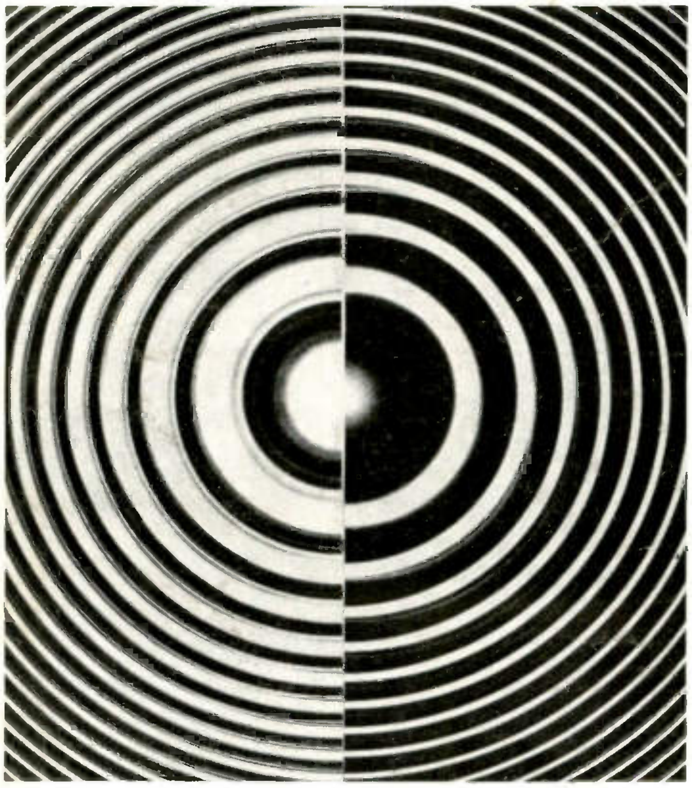
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering
(Including the WAVES AND ELECTRONS Section)

December, 1948
Volume 36 Number 12



First AAAS Photo-In-Science Salon

PRECISION AUTOGRAPH

At the left, interference fringes from green radiation of natural mercury; at the right, corresponding fringes from mercury 198, a man-made stable isotope produced by neutron bombardment of gold in an atomic pile. The mercury 198 spectral wavelengths may be accurate to *one part in a billion* and thus become a new standard of length!

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

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The Institute of Radio Engineers

COMPONENTS FOR EVERY APPLICATION


LINEAR STANDARD
High Fidelity Ideal


HIPERM ALLOY
High Fidelity . . . Compact


ULTRA COMPACT
Portable . . . High Fidelity


OUNCER
Wide Range . . . 1 ounce


SUB OUNCER
Weight 1/3 ounce


COMMERCIAL GRADE
Industrial Dependability


SPECIAL SERIES
Quality for the "Ham"


POWER COMPONENTS
Rugged . . . Dependable


VARITRAN
Voltage Adjustors

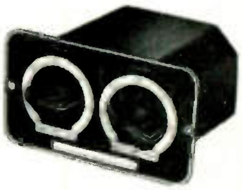

MODULATION UNITS
One watt to 100KW


VARIABLE INDUCTOR
Adjust like a Trimmer


TOROID HIGH Q COILS
Accuracy . . . Stability


TOROID FILTERS
Any type to 300KC


MU-CORE FILTERS
Any type 1/2 — 10,000 cyc.


EQUALIZERS
Broadcast & Sound


PULSE TRANSFORMERS
For all Services


HERMETIC COMPONENTS
Ceramic Terminals


HERMETIC COMPONENTS
Glass Terminals


GRADE 3 JAN
Components

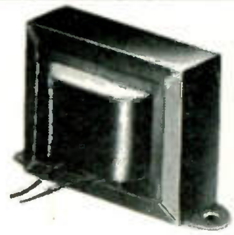

CABLE TYPE
For mike cable line


VERTICAL SHELLS
Husky . . . Inexpensive


REPLACEMENT
Universal Mounting


STEP-DOWN
Up to 2500W . . . Stock

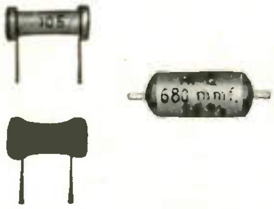

LINE ADJUSTORS
Match any line voltage


CHANNEL FRAME
Simple . . . Low cost

United Transformer Co.

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

Specify HI-Q COMPONENTS

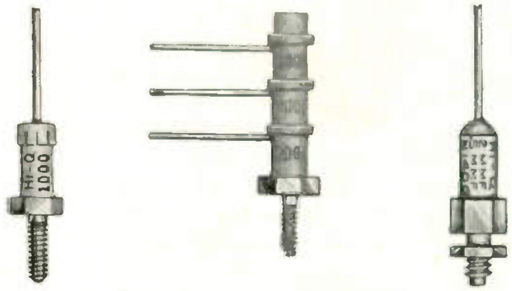


HI-Q TEMPERATURE COMPENSATING CAPACITORS

HI-Q temperature compensating capacitors are available in three types. CN & SI types with capacities from .25 mmf to 1830 mmf and CI types from .25 mmf to 595 mmf with a temperature coefficient range from P 100 to N 1400. All of these **HI-Q** styles are of tubular ceramic construction with pure silver electrodes precision coated. Style SI is insulated with a synthetic coating of Durez, style CN is of Styrene and CI is Steatite covered.

HI-Q GENERAL PURPOSE CERAMIC CAPACITORS

HI-Q General Purpose Ceramic Capacitors readily replace mica and paper condensers of corresponding values. **HI-Q** General Purpose Ceramic Capacitors should not be confused with the **HI-Q** line of close tolerance temperature compensating units. **HI-Q** General Purpose Ceramic Capacitors are available in capacity ratings from 5 mmf to 33,000 mmf.



HI-Q STAND-OFF CAPACITORS

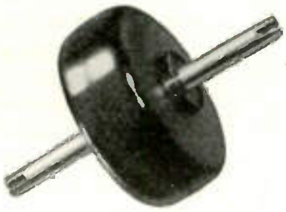
HI-Q "stand-off" capacitors are basically tubular with a screw fixture for mounting to the chassis or common ground. Close coupling and their unique construction make them an excellent choice for by-passing RF in the high frequencies. Standard capacity tolerances are ±10% and ±20% for "stand-off" capacitors and −20% and +30% for multiple tap units. Closer tolerances available wherever economical manufacturing permits. All units flash tested for 1000 volts DC with power factor under 3% maximum and insulation resistance is above 10,000 megohms. All units stamped for capacity.

HI-Q FEED-THRU CAPACITORS

HI-Q "feed-thru" capacitors provide perfect transmission through the chassis or ground, as well as by-passing to ground. The high quality construction of **HI-Q** "feed-thru" capacitors, is extremely rugged and will withstand severe vibration, making them ideal for use in mobile and aircraft applications.

HI-Q HIGH VOLTAGE CAPACITORS

HI-Q HV Capacitors are a sturdy unit, capable of withstanding high voltages, operating at extreme humidity and raised temperatures. They are a natural television component. The basic dielectric is body 20, encased in a low loss, mineral filled bakelite. Available in capacities 50 mmf to 1,000 mmf. Specify desired capacity after type HV when ordering.



HI-Q DISC CAPACITORS

HI-Q Disc Capacitors are high dielectric by-pass, blocking or coupling capacitors. Designed for application where its physical shape is more adaptable than tubular units. The placement of leads is such that close connections are easily made, thus reducing inductance to a minimum, a much desired feature in high frequency designs, such as television and FM. Available in three types: BPD-5: .005 MFD guar. min., BPD-10: .01 MFD guar. min. and BPD-1.5: .0015 MFD guar. min.

WRITE FOR FREE CATALOG

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FRANKLINVILLE, N. Y.

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

PROCEEDINGS OF THE I.R.E., December, 1948, Vol. 36, No. 12. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

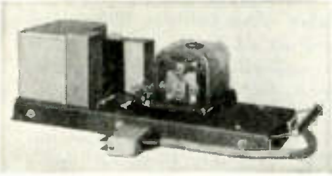
Table of contents will be found following page 32A

the voice that

STEPPING-STONES TO PROGRESS IN MARINE RADIOTELEPHONY

The first ship-to-shore radiotelephone communications were established almost 30 years ago between land stations at Green Harbor, Mass., and Deal Beach, N. J., and the steamers "Ontario" and "Gloucester," operating between Boston and Baltimore.

The "Leviathan" was the first ship to handle radiotelephone messages as a public service to and from land telephones.



This selector set made it possible to dial ships at sea, and eliminated the need for constant monitoring by loudspeaker or headphones.

IT'S COMMONPLACE TODAY to pick up a telephone on shipboard and talk to a business associate on land. But little more than 30 years ago, this was just a dream.

Back in 1915, the spoken voice could travel to far places only by wire. Then telephone scientists developed the radiotelephone, and soon the spoken word was winging its way across the ocean. A further use of this new magic was soon proposed: could not the human voice be sent from shore to ships at sea?

Soon sub-chasers and other small Navy craft were talking to each other over equipment designed by Bell engineers. And in experiments starting in 1919, the men on two coastwise steamers talked through land stations to land telephones of the Bell System.

These early experiments covered fairly short distances. But in the meantime, telephone calls across the Atlantic by radio had become an ordinary occurence. So . . . why not 'phone calls to ships way out in *mid*-Atlantic?

Of course, long-distance ship-to-shore radiotelephony brought up problems of varying distances and directions—problems not encountered in point-to-point transmission. Bell Telephone Laboratories solved these problems with the design of the "Leviathan's" equipment. For the first time, long-range marine radiotelephony became a reality.

Later, Bell Laboratories scientists developed selective ringing, which made it possible to *dial* particular ships at sea. The basic elements of practical marine radiotelephony had now been developed.

BELL TELEPHONE LABORATORIES

World's largest organization devoted exclusively to research and development in all phases of electrical communications.

links the ship and the shore

IN ADDITION TO producing radiotelephone equipment for the largest ocean liners, Western Electric for many years manufactured the 224, 226 and 227 type sets, which brought the benefits of radiotelephone facilities to coastwise vessels and small craft.

These sets provided power capacities ranging up to 100 watts. As the Bell System had tremendously expanded its chain of harbor stations, coastal craft were normally near a shore station. Hence these capacities were ample to maintain contact with land.

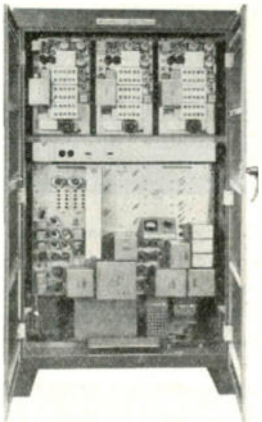
There still existed, however, no equipment specifically designed for tankers, freighters and smaller passenger ships plying the *ocean* lanes. This need has been filled by the introduction of the Western Electric 248A.

This new equipment provides 250 watts of transmitted radio frequency carrier power, resulting in greatly increased range. Provision is made for transmission and reception on the frequencies of the high-seas shore stations (as well as on the coastal harbor and ship-to-ship channels). Because of these two features, a ship equipped with the 248A, at practically any point on world trade routes, can establish contact with a land station.

The 248A combines this advantage with the compactness and simplicity of operation essential on smaller ships.

—QUALITY COUNTS—

THE NEWEST IN MARINE RADIOTELEPHONE EQUIPMENT



Left: Main cabinet of 248A mounting transmitter and three receivers.
Above: Remote control unit.

The long experience of Bell Laboratories and Western Electric in design and manufacture of marine radiotelephone equipment has culminated in the 248A—compact, powerful, simple to operate.

A single cabinet houses the transmitter and three receivers. Each of the three receivers can be tuned to any one of 10 pre-set frequencies; the transmitter to any one of 30. Transfer from one frequency to another is accomplished simply by turning knobs on the remote control panel.

Because three receivers are used, it is possible for the ship to monitor simultaneously on three different channels. The set is designed to permit easy installation of selective equipment to allow dialing the ship from shore stations.

Western Electric

Manufacturing unit of the Bell System and the nation's largest producer of communications equipment.

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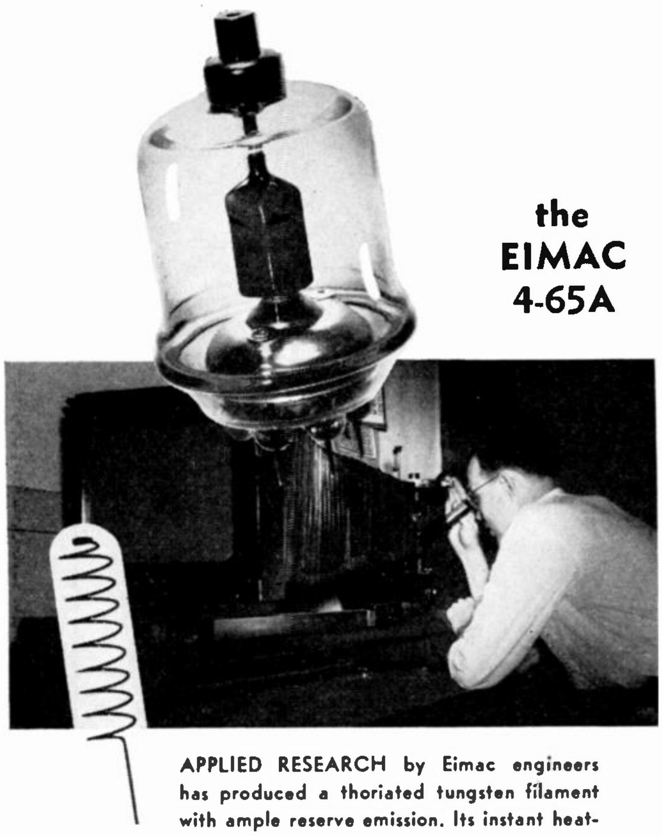
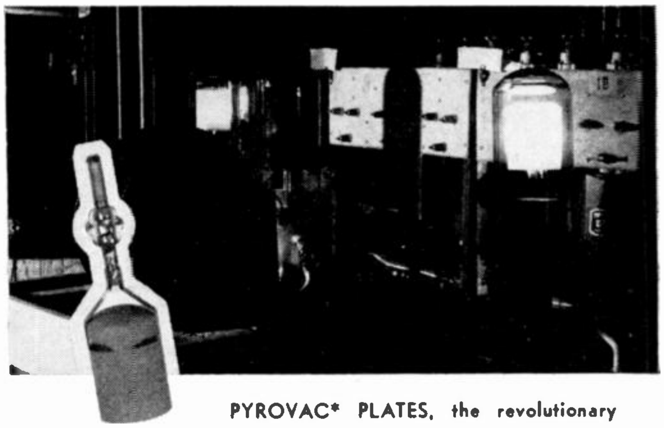
THEY'RE BETTER BECAUSE...

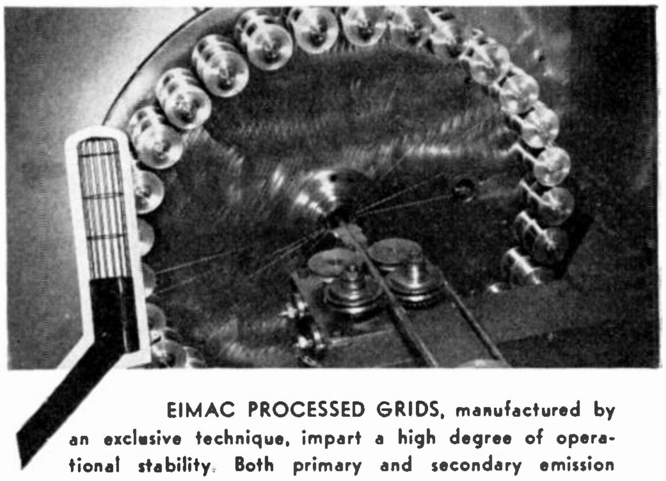


the EIMAC 4-65A

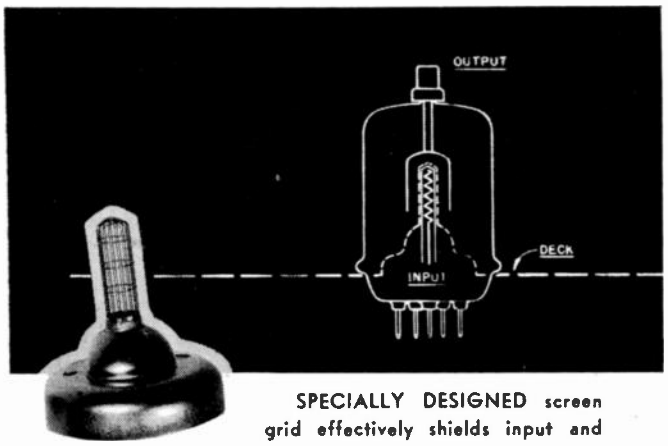
APPLIED RESEARCH by Eimac engineers has produced a thoriated tungsten filament with ample reserve emission. Its instant heating characteristics make the 4-65A well adapted to mobile application.



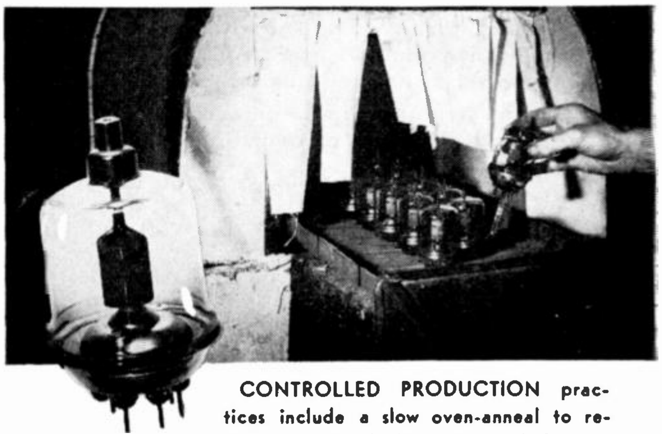
PYROVAC* PLATES, the revolutionary Eimac development, withstand excessive abuse. Manufactured by an advanced technique, these plates can handle momentary overloads in excess of 1000%, consequently they contribute appreciably to the tube's life.



EIMAC PROCESSED GRIDS, manufactured by an exclusive technique, impart a high degree of operational stability. Both primary and secondary emission are controlled.



SPECIALLY DESIGNED screen grid effectively shields input and output circuits, within the tube, without excessive screen power. All internal structures are self supporting without the aid of insulating hardware.



CONTROLLED PRODUCTION practices include a slow oven-anneal to remove the last vestige of residual strains, and four to eight hours of testing under severe VHF conditions.

*Trade Mark Reg. U. S. Pat. Off.

These are but some of the features that combine to make the Eimac 4-65A a better tetrode. It is unexcelled in its category as a power amplifier, oscillator or modulator. For example, in typical operation as a power amplifier or oscillator (class-C telegraphy or FM telephony) one tube with 1500 plate volts will supply 170 watts of output power with less than 3 watts of driving power. A complete comprehensive data sheet on the 4-65A has just been released. Write for your copy today.

EITEL-McCULLOUGH, INC.
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Eimac
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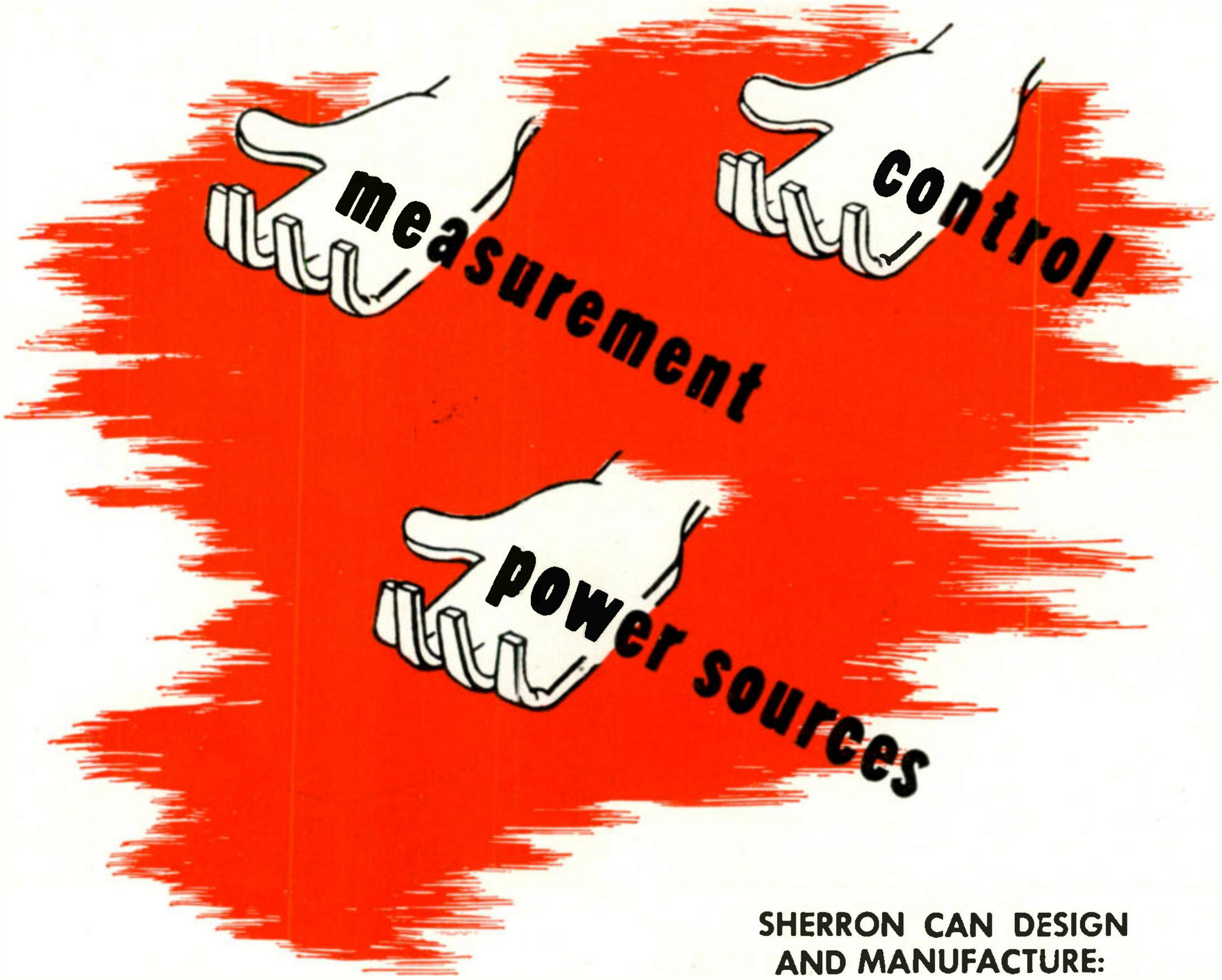
TUBES
The Power for R-F

FIRST AIDS FOR NUCLEAR RESEARCH



CUSTOM BUILT BY SHERRON

Our many years of diversified experience in building custom electronics equipment are now serving in the design, development and production of precision instruments to aid in blazing new trails in nucleonics. Sherron physicists and engineers have at their command the very newest in laboratory resources and facilities to tackle your problems. Let us tell you just how we can work with you.

SHERRON CAN DESIGN AND MANUFACTURE:

COUNTERS: Maximum count as required. Predetermined setting anywhere within the counting range. Resolution in the micro-second region.

COMPUTERS: Mechanical linkages. Electrical analogues of any complexity. Digital computers.

SERVO-MECHANISMS: Control to any desired accuracy. Power as desired. Linear control, logarithmic control.

AMPLIFIERS: R.F., Video, A.F., D.C. to fit any application. Particular emphasis on high gain, high stability characteristics.

OSCILLATORS: All frequencies.

POWER SUPPLIES: Electronic — Regulation ½% and less.

REGULATORS: Electronic — direct or through servo control. Regulation, drift, etc. to specification.

MEASUREMENT—CONTROL: Devices for measuring and control of all parameters capable of being controlled and producing proportional electrical, optical or measuring displacement. Electronic microammeters, radiation counters.

CONTROL OF ACCELERATOR ACCESSORIES: Grouping of controls, supplementary apparatus, and experimental system into a compact versatile unit.

Sherron Electronics
W2XDK

SHERRON ELECTRONICS CO.

Division of Sherron Metallic Corporation

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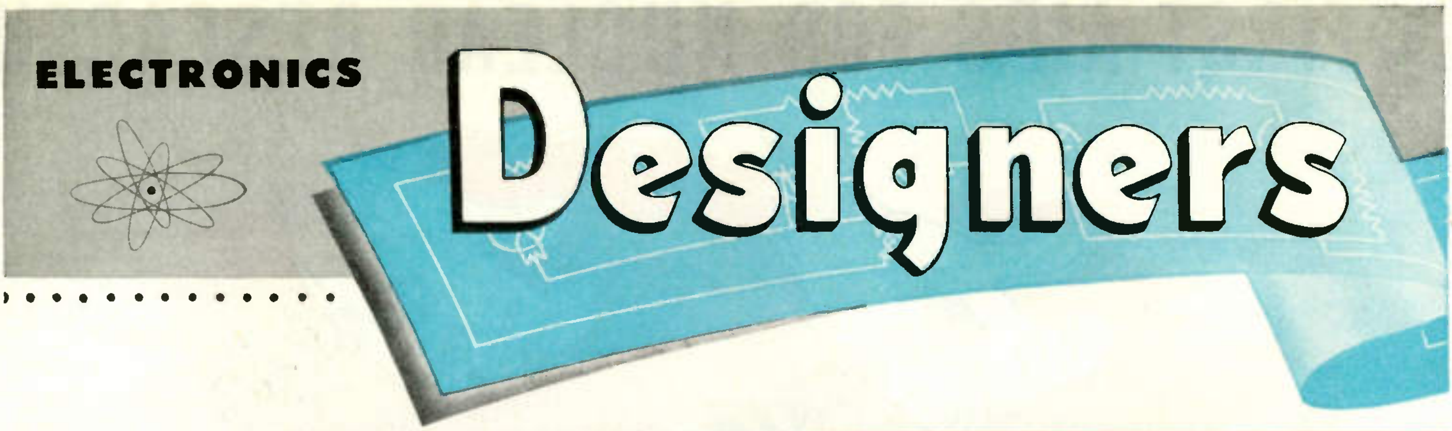


NOW AVAILABLE FOR YOUR COMMERCIAL APPLICATIONS

CAPACITOR PULSE-FORMING NETWORKS

Developed by General Electric and proven by the thousands in the war, these compact units are now available for any commercial use. They find application in radar and industrial equipment where the normal capacitor discharge shape is not suitable and where an impulse having a definite energy content and duration is required. The network consists of one or more equal capacitor sections and the same number of inductance coil sections. Both capacitors and coils are hermetically sealed in the same metal container. Networks are treated with top quality mineral oil to provide stability of capacitance characteristics over a wide range of ambient temperatures. Sizes from which you can make your selection range from a 0.5-kw output rating to 4500-kw. Write for bulletin GEA-4996.

