve Generator—M. L. Kuder. (*Electronics*, vol. 25, pp. 118–124; March, 1952.) Description, with detailed circuit diagram, of equipment for cro display of families of anode characteristics, together with the locus of the load line and coordinates for direct measurement.

621.317.755:621.385.012 1678

Electron Tube Curve Tracer—J. H. Kuykendall. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 9–11, 29; August, 1951.) Description of cro equipment with direct-reading current voltage scales.

621.317.76.029.3 1679

Frequency Comparator—P. Riéty. (*Ann. Télécommun.*, vol. 6, pp. 332–336; November, 1951.) Description of a circuit designed for the calibration of af oscillators at frequencies between 20 cps and 20 kc. A series of subharmonic frequencies is derived from a standard-frequency 1-kc oscillator by frequency division in a multivibrator circuit. The signal of frequency to be measured is made to beat with a suitable harmonic of one of the derived frequencies, a magic-eye or loudspeaker being used as indicator. Calibration points are available at the first 50 harmonics of nine frequencies ranging from 20 cps to 1 kc. A circuit diagram and component values are shown.

621.317.784 1680

Power Meter and Mismatch Indicator—R. G. Medhurst and J. A. Knudsen. (*Wireless Eng.*, vol. 29, p. 112; April, 1952.) A closed expression is given for an integral used by Boff (449 of March) in determining the sensitivity of a pickup loop.

621.396.615.11:534.844.1 1681

Equipment for Acoustic Measurements: Part 3—Acoustic Pulse Measurements—C. G. Mayo, D. G. Beadle and W. Wharton. (*Electronic Eng.*, vol. 23, pp. 424–428; November, 1951.) A tone pulse is radiated into a studio under test, and the sound is picked up by a microphone connected to a triggered-timebase oscilloscope. The equipment weighs <42 lb. Details are given of the circuits used. Triggering pulses at any predetermined interval between 0.5 and 30 sec are provided automatically by a transitron oscillator.

621.396.615.14+[621.396.664:621.397.6 1682

Ultra High Instrumentation—R. G. Peters. (*TV Eng.* (N. Y.), vol. 2, pp. 18–20; November, 1951.) R.C.A. test and measurement equipment described includes a sweep/marker generator and a picture monitor, rf load and wattmeter, and frequency and modulation monitors for television transmitters.

621.396.615.17 1683

Variable-Frequency Clock-Pulse Generator—R. R. Rathbone. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 19–20; August, 1951.) Description of a test unit developed at the Massachusetts Institute of Technology, providing 0.1-μs pulses with repetition frequencies from 0.2 to 4.9×10^6/sec, the frequency being stable to within 20 parts in 10^6. The unit is used for building up complex circuits for pulse operation, and has standard 93-Ω input and output impedances.

621.396.615.17:621.397.6.001.4 1684

Linear Staircase Generator for Television Use—A. M. Spooner and F. W. Nicholls. (*Electronic Eng.*, vol. 23, pp. 481–482; December, 1951.) A cathode-coupled multivibrator is synchronized by line-suppression pulses; at the end of each pulse the multivibrator executes a train of oscillations, controllable in number between 5 and 30. This output is applied to a diode counter circuit, which generates a pulse with the corresponding number of steps. Such a wave form is useful for testing the response of television film recording apparatus.

621.396.933.2.001.4 1685

ILS Field Test Set—C. L. Ellis. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 3–5, 26; November, 1951.) Description of the G-250A set for checking the entire aircraft instrument-landing equipment. It is of rugged, weatherproof construction and can be operated by nontechnical personnel. Three independent crystal-controlled generators provide the marker, localizer and glide-slope frequencies, a choice of 20 being available in the localizer and glide-slope bands. A cycling unit controls the signal modulation sequence and is designed for either automatic or manual operation.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

532.137:621.395.611.62 1686

Vibrating-Plate Viscometer—J. G. Woodward. (*Electronics*, vol. 25, pp. 98–100; February, 1952.) The viscous damping exerted on a plate immersed in a liquid and oscillating in its own plane is measured by an electromechanical transducer.

534.321.9:620.179.1 1687

Ultrasonic Tyre-Testing Equipment—(*Engineer* (London), vol. 192, pp. 565–566; November 2, 1951.) A nondestructive production test developed jointly by the Dunlop and General Electric companies is based on the fact that an internal discontinuity such as imperfect bonding between rubber and fabric gives rise to an air film which reflects ultrasonic waves almost completely. The tyres are immersed in water during test. The generator used is a quartz crystal operating at 50 kc and located 1 inch from the rubber, in the well of the tyre; the power output is about 1 w. See also *Elec. Rev.* (London), vol. 149, pp. 893–894; November 2, 1951.)

621—52:621.396.611.3 1688

Applications of Electrical Methods of Differentiation to Control Problems—J. Rateau. (*Rev. Gén. Élec.*, vol. 60, pp. 451–465; November, 1951.) Analysis of basic differentiating circuits and discussion of their design and application in control mechanisms such as an automatic pilot.

621—526 1689

An Electronic Servo Simulator for Unstable and Open Loop Systems—N. T. van der Walt. (*Electronic Eng.*, vol. 24, pp. 52–57; February, 1952.) The simulator consists of a number of feedback amplifiers in cascade; its response to a square-wave input is viewed on a cro. Inclusion of clamping arrangement, operative during the timebase flyback, ensures a stable return to datum level.

621-57:537.228.1 1690
High-Speed Crystal Clutch—(*Jour. Frank. Inst.*, vol. 252, pp. 427-428; November, 1951.) In a clutch developed by Codier at the National Bureau of Standards, primarily for use in high-speed computers, application of direct voltage to the electrodes of three "bimorph" piezo-electric elements causes them to bend and press the clutch output disk against the rotating input disk.

621.317.083.7 1691
Simultaneous A.M. and F.M. in Rocket Telemetering—W. C. Moore. (*Electronics*, vol. 25, pp. 102-105; March, 1952.) Details of receiver and pressurized transmitter operating on a single carrier frequency of 183 mc and providing two channels, one an AM channel 5 mc wide suitable for rapidly varying phenomena, unusual wave forms, etc., the other a FM channel using a reactance modulator. The AM is effected by screen-grid modulation of the final stage of the transmitter.

621.365.54† 1692
Induction Heating in the Drop-Forging Industry—G. W. Seulen. (*Metal Treat.*, vol. 18, pp. 483-489; November, 1951.) Various types of medium-frequency (i.e. up to 10 kc) generator equipment are discussed and details are given of some German-designed plants.

621.383.001.8:535.61-15:778.37 1693
The Application of Image Converters to High Speed Photography—J. A. Jenkins and R. A. Chippendale. (*Jour. Brit. I.R.E.*, vol. 11, pp. 505-517; November, 1951.) Description of the use of a new type of tube, the Mullard ME 1201, with a Cs-Sb cathode of very low resistance and average sensitivity of 20 μA/lumen to tungsten light at 2,700°K, as a high-speed optical shutter. Exposure times shorter than 10^{-7} sec can be obtained. Numerous test photographs are reproduced.

621.384.6 1694
The Helix as a Linear Accelerator for Protons—D. R. Chick and D. P. R. Petrie. (*Nature* (London), vol. 168, pp. 782-783; November 3, 1951.) A circular waveguide consisting of a closely wound wire helix has been proposed for accelerating protons up to about 20 Mev. To counteract the radial component of electric field due to the wave, which tends to disperse the proton beam, a hollow beam of electrons surrounding the proton beam is suggested.

621.385.833 1695
Ion Image of an Emissive Anode—G. Couchet, M. Gauzit and A. Septier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1087-1090; November 5, 1951.) Report of observations, made with an emission-type electron microscope, of the positive-ion emission from a plane anode heated by electron bombardment. The fluorescent screen is of $CaWO_4$.

621.385.833 1696
The Axial Potential of [electron] Lenses with Grids—M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 298-299; July 23, 1951.) A rigorous expression is derived for the axial potential of a lense consisting of two symmetrical cylindrical or plane equipotential elements with circular holes, and an interposed plane grid at a potential different from that of the other two elements.

621.385.833 1697
Gaussian Elements of [electron] Lenses with Grids—M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1354-1356; November 26, 1951.) Calculations from the formulas previously derived (1696 above) gave results in good agreement with the experimental observations of Knoll and Weichart (4542 of 1938). Both convergent and divergent lenses are treated.

621.385.833 1698
Rigorous Calculation of Typical Electrostatic Electron Lenses—W. Glaser and H. Robl. (*Z. angew. Math. Phys.*, vol. 2, pp. 444-469; November 15, 1951.) Paraxial electron trajectories are determined for configurations approximating to cylindrical lenses.

621.385.833 1699
Newton's Law of the Formation of Images applied to Electron Optics—P. Funk. (*Acta Phys. austriaca*, vol. 4, pp.304-308; December, 1950.) A simpler solution than those of Hutter (1004 of 1946) and Glaser and Lammel (202 of 1942) is presented.

621.385.833 1700
High-Resolution Velocity Analysis with Magnetic Electron Lenses—F. Lenz. (*Naturwiss.*, vol. 38, pp. 524-525; November, 1951.) A method is described for testing lens hv stability and investigating electron velocity losses by observations on a ring-focus line.

621.385.833 1701
Imaging Properties of a Series of Magnetic Electron Lenses—G. Liebmann and E. M. Grad. (*Proc. Phys. Soc.*, vol. 64, pp. 956-971; November 1, 1951.) An investigation of the dependence of the paraxial image-forming properties and the first-order lens aberrations on geometrical design and lens excitation. The range of gap widths, S, investigated was $S/D=0.2$ to $S/D=2$, where D=lens diameter. The field distributions within the lenses were measured; the lens data derived from the measurements are given in a series of graphs applicable to the most common magnetic electron lenses. Application of the results to lens design is discussed.

621.385.833 1702
Technique of Electron Microscopy—D. G. Drummond and G. Liebmann. (*Nature* (London), vol. 168, pp. 819-821; November 10, 1951.) A report of the proceedings at a conference of the Electron Microscopy Group of the Institute of Physics held in St. Andrews University, 19th-21st June 1951, with brief indications of the subject matter of the various papers read.

621.385.833 1703
The Symmetrical Magnetic Electron-Microscope Objective Lens with Lowest Spherical Aberration—G. Liebmann. (*Proc. Phys. Soc.*, vol. 64, pp. 972-977; November 1, 1951.) In the symmetrical magnetic lenses already considered (1702 above) there is an optimum value of the lens excitation parameter which produces minimum spherical aberration. The relation between maximum obtainable axial field strength and the maximum field strength in the pole-piece gap leads to an optimum design for electron microscope objectives giving minimum spherical aberration and maximum resolving power. A modification of this optimum design for practical use is described.

621.387.4 1704
Self-Quenching Counters containing Small Amounts of Polyatomic Constituent—A. D. Krumbein. (*Rev. Sci. Instr.*, vol. 22, pp. 821-827; November, 1951.)

621.395.625.3 1705
Ferrography—R. B. Atkinson and S. G. Ellis. (*Jour. Frank. Inst.*, vol. 252, pp. 373-381; November, 1951.) "Ferrography" is the name given to a new magnetic process, here described, for recording graphic information and reproducing it on paper in visual form. A scanning process similar to that used in facsimile produces an electrical signal which is fed to a magnetic recording head, and a record is made on a film coated with iron oxide. Printed reproductions are made using magnetic inks, either black or colored. The record can be used repeatedly and stored indefinitely.

621.398:621.396.712:621.396.619.13 1706
Remote-Control System for F.M. Broadcoast Stations—P. Whitney. (*Tele-Tech*, vol. 10, pp. 32-35 and 44-45, 80; August and September, 1951.) An effective remote-control system has enabled the WRFL transmitter, installed on a mountain peak, to be controlled from equipment in the studios at Winchester, Va., more than 20 miles away. No operators have been in regular attendance at the transmitter since April 1951. A general description is given of the equipment, with detailed circuit diagrams of the control oscillators, band-pass amplifier, automatic protective circuits and telemetry equipment. The protective equipment was found very necessary owing to frequent interruption of the transmitter power supply due to line surges caused by electrical storms, which operated the main circuit breaker. A different type of breaker, together with a recycling device operating 3-5 times within a few seconds, eliminated this difficulty.

621.52 1707
Fundamentals of Automatic Control. [Book Review]—G. H. Farrington. Publishers: Chapman & Hall, London, Eng., 1951, 285 pp., 30s. (*Electronic Eng.*, vol. 23, p. 453; November, 1951.) Gives a thorough analysis of the fundamentals of process control, but does not deal in detail with servomechanisms as such.

PROPAGATION OF WAVES

538.566 1708
The Jumps of Discontinuous Solutions of the Wave Equation—H. Bremmer. (*Commun. pure appl. Math.*, vol. 4, pp. 419-426; November, 1951.) There is a correspondence between discontinuous changes in an em field, with their associated wavefronts, and the amplitudes of the geometrical-optical approximations which correspond, under steady-state conditions, to ray trajectories orthogonal to the wavefronts. The theory of this correspondence is developed in terms of Dirac's impulse function for any discontinuity connected with the scalar wave equation. Modifications to the theory required in the solution of vector problems related to the application of Maxwell's equations are given.