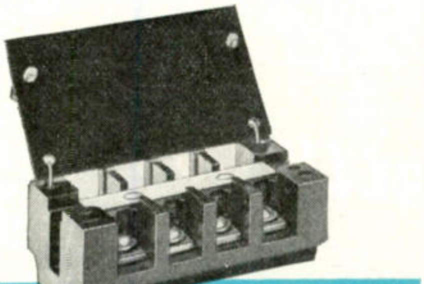
DESIGNED FOR BETTER READABILITY

General Electric's new line of 3½-inch thin panel instruments will save space and add to the appearance of your panels. They're dust-proof, moisture resistant, and vibrations normally encountered in aircraft and moving vehicles have no adverse effects. Especially designed for better readability, the scale divisions stand out by themselves. Lance-type pointers and new-style numbers mean faster reading. Available in square and round shapes, depth behind the panel is only 0.99 inches. Construction is of the internal-pivot type, with alnico magnets for high torque, good damping, and quick response. Check bulletin GEA-5102.

GENERAL ELECTRIC

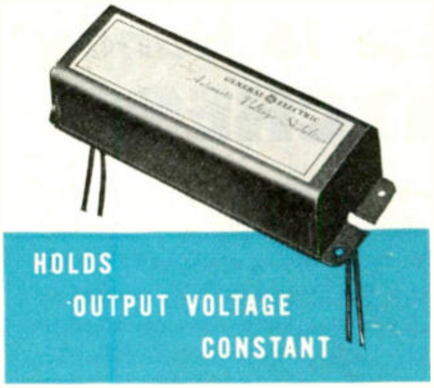
Digest

TIMELY HIGHLIGHTS ON G-E COMPONENTS



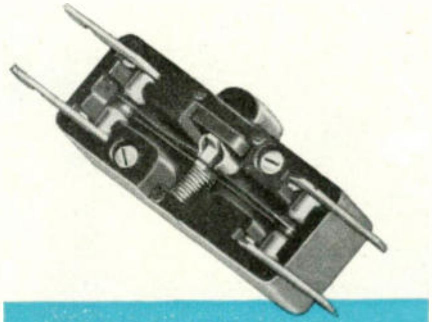
SIMPLIFY CONTROL WIRING WITH THESE TERMINAL BOARDS

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



HOLDS OUTPUT VOLTAGE CONSTANT

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



LOOKING FOR LIGHTWEIGHT SWITCHES?

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

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

FOR YOUR COOLING FANS

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

General Electric Company, Section C642-19
Apparatus Department, Schenectady, N. Y.

Please send me the following bulletins:

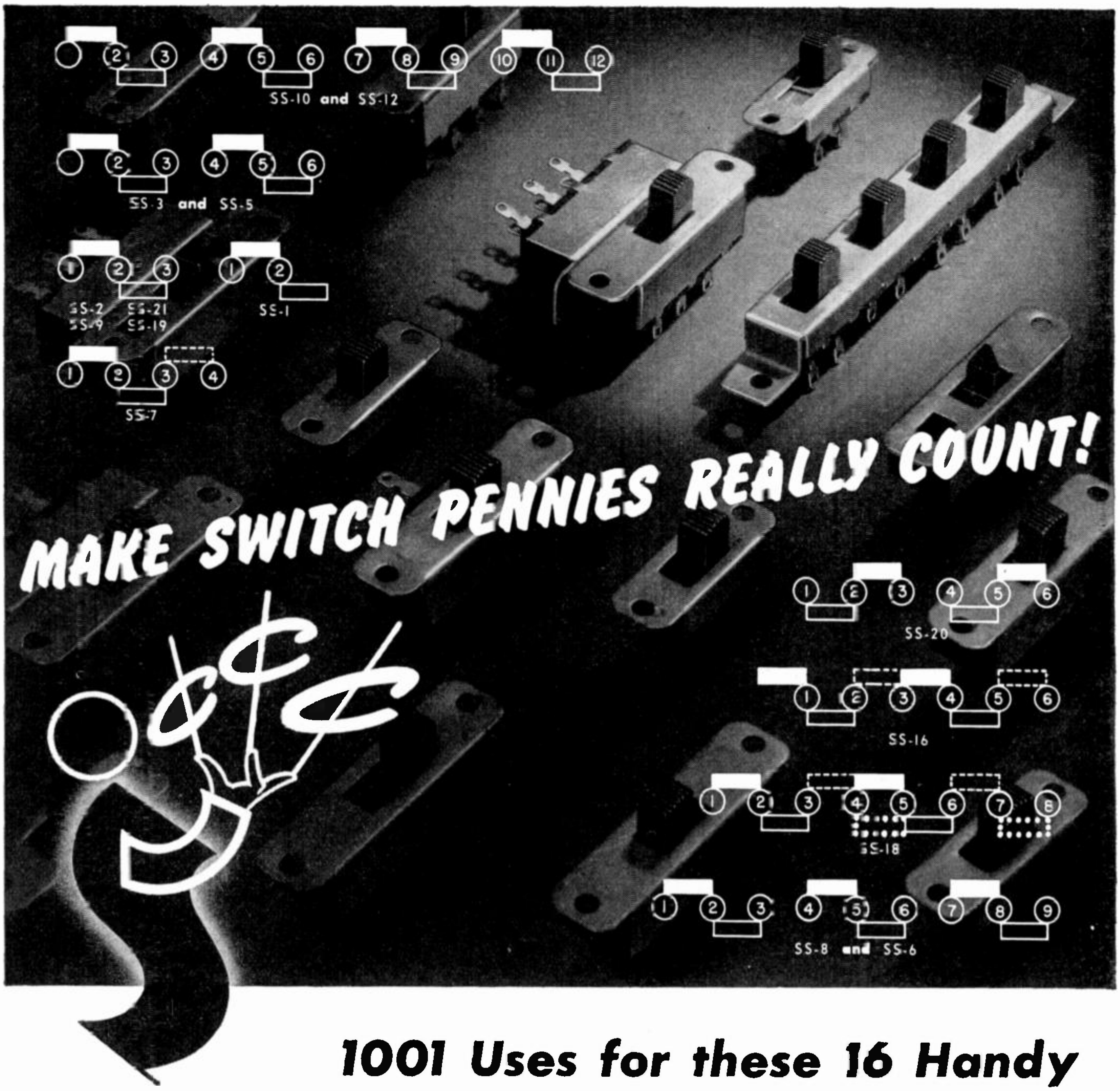
☐ GEA-4996 Capacitor Pulse-forming Networks
☐ GEA-5102 Panel Instruments
☐ GEA-1497C Terminal Boards
☐ GEA-3634B Automatic Voltage Stabilizers
☐ GEA-4888 Switchettes
☐ GEC-219 Unit-bearing Fan Motor

NAME..

COMPANY..

ADDRESS..

CITY.......................................STATE..



1001 Uses for these 16 Handy
SLIDE SWITCHES

OTHER STACKPOLE PRODUCTS

FIXED AND VARIABLE RESISTORS
IRON CORES
POWER TUBE ANODES
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CARBON PILE VOLTAGE REGULATOR DISCS
MICROPHONE CARBONS
SINTERED ALNICO II
PERMANENT MAGNETS
. . . and dozens more

CONTACT CODE

POSITION	1	▬
POSITION	2	▭
POSITION	3	⬚
POSITION	4	⠿

Name the switch contact arrangement you need! From 1 to 6 poles, up to 4 positions, with or without detent, spring return, covers, or other optional features.

Chances are Stackpole can supply exactly the right switch—promptly and inexpensively. 16 standard slide types, each designed for good appearance and real dependability, provide a low cost way of modernizing almost any electrical equipment and adding greatly to its sales appeal. Many economical adaptations can be supplied on special order to large quantity users.

Write for Catalog RC-6

STACKPOLE

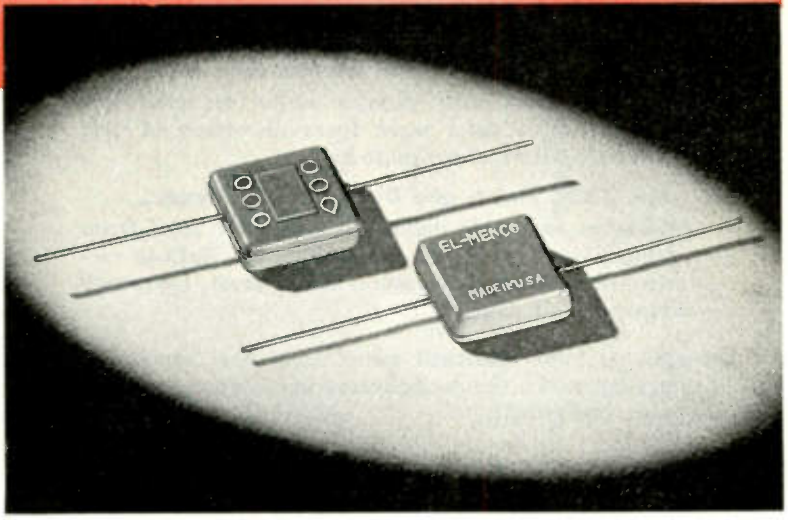
STACKPOLE CARBON CO. • ST. MARYS, PA.

ELECTRONIC COMPONENTS DIVISION

*Because of a nail
a shoe was lost
Because of a shoe
a horse was lost
Because of a horse
a battle was lost
Because of a battle a nation
was lost.

El-Menco

CAPACITORS, like the nail that lost a nation, are small . . . but their importance cannot be overemphasized. For *dependable* components that never "let a product down" — specify El-Menco.



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Our silver mica department is now producing silvered mica films for all electronic applications. Send us your specifications.

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Send for samples and complete specifications. Foreign Radio and Electronic Manufacturers communicate direct with our Export Department at Willimantic, Conn., for information.

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Sole Agent for Jobbers and Distributors in U. S. and Canada.

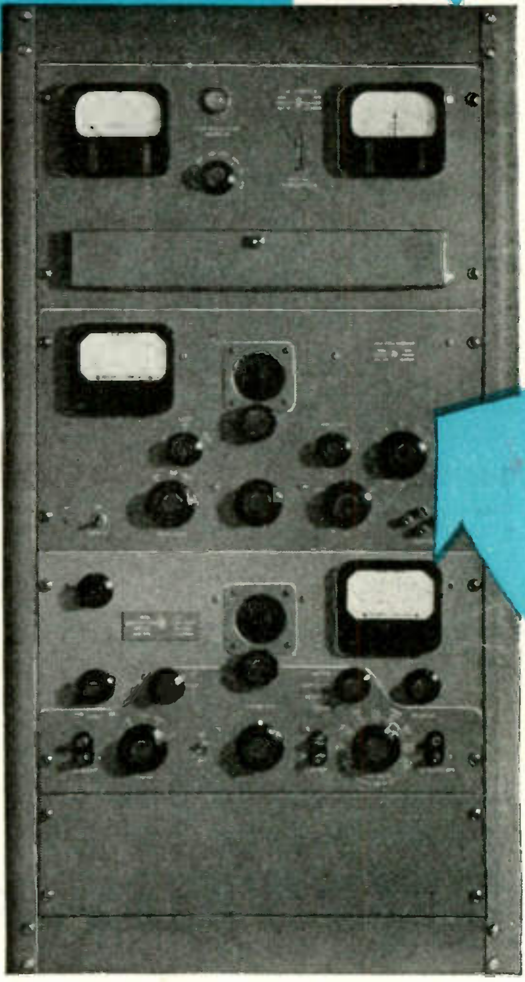
MOLDED MICA El-Menco MICA TRIMMER

CAPACITORS

-hp- Presents...

COMPLETE FM TEST EQUIPMENT

For Broadcast Stations



Here is a complete transmitter maintenance group — providing every measurement necessary for top-flight operation from microphone to antenna! Three fast, accurate precision instruments in one compact whole — specifically designed for years of trouble-free performance — proven in service in radio stations throughout America.

These are the *-hp-* instruments that comprise this group.

1. **-*hp*- 335B Frequency and Modulation Meter.** Continuous measurement of carrier frequency and modulation swing. Low distortion audio output for measuring and monitoring.
2. **-*hp*- 206A Audio Signal Generator.** Provides continuously variable audio frequency voltage having a total wave form distortion of less than 0.1% from 50 cps to 20 kc.
3. **-*hp*- 330C Noise and Distortion Analyzer.** Measures harmonic distortion and noise level from demodulated carrier or audio channels. Built-in-vacuum-tube-voltmeter measures audio level, frequency response and gain.

All instruments have identical panel sizes for convenient mounting in relay racks. Can be delivered in colors and finishes to match your equipment.

GET FULL INFORMATION...WRITE TODAY

HEWLETT-PACKARD COMPANY
1481D PAGE MILL ROAD · PALO ALTO, CALIFORNIA

This -*hp*- Maintenance Group Makes These Essential FM BROADCAST MEASUREMENTS

Carrier Frequency: Continuously monitored with accuracy well within F.C.C. limits.

Modulation Swing: Continuously measured at instrument installation and at control console.

Modulation Limit: Alarm lamp flashes on instrument and console when pre-set level is exceeded.

Aural Monitor: Demodulated signal provides listening check for operator.

Harmonic Distortion: Measured from r-f carrier or audio channel.

Noise: Measured accurately from FM carrier or audio channel.

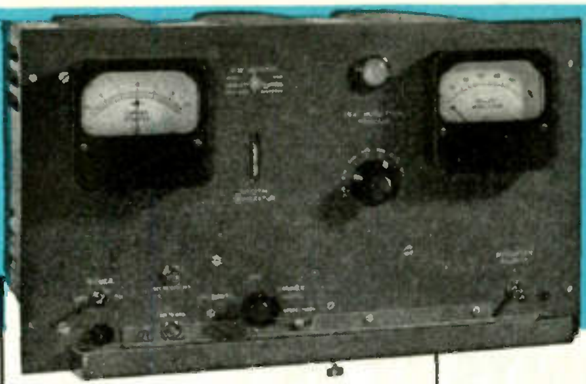
Frequency Response: Overall response, microphone to antenna, of individual units in transmitter set-up.

Audio Transmission: Accurately measures gain of audio channels.

Audio Level: Measured over range from +50 db to −60 db at 600 ohm level.

Equalizer Circuits: Characteristics of circuits and lines can be checked accurately, swiftly.

Oscilloscope Connections: Facilitates visual study of noise and distortion.



-hp- 335B
FM Monitor

Accurate, Stable, Easy to Operate

BRIEF SPECIFICATIONS

Frequency Range: Any single frequency, 88 to 108 mc.

Deviation Range: +3 kc to −3 kc.

Accuracy: Better than ±1000 cps.

Modulation Range: Modulation swing 100 kc. Scale calibrated 100% at 75 kc.

Audio Output: Supplied with 75 microsecond de-emphasis circuit, flat within ½ db of standard curve, 20 cps to 20 kc.

Monitoring Output: 1 milliwatt into 600 ohms, balanced, at 100% modulation.

Size: Panel 10½"x 19". Depth 13".

Precision accuracy, unique stability, new convenience and compact size—those are but a few of the reasons why this *-hp-* 335B is the finest instrument ever developed for FM broadcast monitoring. Here are additional advantages that help make this new *-hp-* instrument an ideal component of the *-hp-* FM group.

Simple to Operate. No adjustments required during operation.

Independent of Signal Level. Readings of frequency or modulation meter are unaffected by variations in transmitter level.

Unusual Stability. Low temperature coefficient crystal in temperature-controlled oven combined with specially developed electronic linear counter circuits provides accuracy far beyond that required. Measurements do not depend on accuracy of conventional discriminator circuits.

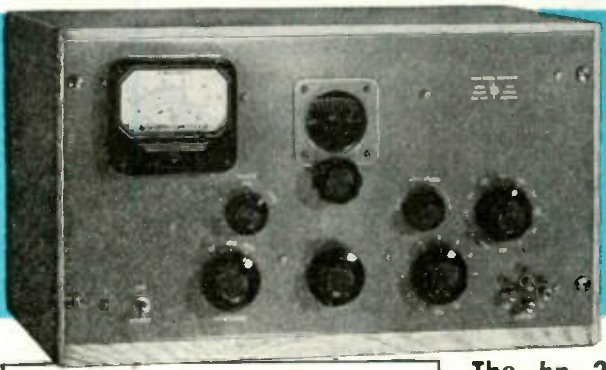
Remote Modulation Meter. Modulation may be monitored at control console or other remote point.

Low Distortion. Audio output for measuring purposes has less than .25% residual distortion.

Low Noise Level. Residual noise and hum in audio output are at least 75 db below 100% modulation.

Meets F.C.C. Requirements.

This instrument is small in size, easy to install, suitable for cabinet or rack panel mounting. Can be furnished to match your transmitter color scheme.



-hp- 206A
Audio Signal Generator

Distortion Less Than 0.1%

BRIEF SPECIFICATIONS

Frequency Range: 20 cps to 20 kc, 3 bands.

Output: +15 dbm to matched resistive loads. 10 volts available for open circuit.

Output Impedance: 50, 150, 600 ohms center-tapped and balanced. 600 ohms single-ended.

Frequency Response: Better than 0.2 db beyond output meter at all levels.

Distortion: Less than 0.1% above 50 cps. Less than 0.25% from 20 cps to 50 cps.

Hum Level: At least 70 db below output signal, or more than 100 db below 0 level, whichever is larger.

Size: Panel 10½"x 19". Depth 13".

The -hp- 206A Audio Signal Generator provides a source of continuously variable audio frequency voltage having a total distortion of less than 0.1%. This feature, combined with high stability, flat frequency response, and great accuracy of output voltage, makes it an ideal component for FM station maintenance. Here are some of this instrument's unusual advantages:

Distortion less than 0.1% between 50 cps and 20 kc.

Continuously variable frequency range, covered in 3 bands, micro-controlled dial, effective scale length 47", ball-bearing smoothness for tuning ease.

Output meter monitors output voltage signal with accuracy of at least 0.2 db.

Special low temperature co-efficient frequency determining elements provide high stability and excellent accuracy over long periods of time.

Precision attenuators vary output signal level in 0.1 db steps over 111 db range.

This new -hp- generator is convenient to use, compact in size. It can be provided for rack or cabinet mounting, in colors matching your installation.

1481

hp laboratory instruments
FOR SPEED AND ACCURACY

NEW ELECTROLYTICS *fully dependable* TO 450 VOLTS AT 85°C

ILLUSTRATIONS
ACTUAL SIZE

for TELEVISION'S exacting applications

Designed for dependable operation up to 450 volts at 85°C. these new Sprague electrolytics are a good match for television's severest capacitor assignments. An extremely high stability characteristic is assured, even after extended shelf life, thanks to a special Sprague processing technique. Greatly increased manufacturing facilities are now available.

Your inquiries concerning these new units are invited.

DEPENDABILITY TO MATCH THESE NEW ELECTROLYTICS! SPRAGUE PHENOLIC MOLDED TUBULARS...

Highly heat- and moisture-resistant—Non-inflammable—Moderately priced—Conservatively rated for —40°C. to +85°C. operation—Small in size—Completely insulated—Mechanically rugged—Thoroughly field-tested
Write for Engineering Bulletin 210A

SPRAGUE ELECTRIC COMPANY · NORTH ADAMS, MASS.

SPRAGUE

Capacitors
*Koolohm Resistors

PIONEERS OF ELECTRIC AND ELECTRONIC PROGRESS

*T. M. Reg. U. S. Pat. Off.

Most prominent position in any parade is

UP FRONT

featuring

"Built-In" CONSTANT VOLTAGE

For the Protection of Our Customers

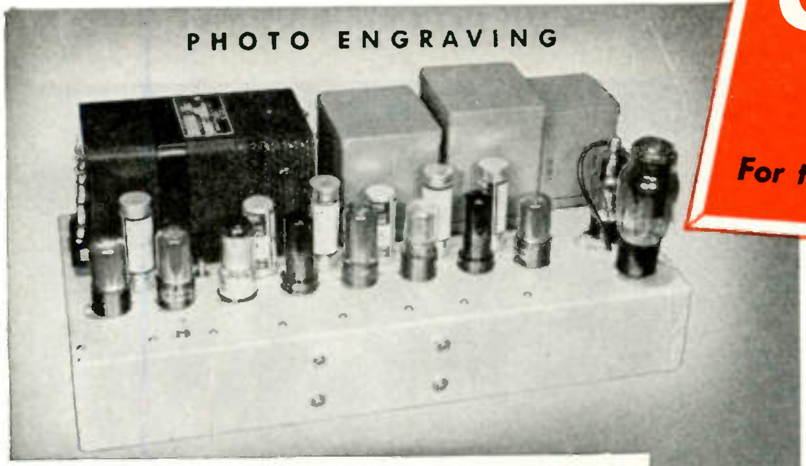
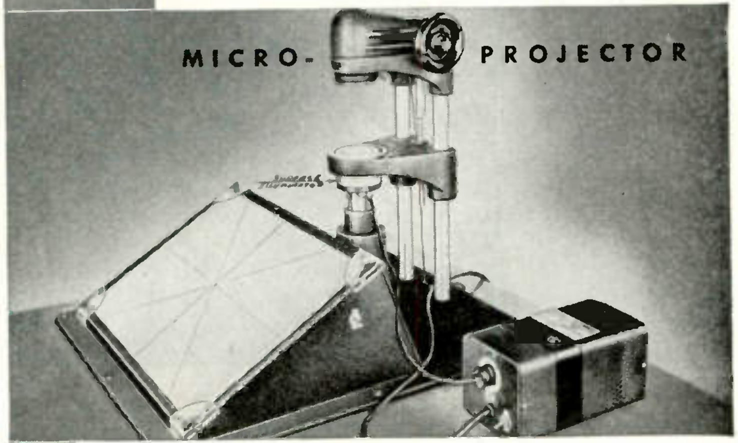


PHOTO ENGRAVING

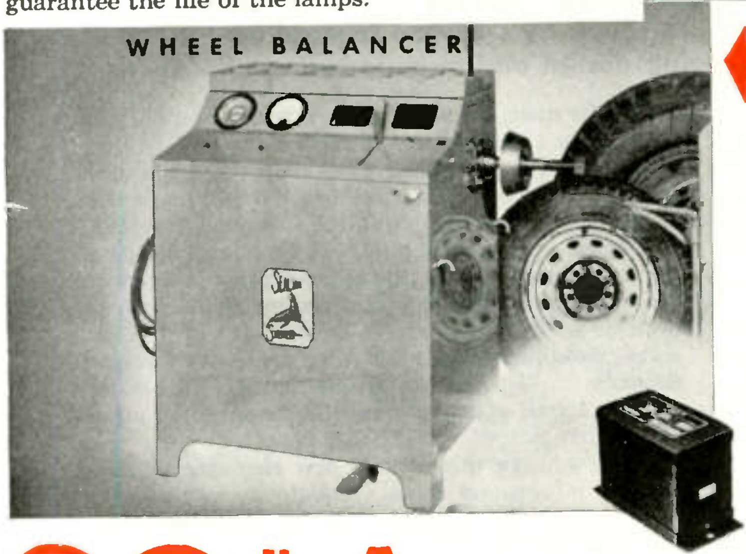
With power shortages playing hob with line voltages all over the country—isn't it about time that you too joined the parade of manufacturers who are featuring constant voltage as a built-in component in their products.

This preamplifier phasing control section of a medium power, low distortion restricted band audio-amplifier employed in a new printing plate engraving system couldn't operate satisfactorily on available line voltages. Robert H. Rigby Corp., solved the problem with a "built-in" SOLA CONSTANT VOLTAGE TRANSFORMER.

Unstable voltages varied the light output essential for satisfactory operation of this precision instrument. High voltages burned out the light source. "Built-in" SOLA CONSTANT VOLTAGE TRANSFORMERS now provide a constant source of light and enable R. S. Wilder Company to guarantee the life of the lamps.



MICRO-PROJECTOR

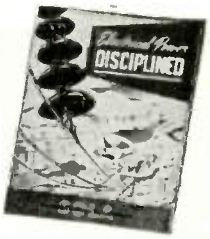


WHEEL BALANCER

The H. C. Schildmeier Co. says, "We have found the SOLA CONSTANT VOLTAGE TRANSFORMER to be the solution to many of our troubles, by maintaining a constant output voltage to actuate a unit that is direct meter reading" . . . a SOLA CV transformer is a built-in component of every Seal Line Balancer produced by this company.

SOLA HANDBOOK
BULLETIN KCV-102

A complete, and authoritative treatise on voltage regulation. Write for your copy.