538.566 1709
The Transport of Discontinuities in an Electromagnetic Field—E. T. Copson. (*Commun. pure appl. Math.*, vol. 4, pp. 427-433; November, 1951.) Generalized solutions of the vector wave equations are obtained and the transport equations are derived, the medium being assumed isotropic, with variable dielectric constant and permeability.

538.566.029.64 1710
Asymptotic Solutions of a Differential Equation in the Theory of Microwave Propagation—R. E. Langer. (*Commun. pure appl. Math.*, vol. 3, pp. 427-438; December, 1950.) "The purpose of this paper is to show that asymptotic formulas for the solutions of a differential equation that is central to the theory of microwave propagation may be readily derived from results that are available in the mathematical literature." The problem of determining the normal modes of propagation in an atmosphere in which the refractive index varies only with the height is first briefly reviewed. The differential equation under conditions applying respectively to the "leaky" and "transitional" modes of propagation is discussed. In each case the results are compared with the analogous results of Pekeris (2211 of 1947), obtained by power-series methods.

621.396.11 1711
The Propagation of E.M. Waves from Land to Sea and vice versa: Part 1—P. Höller. (*Z. angew. Phys.*, vol. 3, pp. 424-432; November, 1951.) An analytical approximation method of investigation is used in which the wave equation is first satisfied while neglecting the boundary conditions, and the boundary conditions are then satisfied while neglecting the wave equation. It is assumed that (*a*) the earth is flat, (*b*) sea and land are individually homogeneous and the boundary is sharp, (*c*) the sea is an ideal conductor, (*d*) the complex

refractive index of land is not very large, so that the simple Weyl solution is applicable, (e) the transmitter height is sufficient for refraction and reflection at the earth's surface to be calculated by geometrical optics, (f) the distances of both transmitter and receiver from the coast are large compared with the wavelength.

621.396.11 1712
Propagation of Very-High-Frequency Radio Waves—E. H. Jones. (*Nature* (London), vol. 168, pp. 870–871; November 17, 1951.) Measurements are reported of the strength of the field at a height of 90 ft. due to an airborne transmitter at a height of 40,000 ft.; operating frequencies of 280.2 and 386.6 mc were used. Observed values are plotted against transmitter/receiver distance and compared with values calculated from conventional ray theory; the agreement is generally better when the earth's radius is taken at its actual value rather than at its four-thirds value. Where the indirect ray was reflected from the sea the observed minima were deeper than the calculated values, the difference corresponding to an apparent increase in reflection coefficient by a factor as great as 2.25 in some cases.

621.396.11:551.510.535 1713
Ionosphere Review: 1951. Greatly Reduced Rate of Decrease in Sunspot Activity and M.U.Fs—Bennington. (see 1605.)

621.396.11:551.510.535 1714
R.F. Time-Delay Measurements—D. Davidson: R. Naismith and E. N. Bramley. (*Wireless Eng.*, vol. 29, pp. 111–112; April, 1952.) Comment on 473 of March and authors' reply.

621.396.81:523.72 1715
Solar Activity and Inospheric Effects—R. E. Burgess and C. S. Fowler. (*Wireless Eng.*, vol. 29, pp. 46–50; February, 1952.) Ionospheric disturbances on short and long waves were investigated by recording the field strengths of two stations on frequencies of 18.89 mc and 191 kc respectively. These results were compared with solar activity in the form of flares and sunspots and with simultaneous recordings of solar noise on frequencies of 30, 42, 73 and 155 mc. The times of commencement of disturbances on short and on long waves and of solar flares were all coincident and varied by up to 3 minutes from the noise bursts which often preceded the other phenomena. 86 per cent of the noise bursts and 50 per cent of the observed flares occurred without any accompanying ionospheric effect.

621.396.812.029.64 1716
Attenuation of Radio Signals caused by Scattering—J. B. Smyth and C. P. Hubbard: A. H. LaGrone. (*Jour. Appl. Phys.*, vol. 22, pp. 1386–1387; November, 1951.) Comment on 2525 of 1951 and reply by one of the authors.

621.396.812.3.029.64 1717
An Experimental Study of Fading in Propagation at 3-cm Wavelength over a Sea Path—D. G. Kiely, and W. R. Carter. (*Proc. IEE*, Part III, vol. 99, pp. 53–60; March, 1952.) The observations were made between July 1950 and January 1951 over an optical path of 10.6 nautical miles, using horizontal polarization. Associated meteorological observations were made at points not far from the transmission path. Typical signal records are shown and the data analyzed in terms of the monthly mean level. No reliable prediction of fading could be made from the simple meteorological observations. The reduction in radar range due to fading reached a maximum in July and August, being up to 20 per cent for 90 per cent of the time. The power level of radar beacons for operation up to the radio horizon is estimated.

RECEPTION

621.396.621:621.396.822 1718
On Determining the Presence of Signals in Noise—I. L. Davies. (*Proc. IEE*, Part III, vol. 99, pp. 45–51; March, 1952.) A theoretical treatment using the concept of existence probability, which is evaluated for a signal whose form is known precisely and for a modulated carrier where only the carrier phase is unknown. The efficiency of any receiver can be derived by comparing existence probabilities at input and output; the practical implications of this are discussed. When applied to the case of a radar signal, the theory shows that the filter giving the optimum range information will also yield the greatest existence information.

621.396.621.54 1719
Calculation of the Sensitivity of Decimetre- and Centimetre-Wave Receivers using Diodes or Detectors as Mixers—H. Behling. (*Arch. elekt. Übertragung*, vol. 5, pp. 489–498 and 561–564; November and December, 1951.) Using the concept of equivalent noise resistance, formulas are developed for calculating the critical sensitivity (corresponding to unity signal/noise ratio); from these and the characteristics of the mixing circuit (mixer tube, first IF tube, IF circuit etc.) it is possible to calculate the oscillator power required to obtain unity signal/noise ratio at the outupt of the IF amplifier. This calculation is made for a crystal mixer, and the influence on the sensitivity of the mixer and IF parameters is examined.

STATIONS AND COMMUNICATION SYSTEMS

517.512.2:621.39.001.11:621.396.67.012.71 1720
Fourier Analysis and Negative Frequencies—M. L. Teles. (*Wireless Eng.*, vol. 29, p. 80; March, 1952.) Discussion on 1422 of June (Shaw).

621.317.35:621.3.015.7 1721
Analysis of Non-Recurrent Pulse Groups—L. S. Schwartz and N. P. Salz. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 8–10; November, 1951.) Graphical determination of the resultant frequency spectrum of non-recurrent groups of regularly spaced pulses such as occur in teletype and pcm transmissions.

621.317.35:621.39.001.11 1722
Signals of Given Duration and Minimum Spectral Width—K. Fränz. (*Arch. elekt. Übertragung*, vol. 5, pp. 515–516; November, 1951.) Using Fourier integrals, an equation is derived whose solution gives the form of signal, for a given duration, for which the energy concentration within a given frequency band has its maximum value. The spectral concentration of energy in a rectangular pulse is very near the maximum. See also 2376 of 1951.

621.39.001.11 1723
The Concept of Information and Transmission Capacity in Communication Technique—H. Weber. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 29, pp. 401–406; November 1951. In German.) Discussion of a coding system for German text, making use of Shannon's theory.

621.39.001.11:519.251.6 1724
Information Theory and Inverse Probability in Telecommunication—P. M. Woodward and I. L. Davies. (*Proc. IEE*, Part III, vol. 99, pp. 37–44; March, 1952.) "The foundations of information theory are presented as an extension of the theory of inverse probability. By postulating that information is additive and taking suitable averages, all the essential definitions of Shannon's theory for discrete and continuous communication channels, with and without noise, are obtained. The theory is based on the idea that receiving a communication, or making an observation, merely changes the relative probabilities of the various possible messages. The whole process of reception can therefore be regarded as a means of evaluating *a posteriori* probabilities, and this leads to the idea that the optimum receiver in any telecommunication problem can always be specified, in principle, by inverse probability. The simplest instance is the correlation receiver for detecting very weak signals in the presence of noise, and its theory is briefly discussed."

621.395.44 1725
Carrier-Frequency Systems free from Linear Distortion—J. Peters. (*Arch. elekt. Übertragung*, vol. 5, pp. 509–515; November, 1951.) Theory of the modulation and demodulation processes in a ssb system is considered; rigorous conditions for freedom from distortion are derived by using the Laplace transformation. A network can be found which will transmit the envelope free from distortion for a sine-voltage input initiated at any phase over the part of the cycle when the voltage is increasing; for initiation at other parts of the cycle the output contains an additional decaying direct component. Distortion-free transmission can still be obtained in this case by a tandem arrangement of circuits designed according to the analysis given.

621.395.44:621.395.97 1726
New Control and Measurement Equipment for the Broadcasting [-network] Repeater Stations of the German Post Office—E. A. Pavel and H. Liersch. (*Fermeldetech. Z.*, vol. 4, pp. 513–518; November, 1951.) Description of station equipment for line tests, control switching and program monitoring, comprising the test rack type 48 and associated loudspeakers. See also 1238 of June (Pavel *et al.*).

621.396.361.1 1727
Radio Communications in the Australian Flying Doctor Service—L. N. Schultz. (*Proc. I.R.E.*, (Aust.) vol. 12, pp. 300–302; October, 1951.) Description of a system which provides subscribers distant more than 25 miles from a telephone with transceiver facilities. Three operating frequencies, of about 2, 4, and 7 mc, are allotted to each of the eight bases together with the associated outposts, the actual frequencies being different for each group. Base-transmitter power ranges from 20 w to 300 w. In addition to normal communication receivers, the bases have a night-alarm receiver. Outpost transmitters have an output of 3 w and are powered by vibrators, except for a few old stations with pedal-driven generators. Outposts use three fixed frequencies together with a variable tuning band for hf reception.

621.396.44:621.315.052.63]+621.317.083.7 1728
Single Sideband Transmission and its Multiple Utilization for Carrier-Current Channels on High-Voltage Power Lines—A. de Quervain (*Brown Boveri Rev.*, vol. 38, pp. 208–219; July/August, 1951.) See also 801 of April (Block).

621.396.5:621.396.931 1729
450-Mc/s Mobile Radio Service—N. E. Wunderlich. (*FM-TV*, vol. 11, pp. 22–25, 38; November, 1951.) Description of equipment for a dispatch system for at least 1,000 taxis operating in Chicago. Five pairs of channels 100 kc wide are used. Eight 15-w or 100-w phm transmissions cover eight zones each roughly 5 miles square. The mobile transmitter-receiver units have a frequency stability of ± 5 parts in 10^6 and can be operated on any of four switch-selected frequencies. Communication on 452 mc is superior to that on 162 mc in built-up areas; the average noise level is 10 db lower and no ignition interference is experienced, while the walls of buildings apparently serve as waveguides, thus reducing attenuation.

621.396.65 1730
New Pennsylvania Turnpike U.H.F. Communications System—D. N. Lapp. (*Elec-*

tronics, vol. 25, pp. 84–87; February, 1952.) Description of a system operating over a 327-mile route between the outskirts of Philadelphia and the Ohio border, with 13 intermediate stations. Frequencies of 953 mc and 960 mc are used along the main route, and frequencies in the band 152–162 mc are used for local communication networks connected to the relay stations.

621.396.65:621.387.4 1731

Use of a Radio Link for studying Coincidences between Pulses from Counters separated by Large Distances—E. Picard, A. Rogozinski and M. Surdin. (*Jour. Phys. Radium*, vol. 12, pp. 854–857; November, 1951.) Equipment is described for a link operating on 10 kmc, with a range of about 20 km. When there is no line-of-sight path between the two stations a frequency of 5 mc is used.

621.396.65.029.63:621.316.726 1732

1400-Mc/s Radiophone—J. B. L. Foot. (*Wireless World*, vol. 58, pp. 132–135; April, 1952.) An experimental radio link is described which uses a DET23 disk-seal triode tube operated to give an output of about 1 w. The range is up to about 30 miles with a signal/noise ratio of about 20 db. Each transmitter oscillator acts also as local oscillator for the double-superheterodyne receiver, a frequency difference equal to the first IF being maintained between the oscillators at the two stations. A "master and slave" system is used to keep this frequency difference constant in spite of oscillator frequency drift, the frequency of the slave station being controlled by an electromechanical method in accordance with the frequency of the signal received from the master station.

621.396.7.029.62+621.396.621]:621.396.619.13 1733

F.M. in Germany—(*Wireless World*, vol. 58, pp. 141–144; April, 1952.) Factors which have influenced the development of the FM vhf broadcasting network in Western Germany are discussed. The disposition of transmitting stations is shown. Some details are given of simple FM receivers which dispense with the use of limiter and discriminator.

621.396.712:621.396.619.13:621.398 1734

Remote-Control System for F.M. Broadcast Stations—Whitney. (See 1706.)