SOLA *Constant Voltage* TRANSFORMERS

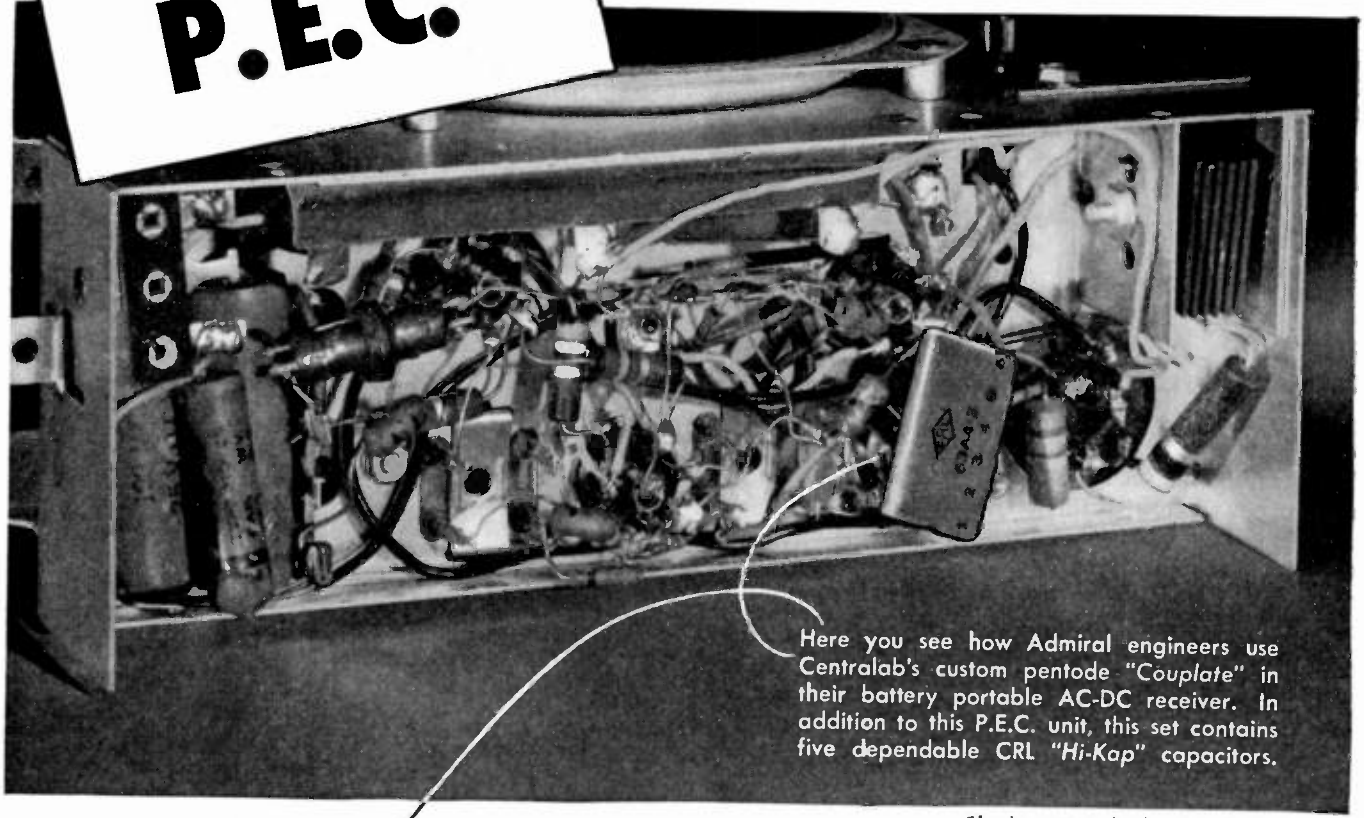
Transformers for: Constant Voltage • Cold Cathode Lighting • Airport Lighting • Series Lighting • Fluorescent Lighting • Luminous Tube Signs
Oil Burner Ignition • X-Ray • Power • Controls • Signal Systems • etc. • **SOLA ELECTRIC COMPANY, 4633 W. 16th Street, Chicago 50, Illinois**

Manufactured under license by: ENDURANCE ELECTRIC CO., Concord West, N. S. W., Australia • ADVANCE COMPONENTS LTD., Walthamstow, E., England
UCOA RADIO S.A., Buenos Aires, Argentina • M. C. B. & VERITABLE ALTER. Courbevoie (Seine), France

PROGRESS REPORT ON

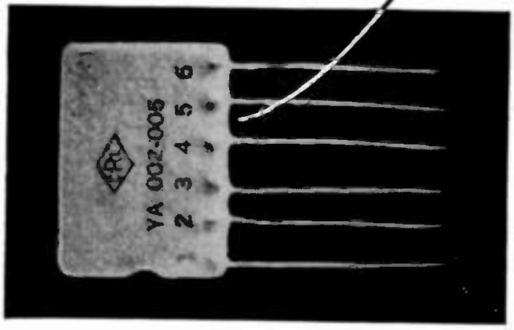
P.E.C.*

How Admiral Radio uses Centralab's *Printed Electronic Circuit* to build finer radios . . . to cut assembling time!

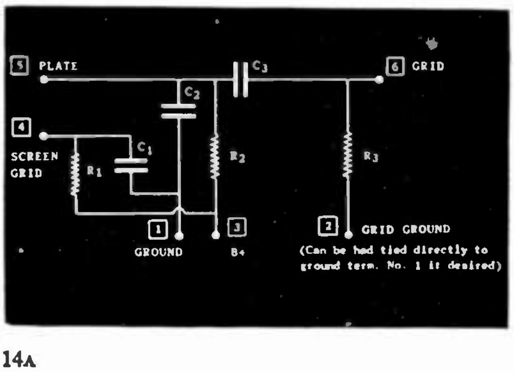


Here you see how Admiral engineers use Centralab's custom pentode *"Couplate"* in their battery portable AC-DC receiver. In addition to this P.E.C. unit, this set contains five dependable CRL *"Hi-Kap"* capacitors.

Chassis courtesy of Admiral Radio Corp.



"COUPLATE" is made of high dielectric Ceramic-X to give long life, low internal inductance, positive resistance to humidity and vibration. A circuit diagram of CRL's *Couplate* is shown below.



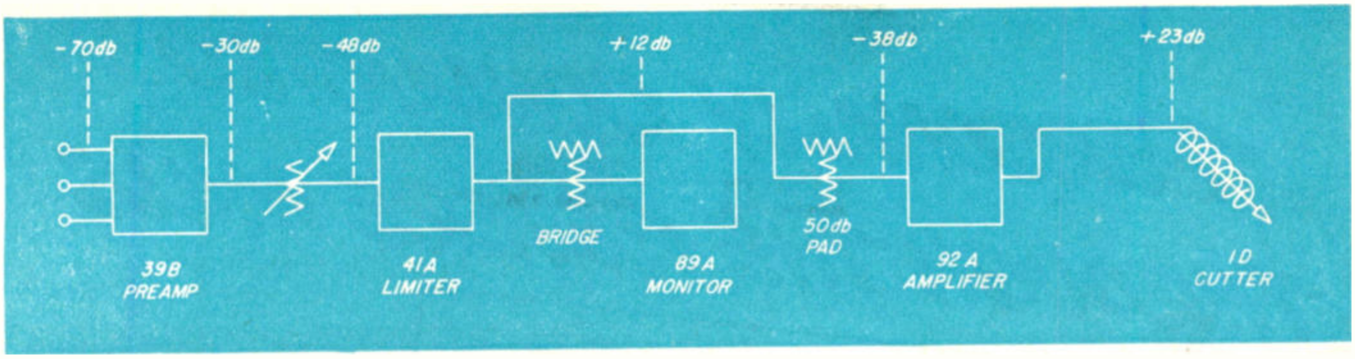
*Centralab's "Printed Electronic Circuit" — Industry's newest method for improving design and manufacturing efficiency!

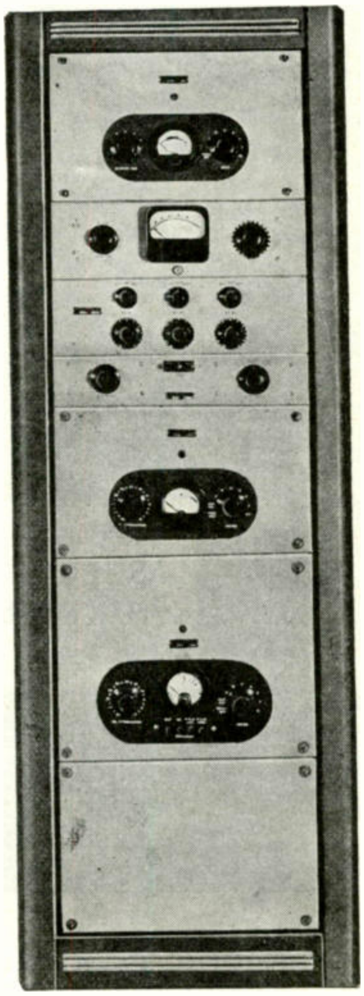
IMAGINE the time, the space, the material you save by using one unit instead of six. That's just what Centralab's amazing pentode *"Couplate"* is doing for Admiral Radio Corporation, Chicago. This complete interstage coupling circuit combines three resistors and three capacitors into one tiny, dependable P.E.C. unit. *"Couplate"* saves time for Admiral by eliminating many assembling operations. It saves space and material by reducing the number of components needed. What's more it improves performance by minimizing the chance of broken or loose connections.

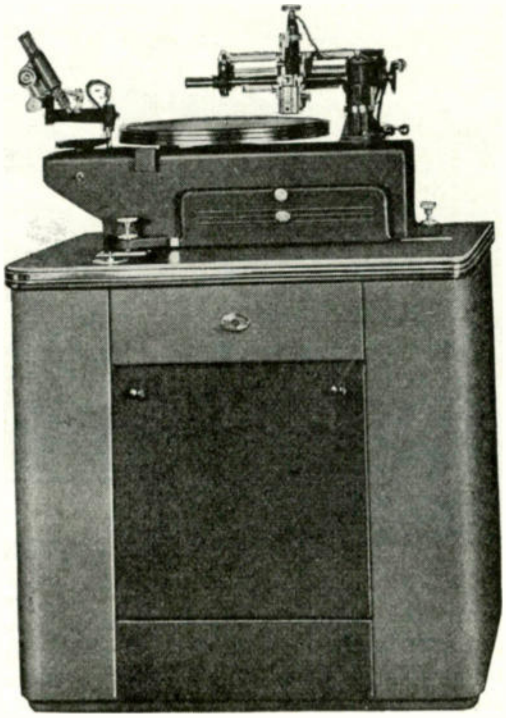
Integral Ceramic Construction: Each *Printed Electronic Circuit* is an integral assembly of *"Hi-Kap"* capacitors and resistors closely bonded to a steatite ceramic plate and mutually connected by means of metallic silver paths "printed" on the base plate.

You'll want to see and test this exciting new electronic development. For complete information about *Couplate,* as well as other CRL *Printed Electronic Circuits,* see your nearest Centralab Representative, or write for Bulletin 999.

LOOK TO **Centralab** IN 1949!
CRL
Division of GLOBE-UNION INC., Milwaukee


-70db
-30db
-48db
+12db
-38db
+23db
39B PREAMP
41A LIMITER
BRIDGE
89A MONITOR
50db PAD
92A AMPLIFIER
1D CUTTER





PERSONALLY CONDUCTED THROUGH THE REVERE MILLS

OFHC COPPER

BECAUSE OFHC Copper looks like any other copper, Revere takes great pains to identify it throughout processing, to see it is not lost track of or mixed up with other types. The obvious thing is to mark each piece, which is done, but markings are obliterated by operations such as rolling, and so Revere goes to the length of assigning special personnel to follow each lot of OFHC Copper from one operation to another, watching carefully to be sure each load is kept intact.

In addition, Revere takes full cognizance of the fact that OFHC Copper for radio purposes must have special qualities. In making anodes, it must be deep drawn, and for the feather-edge seal, it must be capable of being rolled or machined down to .002″/.010″. By carefully controlling mill processing, grain size is kept at or below permissible limits. Freedom from oxygen, and from voids, is guaranteed by the method of casting the bars from which we roll the forms required. In addition, there is an operation which results in Revere OFHC Copper being not just commercially free but *nearly absolutely free* of internal and external defects. This great care in producing copper for radio and radar purposes probably accounts for the fact that Revere is a preferred source of supply.

REVERE PRODUCTS AND SERVICES

All Revere Metals are processed with the care and attention required to assure that they meet all metallurgical and physical specifications. Revere supplies mill products in non-ferrous metals and alloys, and also electric welded and lockseam steel tube. An important part of our service to industry is the Revere Technical Advisory Service, which will gladly collaborate with you on specifications and fabrication methods.

REVERE

COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801
230 Park Avenue, New York 17, New York

• • •

Mills: Baltimore, Md.; Chicago, Ill.; Detroit, Mich.; New Bedford, Mass.; Rome, N. Y.
Sales Offices in Principal Cities, Distributors Everywhere

ALSiMAG

TRADE MARK REGISTERED U S PATENT OFFICE

a superior, low cost alternate for many commonly used materials

● Engineers are often surprised to find that metal, plastic or wood parts can be replaced with AlSiMag components at a saving in cost. At the same time they usually gain highly desirable advantages in product performance. It is natural that a product with the many superior advantages of AlSiMag would be expected to be expensive. The basic materials in AlSiMag are costly.

Automatic and efficient manufacture permits quantity production of AlSiMag parts at low prices. Thus, AlSiMag prices are frequently lower than prices of similar parts in cheaper materials which are more expensive to fabricate.

AlSiMag technical ceramic components are custom made for the individual requirement. AlSiMag is the trade name of a large number of ceramic compositions. The physical characteristics of the various compositions are clearly and accurately listed in the AlSiMag Property Chart, sent free on request.

Our engineers will be glad to submit suggestions on design and give you information on cost if you will submit details of your requirements.

Example: A manufacturer replaced a machined part which cost him 33 cents each with an improved AlSiMag component which cost 14 cents each. AlSiMag engineers cooperated in redesigning this component for maximum usefulness to the customer and minimum production cost—That some engineering cooperation is available to you on request.

47TH YEAR OF CERAMIC LEADERSHIP

AMERICAN LAVA CORPORATION

CHATTANOOGA 5, TENNESSEE

SALES OFFICES: ST. LOUIS, MO., 1123 Washington Ave., Tel: Garfield 4959 • NEWARK, N. J., 671 Broad St., Tel: Mitchell 2-8159 • CAMBRIDGE, Mass., 38-B Brattle St., Tel: Kirkland 4498 • CHICAGO, 9 S. Clinton St., Tel: Central 1721 • LOS ANGELES, 324 N. San Pedro St., Tel: Mutual 9079 • PHILADELPHIA, 1649 N. Broad St.

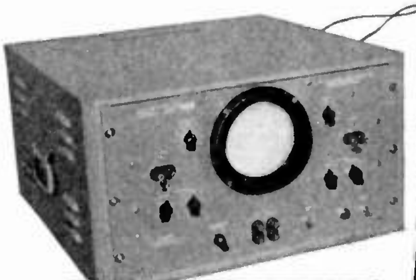
For wider frequency range...top writing rates... increased brightness...it's DU MONT *High-voltage Oscillography*

The basis is the Type 5RP-A Cathode-ray Tube operating at an accelerating potential up to 29,000 volts maximum. This achieves: (1) Greatly increased brightness; (2) Observation or recording of traces hitherto invisible; (3) Vastly increased writing rates even better than 400 inches per microsecond; (4) Optical magnification by projection lenses such as Du Mont Type 2542. Although deflection sensitivities are slightly less than those of low-voltage cathode-ray tubes, high-voltage oscillographs produce smaller spot size and higher brightness, thereby presenting a finer, better resolved trace.

And here's the Du Mont selection of high-voltage oscillographs:

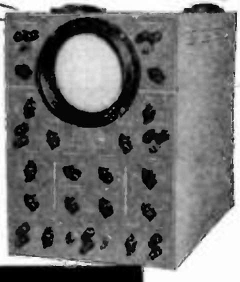
10 CPS to 10 MC

Type 280: A precision time-measuring oscillograph with range of 10 cps to 10 mc. Sweep speeds as high as 0.25 microsecond/in. are available. Duration of any portion of signal measured on 0.25 microsecond/in. sweep to an accuracy of ±0.01 microsecond. Intervals greater than 5 microseconds read on calibrated dial to accuracy of ±0.1 microsecond. Ready application to precise measurement of duration of waveform of various components in the composite television signal. Accelerating potential adjustable from 7,000 to 12,000 volts. Recordable writing rates up to 63 inches per microsecond, with commercially available equipment.



WRITING RATES TO ABOVE 400 IN./MSEC.

Type 281-A: Devoid of internal deflection amplifiers, there are no frequency response limitations within the ratings of its Type 5RP-A tube. Phenomena have been recorded photographically at writing speeds of 85 inches per microsecond. With external power supply (such as Du Mont Type 286-A), photographic writing speeds of over 400 inches per microsecond may be examined. Recommended when oscillographic needs are extremely specialized or too advanced for standard commercial equipment. An accelerating potential as high as 29,000 volts is available with the Types 281-A and 286-A in combination.



D-C to 200 KC

Type 250-H: Covers range from d-c to 200 kc. Potentials containing both d-c and a-c components may be examined. Many special features for general usage include: linear time-base of unusual flexibility; automatic beam control on driven sweeps; internal calibrator of signal amplitude. This is a high-voltage oscillograph with maximum accelerating potential of 13,000 volts. Recordable writing rate of approximately 40 inches per microsecond.

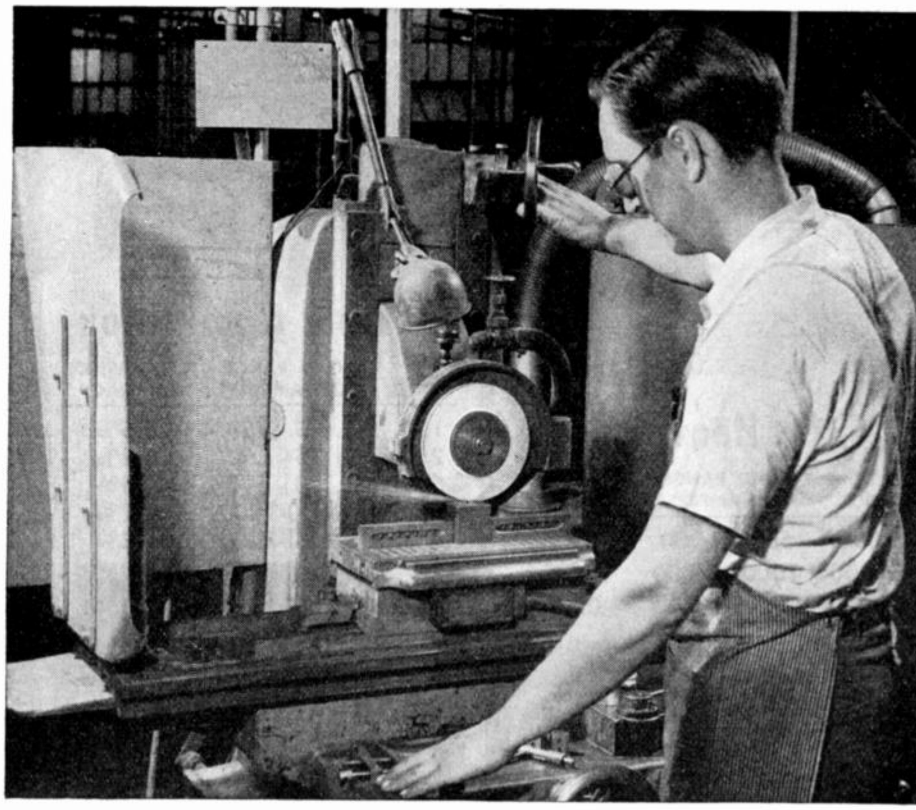
20 CPS-5 MC

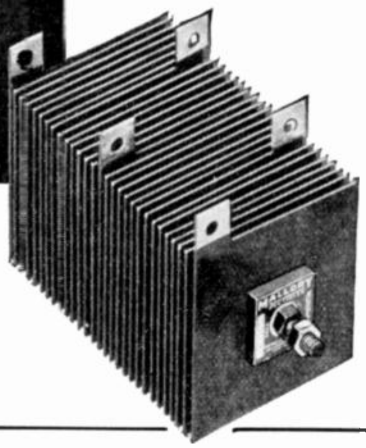
Type 248-A: Frequency range of 20 cps to 5 mc. Specifically intended for investigation of pulses containing high-frequency components of recurrent or transient nature. For this purpose it provides these necessary characteristics: High-frequency recurrent sweeps; short-duration driven sweeps; timing markers; signal delay network. Accelerating potentials up to 14,000 volts at recordable writing rate of approximately 69 inches per microsecond.

LITERATURE ON REQUEST

DUMONT *for Oscillography*

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





"...Worked So Well We've Forgotten About It!"

That's the kind of report we like to hear—because it describes a Mallory Magnesium-Copper Sulfide Rectifier Stack which has served *ten years—in daily operation—with minimum maintenance.* Its job—supplying DC current for the operation of a magnetic chuck on each of five surface grinders such as the one shown above. Its performance—it still gives over 90% of its original efficiency.

This is not surprising. Mallory Magnesium-Copper Sulfide Rectifier Stacks are made rugged—practically immune to damage or abuse. Rectification is confined to core of the stack—the outside fins are for heat dissipation only. No liquids, bulbs or moving parts—nothing to give trouble or wear out. And Mallory MgCuS rectifier stacks are more than "the world's toughest rectifiers"—when subjected to abnormal voltage surges, their rectifying junctions are so made that they actually heal themselves.

No wonder millions are in use. Write for more information or for engineering help.

Check These Features:

- ✓ **Proved long life**
- ✓ **Unaffected by high temperatures**
- ✓ **Withstands abuse and accidental short circuits**
- ✓ **Self-healing rectifying junctions**
- ✓ **Constant output over many years**
- ✓ **Resists harmful atmospheric conditions**
- ✓ **Rugged, all-metal construction**
- ✓ **No bulbs, no brushes, no sparking contacts**

MALLORY MgCuS RECTIFIER STACKS ARE THE WORLD'S TOUGHEST RECTIFIERS

P. R. MALLORY & CO. Inc.
MALLORY
MAGNESIUM-COPPER SULFIDE RECTIFIER STACKS AND POWER SUPPLIES

RECTOPLATER* SUPPLIES—RECTOTRUCK CHARGERS—
RECTOSTARTER* AIRCRAFT POWER SUPPLIES—
RECTOPOWER* SUPPLIES—AUTOMOTIVE BATTERY CHARGERS

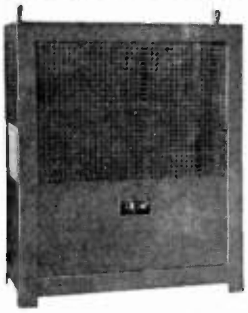
*Reg. U. S. Pat. Off.

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

3-Phase Regulation

MODEL	LOAD RANGE VOLT-AMPERES	*REGULATION ACCURACY
3P15,000	1500-15,000	0.5%
3P30,000	3000-30,000	0.5%
3P45,000	4500-45,000	0.5%

- **Harmonic Distortion on above models 3%. Lower capacities also available.**



Extra Heavy Loads

MODEL	LOAD RANGE VOLT-AMPERES	*REGULATION ACCURACY
5,000†	500 - 5,000	0.5%
10,000†	1000-10,000	0.5%
15,000†	1500-15,000	0.5%

General Application

MODEL	LOAD RANGE VOLT-AMPERES	*REGULATION ACCURACY
150	25 - 150	0.5%
250	25 - 250	0.2%
500	50 - 500	0.5%
1000	100-1000	0.2%
2000	200-2000	0.2%

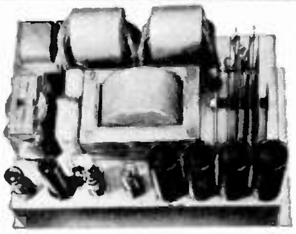
400-800 Cycle Line

INVERTER AND GENERATOR REGULATORS FOR AIRCRAFT.

Single Phase and Three Phase

MODEL	LOAD RANGE VOLT-AMPERES	*REGULATION ACCURACY
D500	50 - 500	0.5%
D1200	120-1200	0.5%
3PD250	25 - 250	0.5%
3PD750	75 - 750	0.5%

Other capacities also available



The NOBATRON Line

Output Voltage DC	Load Range Amps.
6 volts	15-40-100
12 "	15
28 "	10-30
48 "	15
125 "	5-10

- **Regulation Accuracy 0.25% from 1/4 to full load.**

SORENSEN

The First Line of standard electronic AC Voltage Regulators and Nobatrons

GENERAL SPECIFICATIONS:

- Harmonic distortion max. 5% basic, 2% "S" models
- Input voltage range 95-125: 220-240 volts (—2 models)
- Output adjustable bet. 110-120: 220-240 (—2 models)
- Recovery time: 6 cycles: † (9 cycles)
- Input frequency range: 50 to 65 cycles
- Power factor range: down to 0.7 P.F.
- Ambient temperature range: —50°C to +50°C

All AC Regulators & Nobatrons may be used with no load.

*Models available with increased regulation accuracy.

Special Models designed to meet your unusual applications.

Write for the new Sorensen catalog. It contains complete specifications on standard Voltage Regulators, Nobatrons, Increvolts, Transformers, DC Power Supplies, Saturable Core Reactors and Meter Calibrators.

• •

SORENSEN & CO., INC.
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Represented in all principal cities.

QUICK
TRIP
from
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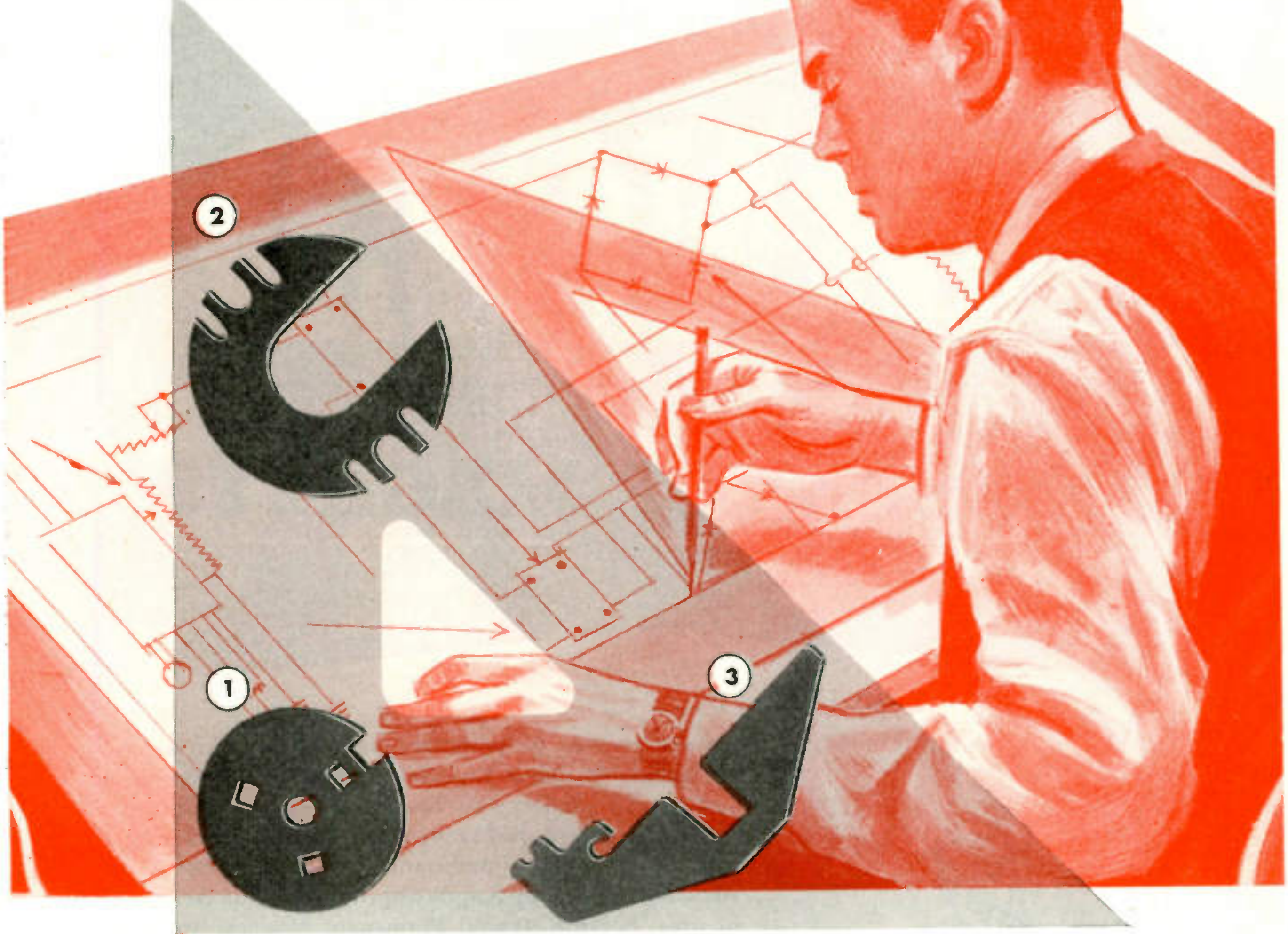
to DELIVERY



As insulating parts and structural members, Taylor Phenol Fibre and Taylor Vulcanized Fibre have literally hundreds of applications in the electrical industry.