621.396.712.029.55 1735

The Allouis-Issoudun H.F. Group of Radiodiffusion Française—A. Gaillard. (*Onde élect.*, vol. 31, pp. 420–433; November, 1951.) Description of equipment for the sw broadcasting service which in 1952 will include thirteen 100-kw transmitters, grouped at two centers. At Allouis one 100/130-kw transmitter operating in the 31, 41 and 49 m bands can be switched to any one of six dipoles oriented in N-S and E-W directions; two other units provide four simultaneous transmissions in the bands from 13 to 49 m from a system of twelve directive rhombic antennas. At Issoudun twelve simultaneous transmissions can be made. Features dealt with include the services now operating, station layout, power supply and tube cooling, circuit arrangement and tubes, and transmission characteristics. A map and chart show the world-wide coverage achieved.

621.396.712(489) 1736

Broadcasting Installations for the Two Programmes in Denmark (*Teleteknik*, (Copenhagen), vol. 2, pp. 207–249; October, 1951.)

- The Planning of the New Broadcasting Stations—G. Pedersen.
- Propagation Conditions for the V.H.F. Range—B. Nielsen.
- The Broadcasting Station at Skive—G. Bramslev.
- The 100-kw Medium-Wave Transmitter at Kalundborg—G. Bramslev.
- Transmitters working on the International Shared Frequencies and New Stations for F.M.—P. Christensen.
- Directive Aerials in Herstedvester—J. Hansen.
- Construction and Erection of Aerial Masts—I. G. Hannemann and B. J. Rambøll.

621.396.931:621.395.635 1737

Selective Calling Applied to Mobile Radio—W. T. Muscio. (*Proc. I.R.E.*, (Aust.), vol. 12, pp. 303–311; October, 1951.) Detailed description of the Selecto-Call system. The fixed station has a choice of eleven frequencies in the range 154–449 cps for AM of a 7-kc subcarrier, the main carrier being frequency modulated. The mobile receiver includes a decoder comprising up to four reeds turned to different transmitter tones; actuation of the needs in a given sequence is required to produce the calling signal.

SUBSIDIARY APPARATUS

621–526 1738

Stabilization of Direct-Current Servomechanisms—M. Cambornac and F. Lajeunesse. (*Onde élect.*, vol. 31, pp. 434–445; November, 1951.) Methods of improving stability are discussed. These include the use of phase-correcting networks and the coupling of an auxiliary dynamo to the servo motor. Circuits and characteristics of different low-power models are given, the response times of which range between 0.01 and 0.2 sec.

621–526:621.396.93 1739

Analysis and Construction of a Position-Fixing Servomechanism—G. Klein. (*Ann. Télécommun.*, vol. 6, pp. 313–324; November, 1951.) The mechanism is designed to operate with the direction discriminator [895 of 1950 (Loeb et al.)] so that a direct reading of azimuth is obtained automatically. A theoretical analysis defines the specifications for the system.

621.311.6 1740

Carrier-Type Regulated Power—J. Houle. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 14–15. .31; November, 1951.) A continuously variable dc output from 0 to 300 v with regulation to within 0.1 v is obtained, using the following principle. An oscillator feeds a small signal to an ac amplifier. In series with the amplifier input is a Ge diode connected so that the dc control signal varies the diode bias current. At the amplifier output, the dc is recovered by means of a rectifier.

621.311.6 1741

Regulated 1600-Ampere Filament Supply—A. W. Vance and C. C. Shumard. (*Electronics*, vol. 25, pp. 122–123; February, 1952.) Description of circuit for the 6-V filaments of some 4,000 tubes used in the Project Typhoon analogue-digital computer of the U. S. Navy.

621.311.6:621.316.722.1 1742

A Stabilized A. C. Supply for Lamps and Valve Heaters—J. C. S. Richards. (*Jour. Sci. Instr.*, vol. 28, pp. 333–335; November, 1951.) Description of a system providing up to 150 w of ac power with the rms voltage stable to within ±0.1 per cent. A saturated diode is used as reference element and a saturable choke as control element.

621.311.62.078.3 1743

A High-Stability High-Voltage Power-Supply Unit—J. Templeton. (*N. Z. Jour. Sci. Tech. B*, vol. 33, pp. 218–223; November, 1951.) Description of a mains-operated unit supplying 1 ma at 3kv steady to within ±1 part in 10^4 over periods of 30 min, and capable of modification to give greater output.

621.314.6 1744

Rectification of Alternating Currents with High Modulation—J. Böhse. (*Arch. elekt. Übertragung*, vol. 5, pp. 363–376; August, 1951.) From an approximate equation for the static characteristic of rectifiers with essentially constant differential slope, equations for the demodulation products (direct-current and modulation-frequency) are derived for the case of high modulation, when the rectifier operates like a tube. The operational parameters, currents, voltages and effective resistances on the input and output sides, as well as the demodulation distortion, are represented by transcendental functions of the phase angle of the current, so that numerical values of the electrical quantities can be taken directly from tables and monograms, whose use is explained by several examples.

621.314.6.012.6 1745

Exact Analysis of the Linear Rectifier Circuit: Part 1—Half-Wave Rectification with with Capacitive Smoothing—H. Niehrs. (*Frequenz*, vol. 5, pp. 273–279; October, 1951.) Formulas are derived for the amplitude and phase of the harmonics and the effective output voltage of a half-wave rectifier with sine-wave input, assuming negative half cycles to be completely blocked and internal resistance of the rectifier to be constant.

621.314.653 1746

The Time Lag of an Ignitron—N. Warmoltz (*Philips Res. Rep.*, vol. 6, pp. 388–400; October, 1951.) Measurements were made of the time lag for igniters of widely different resistance, using (*a*) liquid, (*b*) solid Hg or Sn cathodes. The effect of the gas pressure in the tube was also investigated. The results favor the thermal theory of Mierdel.

621.316.722.1 1747

An Electronic Voltage Stabilizer with Self-Regulated Heater Supply—C. Morton. (*Electronic Eng.*, vol. 24, p. 65; February, 1952.) The heaters of the tube cathodes are connected in series with the load, thus constituting part of the output circuit across which the stabilized voltage is developed.

621.316.722.1 1748

Voltage Stabilization: Demands and Methods—A. J. Maddock. (*Jour. Sci. Instr.*, vol. 28, pp. 325–333; November, 1951.) Typical cases are discussed in which stabilized supplies are required. The principal types of stabilizer are described and details are given of their performance and their voltage and power ratings. Mains generator control is not considered.

TELEVISION AND PHOTOTELEGRAPHY

621.397.335:535.623 1749

N.T.S.C. Color-TV Synchronizing Signal—R. B. Dome. (*Electronics*, vol. 25, pp. 96–97; February, 1952.) Discussion of the synchronizing signal required in the field tests on the band-sharing system (1750 below) in which a local oscillation synchronized to the frequency of the color subcarrier is used for demodulating the color signal at the receiver. The signal chosen is a train of about 10 cycles of the color subcarrier frequency, timed to occur about midway between the end of the horizontal synchronizing pulse and the end of the blanking pulse.

621.397.5:535.623 1750

Principles of N.T.S.C. Compatible Color Television—C. J. Hirsch, W. F. Bailey and B. D. Loughlin. (*Electronics*, vol. 25, pp. 88–95; February, 1952.) Discussion of the specifications formulated by the U. S. National Television System Committee to govern field tests on the system in which a separate color signal is transmitted simultaneously with a monochrome signal on a separate subcarrier within the 4-mc channel carrying the monochrome signal [826 of April (Loughlin)]. Factors affecting the choice of the color-subcarrier frequency and the modulation system are considered.

621.397.6:621.396.664]+621.396.615.14 1751
Ultra High Instrumentation—Peters. (See 1682.)

621.397.611/.621].2 1752
Scanning-Current Linearization by Negative Feedback—A. W. Keen. (*Jour. Telev. Soc.* vol. 6, pp. 308–315; October/December, 1951.) "An introductory qualitative survey of the known methods of applying negative feedback to the problem of linearizing the output current of scanning systems needed in television transmitting and receiving equipment. Practical circuit developments are divided into two categories according as the fundamental process, termed 'current integration,' is separable into two component operations, viz., voltage integration and voltage-to-current conversion, which are carried out consecutively by separate feedback systems connected in cascade, or performed by a single system subject to over-all feedback."

621.397.611.2 1753
Paraxial Image Formation in the "Magnetic" Image Iconoscope—J. C. Francken and R. Dorrestein. (*Philips Res. Rep.*, vol. 6, pp. 323–346; October, 1951.) Describes a method of computing electron trajectories near the axis of a system of es and em fields having rotational symmetry. Numerical results are given for a case which approximates to conditions in the "magnetic" image iconoscope. The mechanism of image formation is discussed; it differs considerably from that in ordinary magnetic lenses.

621.397.611.2 1754
The Image Iconoscope, A Camera Tube for Television—P. Schagen, H. Bruining and J. C. Francken. (*Philips Tech. Rev.*, vol. 13, pp. 119–133; Novmeber, 1951.) The different types of television camera tube are discussed and a detailed description is given of the Philips image iconoscope Type 5854. The Cs-Sb-O photocathode gives an output of about 45 μA/lumen when illuminated by an incandescent lamp with color temperature 2,600°K. Magnetic focusing and deflection are used for the scanning beam, the resolution being 900–1,000 lines at the middle of the target and as high as 700 lines at the edges. The mica target is 25 μ thick and has a thin coating of MgO to increase secondary emission. See also 1452 of June (Francken).

621.397.611.2 1755
Television Camera Tube—R. Barthélemy. (*Onde élect.*, vol. 31, pp. 415–419; November, 1951.) Theory previously given (1499 of 1951) is discussed with reference to (*a*) the use of a thin-film target with its potential adjusted by an auxiliary electron stream, (*b*) the optimum thickness of the self-polarized target film. A few performance details are given of the super-iconoscope with es deflection and self-polarized target.

621.397.62 1756
Basic Circuit Description of a R.C.A. Television Receiver—*Radiotronics*, vol. 16, pp. 211–224 and 228–243; October and November, 1951.) Description of the R.C.A. Victor 630TS receiver, with full circuit details and analysis of the functions of the various circuits.

621.397.62 1757
Wide Angle Deflection Yokes—H. E. Thomas. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 3–6; September, 1951.) A general discussion of factors to be considered in designing systems with deflection angles up to 90° and with low distortion. The length of the yoke is determined as a compromise between the requirements for high sensitivity and those for avoidance of neck shadow. Auxiliary magnetic devices for eliminating neck shadow are described. Methods of obtaining an optimum relation between spot distortion and pattern distortion are indicated.

621.397.62:[535.623+535.61–29 1758
C.B.S. Columbia—First Commercial Color plus Black-and-White Set—I. J. Melman, E. S. White and S. Cuker. (*Radio-Electronics*, vol. 23, pp. 24–27; November, 1951.) Description of a receiver for 525-line black-and-white or 405-line full-color pictures. Normal circuits are used together with a color scanning disk with silent motor drive and associated disk-control circuit.

621.397.62:621.396.662 1759
Concentric-Lines tune U.H.F. Channels—E. E. Harries and M. Cawein. (*Electronics*, vol. 25, pp. 108–112; February, 1952.) Circuit and construction details are given of a converter for use with a conventional vhf television receiver; a three-section tuner of Inductuner type is used, and the frequency range 470–890 mc is covered. The circuit consists of preselector, crystal mixer and oscillator, followed by an IF stage. The noise figure of the converter is discussed.

621.397.62:621.398 1760
Remote Controls for TV promote Viewer Comfort—R. F. Scott. (*Radio-Electronics*, vol. 23, pp. 28–31; November, 1951.) Description of electromechanical and electronic systems incorporated in various television receivers which permit control from a convenient viewing point some distance from the set.

621.397.621:621.316.721.078.3 1761
Stabilizing Vertical-Deflection Amplifiers—W. B. Whalley, C. Masucci and K. Hillman. (*Electronics*, vol. 25, pp. 116–117; March, 1952.) Application of inverse feedback to the vertical-deflection amplifier makes vertical linearity and picture-height stability practically independent of tube transconductance.

621.397.621:621.317.35 1762
Television Picture Line Selector—J. Fisher. (*Electronics*, vol. 25, pp. 140–143; March, 1952.) Description of equipment enabling examination of the video wave form in a single selected scanning line. The oscilloscope is triggered with a single horizontal-synchronization pulse which precedes the line to be observed. Application to measurement of frequency response and transient response of television cameras and picture-generating devices, such as the monoscope and flying-spot scanner, is described.

621.397.621.2 1763
Evaluating Performance of TV Picture Tubes—J. Green (*Electronics*, vol. 25, pp. 124–129; February, 1952.) Methods and apparatus are described for accurately determining the vertical and horizontal dimensions of the spot, using special rasters; the significance of the measurements in relation to the design of the various parts of the cr tube is discussed.

621.397.621.2:621.385.832 1764
TV picture Tubes with Iron Envelopes—Szegho and Pohl. (See 1790.)