Not the least of their advantages is the speed and versatility of fabrication. Sheets, rods, and tubes of Taylor Laminated Plastics machine with such ease, and such precision, that parts can usually be delivered to stock rooms well in advance of requirements . . . helping to solve many a production headache.

If you do your own fabricating, Taylor can supply you with Phenol Fibre, Vulcanized Fibre, or special formulations . . . *and* with valuable advice to increase the speed of your production.

1. **Contact insulation washer, stamped from Taylor Phenol Fibre sheet.**
2. **Switch insulator, stamped from Taylor Vulcanized Fibre sheet.**
3. **Support member, stamped from Taylor Phenol Fibre sheet.**

If you seek a source of supply for finished parts, Taylor again is your answer. Taylor's completely equipped Fabricating Service is always at your call.

Whatever your problem, mechanical or electrical, our engineers will be glad to tell you exactly what Taylor can do for you. Write today, sending sketch or blueprint.

TAYLOR FIBRE COMPANY

LAMINATED PLASTICS: PHENOL FIBRE • VULCANIZED FIBRE • Sheets, Rods, Tubes, and Fabricated Parts

NORRISTOWN, PENNA. **Offices in Principal Cities** **Pacific Coast Plant: LA VERNE, CAL.**

THERE'S PROFIT FOR YOU IN

THE TIME AND MONEY-SAVING QUALITIES OF

ARNOLD

PERMANENT MAGNETS

Several avenues of profit are open to you in Arnold Permanent Magnets. You can improve the performance and overall efficiency of equipment. You can increase production speed, and in many cases reduce both weight and size. And most important, you can maintain these advantages over any length of production run or period of time, because Arnold Permanent Magnets are completely quality-controlled through *every* step of manufacture—from the design board to final test and assembly. You'll find them unvaryingly uniform and reliable in every magnetic and physical sense.

It's our job to help you discover and then fully attain these benefits. Arnold Products are available in all Alnico grades and other types of magnetic materials—in cast or sintered forms, and in any size or shape required. Our engineers are at your command—check with our Chicago headquarters, or with any Allegheny Ludlum branch office.

W&D 1298

THE ARNOLD ENGINEERING CO.

Subsidiary of **ALLEGHENY LUDLUM STEEL CORPORATION**

147 East Ontario Street, Chicago 11, Illinois

Specialists and Leaders in the Design, Engineering and Manufacture of PERMANENT MAGNETS

Optical Microscope IS ANOTHER PRECISION INSTRUMENT EMPLOYED BY SYLVANIA TO ASSURE TUNGSTEN AND SPECIAL ALLOY WIRE PERFECTION ★ ★ ★

Studying the crystals in a magnified section of Sylvania tungsten and special alloy wire is just one more phase of Sylvania's never-ending efforts toward higher and higher quality in radio and electronic tubes.

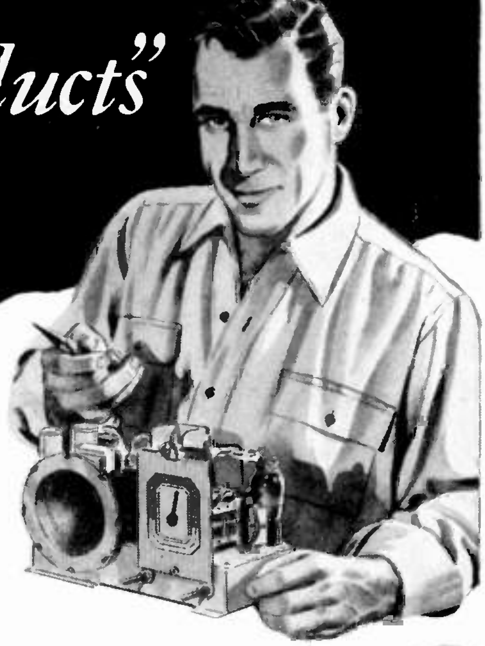
With the optical microscope shown above, the image of the wire section, magnified as much as 2,000 times, can be projected on the ground glass seen to the left. To the metallurgist who uses this instrument, the sizes, shapes and distribution of the crystals in filamentary and heater wires are extremely important. On the microscope, he examines and studies specific crystal features that are essential for long tube life.

Sylvania's research metallurgy facilities are in constant touch with the Sylvania wire plant in Towanda, Pa., and special alloy wire plant in Warren, Pa., to assure superlative products. Sylvania Electric Products Inc., 500 Fifth Avenue, New York 18, N. Y.

SYLVANIA ELECTRIC

RADIO TUBES; CATHODE RAY TUBES; ELECTRONIC DEVICES; FLUORESCENT LAMPS, FIXTURES, WIRING DEVICES; PHOTOLAMPS; LIGHT BULBS

"I like the DEPENDABILITY of OHMITE Products"



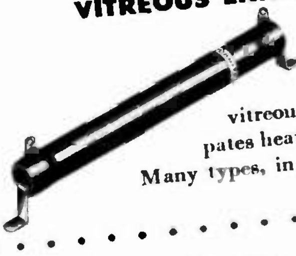
Among radio engineers everywhere—there's a definite preference for Ohmite resistance products. These men know—from experience—that Ohmite rheostats, resistors, and chokes provide long, trouble-free service.

Here's the reason why you get extra performance. Every Ohmite product is designed and constructed to stand up under severe operating conditions. Every unit is built to withstand the effects of shock, vibration, temperature extremes, altitude, and humidity. Make sure you get the benefit of this unfailing dependability. Ask for Ohmite products by name.

CLOSE CONTROL RHEOSTATS

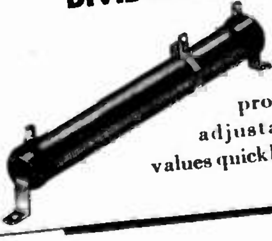
Here is the most extensive line of rheostats offered today . . . 10 sizes, from 25 to 1000 watts, with many resistance values in each size. All-ceramic construction. Windings are locked in vitreous enamel.

VITREOUS ENAMELED RESISTORS



Wire wound on a ceramic core, rigidly held in place, insulated and protected by vitreous enamel. Even winding dissipates heat rapidly—prevents hot spots. Many types, in ratings from 5 to 200 watts.

DIVIDOHM ADJUSTABLE RESISTORS



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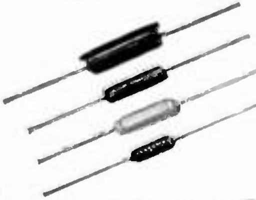

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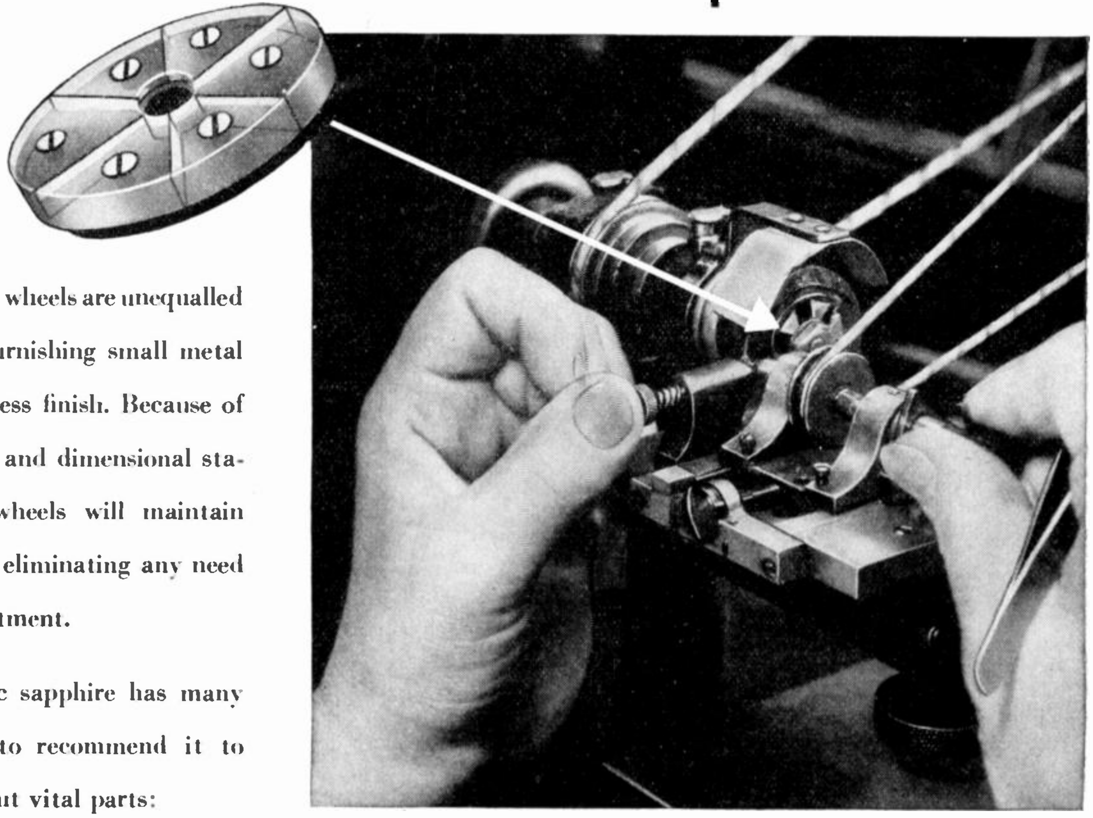

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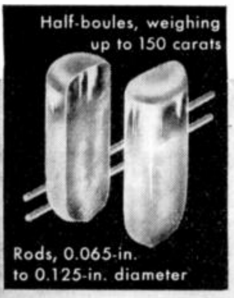

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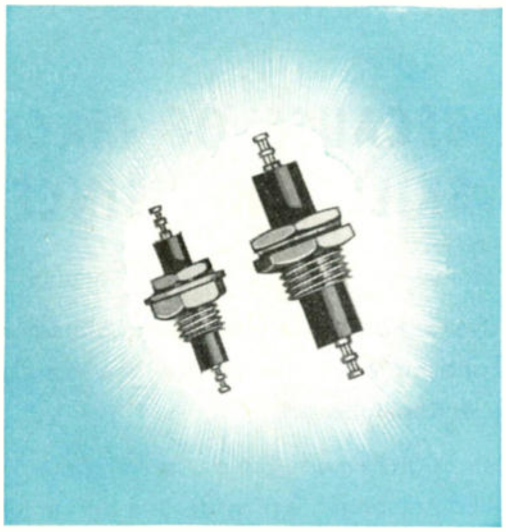

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C.T.C. 1795B Insulated Feed-Thru Terminals fulfilled every requirement. Design-features like these show you why: *Rugged construction that withstands loosening under vibration or shock . . . approved phenolic insulating material, JAN type LTS-E-4 . . . brass bushings, cadmium plated . . . brass thru-terminals, silver plated for easy soldering.*

SPECIFICATIONS

The 1795B mounts in a ¼" hole, and has an over-all length of approximately ⅞". C.T.C. Feed-Thru Terminals are available in additional sizes. The 1795A is similar to the 1795B, but with an over-all length of 1". Also similar in design and function are X1771A and X1771B, but larger in size and mounting in a ⅜" hole. Breakdown voltages, at 60 cycles R.M.S., are:

1795A . . .	3800V	X1771A . . .	8200V
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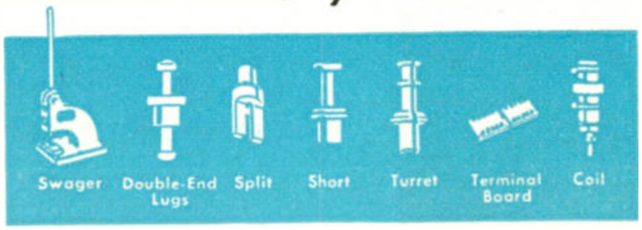

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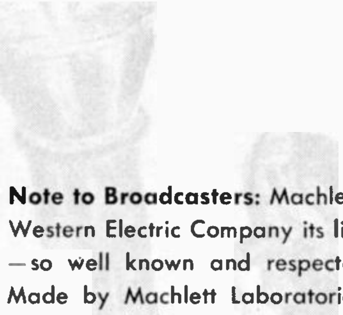
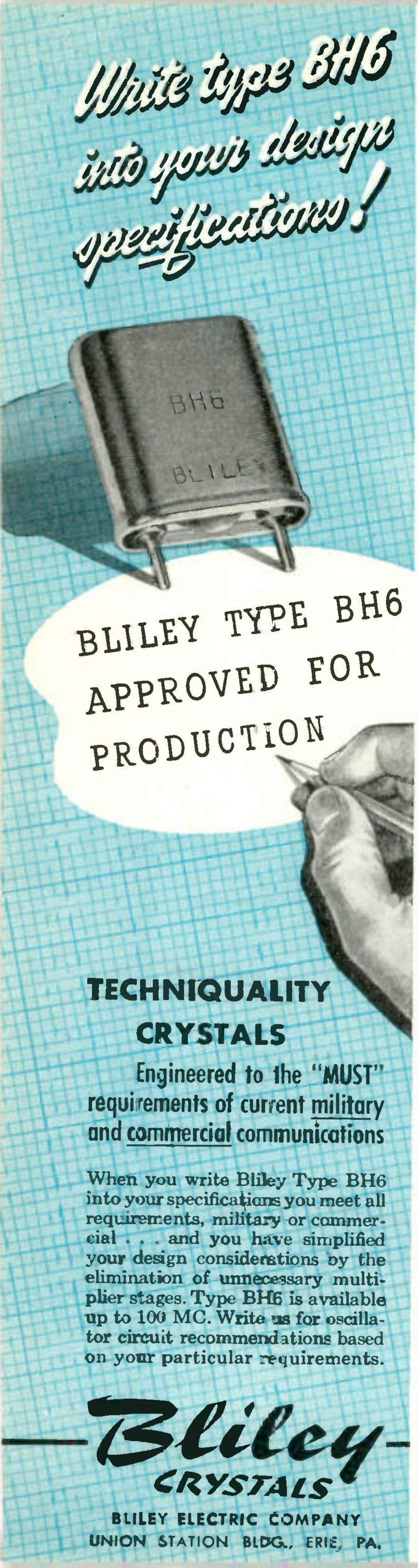


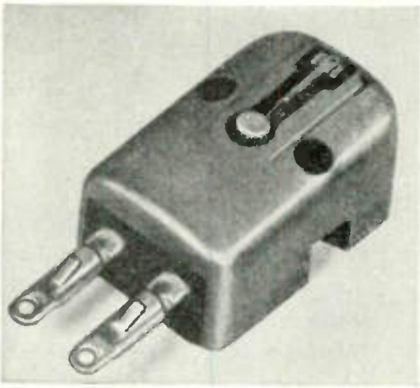


NEWS and NEW PRODUCTS

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

Microgroove

This fall our industry has seen one outstanding development in the home entertainment field, that interests all engineers conscious of their need to keep abreast of the radio field. The vast amount of interest in those markets where it has been shown, has caused many firms to announce materials and accessories for the reproduction of the new long-playing Columbia Microgroove records. Changers, pick-ups, new cartridges, and similar accessories are flooding the market, not to overlook the adaptation of this feature to the completed sets offered the buying public. When far enough along in the development work Columbia records asked the co-operation of Philco to make a player attachment, with slow speed motor and light crystal cartridge equipped playing arm. Since then many other manufacturers have added these and other components to their lines.



General Electric, Electronics Division, Electronics Park, Syracuse, N. Y., who caused the largest stir in the phonograph-reproducer field, by the introduction about 18 months back, of their D1 RM 66 variable-reluctance cartridge, has entered the microgroove field with the announcement of the MICRO-GROOVE MODEL featuring a 1-mil-radius stylus, low mass of the moving system, and high compliance, to meet the requirements of the new records. Physically, the unit is one-third smaller than the predecessor model for standard records. All of the salient features such as quietness, absence of needle talk, long record life, etc., have been brought into this new design.

Gray Research & Development Co., 16 Arbor Street, Hartford 1, Conn., have introduced their version of the older GE cartridge, to which has been added a new moving system, of extreme lightness of mass, high tracking compliance, and proper stiffness, together with a diamond stylus of 1-mil radius to fit the shape of the new grooves. New damping blocks also have been fitted to Gray cartridge.

Webster-Chicago, 4245 N. Knox Ave., Chicago, Ill., has announced as available two models of a record changer capable of playing both 33 and 78 rpm records, equipped with a special needle in their "Tilt-o-Matic" tone arm. These changers play both types of records on an automatic or manual control basis, and afford the added advantage of disengaging the driving rubber idler wheels when in the off position, thus preventing the idlers from forming bumps or flats, and giving quieter operation.

Zenith Radio Corp., 6001 W. Dickens Avenue, Chicago 39, Ill., has announced that the console models of their fall line have microgroove feature added. Magnavox, and many more of the set makers have jumped aboard the bandwagon, to offer the buying public the means with which to play this latest development, in the home entertainment field.

Without philosophizing on the merits of this new offering, it is interesting to see the way in which manufacturing ingenuity and speed of production can accrue to fill a market readymade for it, when some enterprising firm takes it upon itself to pioneer a new product.

Recent Catalogs

National Co. 61 Sherman Street, Malden, Mass. announces a new television receiver, with 7" tube, covering all 13 channels, scheduled for fall marketing. Techniques developed by this renowned company in stable communication receiver development find their way into the video sets they will offer. A novel innovation is the employment of dual 6" oval loudspeakers flanking the video tube, for more realistic sound reproduction. Full particulars will be supplied for the asking.

Lenkurt Electric Co., 1129 County Road, San Carlos, California, announce a new catalogue comprising a comprehensive listing of the carrier-current telephone and telegraph equipment of their manufacture, form CX-42. In addition to the communication equipment described, ringing, dialing, and telemetering apparatus are covered, as well as selected test equipment, especially designed for operational servicing of carrier-current apparatus, is displayed and described. Comprehensive bulletins on each system are also available from the company.

Metalace Corp., 2101 Grand Concourse, New York 53, N. Y., have announced through a descriptive sheet their all-purpose antenna mounting base for affixing an antenna to a chimney or roof parapet. With the increased number of FM and video antennas finding their way to the rooftops of apartment houses and rural homes, such an accessory will lend itself readily to the safe and workmanlike installation of the antenna supporting pole. Simple tools alone are needed to make the installation, following the complete instructions furnished with the new stand.

(Continued on page 48A)

Representative RCA types standardized for future equipment designs

Tubes for today and tomorrow

HERE ARE power tubes, phototubes, and c-r tubes to serve the major requirements of equipment manufacturers for a long time to come. The tubes listed are those you can depend on now, and for your future designs.

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Gas Types	1P41	921	927	930
Vacuum Types	922	929		
Multiplier	931-A			

GAS TUBES

Thyratrons	2D21	3D22	884	2050	5563	
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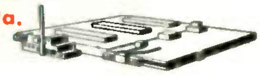

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RADIO CORPORATION of AMERICA

HARRISON, N. J.

PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 36 *December, 1948* NUMBER 12

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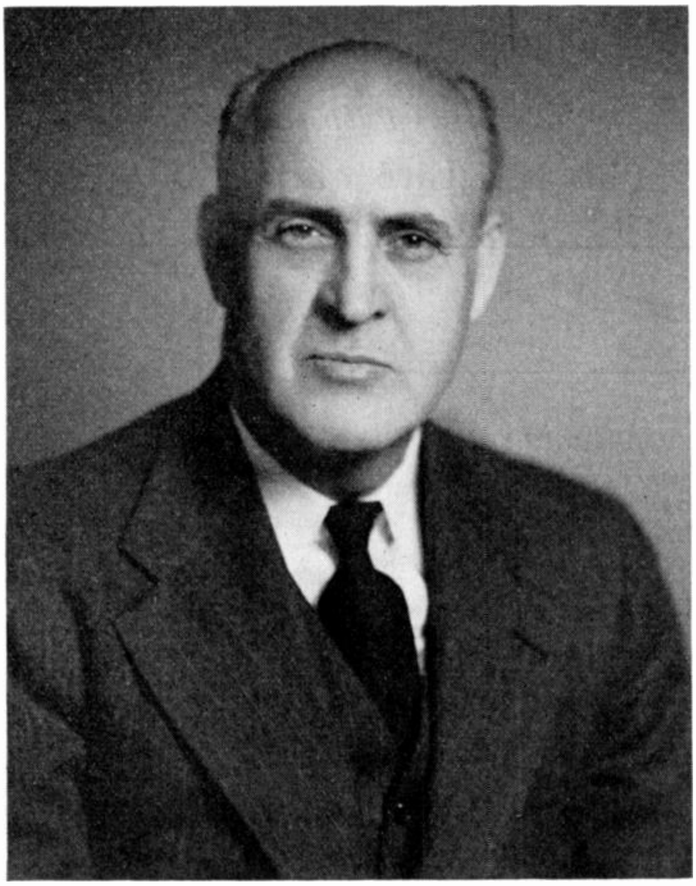

Frederic Stanley Howes

DIRECTOR, 1948

Frederic Stanley Howes, associate professor of electrical engineering at McGill University in Montreal, was born on July 25, 1896, at Paris, Ont., Canada. Graduating from high school in 1916, he immediately enlisted in the Canadian Army and served as a signaller until his discharge at the end of the First World War. He thereupon entered industry and worked as a sheet-metal layout man until 1920, when he matriculated at McGill University in order to complete his interrupted studies.

In 1924 Mr. Howes received the B.Sc. degree from McGill and was then appointed an instructor there. He continued in that post until 1927, having been awarded the M.Sc. degree the year before. His graduate work was finished in England where the City and Guilds Institute, Imperial College, University of London, gave him the doctorate in electrical communication engineering following a two-year period of study there.

Immediately afterward, in 1929, Mr. Howes rejoined McGill as an instructor and has taught there since with but one interruption—after becoming an Associate of the IRE in 1937, he spent the following year, 1938–1939, in studying advanced communication engineering at the University of California in Berkeley, U. S. A.

In 1940, shortly after Canada entered the War, Dr. Howes was appointed a civilian instructor at the RCAF's No. 1 Wireless School at McGill. Meanwhile, during the summers of 1941 and 1942 he did war development work at the Northern Electric Company in Montreal. In 1942 he organized graduate evening courses in communication engineering at McGill. The year 1943 heralded Dr. Howes appointment to an assistant professorship; his opening of a private consulting practice as a radio engineer specializing in broadcast antenna design and related problems; and his being transferred to the rank of IRE Member and advancement to Senior Member.

Dr. Howes became an associate professor in 1946. Currently, in addition to his lecture work, he is in charge of the radio engineering laboratories at the University and is supervisor of graduate evening classes at McGill and Ottawa both.

Elected Chairman of the Montreal Section for the year 1945, Dr. Howes was also appointed Chairman of the IRE Canadian Council, a position which he still holds. He is the Canadian IRE representative on the administrative board of the Canadian Radio Technical Planning Board and also represents the IRE on the Canadian Council of Professional Engineers and Scientists. Chairman of the latter organization, which is composed of the Presidents of eleven national engineering and scientific organizations, Dr. Howes also is a member of the Council of the Corporation of Professional Engineers of Quebec.

There has been a wealth of sometimes heated argument as to the relative opportunities for engineers in private versus governmental employ. One side of this question is clearly and forcefully presented, with substantiating reasons, by the writer of the following guest editorial—a member of the IRE Board of Editors and a skilled communications engineer who, after seventeen years of experience in private industry, has joined the staff of the Sandia Base Branch of the Los Alamos Scientific Laboratory.—*The Editor.*

Are You Satisfied?

C. W. CARNAHAN

To those radio engineers who are happy in their present work, this editorial can have little interest. It is addressed primarily to those who had real jobs to do during the war, and who now find their return to peacetime pursuits something of a let-down.

This group will remember that their war labors, while long and arduous, were vital and exciting, and that they demanded the utmost of their skill and knowledge. Things were built in the way they should be built, neglected text books had to be dug into again, and sudden accretions of knowledge in entirely new fields had to be gained. Results were measured in terms of lives and victories, not in figures on a quarterly statement.

Since the end of the war, however, many engineers have returned to their prewar occupations, where their technical achievements now may consist of such exciting things as saving ten cents on that new receiver, or making a further compromise with quality to beat that other outfit. They have returned to the normal battle with the sales department, whose decisions and salaries invariably outrank their own. Many of them look back with longing to the days when vice-presidents made fair expediters, and sales departments, while drawing those fat salaries, of course, at least had to listen to the engineers without any backtalk. In their leisure, they now read in the PROCEEDINGS about a rapidly expanding art which is growing beyond them.

Continued government spending in the postwar period has already had a considerable effect in those radio industries which existed before the war. The unfilled demand for engineers in government work has made more valuable commodities out of those still in private industry. This point management usually concedes with reluctance, being prone to indicate to dissatisfied engineers the horrors of governmental work, the insecurity, the red tape, bureaucracy, restriction of publications, and so forth. Those old timers who remember the security and other advantages offered by private industry in the thirties will be able to assess these arguments at their true worth.

Financially, the average engineer will not be any more underpaid in government work than he is in private industry. As for job security, there are more problems now than there are men to solve them, and new ones arise every day Furthermore, the advances in knowledge and new techniques resulting from this work will create new jobs in private industry for those who can fill them. There are publication restrictions, of course, but to compensate there are interesting, demanding work on the frontiers of radio knowledge, unlimited facilities to work with, and no sales department on the engineer's neck. What more can a good engineer want?

So far, this has been an appeal to the self-interest of the individual engineer. There is, however, a larger side to this. Good men must be found for the unfilled government jobs now available, if our stated objectives of national security are to be attained. If a fair percentage of those who performed so splendidly during the war will take up where they left off, this need will be met.

A Digital Computer for Scientific Applications*

C. F. WEST†, MEMBER, IRE, AND J. E. DeTURK†

Summary—During the past two years development has been initiated on several large-scale automatic digital computing machines, both in this country and abroad. The present paper is concerned with the over-all organization of one such machine. A logical division of the machine into four major components is described, and the machine performance is interpreted in terms of these component functions. The electronic techniques used to accomplish the storage, transmission, and arithmetic manipulation of numbers, together with certain methods used for control of the computer, are briefly discussed. Although the paper is concerned with the design of a particular machine, it is felt that the design problems and engineering techniques are applicable to most large-scale computing machines.