621.397.8 1765
Television Ghosts. Effect of Multi-path Propagation in Hilly Country—J. A. Hutton. (*Wireless World*, vol. 58, pp. 84–88; March, 1952.) An investigation of reception with twelve different antennas in hilly country round the Holme Moss transmitter. A standing-wave field due to reflection from hills two or three miles away may give rise to positive or negative ghost images according to the time delay and the modulation amplitude. The ghost image is least apparent when the antenna is between a node and antinode of the standing-wave field. The double-H antenna systems consisting of two H antennas $\lambda/2$ apart, with $\lambda/4$ spacing of the dipoles, was found the best of the antennas tested.

621.397.822:621.397.2 1766
Random Noise. Rate of Occurrence of Peaks—V. J. Francis. (*Wireless Eng.*, vol. 29, pp. 37–40; February, 1952.) The number of peaks occurring above various amplitude levels on a noise trace is calculated and the results are compared with those of laboratory experiments on a system simulating the London-Birmingham television radio relay link. The results are related to the perceptible level of noise peaks on actual television pictures.

621.397.822.1 1767
Observer Reaction to Video Crosstalk—A. D. Fowler. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 57, pp. 416–424; November,1951.)

TRANSMISSION

621.396.615.16.029.55:621.396.933 1768
Brown Boveri Transmitters in the Service of Civil Aviation—A. Vincre. (*Brown Boveri Rev.*, vol. 38, pp. 220–226; July/August, 1951.) SW transmitters installed for the French Ministry of Transport and Public Services meteorological service and for Air France overseas communications are described.

TUBES AND THERMIONICS

537.533.8 1769
The Angular Distribution of the Secondary Electrons of Nickel—J. L. H. Jonker. (*Philips Res. Rep.*, vol. 6, pp. 372–387; October, 1951.) Description of the measurement tube, and graphical presentation of results obtained for the distribution of the secondary electrons as a function of the angle of incidence and the voltage of the primary electrons.

621.383:546.482.21 1770
Photoelectric Cells using Activated Cadmium Sulphide—P. Goercke. (*Ann. Télécommun.*, vol. 6, pp. 325–331; November, 1951.) The photoelectric properties of DdS may be enhanced by addition of traces of Cu or Ag. The influence of this impurity content on spectral sensitivity, electrical resistance and noise figure is studied.

621.383.4:546.817.231 1771
The Long-Wave Limit of Infra-Red Photoconductivity in PbSe—A. F. Gibson, W. D. Lawson and T. S. Moss. (*Proc. Phys. Soc.*, vol. 64, pp. 1054–1055; November 1, 1951.) Measurements of the photoconductivity of a PbSe photo-diode, consisting of a *p*-type crystal with a tungsten whisker, show that the long-wave limit, defined as 50 per cent decrease from the maximum sensitivity, was 4.7 μ at room temperature and 6.8 μ at 90°K. Another cell with a periclase window had a long-wave limit of 8.1 μ at 20°K.

621.384.5:621.316.722 1772
The Characteristics of some Miniature High-Stability Glow-Discharge Voltage-Regulator Tubes—F. A. Benson. (*Jour. Sci. Instr.*, vol. 28, pp. 339–341; November, 1951.) Three types of tube were studied. The results of short- and long-term tests to determine striking-voltage and running-voltage variations are presented. Values are also given for temperature coefficient of running voltage and for the magnitudes and durations of the initial drifts. See also 3159 of 1951.

621.385 1773
The Measurement of Microphony in Valves—R. Bird. (*Electronic Eng.*, vol. 23, pp. 429–431; November, 1951.) Arrangements are described for investigating microphony in

a tube by using it as the input tube of an af amplifier and locating it in the sound field of a loudspeaker fed by the amplifier. Optical and electrical methods of detecting the vibrating elements are discussed.

621.385-71 1774
Electron-Tube Heat-Transfer Data—B. O. Buckland. (*Elec. Eng.* (N. Y.), vol. 70, pp. 962-966; November, 1951.) Essentials of 1951 A.I.E.E. Summer General Meeting paper. The method of calculating heat flow by means of equivalent electrical circuits is considered. The effects of temperature differences and cooling-fin shape on radiation- and air-cooled tubes are discussed and design considerations for water-cooled and forced-air-cooled tubes are summarized. Graphs are given from which the data necessary for the design of cooling systems may be determined.

621.385:537.525.92 1775
Two-Way Space-Charge Flow with Plane Electrodes—K. Müller-Lübeck. (*Z. angew. Phys.*, vol. 3, pp. 409-415; November, 1951.) The potential in the space-charge field in the presence of electrons and positive ions was found graphically by Langmuir (1929 Abstracts, p. 511); a rigorous analytical solution is now derived for this potential.

621.385.004.15 1776
The Technique of Trustworthy Valves—E. G. Rowe (*Jour. Brit. I.R.E.*, vol. 11, pp. 525-540; November, 1951. Discussion, pp. 540-543.) A survey of progress in the design, manufacture and testing of radio tubes to ensure high reliability. Failures occurring in manufacture and during the subsequent life of the tube are discussed, and the need for increased cooperation between user and manufacturer is stressed.

621.385.004.15 1777
A Survey of Quality and Reliability Standards in Electronic Valves for Service Equipment—G. L. Hunt. (*Jour. Brit. I.R.E.*, vol. 11, pp. 519-524; November, 1951. Discussion, pp. 540-543.) Conditions of use of tubes in Service equipment are described and also methods adopted for ensuring supplies of satisfactory tubes.

621.385.012:621.317.755 1778
Electron Tube Curve Tracer—Kuykendall. (See 1678.)

1779
621.385.029.6:538.311:621.318.423:515.647.1
Properties of the Electromagnetic Field of Helices—Roubine. (See 1580.)

621.385.029.63/.64 1780
Equivalent Temperature of an Electron Beam—M. E. Hines: P. Parzen. (*Jour. Appl. Phys.*, vol. 22, pp. 1385-1386; November, 1951.) Comment on 2580 of 1951 and reply by one of the authors.

621.385.032.216 1781
The Life of Oxide Cathodes in Modern Receiving Valves—G. H. Metson, S. Wagener, M. F. Holmes and M. R. Child. (*Proc. IEE*, Part III, vol. 99, pp. 69-81; March, 1952. Discussion pp. 82-87.) A summary of existing information, with some results of original research. If mechanical faults, the effect of gas on cathode emission, and excessive interface feedback can all be avoided, tube life is probably limited only by evaporation of the activated oxide. 24 references.