I. INTRODUCTION

THE TERM "digital computer" applies to a calculating machine in which a number can assume only a discrete value the precision of which is determined by the number of digits used for its representation. A desk calculator is a digital (or discrete-variable) computer, whereas a slide rule is a continuous-variable computer. A large-scale digital computer is a machine not only capable of digital computation, but one which can perform long sequences of computations in accordance with a pre-established program of operation. Machines of this type are also referred to as sequence-controlled calculators. Such a computer is capable of solving complex problems involving thousands or millions of individual arithmetic operations without the intervention of a human operator.

Large-scale computers may be divided roughly into two categories: Scientific machines which are designed to perform large numbers of calculations based upon relatively few input data and yielding relatively few output data, and statistical machines of which the opposite is true. The calculator to be described is intended primarily for scientific applications. Some of the types of problems which the present machine is intended to solve are the following:

1. The systematic handling of linear arrays.
2. Solution of the partial differential equations of hydrodynamics.
3. Fourier analysis and synthesis.
4. Applications of electromagnetic theory.
5. Study of shock waves.
6. The solution of nonlinear differential equations.
7. The problem of systematic sorting.

The only true computing operations which the machine can perform are the basic arithmetic operations of addition, subtraction, multiplication, and division. Before the above complex problems can be presented to the machine for solution, they must be reduced to arithmetic processes through application of the methods of numerical analysis. These arithmetic routines must then be expressed in terms of coded commands which the machine is capable of following.

II. MACHINE ORGANIZATION

Fig. 1 is a block diagram showing the principal components of a large-scale digital computer. The arithmetic unit is the only true computing unit in the machine. That is, it is the only one capable of generating new numbers. The internal (or high-speed) memory is a storage place for numbers and commands. During computation, numbers which serve as operands are transferred from the internal memory to the arithmetic unit, where the arithmetic operations take place. The result of each arithmetic operation is returned to the internal memory. The central control unit of the machine governs the exchange of numbers between the internal memory and the arithmetic unit. Central control governs this exchange in accordance with orders or commands which are also located in the internal memory. For each arithmetic operation, the central control must select two operands from the internal memory, and must supply these to the arithmetic unit. It must designate to the arithmetic unit which operation (e.g., addition, division) is to be performed, and must transfer the result of the operation to a selected memory position. Central

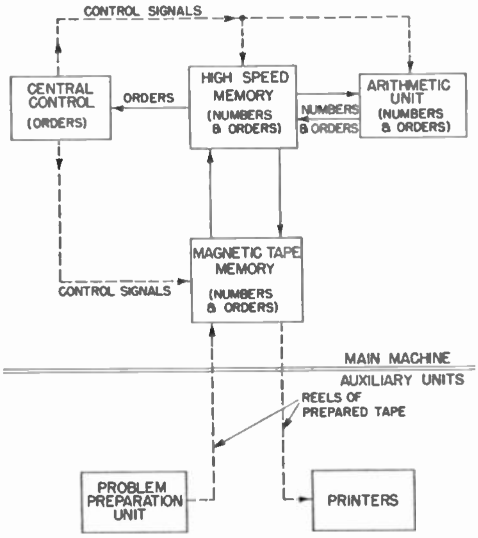

Fig. 1—Block diagram of the digital computer.

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* Decimal classification: 621.375.2. Original manuscript received by the Institute, May 19, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 25, 1948.

This paper describes preliminary computer design studies conducted for the National Bureau of Standards under Contract CST 8092.

† Raytheon Manufacturing Company, Waltham, Mass.

control then initiates the next operation by selecting from the internal memory the next command.

The magnetic memory units are used to supplement the internal memory. These units store numbers and orders on magnetic tape. The speed of operation of the magnetic units is considerably less than that of the internal memory, but the storage capacity is many times greater. These units also serve as input-output devices for the computer.

The page printers and problem-preparation unit shown in Fig. 1 are auxiliary units which are not directly connected to the main part of the machine, but which communicate with the computer by means of the magnetic memory units. The problem-preparation unit consists of a manually operated keyboard which is used to record initial numbers and commands on magnetic tape. This device makes use of additional magnetic storage containing the commands for frequently used computing routines. Thus, certain complete routines may be introduced into the computer by a single manual operation. The page printers are electrically operated typewriters which respond to signals recorded on magnetic tape. They are used to record the final results of computation.

Because one command is required for each arithmetic operation, it might seem that a prohibitive number of commands would have to be introduced into the machine in order to direct the solution of a relatively simple problem. This is not the case. The iterative methods of numerical analysis involve the repeated performance of computing routines. When a routine is repeated, the commands governing the computation may differ from those of the previous cycle only with respect to some systematic pattern of variation. By storing commands in the internal memory and by the use of suitable schemes for their coding, they may be introduced into the arithmetic unit and modified by addition or subtraction. As an example of the effectiveness of this process, the total number of commands which must be supplied to the machine to obtain all of the roots of a polynomial equation should not exceed fifteen. This number is independent of the degree of the polynomial.

The complexity of the problems which a machine can solve efficiently is limited both by computation speed and memory capacity. For example, partial differential equations in three dimensions and time may require a total storage of 10^6 numbers and may involve 10^{10} arithmetic operations. In the present machine, the internal memory has a capacity of approximately 4000 numbers, and the permanent storage medium associated with each magnetic memory unit has a capacity of 200,000 numbers. The over-all speed of the computer depends primarily upon the time required to perform the basic arithmetic operations and the time required to select a number from the internal memory. In this machine, 900 arithmetic operations together with the associated memory selections are performed each second.

Since the entire function of the machine is to carry out numerical computation in accordance with coded commands, the representations of numbers and commands are basic elements of the machine design. Numbers may be represented in a variety of ways, depending upon both the mathematical and the physical means employed.

Mathematical representations may employ different number bases. Thus an n-digit decimal number (base ten), as conventionally written, is a shorthand expression for the quantity.

$$A_n\cdot 10^n+A_{n-1}\cdot 10^{n-1}+\cdots+A_j\cdot 10^j+\cdots+A_0\cdot 10^0 \qquad (1)$$

where the integer coefficients A_j are the digits of the number. Any A_j in the decimal system may take on a value from 0 through 9. When a number is represented to the base X, it is still written as a sequence of digits A_j, but these now are interpreted as meaning

$$A_n\cdot X^n+A_{n-1}\cdot X^{n-1}+\cdots+A_j\cdot X^j+\cdots+A_0\cdot X^0 \qquad (2)$$

and any digit A_j can now take on only the values 0 through $X-1$.

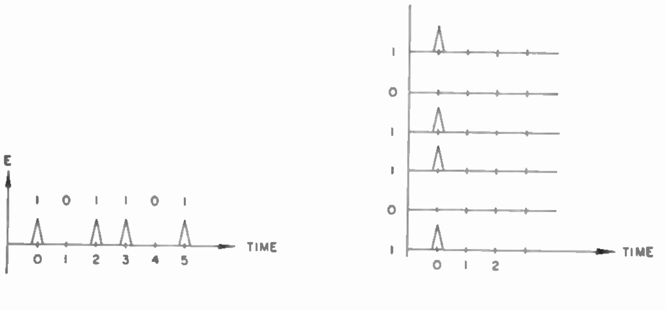
In addition to the decimal notation, this paper will refer to the binary scale of notation; i.e., $X=2$ in (2) above. Here the only possible digits are 0 and 1. The binary equivalents of the decimal numbers 0 through 15 are shown in Table I.

TABLE I
DECIMAL BINARY EQUIVALENTS

Decimal	Binary	Decimal	Binary
0	0000	8	1000
1	0001	9	1001
2	0010	10	1010
3	0011	11	1011
4	0100	12	1100
5	0101	13	1101
6	0110	14	1110
7	0111	15	1111

The physical representation of a number to the base X requires a physical representation for each of the possible digits (0 through $X-1$) in each of the n-digit columns. That is, each number is denoted by a particular selection from X^n physical states. For a given number base, several physical representations are possible, depending upon the number of temporal and spatial selections used to designate the number. Fig. 2 shows two ways in which the binary number 101101 may be represented. At (a) the number is being transmitted serially on a single wire and the representation is entirely temporal. At (b) the number is being transmitted in parallel on six wires and the representation is entirely spatial. In either case, the digit 1 is represented by the presence of a pulse, and the digit 0 by the absence of a pulse.

Because the rules of arithmetic are simpler in the binary notation than in any other base, this notation is used in the present machine. Binary-decimal conversion is required at the input and output of the machine. This may be justified in a calculator for scientific problems because of the large amounts of calculation that are done with comparatively few initial data.



(a) (b)

Fig. 2—Serial and parallel representation of the number 101101. (a) Serial transmission on one wire. (b) Parallel transmission on six wires.

When discussing machine operation, it is convenient to speak of a composite pulse group of fixed length as a "word." A word in the present machine contains 45 pulse positions or binary digits which are transmitted between units in a serial manner. Two kinds of words are shown in Fig. 3. These are: (1) a "number word," which contains the absolute value of a number, its sign, and auxiliary digits used in checking; and (2) a "command word," which contains coded pulse groups capable of governing the machine operation.

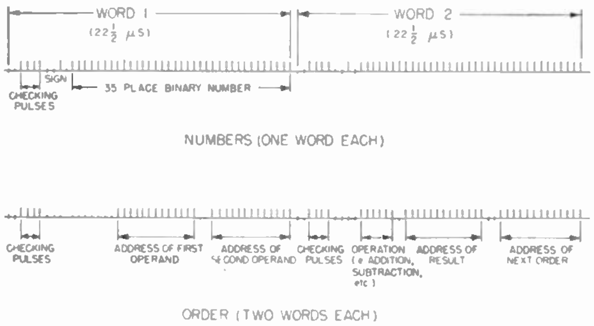


Fig. 3—Allocation of information in number and order words.

The machine cycle is the basic unit of computing operation, and in this machine is approximately 1 millisecond in duration. During this cycle four distinct events take place. (1) Two operands are selected from the internal memory and are sent to the arithmetic unit. (2) The arithmetic unit performs the desired operation. (3) The result of the operation is sent back to the internal memory. (4) The command which governs the next operation is selected from the memory.

A command is required for each machine cycle and contains all the information necessary for the performance of the cycle, namely, the locations of the operands, the specification of the operation, the location which is to receive the result, and the location of the next command. Locations within the memory are termed addresses, and are specified by binary numbers which identify consecutively all of the storage positions of the memory. Operations are also specified by coded pulse groups.

Besides addition, subtraction, etc., several nonarithmetic operations are required during the solution of most problems. The transfer operation serves to transfer a word between different memory locations, or between the internal memory and one of the magnetic memory units. The substitution operation is used to modify a command word by adding or subtracting from one or more of the addresses contained within the command.

The branch operation allows the machine to choose between two computing routines on the basis of the results of past computation. The command governing the branch operation contains the locations of two commands, only one of which is to be chosen to govern the next computing cycle. This choice is determined by the sign of the inequality of the two operands. As an example of the use of the branch operation, consider the solution of a polynomial equation. An approximate root of the equation is calculated by means of computing routine A. The difference between this approximation and the last approximation is obtained. If this difference is greater than some preassigned number, the next command selected is the first command of routine A, and its selection results in the calculation of a nearer approximation. If the difference is less than the pre-established tolerance, the next command selected is the initial command of routine B, which initiates reduction of the degree of the polynomial in preparation for the calculation of the next root.

III. INTERNAL MEMORY

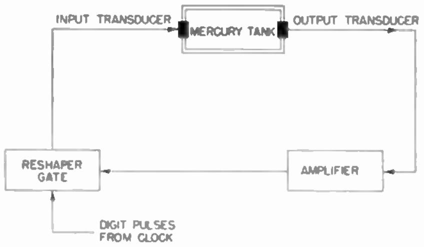


Fig. 4—Delay-line memory unit.

The internal memory must be capable of storing a large number of words with short access time. The stored information must be easily erasible. In the present machine, the internal memory makes use of the acoustic delay line as the storage mechanism. Fig. 4 is a block diagram of an acoustic delay line in which mercury is the acoustic medium. Assume that the line contains some pulse configuration at a given instant of time, and that the configuration is propagating down the line at the velocity of sound in mercury. Each pulse received at the output transducer is amplified and applied to a gate or reshaper which allows a new digit pulse from a continuous pulse source to be fed to the input transducer. Thus, the pulse configuration will circulate around the closed loop indefinitely without progressive degeneration of the pulse shapes or the pulse spacing.

The number of pulse positions which a line contains (i.e., its storage capacity) is proportional to the time delay of the line and the frequency of the continuous pulse source. The former is limited by the attenuation of the mercury and the transducers; the latter, by the bandwidth which can be attained around the circulation path and the dependence of attenuation on frequency.

The access time of the delay line is equal to the delay time. For a given memory capacity, the access time may be decreased by decreasing the length of each line and increasing the total number of lines while the repetition rate is held constant. Reduction obtained in this manner is costly, for the total memory equipment is primarily proportional to the number of circulation paths and is only secondarily influenced by the attenuation per path. The access time can be reduced more effectively by increasing the pulse-repetition rate and shortening the line, while holding constant the number of lines and the pulse capacity of each.

The present machine design makes use of 255 acoustic delay lines, each capable of storing 16 information words and one additional word used for checking. The total memory capacity is, therefore, 4080 words. The delay lines operate at a pulse-repetition rate of 2 Mc, and the digit pulses are amplitude-modulated upon a 30-Mc carrier. The memory access time is about 380 microseconds.

Fig. 5 is a block diagram of a delay line showing the use of additional gates for reading (i.e., taking information from the line), writing (i.e., putting information into the line), and erasing. The erase gate and the write gate are normally connected together, so that a word is erased only when a new word is written into the line.

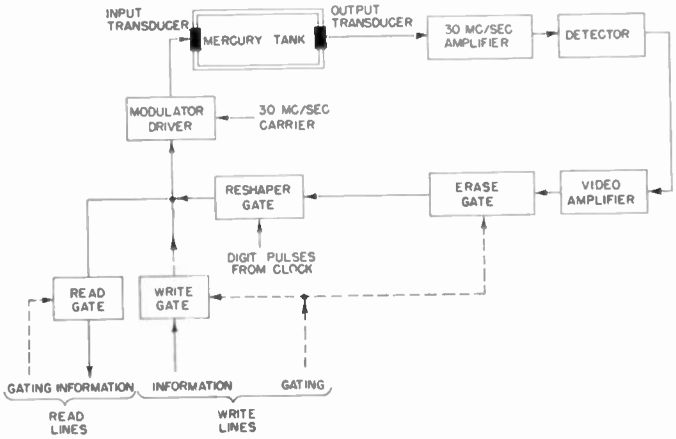


Fig. 5—Organization of memory-delay-line circuits.

Continuous circulation through a delay line requires that the delay time be a constant integral multiple of the period of the continuous pulse source used for reshaping. Because the acoustic velocity in mercury is temperature-dependent, some means of temperature control is required. The temperature coefficient of acoustic velocity is such that a 1 per cent change in delay time results from a temperature change of 30°C. The temperature effect may be expressed more conveniently by the equation

$$\Delta T = \frac{3000\eta}{N}\ {}^{\circ}\mathrm{C}. \qquad (3)$$

where ΔT is the permissible temperature variation in degrees centigrade, N is the total number of pulse positions in the line, and η is the fraction of a pulse period by which the delay can change and still allow reshaping of the circulating pulses. By using sharp pulses in the reshaping operation, the value of η may be made as high as 0.75 without difficulty. In the present design, each delay line contains 765 pulses, and the corresponding permissible temperature gradient is approximately 3°C.

Temperature gradients in the memory can be controlled, while reasonable equipment accessibility is maintained by subdividing the memory into groups of lines. Each group may consist of several (say, 6 to 10) independent acoustic paths operating within a single container of mercury. The relatively high thermal conductivity of mercury is effective in keeping the gradients between paths small. Each container is then supplied with an independent temperature-control mechanism which maintains the temperature of the pool constant.

Fig. 6 is a photograph of a mercury pool contained in a stainless-steel tank less than 7 inches long, 2¼ inches high, and 2¼ inches wide. Three acoustic paths operate within the pool. Each path is multiple-reflecting and consists of three travels the length of the container. The circulation time of each path is 320 microseconds. This pool has been used in a prototype memory unit

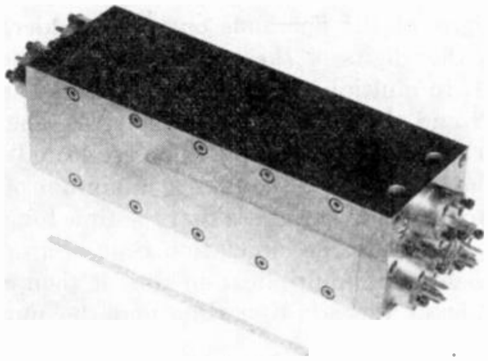


Fig. 6—Photograph of a mercury pool contained in a stainless-steel tank, with three acoustic paths operating within the pool.

with temperature control used to maintain synchronization between the delay line and a fixed-frequency pulse source.

IV. Arithmetic Unit

The basic arithmetic operation is addition. Other arithmetic operations, though coded as single operations, are compounded of successive additions performed under the local control of the arithmetic unit. Accordingly, the mechanism of addition largely determines the manner in which other operations are carried out. Addition may be performed serially or in parallel. In a serial adder the two operands are added one digit at a time, commencing with the lowest-order digit. In a parallel adder, all of the digits of one operand are simultaneously added to the digits of the other operand, thereby generating all the digits of the sum at once.

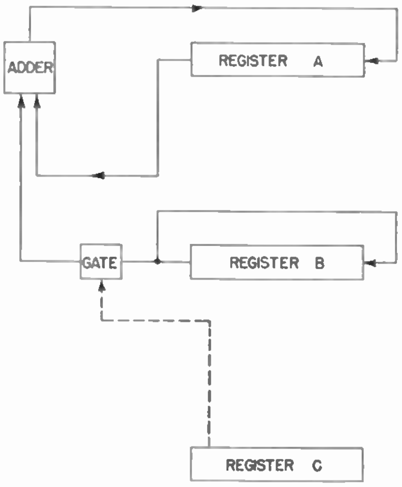


Fig. 7—Serial-adder operation.

Fig. 7 is a block diagram indicating the operation of a serial adder. The two operands are assumed to be in registers A and B which may be delay lines. The successive digits of the operands enter the adder, which generates the digits of the sum and enters these into register A. In multiplication, the multiplicand stands in register B and the multiplier in register C. The partial sums generated during multiplication are stored in register A. The serial adder requires a minimum of equipment. Its chief disadvantage is that the time for addition cannot be less than the circulation time of an operand in its register. The multiplication time is then equal to the product of the addition time and the number of digits used.

Fig. 8 is a block diagram of a parallel adder. The addend stands in register A and the augend in register B. Upon application of a control pulse, all columns of the addend are simultaneously added to the augend and the sum is left standing in the sum-augend register. In principle, the addition is completed in one pulse time, although practical considerations generally require that the process take from 5 to 10 pulse times.

Speed is obtained at the expense of equipment in the parallel adder, for the basic columnar adding circuit must be repeated for each digit column of the addend and augend. A further advantage of the parallel over the serial adder is that the parallel unit requires less complex control circuits to perform operations compounded of repeated additions.

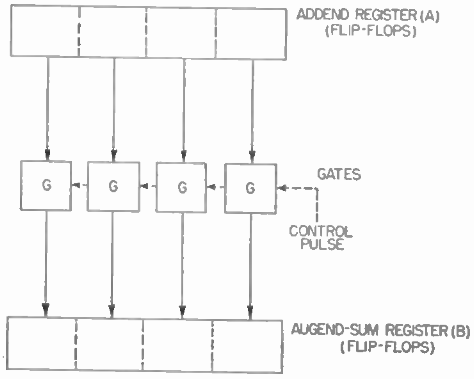


Fig. 8—Parallel-adder operation.

With either type of adder several variations are possible, depending upon the method used to transfer carries from one column to another. For example, a parallel adder may first perform the addition without carries and later add the carries to the result. The addition of carries may generate new carries which must be added in again. Such sequential adding of carries increases the time required for an addition. Simultaneous addition of carries may be performed by adding all carries to the augend in one operation. In this case, the addition speed is limited only by the propagation times of the carry pulses and the add pulses through the necessary circuits.

Before discussing the adder circuits in detail, it is worth while to describe a shift register. This is a simpler device than the parallel adder, but involves similar techniques. The register may be used to convert between serial and parallel number representation, to change the repetition rate of a serially transmitted number, and to act as buffer storage for numbers entering the arithmetic unit.

The register consists of a chain of flip-flops, one for each digit to be stored, as shown in Fig. 9. One plate of each flip-flop is connected to the grid of an adjacent flip-flop through an electrical delay network. Provision is made for applying a reset pulse simultaneously to all flip-flops in the register.

If any configuration of ones and zeros stands in the register and a reset pulse is applied to all flip-flops, those which stand at 1 undergo a change of state. This change produces a pulse at one of the plates which is momentarily stored in the delay network associated with that plate. After the reset pulse has passed, every pulse which entered a delay network emerges and triggers the flip-flop immediately to the right of it. In this manner the configuration of ones and zeros is shifted one column to the right.

In Fig. 9 a three-digit word is shown serially entering the shift register. A train of reset pulses is applied to the shift register at the same repetition rate as the digits of the entering word. The reset pulses must be advanced, with respect to the applied digit pulses, to avoid interference between the two pulse trains. Once the word stands in the register it may be read out in parallel from the plates of the flip-flops, or it may be read out serially at some other repetition rate by applying reset pulses of the desired frequency. Fig. 10 is a circuit diagram of two columns of a shift register which operate reliably at pulse rates up to 2 Mc.

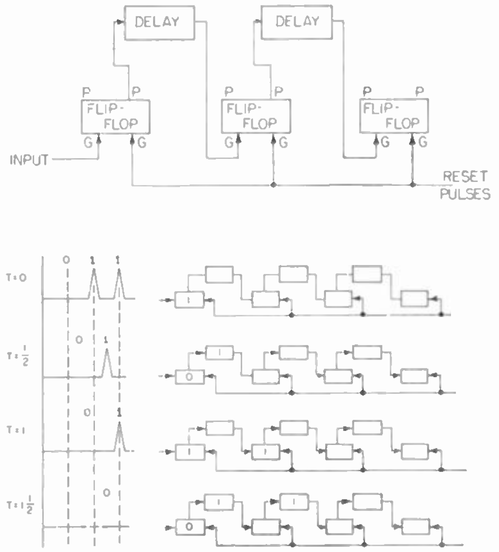


Fig. 9—Shift-register operation.

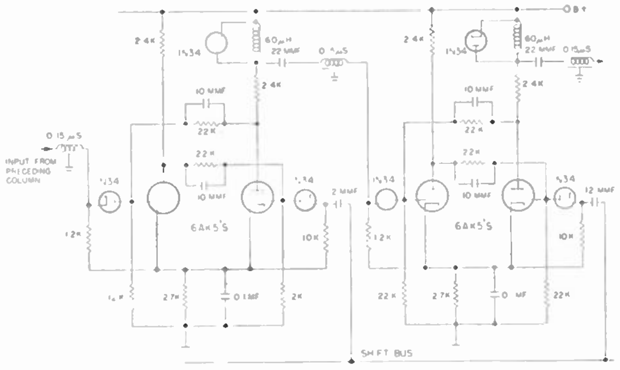


Fig. 10—Shift-register column.

The present machine makes use of a parallel adder with simultaneous carry, as shown in Fig. 11. In this circuit the addition (without carry) of the addend to the augend occurs first, and is followed by the simultaneous addition of all carries. The operands are assumed to stand in the addend (*A*) and the augend-sum (*B*) registers. An add control pulse is applied to the gates *G1*. In each column where the addend contains a 1, the control pulse passes through the normally closed gate *G1* and changes the state of the corresponding column of the augend-sum register. Following the addition without carry, a carry pulse is applied to the normally closed gates *G2*. Gates *G4* sense the digits standing in the *A* and *B* registers. In each column where a 1 stands in the addend register and a zero in the augend-sum register, gate *G4* opens gate *G2*. The applied carry pulse passes through *G2* to an electrical delay network, as well as to gate *G3*, through the phase inverter *I*. Gate *G3* is open if a 1 stands in the augend-sum register, and the carry pulse passes *G3* to the next higher-order column, etc. Thus, a carry pulse which is initiated in any column may pass several gates *G3* and be applied to several higher-order columns. Each time a carry pulse passes a gate *G3*, it is applied to the electrical delay network associated with the next column. The propagation of the carry pulses does not immediately cause a change of state in any of the flip-flops, but rather serves to introduce the carry pulses into the appropriate electrical delay networks in accordance with the configurations of the numbers in the *A* and *B* registers. After propagation of the carry pulse, all pulses applied to the delay networks emerge and change the states of the associated augend-sum flip-flops. The true sum then stands in the lower register.

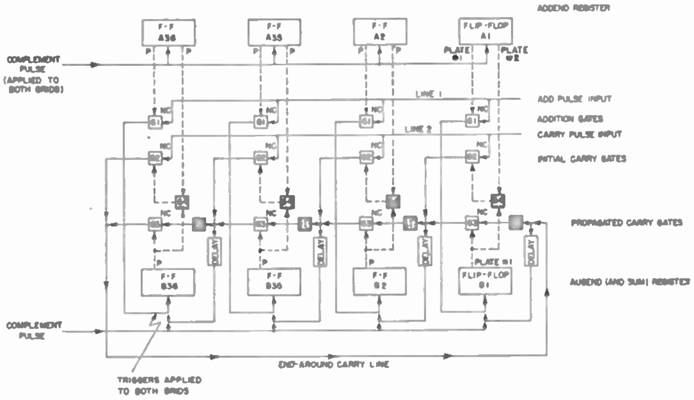


Fig. 11—Parallel adder employing simultaneous carry.

Numbers may be placed in the adder registers by the use of shift-register techniques, in which case additional gates and delay circuits are associated with the register flip-flops so that these may function as a shift register during the introduction of a number.

The arithmetic unit of the computer requires numerous control circuits in addition to the basic adder. These are necessary to perform the compound arithmetic operations of multiplication and division, as well as the logical operations described previously.