621.385.032.216:539.16 1782
Contribution to the Study of Electronic Tubes by the Use of Radioactive Elements—J. Debiesse and G. Neyret. (*Le Vide*, vol. 6, pp. 1098-1102; November, 1951.) Radioactive isotopes are used for getter and cathode materials to facilitate the investigation of the distribution and migration of the materials and the origin of the electron emission. Radioactivity of the metal base in general reduces the thermionic emission from the cathode. See also 2301 of 1951 (Debiesse *et al.*)

621.385.2 1783
Influence of Initial Velocities on Electron Transit Times in Diodes—J. T. Wallmark. (*Phys. Rev.*, vol. 84, p. 598; November 1, 1951.) Barut (2057 of 1951) has described a method for calculating the transit time of electrons in diodes with partial space-charge, assuming a uniform initial-velocity distribution. A first-order perturbation method has been applied to the calculation of transit times for electrons with nonuniform initial-velocity distribution. Curves are given showing the spread in transit time (*a*) for a diode with anode voltage 100 v and anode-cathode distance of 5 mm, as a function of current density in the diode for a difference in initial velocity of 0.1 v (roughly corresponding to conditions for an oxide cathode at 1,100°K), (*b*) for a reflected beam under the same conditions, the spread in transit time being in this case much reduced. A complete report is in preparation.

621.385.2 1784
The Transit-Time Effect in a Cylindrical Diode—Way Dong Woo. (*Jour. Appl. Phys.*, vol. 22, pp. 1333-1339; November, 1951.) The impedance offered to a small superposed alternating current by a space-charge-limited cylindrical diode is expressed in terms of the variational anode conductance, the transit-time angle, and a group of constants which are functions of the ratio of the anode and cathode radii. The result covers the range of transit-time angle up to π radians and of the ratio of the anode and cathode radii from unity up to 100.

621.385.2:546.289]+631.314.7 1785
Germanium Crystal Valves—B. R. A. Bettridge. (*Electronic Eng.*, vol. 23, pp. 414-417; November, 1951.) Characteristics of Ge diodes are outlined, and television circuit applications described. Ge triodes are mentioned briefly.

621.385.2:621.318.572 1786
R. F. Bursts actuate Gas-Tube Switch—H. J. Geisler. (*Electronics*, vol. 25, pp. 104-105; February, 1952.) Description of technique for using simple gas-filled diodes as switches in storage and other circuits of electronic computers. Pulses of rf voltage applied to external metal bands round the diode envelopes cause the diodes to strike at reduced dc voltage. Examples of circuit applications are given.

1787
621.385.5.032.212:[621.318.572+681.142
The Single-Pulse Dekatron—J. R. Acton. (*Electronic Eng.*, vol. 24, pp. 48-51; February, 1952.) Description of a new type of gas-filled cold-cathode counting tube which differs from the earlier dekatrons [2066 of 1950 (Bacon and Pollard)] in requiring only a single input pulse to move the cathode glow on a complete step. The tentative specification is given for the GO1OD development tube, which is reliable for pulse rates up to 20,000 per sec. Input and output circuits are discussed in relation to the nature of the pulses dealt with.

621.385.832 1788
Cathode-Ray Tubes. A Review of Progress—L. F. Broadway. (*Proc. IEE*, Part I, vol. 98, pp. 316-320; November, 1951.)

621.385.832.:621.318.572 1789
New Electronic Tubes employed as Switches in Communication Engineering: Part 2—Switch Tubes—J. L. H. Jonker and Z. van Gelder. (*Philips Tech. Rev.*, vol. 13, pp. 82-89; October, 1951.) Experimental multicontact tubes are described in which the "contacts" are effected by means of secondary emission, the primary electron beam being directed electrostatically on to the various secondary-emission elements. The use of a ribbon-shaped primary beam permits currents of several milliamperes with voltages of 200 to 300 v, and tube dimensions can be kept small. Part 1: 1173 of May.

621.385.832:621.397.621.2 1790
TV Picture Tubes with Iron Envelopes—C. S. Szegho and R. G. Pohl. (*TV Eng.* (N.Y.), vol. 2, pp. 8-9, 27; November, 1951.) An envelope with an iron cone and Cr/Fe-alloy beads for sealing the cone to the glass parts is cheaper to make than one with a Cr/Fe-alloy cone. As a further development the Cr-Fe-alloy beads were eliminated and screen glasses were developed suitable for sealing to the iron cones, using an intermediate glaze at the sealing area; technique for this operation is described. The requirements for the neck glass are also discussed.

621.396.615.141.2:537.533.8 1791
Influence of Secondary Emission on the Oscillation Process in Whole-Anode Magnetrons—F. W. Gundlach and K. Schörken. (*Z. angew. Phys.*, vol. 3, pp. 416-424; November, 1951.) Measurements are reported of the cathode resistance, the back-heating current and the hf voltage of the oscillating magnetron as functions of anode voltage. Oscillation is maintained even when the external cathode-heater circuit is completely cut off, the cathode temperature being then such that no appreciable thermionic emission can occur. Analysis of the results leads to the conclusion that the magnetron is maintained by secondary emission.

621.385+621.396.6 1792
Introduction to Electronic Circuits. [Book Review]—R. Feinberg. Publishers: Longmans, Green & Co., London, Eng., 163 pp., 18s. (*Wireless Eng.*, vol. 29, p. 113; April, 1952.) "The treatment is, on the whole, satisfactory although some will find it rather compressed."

MISCELLANEOUS

061.3:621.396.029.63 1793
Report on the First I.R.E. U.H.F. Symposium—B. M. Ely. (*TV Eng.*, (N. Y.), vol. 2, pp. 10-13, 28; October, 1951.) Subjects dealt with at this symposium, held in Philadelphia, included transmission tests at 850 mc, apparatus for frequency and impedance measurements, a side-fire helical transmitting antenna, and the design of uhf receivers.

6:061.4 1794
British Instrument Industries Exhibition—London, 1951—M. W. Thring. (*Jour. Sci. Instr.*, vol. 28, pp. 293-300; October, 1951.) The development of the scientific instrument industry in Britain is reviewed, and some of the more important exhibits at the first exhibition are discussed.

621.396 1795
Radio Handbook. [Book Review]—Publishers: Editors and Engineers Ltd. Santa Barbara, California, 13th ed., 736 pp., $6.00. (*Electronics*, vol. 24, pp. 322, 326; December, 1951.) "Progressive and thorough in its coverage of the newest and most helpful developments."

621.396 1796
Radio Amateur's Handbook. [Book Review]—Publishers: American Radio Relay League, West Hartford, Conn., 20th ed., 618 pp., $3.00. (*Electronics*, vol. 24, pp. 322, 326; December, 1951.) "Contains what is still probably the most complete listing of communication types tubes to be found anywhere."