V. Central Control

The basic cycle of machine operation requires supervision of the following processes: selection of the operands from the memory, specification of the operation to be performed, disposition of the result, and the selection from memory of the next command. Many possible organizations of the central control unit exist, depending upon the time and manner in which the above steps are carried out. For example: (1) the selection of the two operands may occur sequentially or simultaneously; (2) the selection of operands may occur while previous operands are being processed in the arithmetic unit, or after the arithmetic process has been completed; (3) the disposition of a result may or may not occur simultaneously with some of the other steps; (4) part of or all of a command may be selected in one step; (5) a command may be selected in accordance with information contained in the previous command, or all commands may be located consecutively in the memory; (6) the control sequence may constitute a fixed cycle in which each step or combination of steps is allocated a fixed amount of time, or a variable cycle in which each step proceeds as soon as the previous step has been completed. There is no known optimum organizational pattern for the control process. Compromises must be made between speed of operation, simplicity of operation, and economy of equipment. By performing a large number of steps in parallel, the speed, cost, and complexity of the machine are all increased.

The control system of the present machine involves fixed-cycle, dual-selection operation. Each command is selected on the basis of information contained within the previous command.

A command consists of two words which are stored in adjacent positions in the internal memory. Referring to Fig. 3, it will be seen that each word of a command contains two address positions and a group of checking pulses. The second command word also contains an operation code. For most operations, address positions 1 and 2 contain the addresses of the two operands, address position 3 contains the address of the result, and address position 4 contains the address of the next command. Two exceptions to the above occur: (1) Certain operations (e.g., transfer) require but one operand. This appears in the first position, and the second position is then vacant. (2) In the branch operation the addresses of two commands appear in the third and fourth positions. The branch operation chooses which of the commands will be used in the next computing cycle.

The fixed-operation cycle is called the machine cycle, and is approximately 1 millisecond in duration. It is composed of three equal parts called major cycles which are equal in duration to the circulation time of a memory delay line, and which are synchronous with the memory circulation. During the first major cycle, the command governing the cycle is selected from the memory, and simultaneously the result of the previous computation is transferred to the memory. The selections made at this time are governed by the third and fourth addresses of the previous command. During the second major cycle, the two operands specified by the new command are selected from the memory, and the operation code is transmitted to the arithmetic unit. During the third major cycle, the arithmetic operation occurs.

From the above discussion it is apparent that the central control unit must contain registers for storing a complete command, and two selection circuits for the simultaneous selection of two memory locations. Since the operands selected from the memory may arrive at central control during any part of the selection cycle, the central control should also have one-word storage registers for holding the operands until they are to be transmitted to the arithmetic unit. A one-word register for the result is also required. The simplified organization of the central control is shown in Fig. 12.

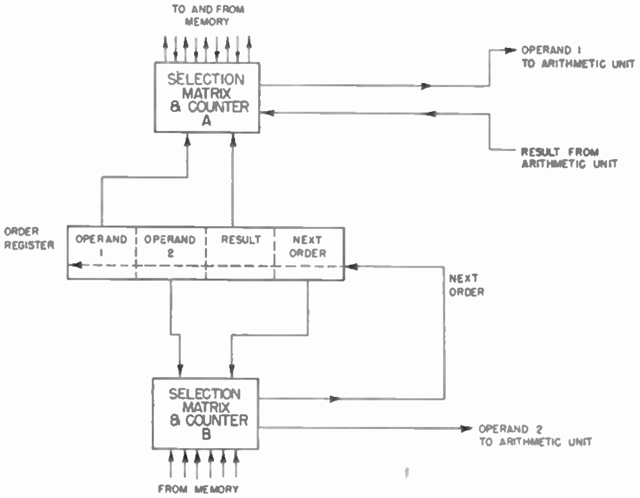


Fig. 12—Simplified organization of central control.

The operation of central control while making a given selection from the internal memory involves a spatial selection of one of 255 delay lines and a temporal selection of one of 16 word positions within the line. The address of any memory position is specified by a 12-digit binary number, the lower-order four digits of which refer to the temporal selection, while the upper-order eight digits refer to the spatial selection.

A selection of one of 2^n gating lines in accordance with an n-digit binary number governing the selection can be effected through the use of a diode matrix. Fig. 13 shows an elementary matrix in which a two-digit binary number contained in two flip-flops selects one of four gating tubes. The principle of operation depends upon the fact that the impedance across a parallel combination of diode gates varies only slightly with the

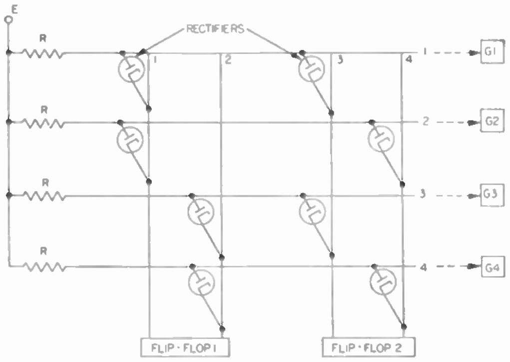


Fig. 13—Rectifier matrix.

number of gates closed if at least one gate is closed, but changes abruptly when all gates are opened. The rectifier matrix of Fig. 13 can be extended in both directions to accommodate as many binary digits as desired.

VI. MAGNETIC MEMORY

The magnetic memory units serve to augment the internal memory of the computer, to introduce initial data and commands into the machine, and to record the results of computation. These units make use of magnetic tape as a permanent continuous-storage medium. Each unit has a capacity of approximately 200,000 words, and is capable of operating at a maximum rate of 500 words per second.

Communication between the magnetic-tape units and the other parts of the machine is under the jurisdiction of the central control. Each unit has assigned to it an address which is of the same form as an internal memory address, except that the address in this case includes one of the three operation codes: read, write, and hunt. The read and write codes are always interpreted as applying to the next consecutive word position on the magnetic tape. The hunt code causes the tape unit to hunt for a particular word position, the number of which is supplied by central control.

Because the central control may call upon a magnetic memory unit at highly irregular rates, it is not feasible to propel the tape in response to each individual command. Instead, information is recorded on the tape in blocks of 16 words each and the tape mechanism always operates to read or write an entire block. A mercury-delay-line reservoir consisting of two 16-word delay lines is associated with each tape unit, as shown in Fig. 14. During a sequence of writing operations, the words to be written are accumulated in the consecutive word positions of this reservoir. When 16 words have accumulated in either delay line, the tape mechanism is actuated by local controls, and the contents of the reservoir are transferred to the tape. During a sequence of reading operations, the reservoir is filled from the tape, and as soon as one of the delay lines has been emptied by central control, the tape starts and refills the delay line.

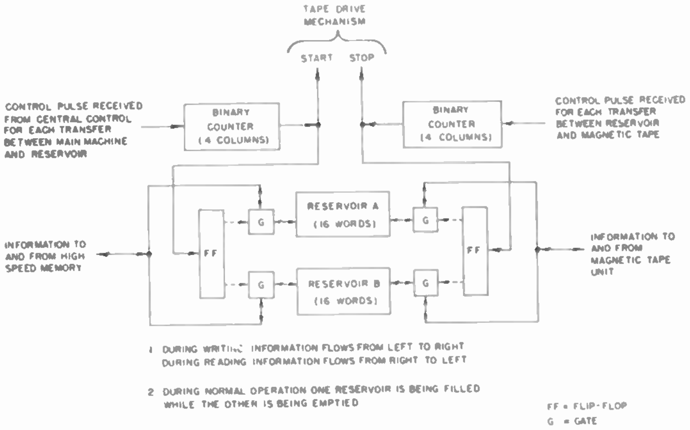


Fig. 14—Organization of magnetic-memory-unit reservoirs.

The recording medium is $\frac{1}{4}$-inch-wide magnetic tape on which five parallel channels are recorded. Four of these contain information, and the fifth serves as a marker or control channel. The marker channel contains markers which identify the beginning and end of each block of 16 words, as well as binary numbers which consecutively identify the blocks.

The maximum speed of operation of the magnetic memory units depends upon the pulse-repetition rate per channel and the time required for starting and stopping the tape. With operation rates of several hundred words per second, it may be necessary to accelerate the tape at several thousand inches/sec/sec. Fig. 15 shows the arrangement of a tape-drive mechanism which permits rapid acceleration. A magnetic clutch and drive capstan is used to propel a short segment of tape which is in contact with the recording heads. Spring slack absorbers are used to average the motion of the active portion of the tape and to control servomechanisms which drive the take-up reels.

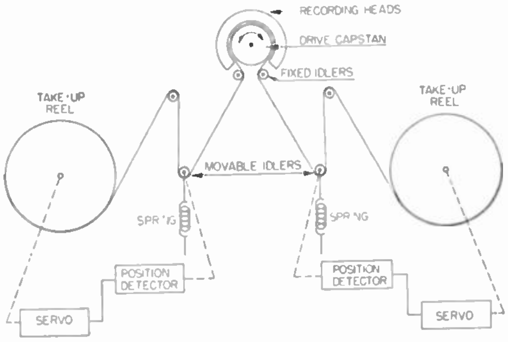


Fig. 15—Input-output drive mechanism.

VII. CHECKING

It is desirable that a computer be self-checking in order to minimize the number of undetected errors. The error-checking mechanism should be highly diagnostic, so that the location of a defective part is indicated whenever an error occurs. By incorporating diagnostic checking equipment in the design, the trouble shooting of equipment failures is greatly facilitated and the per cent of the total time during which the machine is operative is proportionately increased.

Numerous methods for checking digital computers have been suggested. Some of these include the simultaneous operation of duplicate equipments, the repetition of each operation or its inverse on the same equipment, and the use of operational checks which involve the periodic running of a test problem.

Checking by duplication of equipment may be undesirable for several reasons. The total amount of equipment is doubled, thereby doubling the cost as well as the total frequency of failures. Repeating each operation or its inverse reduces computation speed by one-half. It also seems doubtful that checking schemes based on the repetition of each operation can be made sufficiently diagnostic to indicate the location of a failure. The periodic running of a test problem has many of the disadvantages cited above concerning loss of speed and lack of diagnostic information. The existence of a time lag between an error and its detection makes it difficult to determine the exact point in the computing routine at which failure occurred, and consequently increases the difficulty of resuming operation. It is possible that intermittent errors might escape detection altogether. It may also be difficult or impossible to construct a test problem which completely checks the machine (e.g., tests all storage positions or all control circuits).

A general system of checking developed by Bloch,[1] known as the method of weighted counts, is believed to offer numerous advantages in regard to economy, reliability, speed, and diagnostic ability. A particular adaptation of this theory is used to check memory storage, word transfers, and arithmetic operations in the present machine.

The successive digits of a binary number may be said to be valued or weighted with the value 2^n where n is the number of the column. The conventional value of the number is then the sum of the weights of those columns in which 1's appear. A different number can be obtained by weighting the columns in a different manner. For example, successive columns may be given the weights 1, 2, 4, 1, 2, 4, etc. The new number is the sum of the weights off all columns in which the digit 1 occurs. This new number (or rather, the lowest-order four digits of it) is called the weighted count of the original number. Each number stored in the computer has its weighted count stored with it (note the group of checking pulses in Fig. 3). A number can be checked whenever desired by formulating a new weighted count and checking this for identity with the weighted count carried with the number. Such a check is performed each time a number is transferred from one location to another.

It can be shown that the basic arithmetic processes can be checked by means of the weighted count. Such checks are possible because the sum (difference, product) of the weighted counts of two numbers has a known relation to the weighted count of the sum (difference, product) of the numbers.

The weighted-count system of checking storage and transfers is operative throughout the machine. When initial input data and commands are manually recorded in the problem preparation unit, weighted counts are generated simultaneously with the depression of the recording keys. Thenceforth, each word contains its own weighted count. Each transfer of a word from the time of manual recording up to and including the time when results are printed by a page printer is checked by the weighted-count process.

Additional checking methods are used to verify the selection of memory positions, and the transfer of operation codes. A selection from the memory involves a temporal selection from among the 16 information word positions in each line and a spatial selection from among the 255 delay lines. Each delay line has a seventeenth word position which contains the binary number of the line. Whenever a line is selected, the number of the line is taken from this seventeenth word position, and is compared for identity with the number of the line as it occurs in the command governing the selection. An additional delay line is used to check temporal selections. This line operates in synchronism with the 255 information lines. The 16 information word positions of the extra line contain the binary numbers from zero to fifteen. Each time a word is taken from an information line, a word is simultaneously gated out of the extra delay line, and the number so obtained is compared for identity with the word position as it appears in the command governing the selection. In this way, each selection made from the memory is checked during the selection cycle in which it occurs.

The transmission of operation codes from the central control to the arithmetic unit is checked by incorporating a code generator in the latter unit. For each arithmetic operation the code generator returns an operation code to central control for checking purposes.

Although the checking system is elaborate, in the sense that each function of the computer is checked during each operating cycle, the equipment necessary to provide such checking does not exceed 20 per cent of the total equipment in the computer. The checking equipment has associated with it a set of controls and neon lights which operate to stop the machine in the event of an error and to indicate the location of the fault. The actual numbers and commands which were being processed when the error occurred are displayed to the machine operator. By virtue of these diagnostic aids, a defective part or subassembly can be quickly located and replaced, in many cases without loss of information or the disruption of the computing routine.

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Signal-to-Noise Ratio in AM Receivers*

EUGENE G. FUBINI†, SENIOR MEMBER, IRE, AND DONALD C. JOHNSON‡

Summary—**Experimental tests have been made to determine the effect of linear detectors upon the signal and the signal-to-noise ratio obtained from the demodulation of an rf carrier. The experimental data confirm theoretical results and indicate that:**

1. The concept of excess noise figure must be used with care. It cannot be employed to calculate the noise at the output of a receiver unless due consideration is given to the presence of a nonlinear device following the if stage. The reason for this is the effect of the presence of a carrier on the noise output of a detector. This noise increases when a carrier is present; the maximum increase is 4 to 7 db, depending upon the shape of the if filter.

2. For sine-wave amplitude modulation and if bandwidths at least 3 or 4 times larger than the af bandwidth, a universal curve can be given that shows the relation between the signal-to-noise ratio at the output and the carrier-to-noise ratio at the input of a second detector.

3. If two AM carriers are simultaneously present at the input of a linear second detector, this discriminates against the modulation of the weaker carrier.

INTRODUCTION

THE INFLUENCE of the second detector on the performance of a system involving the use of modulated radio frequencies is well known for the cases of frequency modulation and pulse modulation, while only recently has a substantial amount of work been done for the case of amplitude modulation. It has often been assumed, for instance, that an AM receiver will furnish at the output of its audio or video stage a signal-to-noise ratio that does not depend upon the bandwidth of the if stage; this assumption has led to the use of the AM receiver as a standard for the measurement of the "improvement" due to FM or other types of modulation. On the other hand, in some cases (radar receivers, for instance), the if bandwidth of AM receivers has been considered as one of the essential parameters that determine the performance of the set. It is also well known that FM systems discriminate against interfering signals in favor of the strongest signal present, but it is seldom mentioned and often forgotten that a similar phenomenon may well occur in any nonlinear system. The so-called "linear detector" used in AM receivers is by definition a detecting—and, therefore, a nonlinear—device; it is not surprising, therefore, to find that, as in FM, AM linear detectors do discriminate in favor of strong signals.

It may be useful to consider in detail the very common assumption that the output signal-to-noise ratio of AM communication receivers is independent of the if bandwidth. Consider a linear detector that follows an if stage and precedes an audio amplifier. At the input of the detector, there is a modulated carrier and some noise; at the output, there will be an af signal and some noise. If the carrier is strong, the spectrum of this noise is practically uniform because the beats between noise components are negligible when compared with the beats between the carrier and the noise, and the latter are uniformly distributed from zero cps to a frequency equal to half the if bandwidth.[1] Since af stages pass only part of this uniform spectrum, it is obvious that, when the carrier is strong, the noise at the output will depend only upon the bandwidth of the audio filter and upon the noise density; it will be independent of the if bandwidth.

In low-frequency applications, the if bandwidth is usually of the order of the bandwidth of the audio or video stage; for acceptable service, the signal at the output of the detector must be several db above the noise level. This condition usually requires that the carrier at the input of the detector be several db above the noise level. For this reason, the assumption that the signal-to-noise ratio at the output is independent of the if bandwidth holds in most of the low-frequency applications. At very high frequencies (around 100 Mc or more), however, the if bandwidth can well be 100 times wider than that of the audio or video stages, and the noise level at the input of the second detector will not necessarily be small compared with the signal, even when this is large enough for acceptable service.

These questions then arise: How strong does the signal need to be, as compared with the noise, in order to make the output signal-to-noise ratio independent of the if bandwidth, and what happens if the signal is not strong enough?

This paper will deal first with the effect of the presence of an unmodulated carrier on the noise at the output (audio or video) of an AM receiver. Modulated carriers are then considered, and the amplitude of the signal and its relation to the noise level are discussed.

The theory of detection of rf signals in the presence of noise for different spectral shapes, percentages of modulation, and detector characteristics has been rigorously developed by Middleton.[2]

EFFECT OF AN UNMODULATED CARRIER ON THE OUTPUT NOISE

Consider a detector with a noise voltage applied at its input and consider the noise as a voltage caused by the presence of many components comparable in ampli-

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* Decimal classification: R261.51×R361.211. Original manuscript received by the Institute, February 20, 1948; revised manuscript received, June 28, 1948. This work was done under Navy Contract NOa(s)-8575, under sponsorship of the Office of Naval Research.

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[1] These considerations are only qualitatively correct; they assume, for instance, a carrier centered with respect to a symmetrical and uniform (square) if filter.

[2] David Middleton, "Rectification of a sinusoidally modulated carrier in the presence of noise," PROC. I.R.E., this issue, pp 1467–1477.

tude and random in phase. Assume now that a very small if carrier is introduced at the input of the second detector and that its amplitude is comparable with that of one of the noise components. The detector will treat the carrier as if it were just another noise component. To the noise beats existing in the absence of the carrier, new beats will be added with an amplitude roughly proportional to the carrier amplitude c. Since all these beats are random with respect to each other, the total noise power n^2 in the presence of the carrier can be expressed as

$$n^2 = n_0^2 + k^2c^2,\ k^2c^2 \ll n_0^2 \tag{1}$$

where n_0^2 is noise power in the absence of the carrier, k^2 is a proportionality constant, and c^2 is the carrier power. Fig. 1 represents (1) graphically; n^2 (in db referred to n_0^2) is plotted against c^2 (in db referred to n_0^2/k^2).

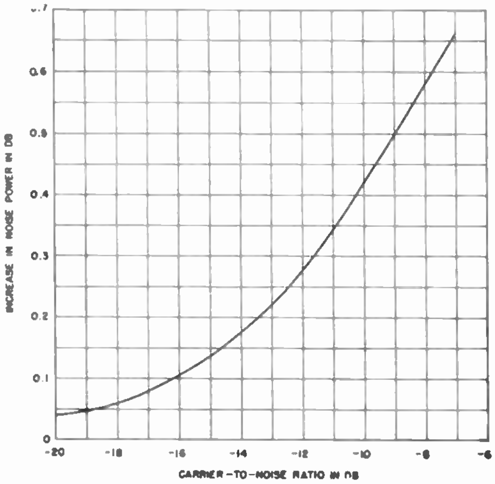

Fig. 1—Increase in noise power at the output of an audio- or video-frequency amplifier due to the presence of a carrier at the input of the second detector. This curve is based on the broad assumptions described in the text; it is not quantitatively accurate.

It is obvious that the preceding assumptions and the resultant Fig. 1 are only qualitatively accurate, but they are helpful in visualizing how the noise at the output of an AM receiver increases when a small if carrier is present at the input of the second detector.[3] It is easy to realize that the noise at the output will soon level off at a few db over its no-carrier value. It is well known, in effect, that the if voltage in a receiver is independent of the amplitude of the local-oscillator voltage if the latter is large enough. Similarly, the noise in the audio or video stage must be independent of the if carrier amplitude if the latter is large enough. It can be said roughly that when the if carrier voltage is large the carrier-noise beats "ride" on top of the carrier, and the noise voltage, after detection, is then obviously not a function of the carrier level.

The results of the experiments and of the rigorous theory are plotted in Fig. 2. Two curves give the theoretical values for an "optical"[4] spectral shape and for a uniform (square) spectral shape. The other four curves give the experimental data on the amount of noise power present at the output of an af amplifier for two if bandwidths (4.2 and 1.7 Mc) and two af bandwidths

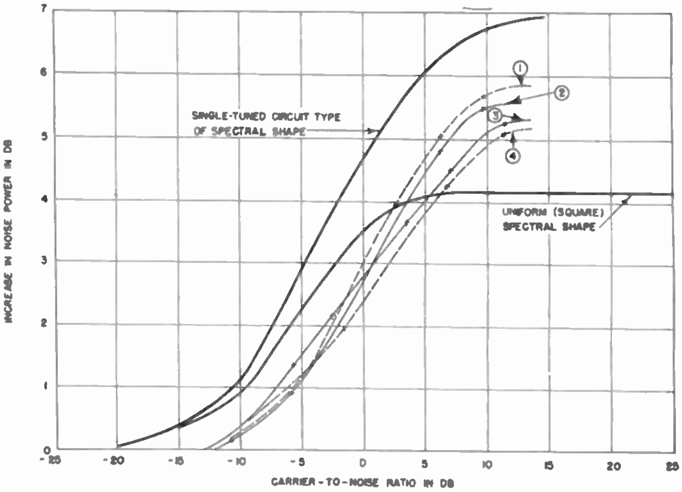

Fig. 2—Comparison between experimental and theoretical results on the increase in noise power at the output of an audio- or video-frequency amplifier due to the presence of a carrier at the input of the second detector preceding it. The two heavy lines are the theoretical curves for "optical" and uniform spectral shapes. The other four curves represent experimental results for the following if and af bandwidths:

Curve 1—4.2-Mc if, narrow af;
Curve 2—4.2-Mc if, wide af;
Curve 3—1.7-Mc if, narrow af;
Curve 4—1.7-Mc if, wide af.

The increase in noise power is referred to the noise present at the output when no carrier voltage is present.

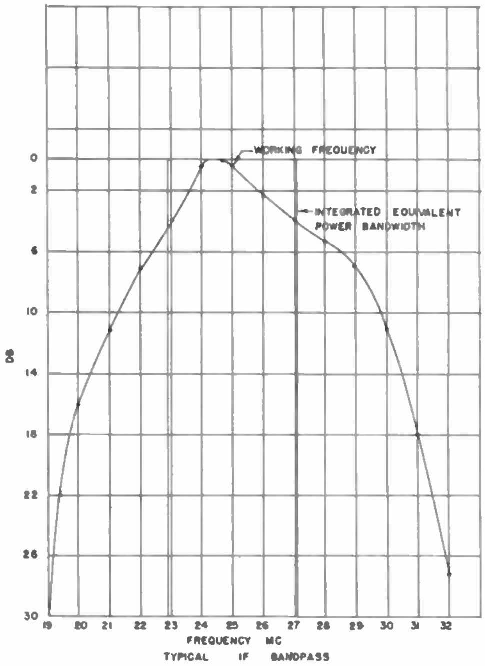

Fig. 3—Typical shape of if selectivity curve showing integrated equivalent power bandwidth.

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[3] This increase is completely unrelated to the effect of the first detector on the noise present in the if stage

[4] "Optical spectrum" is a term introduced by J. H. Van Vleck of Harvard University to represent a spectral shape equal to the selectivity curve of a single-tuned circuit.

(roughly, from 100 to 1100 cps and from 100 to 3700 cps). It appears clear from Fig. 2 that, according to theory, the spectral shape has a definite effect on the total increase in noise output between the no-carrier condition and the infinite-carrier condition (Fig. 3 shows a typical shape of the if stages used in the tests). It must be emphasized, however, that, although this increase depends upon the selectivity curve of the if stage, it does not depend, for a given spectral shape, upon the actual width of the if band. The reason behind this result is the normalizing effect obtained by referring the ordinates in Fig. 2 to the noise at the output when no carrier is present. The theoretical curve given for the case of uniform spectral shape is exactly the same as that given by Bennett.[5]

The application of this theory is limited to the case in which the if bandwidth is substantially larger than the af bandwidth. The theoretical results obtained by Middleton show that a ratio of 3 or 4 between half of the if bandwidth and the af bandwidth is sufficient to guarantee satisfactory accuracy.[6] The theoretical and experimental curves show a general trend common to both, but there is a discrepancy for the case of low-input carriers that cannot be easily explained.

The carrier-to-noise ratio at the input of the second detector was determined experimentally by a method suggested by North.[7] It is based on the principle that a dc voltage measured across a diode detector whose time constant is equal to the reciprocal of the if bandwidth will increase by a factor of 3 db if a nonmodulated if carrier is applied whose rms voltage is equal to the rms noise voltage. The results are independent of the spectral shape.

In calculating the useful range of some AM communication systems, it will be worth while to remember that the noise increases from 4 to 7 db when a carrier is present.

LINEAR DETECTION OF AN RF CARRIER IN THE PRESENCE OF NOISE

The rigorous theory shows that *any* detector will act as a square-law detector for carriers whose amplitudes are very small in comparison with the noise level. It is the purpose of this section to give experimental confirmation of this theoretical conclusion and to offer some nonrigorous explanations. Referring to (1) and considering the case in which the carrier is 100 per cent modulated, one finds that the noise varies with the modulating voltage. It can be assumed to a first approximation that the fundamental component m of the modulation is proportional to the difference between the noise voltages when the carrier is present and when the carrier is not present. We obtain, therefore, the approximate formula

$$\begin{aligned} m &= h\left(\sqrt{n_0^2 + k^2c^2} - n_0\right) \\ &\cong h\left[n_0\left(1 + \frac{k^2c^2}{2n_0^2}\right) - n_0\right] \\ &= \frac{hk^2c^2}{2n_0} \qquad (2) \end{aligned}$$

where m is the fundamental audio- or video-frequency component at the output, and h is a proportionality constant.

Equation (2) indicates that when the signal is small the detector follows a square law. Although this reasoning is only qualitatively accurate, the experiments and the rigorous theory check the conclusions. It is clear from Fig. 4 (which represents the result of a typical experiment) that the slope of the curve is 1 for large carrier inputs and about 2 for small carrier inputs, confirming the fact that the detector acts as a square-law detector for small carrier inputs.

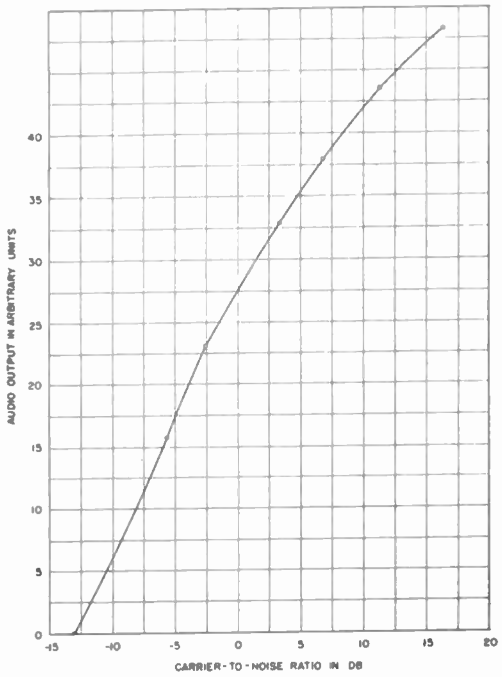

Fig. 4—Experimental determination of the effect of noise on the characteristic of a linear detector. The ordinates represent the af voltage obtained after a sine-wave-modulated rf voltage was detected and filtered. The curve shows that, for low carrier inputs, a linear detector becomes a square-law detector.

The above result (or, for that matter, any of the results presented in this paper) must not be attributed to the fact that any linear detector realizable in practice becomes nonlinear if the voltage applied to it is small enough. The experiments were conducted with a type 9005 diode detector, which is linear when the voltage applied is greater than 0.2 volt (see Fig. 9). The noise

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[5] W. R. Bennett, "Response of a linear rectifier to signal and noise," *Bell Sys. Tech. Jour.*, vol. 23, pp. 97–113; January, 1944.

[6] D. Middleton, "The response of biased, saturated linear and quadratic rectifiers to random noise," *Jour. App. Phys.*, vol. 17, p. 789, Fig. 11; October, 1946.

[7] D. O. North, unpublished synopsis of "The modification of noise by certain nonlinear devices," presented at the 1944 IRE Winter Technical Meeting, New York, N. Y., January 28, 1944.

at the output of the if amplifier in the absence of a carrier was greater than 0.7 volt, so that for all practical purposes the detector was operated well in the linear region of its characteristic, even with the smallest carrier input.

THE SIGNAL-TO-NOISE RATIO

Comparing Fig. 2 and Fig. 4, one sees that, for large signals, the noise is constant, and that the audio or video output increases linearly with the carrier input. This means that the signal-to-noise ratio in this region will also increase linearly with the input; in other words, an increase of, say, 3 db in the carrier input corresponds to a 3-db improvement in signal-to-noise ratio at the output. Again comparing Fig. 2 and Fig. 4, one finds that for very small carrier inputs the noise is practically constant, and the audio- or video-frequency signal is a quadratic function of the carrier input. If, therefore, one plots signal-to-noise ratio in db versus carrier input in db, one would expect to obtain a curve that, for large signals, would be asymptotic to a line with a slope of 1 and, for small signals, to a line with a slope of 2.

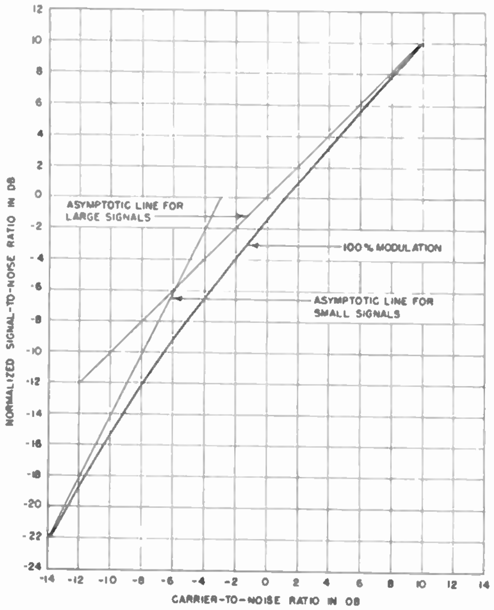

Fig. 5—Universal curve for determining the signal-to-noise ratio at the audio or video output of an AM receiver. The curve is valid for sine-wave modulation when the af bandwidth is much (at least 3 or 4 times) smaller than the if bandwidth. The signal-to-noise ratio is normalized to take into account the different if to af bandwidth ratios of different receivers; each ordinate represents the calculated value of the signal-to-noise ratio at the output in db minus 10 log (if bandwidth)/(2×af bandwidth). Modulation coefficients m different from 1 per cent can be taken into account with good approximation (errors less than 0.3 db in all practical cases) by decreasing the ordinates by 20 log $1/m$.

The rigorous theory confirms this result, and the curve resulting from it (for the "optical spectral" case) is shown in Fig. 5. It is important to note that Fig. 5 represents a universal curve that is valid for any if and af bandwidth so long as the if bandwidth is at least 3 or 4 times larger than the audio- or video-frequency bandwidth. The carrier input is plotted in db referred to the input noise power, and the signal-to-noise ratio is plotted in "normalized" db. Normalization is necessary in order to make the curve universal: the signal-to-noise ratio at the output is plotted in such a way as to take into account the different if and af bandwidths of various receivers; a number equal to

$$10 \log \frac{\text{if bandwidth}}{2 \times \text{af bandwidth}}$$

is subtracted from the calculated or measured value of db before it is plotted in the diagram. In practice, if one dealt, for example, with a receiver whose if bandwidth was 100 kc and whose af bandwidth was 3 kc, one would have to *add* to the value of signal-to-noise ratio read as an ordinate, a number equal to 10 log 100/6 = 12.2 db. In general, one should always *add* to the ordinate read in the diagram the number of db corresponding to the ratio given above.

Before discussing the experimental results, it may be useful to examine in detail the consequences of the theory. Consider two receivers that are exactly alike except for the fact that the excess noise figure is 3 db worse for the first receiver than for the second. This would mean that, for the same carrier input if, the value read as an abscissa would be 3 db higher for the better receiver than for the worse one; from Fig. 5, it appears that the signal-to-noise ratio would also be 3 db worse in the first receiver when the carrier input was large enough, but that it might be up to 6 db worse if the carrier input was very small.

If one considers two receivers that are otherwise identical but whose af stages have bandwidths in the ratio of 2 to 1, one will find that, irrespective of the values of the carrier input, the signal-to-noise ratios at the output will always differ by 3 db, provided, of course, that the if bandwidth is substantially larger than the af bandwidth.

It is more difficult to visualize the case of two receivers identical in every respect but having different if bandwidths. This difficulty occurs because changing the if bandwidth introduces changes both in the abscissa and the ordinates of the diagram. For instance, doubling the if bandwidth, while leaving the signal unchanged, means shifting the reading on the abscissa 3 db to the left and adding to the reading on the ordinate 3 db more than in the preceding case. If the carrier input is large enough, this does not result in any change in the output signal-to-noise ratio. However, if the signal is small enough, the signal-to-noise ratio at the output may decrease as much as 3 db. Consider, for instance, a receiver with an if bandwidth of 100 kc and an af bandwidth of 3 kc, and consider the case in which the carrier-to-noise ratio at the input of the second detector is −8 db; the normalized signal-to-noise ratio at the output, as read in Fig. 5, is about −12.2 db. The actual signal-to-noise ratio is, therefore, equal to −12.2+10 log 100/6 = 0 db.

Now suppose that the if bandwidth of the receiver is doubled and the carrier input remains the same. The carrier-to-noise ratio becomes −11 db (instead of −8 db) and the normalized signal-to-noise ratio, as read in Fig. 5, becomes about −16.8 db (instead of −12.2 db). The actual signal-to-noise ratio in this case will be $-16.8+10 \log 200/6=-1.6$ db.

In the region of small carrier inputs, therefore, the effect of doubling the if bandwidth is to impair the signal-to-noise ratio by amounts that increase from 0 to 3 db with decreasing carrier input.

Experimental checks of the theoretical results have been made at 30 Mc, using three different if bandwidths (12.6, 4.2, and 1.7 Mc) and two af bandwidths (1060 cps and 3600 cps). The results of the experiments are shown in Fig. 6. It is apparent that the signal-to-noise ratio measured for small carrier inputs is almost always a little worse than that predicted by the theory. This is only partially due to the assumption made in the theory that the spectral shape is "optical"; the differences are, at any rate, of the order of the experimental errors made in calculating the if and af bandwidths, in measuring the percentage of modulation, etc.

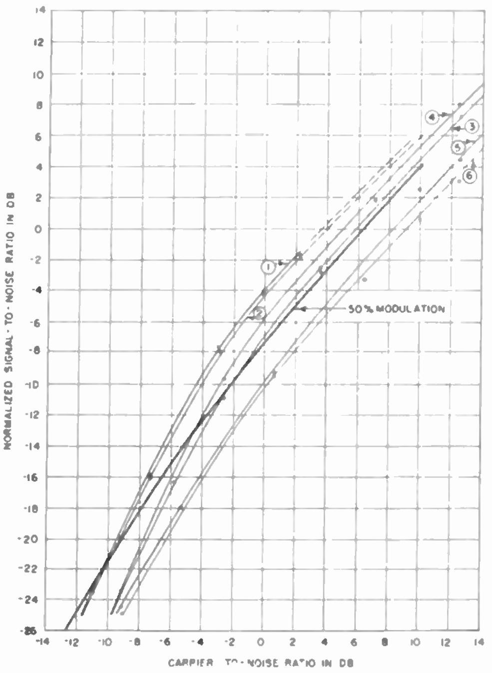

Fig. 6—Comparison of theoretical and experimental data. The 50 per cent modulation curve is obtained by means of a 6-db shift of the 100 per cent modulation curve in Fig. 5. The other six curves represent experimental results for the following if and af bandwidths:

Curve 1—12.6-Mc if; wide af
Curve 2—12.6-Mc if; narrow af
Curve 3—4.2-Mc if; narrow af
Curve 4—4.2-Mc if; wide af
Curve 5—1.7-Mc if; narrow af
Curve 6—1.7-Mc if; wide af

The modulation percentage used in the experiments was 50 per cent. The ratio between the wide and the narrow af bandwidths corresponds to a 5.3-db difference in the signal-to-noise ratio.

Discrimination Against Weak Signals

Fig. 4, which shows that a linear detector becomes a square-law detector when the signal is small enough, could be interpreted as stating that a linear detector discriminates against the signal in favor of the noise when the signal level is below the noise level. On the same basis, one would expect that, if two equally modulated rf carriers were present at the input of a linear detector, the ratio of the amplitudes of the two corresponding audio frequencies would not be equal to the ratio of the two carrier amplitudes; in other words, one would expect that, if two signals were present, the linear detector would discriminate in favor of the stronger signal.

This conclusion can be shown to be correct. Two signals were introduced at the input of an if amplifier, one with sine-wave modulation and the other unmodulated. When the unmodulated signal had an amplitude greater than that of the modulated signal, the af voltage at the output decreased—the same effect as that due to overload in an average amplifier. The voltages utilized, however, were such that both the detector and the amplifier were still working in the linear portion of their characteristics. The results of the experiment are plotted in Fig. 7.

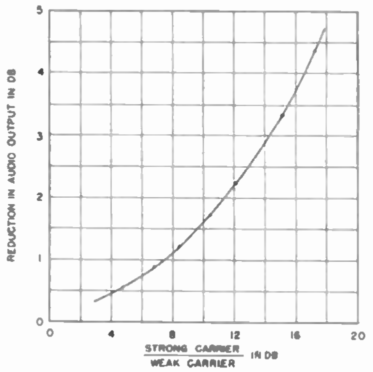

Fig. 7.—Experimental data showing how the audio output resulting from the detected modulation of one carrier is reduced by the simultaneous presence of a stronger carrier at the second detector. (That part of the curve for values of the abscissa greater than 12 db may have been somewhat influenced by a slight nonlinearity in the amplifiers.)

Conclusions

It has been shown that it is possible to predict to a reasonable approximation the signal-to-noise ratio at the output of an AM receiver as a function of the if and af bandwidths and the carrier level at the input. This paper has dealt exclusively with sine-wave modulation and with if bandwidths substantially larger than af bandwidths. Experiments confirm the validity of a universal curve calculated from Middleton's theory, which gives the desired relations for any ratio of if to af bandwidths greater than 3 or 4. The following conclusions can be drawn:

1. There is a form of threshold for AM receivers; below this threshold, the influence of the if bandwidth on the performance of the receiver becomes increasingly important. When the carrier rms voltage is equal to the noise rms voltage, the influence of the if bandwidth is reasonably small (a 2 to 1 change in the if bandwidth

will cause about a 1-db change in the output signal-to-noise ratio).

2. The excess noise figure measured after the second detector of a receiver is a function of the carrier amplitude; care should be used, therefore, in applying the concept of excess noise figure to the calculation of noise output of a receiver.

3. The noise level increases in the presence of a carrier by an amount that is a function of the shape of the if selectivity curve, but that is independent of the values of the if and the af bandwidths.

4. An AM receiver is not a linear device. For example, if two rf signals of different amplitudes but with the same percentage of modulation are present at the input of a second detector, the corresponding audio- or video-frequency voltages, after detection and filtering, are not in the ratio of the two rf carrier amplitudes (unless this ratio is 1 to 1). The detector always discriminates in favor of the modulation of the stronger carrier.

APPENDIX—EXPERIMENTAL PROCEDURE

The experiments were conducted at 30 Mc, using an if amplifier and a standard-signal generator. The detector was followed by an af preamplifier, a filter, and a power amplifier that was terminated with a thermocouple and a wave analyzer in parallel. (See Fig. 8.)

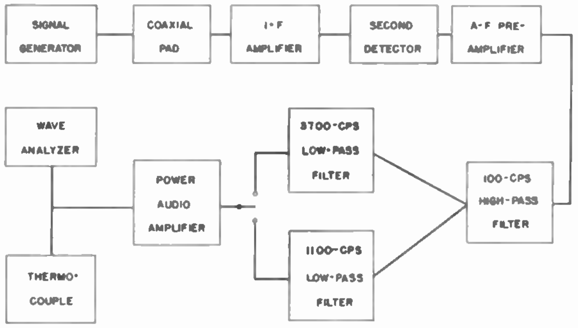

Fig. 8—Block diagram of the experimental setup used for the measurements.

A General Radio signal generator type 805A provided the if signal source. This signal generator gives a modulated or unmodulated rf output continuously variable from 0.1 microvolts to 2 volts and 400-cps modulation variable from 0 to 100 per cent. A Western Electric coaxial pad was used between the generator and the input of the if amplifier to give an attenuation of 20 db. The if strip, which used type 6AC7 tubes, was originally built for a type SPR-2 receiver. It was modified for single-tuned interstage coupling. The if strip was followed by a type 9005 diode detector which was, in turn, followed by a two-stage af preamplifier, feeding into a filter system consisting of a high-pass filter with a 100-cps cutoff and a low-pass filter with either a 1100-cps or a 3700-cps cutoff. The output of the filter system was fed into a power amplifier, and the output of the amplifier was fed to a General Radio wave analyzer type 736A connected in parallel with a Western Electric vacuum thermocouple type 20A.

The linearity of the diode detector was tested by measuring dc output versus rf input (Fig. 9); the diode was found to be linear down to an input as low as 0.2 volt. Since the dc reading due to noise at the output of the diode was never allowed to go below 0.7 volt, the diode for all practical purposes was operated on the linear portion of its characteristic. The bandwidth of the af filter system was calculated by graphical integration of the power versus frequency curve, and the ratio between the two af bandwidths corresponded to 5.3 db. The linearity of the if amplifier for large signals was checked by feeding a modulated rf signal into the input of the if amplifier, reducing the modulation, and increasing the amplitude to maintain a constant output. The exact value of each if bandwidth of the three amplifiers used was calculated by graphically integrating the area under the if selectivity power curve and by referring the equivalent width to the amplitude of the curve at the frequency used for the measurements (see Fig. 3).

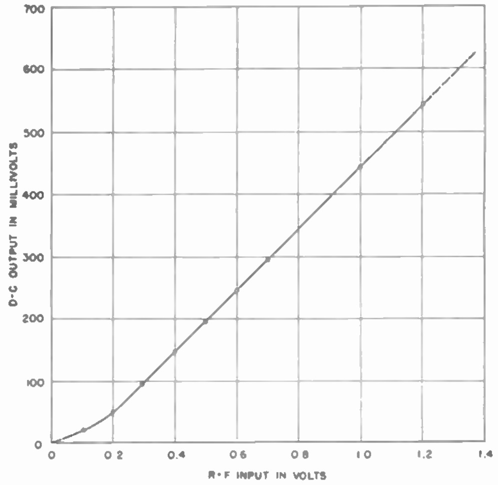

Fig. 9—Characteristic of the type 9005 diode, showing the linearity of the characteristic down to input voltage of the order of 0.2 volt.

The signal-to-noise ratios were measured as follows. For each value of rf input, the audio signal at 400 cps was read on the wave analyzer, and the signal-plus-noise power was simultaneously read on the thermocouple. The modulation percentage was 50 per cent. The modulation was then turned off, and the noise power alone was measured with the thermocouple. For each carrier input level, the measurements corresponding to the narrow and the wide af bandwidths were taken consecutively. Conversion factors between the thermocouple and the wave analyzer were determined, using the modulated-output reading on the thermocouple when noise was negligible. The signal output voltages read on the wave analyzer were then converted to thermocouple units of power, and the signal-to-noise ratio was computed.

Rectification of a Sinusoidally Modulated Carrier in the Presence of Noise*

DAVID MIDDLETON†, MEMBER, IRE

Summary—**The low-frequency output, signal and noise, is determined following a general νth-law ($\nu>0$) half-wave rectification of a sinusoidally modulated carrier. The spectral width of the noise before rectification is assumed to be much narrower than its mean frequency f_0, the normal condition for reception. Attention is focused on one of the harmonics of the rectified signal, generally the first, whose modulation frequency is small compared to the if frequency f_0 and to the width of the noise band. Special attention is given to the important cases of linear ($\nu=1$) and quadratic ($\nu=2$) detection, and audio signal-to-noise ratios s_a/n_a are calculated for them. The noise passed by the audio filter is found to depend on the spectral shape of the if; for the three types of if filter specifically considered, namely, rectangular, gaussian, and "optical," the optical yields the least, the rectangular the most noise; provided, of course, that the mean input signal-to-noise power p is of the order of unity or less, the customary region of interest. When $p \ll 1$, all half-wave detectors ($\nu>0$) behave like the quadratic device. Modulation ($0 \leqq \lambda \leqq 1$) does not radically affect the output noise power, particularly when ($p>1$), and for very large signals, $p\rightarrow\infty$, s_a/n_a is directly proportional to the input carrier amplitude, is independent of if filter width, and is only slightly dependent on filter shape. The best if filter is one whose response is the modulus of the Fourier transform of the signal, as in the radar case of pulsed signals. The theory is in good agreement with the experimental results for the linear rectifier, and a number of theoretical curves illustrating the variation of the audio ratio s_a/n_a with p are also included.**

I. INTRODUCTION

IN RECEIVERS the observability of a signal is limited by the accompanying noise, originating either in the receiver elements themselves or coming in with the carrier from external sources. Now, in all cases the complicating factor is the nonlinear device, the second detector, by which demodulation of the carrier is accomplished.[1] Signal and noise enter this element simply as the sum of their instantaneous amplitudes, but leave it in quite a different manner. The nonlinearity of the detector produces beats between components of the noise ($n\times n$), the signal and noise ($s\times n$), and the signal with itself ($s\times s$). What then appears is no longer the "pure" noise and "pure" signal that entered originally, but noise modified by the presence of the signal, and more significantly, especially for small signals, signals that are altered by the noise. Under these circumstances it has been found that a useful criterion by which to estimate the performance is the signal-to-noise ratio, obtained at the output in a manner appropriate to the type of receiver in question. This ratio, in turn, may be related to the incoming signal and noise in a fashion depending on the filter response, detector characteristic, and mode of presentation of the apparatus.

Our present problem concerns the rectification of a sinusoidally modulated carrier, accompanied by random noise. The bandwidth ($\sim\Delta f_{\text{noise}}$) of the if stage is taken to be much larger than the modulation frequency f_A and much less than the mean frequency of the band f_0, so that the low-frequency output of the detector may be easily separated from the higher spectral zones generated by the cross-modulation of the components of the input (see Fig. 1(a)). Of this low-frequency output, only the spectral region in the neighborhood of $f=0$ is of interest (Fig. 1(b)), since it contains the harmonic (first or higher) of the modulation, which constitutes the detected signal. A filter Δf cps wide is employed which allows only this harmonic (lf_A, $l=1, 2, 3 \cdots$) and the rectified noise in the region $lf_A-\Delta f/2$ to $lf_A+\Delta f/2$ cps to pass. Our output signal-to-noise ratio may then be defined as: the rms value of the lth component of the modulation divided by the rms value of the noise transmitted by the selective audio filter. (It should be noted that the audio filter Δf may be either the natural response of the ear[2] or the response of an electrical filter, whichever is the narrower at this frequency.) This criterion is precisely the same as that for the aural detection of radar signals.[2]

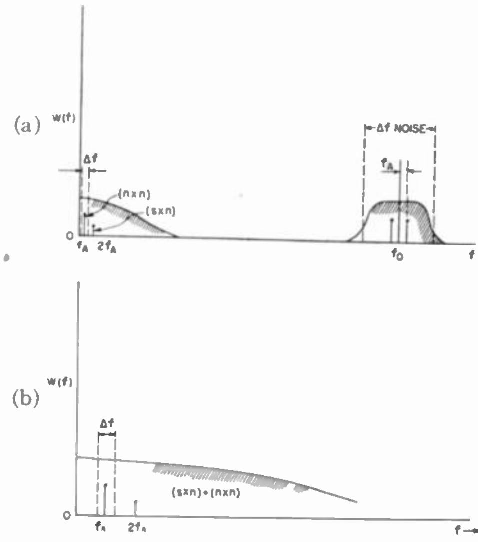

Fig. 1—Spectral distribution of the input and output power after (a) half-or full-wave quadratic detection; (b) low-frequency output.

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* Decimal classification: R134×R161.6. Original manuscript received by the Institute, February 20, 1948; revised manuscript received, June 1, 1948.

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[1] Although the first detector or mixer is also a nonlinear element, the presence of a beating-oscillator voltage, large compared with either signal or noise, makes its operation effectively linear in that there is no mixing of signal and noise.

[2] J. H. Van Vleck and D. Middleton, "A theoretical comparison of the visual, aural, and meter reception of pulsed signals in the presence of noise," *Jour. Appl. Phys.*, vol. 17, pp. 940–971; November, 1946. See part I, sec. 1.

Ragazzini[3] has obtained an approximate expression for the output power spectrum when a carrier modulated by a sine wave is impressed on a linear rectifier. His result for the low-frequency part of the output continuum is, as Rice[4] has shown, quite accurate for small indices of modulation, and particularly when the amount of noise relative to the signal is small. Work on related lines by Bennett[5] also bears out this observation. The scope of the present paper is considerably wider than previous efforts, covering all degrees of modulation (0–100 per cent) and all ($\nu > 0$) types of unsaturated half-wave rectifiers. The more precise technique of the present treatment, following the method of Rice,[6] is obtained from generalizations developed by the author in a recent paper.[7] Our purpose, then, is to determine the noise, the audio signal, and signal-to-noise ratios after rectification by a half-wave νth-law detector, and their dependence on input signal, degree of modulation, if spectral shape, etc. Detailed calculations are given for the two important cases of linear and quadratic detection, and three specific if filter responses are considered: rectangular, gaussian, and "optical."

To illustrate the approach, we examine first the simple and important problem of small-signal, full-wave quadratic rectification.

II. SMALL-SIGNAL, FULL-WAVE QUADRATIC RECTIFICATION

Following the notation of footnote reference 2, we write, for the amplitude factor of the if carrier entering the second detector,

$$\sqrt{2}\, s_F(t) = A_0(t) = s_F\sqrt{2}\,(1 + \lambda \cos \omega_A t),$$
$$\omega_A = 2\pi f_A, \qquad 0 \leqq |\lambda| \leqq 1 \tag{1}$$

where λ is the modulation index and ω_A is the (angular) frequency of modulation. The actual wave leaving the if is, exclusive of the noise,

$$A_0(t) \cos \omega_0 t = s_F\sqrt{2}\,(1 + \lambda \cos \omega_A t) \cos \omega_0 t$$
$$(\omega_0 = 2\pi f_0), \tag{2}$$

so that s_F is the rms amplitude of the signal for no modulation ($\lambda = 0$); ω_0 is the angular frequency of the carrier, which is also the same as the central frequency of the noise band, since in all our work the if filter is assumed so tuned that the signal is in the center of the band. Then, the mean total signal power is clearly

$$P_s = s_F^2(1 + \lambda^2/2), \tag{3}$$

of which $\lambda^2 s_F^2/2$ is attributable to the modulation.

Now, from equation (16) of footnote reference 2 or from equation (8.20) of footnote reference 7, the low-frequency correlation[8] function after full-wave quadratic rectification is, exclusive of the dc component due to the noise terms $(n \times n)$ and $(s \times n)$,

$$R_0(t) = \beta^2\{\psi_0(t)^2 + 2\psi_0(t)\overline{s_F(t_0')s_F(t_0' + t)}$$
$$+ \overline{s_F(t_0')^2 s_F(t_0' + t)^2}\} \tag{4}$$

where the bars indicate the average over the phases (proportional to, t_0') of the modulation, and $\psi_0(t)$ is the correlation function of the if noise band, central frequency removed. (The actual correlation function is $\psi(t) = \psi_0(t) \cos \omega_0 t$. See (11)–(13) of footnote reference 2, and consult Table I for specific examples.) We note that

$$\psi_0(t) = \psi r_0(t) \equiv n_F^2 r_0(t); \qquad n_F^2 \equiv \psi = \psi(0) \tag{5}$$

where $r_0(t)$ is the (normalized) correlation of the input noise referred to the frequency f_0. The quantity β is a scale factor that has the proper dimensions so that, if the incoming wave is a voltage on the grid of the detector, it appears as a current in the output. (It may also appear as any other suitable quantity, depending on the circuit, with the proviso that the circuits in question are primarily resistive, i.e., have only small reactive components over the frequency range of operation. This is to insure the one-valuedness of the dynamic output versus input path.)

The first term in (4) represents noise alone $(n \times n)$, the second term gives the cross-modulated noise $(s \times n)$ produced when the signal and noise components beat together, while the final term is the contribution due to the modulation itself. The average of the third term in (4) is readily shown to be

$$\overline{s_F(t_0')^2 s_F(t_0' + t)^2}$$
$$= s_F^4\left\{\left(1 + \frac{\lambda^2}{2}\right)^2 + 2\lambda^2 \cos \omega_A t + \frac{\lambda^4}{8} \cos 2\omega_A t\right\}. \tag{6}$$

The mean power in the harmonic ($f = f_A$) we wish to hear is, then,

$$P_A = 2\beta^2\lambda^2 s_F^4 \equiv s_a^2. \tag{7}$$

Since the noise is assumed to be spectrally wide compared to the modulation frequency, i.e., $\Delta f_{\text{noise}} >> f_A$ (where Δf_{noise} is the width between half-power points of the if noise power spectrum (see Fig. 1)) we may take as the spectrum of the noise output in the neighborhood of f_A the value of the spectrum at $f = 0$. The well-known transform relations between the mean power spectrum and its correlation function[2,4,6,7] permit us to write for the mean noise power transmitted by the audio (or aural) filter of width[9] Δf

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[3] J. R. Ragazzini, "The effect of fluctuation voltages on the linear detector," PROC. I.R.E., vol. 30, pp. 277–288; June, 1942.

[4] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 24, pp. 46–108; January, 1945. See discussion following equation (4.10–11).

[5] W. R. Bennett, "Response of a linear rectifier to signal and noise," *Jour. Acous. Soc. Amer.*," vol. 15, pp. 164–172; January, 1944.

[6] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 23, pp. 282–332; July, 1944; and Part IV of reference 3.

[7] D. Middleton, "Some general results in the theory of noise through non linear devices," *Quart. Appl. Math.*, vol. 5, p. 445; January, 1948.

[8] The correlation function is the Fourier transform of the mean power spectrum. See equation (11).

[9] The audio filter is considered rectangular, of width Δf, which is not a very critical assumption.

$$n_a^2 = [W_0(f)\Delta f]_{f=0} = 4\Delta f\left(\int_0^\infty R_0(t)\cos\omega t dt\right)_{\omega=0}$$

$$= 4\beta^2\Delta f\left\{\int_0^\infty \psi_0(t)^2\cos\omega t dt + 2s_F^2\int_0^\infty \psi_0(t)\overline{(1+\lambda\cos\omega_A t_0')(1+\lambda\cos\omega_A[t_0'+t])}\cos\omega t dt\right\}_{\omega=0}. \tag{8}$$

Because $\Delta f_{\text{noise}} >> f_A$, it is easily shown for any particular case that $\psi_0(t)$ falls off to zero so rapidly for $t>0$ that we may set $\cos\omega_A t = 1$ as well as $\cos\omega t = 1$, i.e., for $\omega = 0$. Then the mean audio noise power (8) becomes

$$n_a^2 = 4\beta^2\Delta f n_F^4\left\{\int_0^\infty r_0(t)^2 dt + (2+\lambda^2)p\int_0^\infty r_0(t)dt\right\} \tag{9}$$

where $p \equiv s_F^2/n_F^2$; see (5).

The rms signal-to-noise ratio follows from (7) and (9), and is

$$\frac{s_a}{n_a} = \frac{\lambda p}{(2\Delta f)^{1/2}}\left\{\int_0^\infty r_0(t)^2 dt + (2+\lambda^2)p\int_0^\infty r_0(t)dt\right\}^{-1/2}. \tag{10}$$

As will be shown later (Part VI), this result also holds exactly for the half-wave quadratic detector ($\nu = 2$).

Before examining any special input noise spectra, we observe at once that the values of the integrals in (10) depend on the spectral shape of the if, and this dependence may be expected to differ for $(n\times n)$ and $(s\times n)$ terms by a not entirely negligible numerical factor. Physically, the argument goes something like this: The spectral contribution near $f=0$ arises from beats between closely adjacent components; hence, the power in beats of this kind produced at high frequencies, i.e., out on the "tails" of the if spectrum, will depend on the power density in their neighborhood, and hence the spectral shape of the if output noise will determine in varying degrees the spectral ordinate near $f=0$ after rectification, as it may be considered as the sum (properly weighted in accordance with the law of the detector; see Section V) of these beats produced in the process of detection. Thus, for a rectangular if spectrum we will get one spectral density near $f=0$; for the gaussian, another.

Table I below gives the salient properties of three common spectral distributions; the correlation functions are calculated from the relations[2,4,6,7]

$$R(t) = \int_0^\infty W(f)\cos\omega t df,$$

$$\left(\text{with } W(f) = 4\int_0^\infty R(t)\cos\omega t dt,\quad \omega = 2\pi f\right) \tag{11}$$

where use is made of the fact that the central noise frequency $f_0 >> \Delta f_{\text{noise}}$, so that the if correlation function may be written

$$\psi(t) = n_F^2 r_0(t)\cos\omega_0 t. \tag{12}$$

The various spectral parameters ω_F as determined by the normalization are related by

$$(\omega_F)_{\text{opt.}} = \frac{1}{\sqrt{\pi}}(\omega_F)_{\text{gauss}} = \frac{2}{\pi}(\omega_F)_{\text{rect.}} \tag{13}$$

The audio signal-to-noise ratio (10) then becomes explicitly

$$\frac{s_a}{n_a} = \frac{\lambda p}{\sqrt{2}}\left[\frac{(\omega_F)_{\text{opt.}}}{\Delta f}\right]^{1/2}\left\{\underset{(n\times n)}{\begin{pmatrix}1\\ 1/\sqrt{2}\\ 1/2\end{pmatrix}} + \underset{(s\times n)}{\begin{pmatrix}1\\ 1\\ 1\end{pmatrix}}p(2+\lambda^2)\right\}^{-1/2} \tag{14}$$

TABLE I

Type of if spectrum	Analytic form	Low-frequency correlation $r_0(t)$	$n_F^{2*} \equiv \psi$	Half-width $[\Delta f_{\text{noise}}/^2]$ vs f_F	$(n\times n)$ $\int_0^\infty r_0(t)^2 dt$	$(s\times n)$ $\int_0^\infty r_0(t)dt$
Rectangular	$W(f) = 0, f > f_0 + f_F$ $f < f_0 - f_F$ $= W_0, f_0 - f_F < f$ $< f_0 + f_F$	$\frac{\sin\omega_F t}{\omega_F t}$	$\frac{W_0\omega_F}{\pi}$	$f_F = \frac{\Delta f_n}{2}$	$\frac{\pi}{2\omega_F}$	$\frac{\pi}{2\omega_F}$
Gaussian	$W(f) = W_0 e^{-(\omega-\omega_0)/\omega_F^2}$	$e^{-\omega_F^2 t^2/4}$	$\frac{W_0\omega_F}{2\sqrt{\pi}}$	$f_F = 1.15\times\frac{\Delta f_n}{2}$	$\frac{1}{\omega_F}\sqrt{\frac{\pi}{2}}$	$\frac{\sqrt{\pi}}{\omega_F}$
"Optical"**	$W(f) = \frac{W_0}{1+\left(\frac{\omega-\omega_0}{\omega_F}\right)^2}$	$e^{-\lvert\omega_F t\rvert}$	$\frac{W_0\omega_F}{2}$	$f_F = \frac{\Delta f_n}{2}$	$\frac{1}{2\omega_F}$	$\frac{1}{\omega_F}$

* These spectra are normalized so that $n_F^2 = \psi$ is the same in all cases and W_0, the maximum spectral intensity, is likewise kept unchanged; ω_F is accordingly varied.

** This is equivalent to the selectivity curve of a single-tuned circuit.

where the parantheses ( ) apply downward in the order given by Table I; (14) is illustrated in Figs. (9–11) for the three types of if spectrum, respectively, by the curves for quadratic rectification. Note that, of the three, the optical spectrum, corresponding to a single-tuned circuit, gives the largest signal-to-noise ratio.

III. GENERAL HALF-WAVE THEORY: SIGNAL

We proceed now with the more difficult problem of the general half-wave detector. First let us consider the output signal; this may be obtained from the appropriate part of the correlation function after rectification, which contains only frequencies of the modulation, their harmonics, and the dc contribution from $(n\times n)$ and $(s\times n)$ products. From equations (3.3) of footnote reference 7 we find the expression to be

$$R_0(t)_{(s\times s)} = \frac{\beta^2\Gamma(\nu+1)^2}{4\Gamma(\nu/2)^2(\nu/2)^2}\left(\frac{\psi}{2}\right)^{\nu}\left\{{}_1F_1(-\nu/2;1;-p_1)\,{}_1F_1(-\nu/2;1;-p_2)\right\}_{av.}, \tag{15}$$

where p_1 is the instantaneous input signal-to-noise power ratio $p(t)$ at time t_0', and p_2 is this ratio at time t later. Specifically, $p(t)$ is here

$$\begin{aligned} p(t) &= (1+\lambda\cos\omega_A t)^2 s_P^2/n_P^2 \\ &= p(1+\lambda\cos\omega_A t)^2, \quad 0 \leqq \lambda \leqq 1. \end{aligned} \tag{16}$$

The quantity ${}_1F_1$ is the confluent hypergeometric function, some of whose more important properties are mentioned in Appendix III of footnote reference 7. The average, as in (4) and (8), is to be performed over all phases of the modulation. Observe that when $\nu=2$, (15) reduces to (4), the numerical factor $\frac{1}{4}$ appearing in the former because only half the wave is transmitted, and that subject to square-law rectification.

The direct determination of averages in (15) is very laborious in most instances. However, we can get around this difficulty by observing that the form of the dc output $s_v(t)$ with carrier modulation is the same as that of the dc without modulation, s_v, except that the former fluctuates, adiabatically compared with the if and higher frequencies. Analytically, this is equivalent to setting $p_1=p_2=p(t)$ in (15) and taking the square root; the result is

$$s_\nu(t) = \frac{\beta\Gamma(\nu+1)}{2\Gamma(\nu/2+1)}\left(\frac{\psi}{2}\right)^{\nu/2}\left\{{}_1F_1(-\nu/2;1;-p(t)) - 1\right\}, \tag{17}$$

exclusive of dc due to $(n\times n)$ products. The mean dc response (due to the signal) may be safely discarded since it is an unimportant part of the low-frequency output not passed by the video or audio filters. Expansion of $s_v(t)$ in a Fourier series yields

$$\begin{aligned} s_\nu(t) &= \sum_{l=-\infty}^{\infty} a_l e^{i\omega_A l t}, \quad \text{with} \\ a_l &= \frac{1}{T_0'}\int_{-T_0^{1/2}}^{T_0^{1/2}} s_\nu(t)e^{-i\omega_A l t}dt \end{aligned} \tag{18}$$

where T_0' $(=2\pi/\omega_A)$ is the period of the modulation. The rms amplitude of the lth component, which is under observation, is

$$s_a = \sqrt{2}\,|a_l| = \frac{\sqrt{2}}{T_0'}\left|\int_{T_0^{1/2}}^{T_0^{1/2}} s_\nu(t)e^{-il\omega_A t}dt\right|, \quad l \geqq 1. \tag{19}$$

Since we observe only one component, phase considerations are not important (not so if we look at a radar screen, for example, where phase relations between components of the modulation are essential for our information, which is obtained from the *envelope* of the disturbance). From (17) and (19), the rms audio signal amplitude may be written

$$s_a = \frac{\beta\Gamma(\nu+1)}{\sqrt{2}\,\Gamma(\nu/2+1)}\left(\frac{\psi}{2}\right)^{\nu/2}\left|\frac{1}{2\pi}\int_{-\pi}^{\pi}[{}_1F_1(-\nu/2;1;-p(\theta)) - 1]e^{-il\theta}d\theta\right|, l \geqq 1. \tag{20}$$

Instead of expanding the hypergeometric function, let us approximate in an ad hoc fashion, according to the procedure of footnote reference 2, part II, II(d), which consists in replacing the infinite series by a polynomial of two terms.[10] After this has been done and (16) has been used for $p(t)$, we obtain finally for the first harmonic $(l=1)$, which is of primary interest,

$$(\nu=1):\quad s_a = \frac{\beta\psi^{1/2}p\lambda}{\sqrt{4\pi}}(1 - 0.2289p^{1/2}[1+\lambda^2/4]), \qquad p(1+\lambda)^2 \leqq 10, \tag{21a}$$

$$(\nu=2):\quad s_a = \frac{\beta\psi p\lambda}{\sqrt{2}} \qquad p \geqq 0. \tag{21b}$$

Figs. 2 and 3 illustrate the variation of s_a when $p \leqq 2.5$. and λ takes on all permissible values, with $\nu=1$ or $\nu=2$. As we expect, for small signal-to-noise ratios the linear and quadratic responses are nearly the same, due to modulation suppression in the former case. There, the noise reduces the signal strength. However, as p increases, the noise in turn is progressively suppressed by the signal, and the audio signal amplitude becomes more nearly proportional to $s_P\lambda$, as compared to $s_P^2\lambda$

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[10] Details are available in Technical Report No. 45, Cruft Laboratory, Harvard University, Cambridge, Mass.

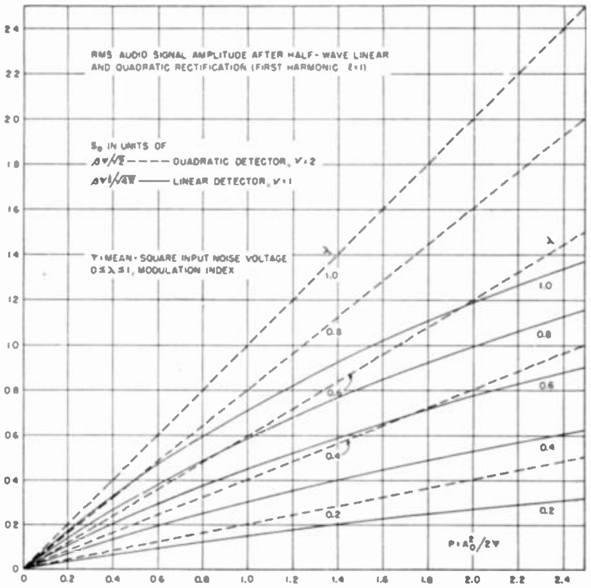

Fig. 2—Rms audio signal amplitude after half-wave linear or quadratic rectification as a function of $p=s_F^2/n_F^2$.

for the square-law device. These effects are naturally enhanced for the higher percentages of modulation, for then the signal is, on the average, greater by a factor $(1+\lambda^2/2)^{1/2}$.

When $p\rightarrow\infty$, our expression (20) for the rms audio signal is easily modified with the aid of the asymptotic form of ${}_1F_1$ (see A (3.3) of footnote reference 7) to give finally the important result for the first harmonic, namely,

$$s_a\Big|_{p\rightarrow\infty} \cong \frac{\beta\Gamma(\nu+1)\lambda\nu s_F^{\nu}}{2^{(\nu+3)/2}\Gamma(\nu/2+1)^2}\,{}_2F_1\left(\frac{1-\nu}{2},\ \frac{2-\nu}{2};\ 2;\ \lambda^2\right), \qquad (l=1), \tag{22}$$

and in particular, for the linear and quadratic detectors, (22) becomes

$$(\nu=1):\quad s_a \cong \frac{\beta\lambda s_F}{\pi}; \qquad (\nu=2):\quad s_a \cong \frac{\beta\lambda s_F^2}{\sqrt{2}}\cdot \tag{23}$$

As we would expect, the noise is no longer effective in modifying the signal amplitude.

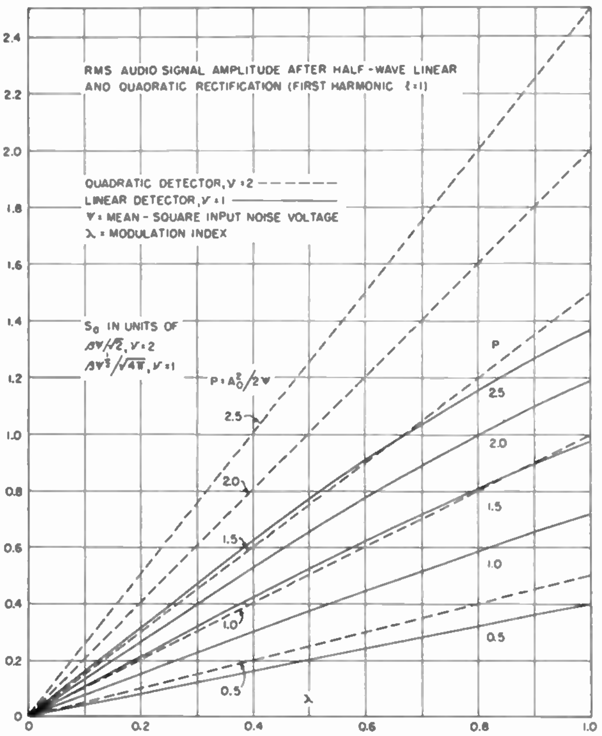

Fig. 3—Same as Fig. 2, as a function of the modulation index.

IV. General Half-Wave Theory: Noise

In order to obtain signal-to-noise ratios, we must determine the mean-square or the rms noise current or voltage passed by the detector and its associated filters. This is a more difficult problem for noise than for signals, since the mixing of the components of the modulation with those of the noise, as well as cross-modulation of the noise with itself, have to be calculated. The results themselves are more complex and less susceptible to facile approximation than those for the signal, although simplifying approximations may still be made, but these require a narrower range of values for such variables as the input signal-to-noise ratio p, and modulation index λ.

The low-frequency correlation function $R_0(t)$ for the noise output forms the starting point of this part of our analysis. Footnote reference 7 then enables us to write[11] the $(n\times n)$ part of the correlation function as

$$R_0(t)_{(n\times n)} = C_\nu^2\left\{r_0(t)^2\left[{}_1F_1(1-\nu/2;\,1;\,-p_1)\,{}_1F_1(1-\nu/2;\,1;\,-p_2)\right]_{\text{av.}}\right.$$
$$\left.+\ \Gamma(\nu/2)^2\sum_{q=2}^{\infty}\frac{\left[{}_1F_1(q-\nu/2;\,1;\,-p_1)\,{}_1F_1(q-\nu/2;\,1;\,-p_2)\right]_{\text{av.}}r_0(t)^{2q}}{(q!)^2\Gamma(1-q+\nu/2)^2}\right\}, \tag{24}$$

and for the contribution arising from the beats between signal and noise

$$R_0(t)_{(s\times n)} = C_\nu^2\left\{2r_0(t)\left[(p_1p_2)^{1/2}\,{}_1F_1(1-\nu/2;\,2;\,-p_1)\,{}_1F_1(1-\nu/2;\,2;\,-p_2)\right]_{\text{av.}}\right.$$
$$\left.+\ \Gamma(\nu/2)^2\sum_{m=1}^{\infty}\sum_{q=1}^{\infty}\frac{2r_0(t)^{m+2q}\left[(p_1p_2)^{m/2}\,{}_1F_1(m+q-\nu/2;\,m+1;\,-p_1)\,{}_1F_1(m+q-\nu/2;\,m+1;\,-p_2)\right]_{\text{av.}}}{q!(q+m)!(m!)^2\Gamma(1-m-q+\nu/2)^2}\right\} \tag{25}$$

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[11] The low-frequency correlation function follows at once from (3.3), footnote reference 7, when l is put equal to zero. Equation (4.4) ibid., with $b_0=0$ for the half-wave condition, gives us $H_{m,m+2q}$ if we let $n=m+2q$. The result is the correlation function consisting of $(n\times n)$ and $(s\times n)$ terms, the former appearing if $m=0$, $q\geq1$, and the latter if $m\geq1$, $q\geq1$. The dc noise output is obtained when $m=q=0$.

where

$$C_\nu^2 = \frac{\beta^2\Gamma(\nu+1)^2}{4\Gamma(\nu/2)^2}\left(\frac{\psi}{2}\right)^\nu. \tag{26}$$

Observe that when $p \to \infty$, either because the signal relative to the noise is very large or the noise is vanishingly small while the signal remains constant in strength, the low-frequency correlation function assumes the following forms, in the important special cases of linear and quadratic rectification discussed here, specifically:

$$\begin{aligned}(\nu = 1):\ R_0(t)_{(n\times n)} &\cong \frac{\beta^2\psi p^{-1}}{8\pi^2} r_0(t)^2{}_2F_1(1;3/2;1;\lambda^2);\\ R_0(t)_{(s\times n)} &\cong \frac{\beta^2\psi}{\pi} r_0(t),\end{aligned} \tag{27a}$$

and

$$\begin{aligned}(\nu = 2):\ R_0(t)_{(n\times n)} &\cong \frac{\beta^2\psi^2}{4} r_0(t)^2;\\ R_0(t)_{(s\times n)} &\cong \frac{\beta^2\psi^2 p}{4}(2+\lambda^2)r_0(t).\end{aligned} \tag{27b}$$

We notice from (24) and (25) that the $(n\times n)$ noise is *always* (all $\nu>0$) suppressed relative to the $(s\times n)$ disturbance, by a factor $p^{-1}(\lambda=0)$ at least.

When $\nu=2$, we have the familiar half-wave quadratic detector, for which (24) and (25) become particularly simple, namely, one-quarter of the first two terms of (4), respectively. In fact, when ν is even, (24) and (25) assume simple forms, since the various series terminate after $\nu+2-2m-2q \leqq 0$ terms; ν odd, however, is not so obliging. The series do not terminate, and we must investigate them to see in what way it may be possible to evaluate the averages and reduce the number of terms.

First, let us consider the case of no modulation, $p(t)=p$, when ν is odd and, in the most important instance, unity. Numerical examination of (24) when $t=0$ shows that the error is greatest (9.3 per cent) in omitting the series $(q\geqq 2)$ for small values of $p(<1)$, and that this error steadily decreases as $p\to\infty$. The limiting expression for (24) is

$$(\nu = 1):\quad R_0(t)_{(n\times n)} \cong \frac{\beta^2\psi}{8\pi}\left\{\frac{r_0(t)^2p^{-1}}{\pi} + \sum_{q=2}^{\infty}\frac{\pi p^{1-2q}r_0(t)^{2q}}{(q!)^2\Gamma(3/2-q)^4}\right\}_{p\to\infty}, \tag{28}$$

indicating that on the whole the $(n\times n)$ contribution is suppressed as p^{-1}, $p\to\infty$, but that the higher terms $(q\geqq 2)$ go out at least p^2 as rapidly as the leading term $(q=1)$.

A similar analysis for (25) shows that in this instance the absolute error arising from the omission of the series $(m\geqq 1,\ q\geqq 1)$ starts at zero for $p=0$, increases to a maximum of about 0.09 at $p\sim 0.7$, and then falls off uniformly with increasing p. The limiting expression for (25) becomes

$$(\nu = 1):\quad R_0(t)_{(s\times n)} \cong \frac{\beta^2\psi}{8\pi}\left\{\frac{8r_0(t)}{\pi} + \sum_{m=1}^{\infty}\sum_{q=1}^{\infty}\frac{2\pi p^{1-m-2q}r_0(t)^{2q+m}}{q!(q+m)!\Gamma(3/2-q)^2\Gamma(3/2-m-q)^2}\right\}_{p\to\infty}, \tag{29}$$

from which it is evident that the series terms, representing higher-order $(s\times n)$ modulation products, are suppressed at least as p^{-2} versus p^0 of the leading term. In other words, when s_F^2 exceeds n_F^2 $(=\psi)$ sufficiently, there is still noise in the low-frequency output, but this noise is directly proportional to the original noise leaving the if and quite independent of signal strength.

The total per cent error in omitting the series terms in $R_0(t)_{(n\times n)}$ and $R_0(t)_{(s\times n)}$ is found to be 8.5 per cent, or less, as p increases, in the instance of the mean noise power $(t=0)$, and is even smaller for the various spectra; for example, (<4 per cent) for the optical if when $p=0$. It has also been found that the error is somewhat reduced when there is modulation. In any case, for most purposes we see that it is quite safe to neglect the higher noise components $(m\geqq 0,\ q\geqq 2$, and $m\geqq 1,\ q\geqq 1)$. There then remains the calculation of the averages of the leading terms of (24) and (25). With modulation the exact treatment is also tedious but, fortunately, unnecessary for input signal-to-noise ratios less than about 2, as we may then approximate the hypergeometric functions in the leading terms in the same manner as was used for the signal (see Part III). Very large values of $p(t)$ permit us to use the asymptotic development in connection with (24) and (25), but for intermediate values there appears to be no escaping the direct development in moments $[p_1^{k_1}p_2^{k_2}]_{\text{av}}$. Again, fortunately, the region of greatest interest is usually that for which $p<2$.

Observing that we may set $t=0$ in p_2, i.e., $p_2=p_1=p(t_0')$ (but not in $r_0(t)$!), with negligible error, in the first terms of (24) (25), because $r_0(t)$ falls off to zero so rapidly when $t>0$ compared to $\cos\omega_A lt$, we obtain finally the correlation functions for the noise

$$\begin{aligned}(\nu = 1):\quad R_0(t)_{(n\times n)} &\doteq C_\nu^2 r_0(t)^2\sigma_1(p,\lambda;\nu)\\ &\doteq C_1^2 r_0(t)^2\left\{1 + 2\alpha_1\overline{p} - 2\gamma_1\overline{p^{k_1}} + \alpha_1^2\overline{p^2} + \gamma_1^2\overline{p^{2k_1}} - 2\alpha_1\gamma_1\overline{p^{k_1+1}}\right\},\end{aligned} \tag{30}$$

and

$$\begin{aligned}(\nu = 1):\quad R_0(t)_{(s\times n)} &\doteq C_\nu^2 r_0(t)\sigma_2(p,\lambda;\nu)\\ &\doteq C_1^2 r_0(t)2\left\{\overline{p} + 2\alpha_2\overline{p^2} - 2\gamma_2\overline{p^{k_2+1}} + \alpha_2^2\overline{p^3} + \gamma_2^2\overline{p^{2k_2+1}} - 2\alpha_2\gamma_2\overline{p^{k_2+2}}\right\},\end{aligned} \tag{31}$$

with $p(t)_{\max} \leqq p$, the upper limit for which our approximation of the hypergeometric functions are valid.

In the specific instance of the sinusoidally modulated carrier, (16) describes $p(t)$, and from (2.30) of footnote reference 7 we observe that the average value of $p(t)^\mu$ becomes

$$p(t)^\mu = p^\mu {}_2F_1(-\mu, -\mu + 1/2; 1; \lambda^2). \tag{32}$$

Then we may write, finally,

$$\begin{aligned}\sigma_1(p, \lambda; 1) = \{&1 - p(1 + \lambda^2/2) + 0.334p^{3/2}(1 + 3\lambda^2/2)\\ &+ p^2(1 + 3\lambda^2 + 3\lambda^4/8)/4\\ &- 0.167p^{5/2}(1 + 5\lambda^2 + 15\lambda^4/8)\\ &+ 0.0279p^3(1 + 15\lambda^2/2\\ &+ 45\lambda^4/8 + 5\lambda^6/16)\},\\ \sigma_2(p, \lambda; 1) = 2p\{&(1 + \lambda^2/2) - p(1 + 3\lambda^2 + 3\lambda^4/8)/2\\ &+ 0.131p^{3/2}(1 + 5\lambda^2 + 15\lambda^4/8)\\ &+ p^2(1 + 15\lambda^2/2 + 45\lambda^4/8 + 5\lambda^6/16)\\ &- 0.0327p^{5/2}(1 + 21\lambda^2/2 + 105\lambda^4/8\\ &+ 35\lambda^6/16) + 4.29 \cdot 10^{-3}p^3(1 + 14\lambda^2\\ &+ 105\lambda^4/4 + 35\lambda^6/4 + 35\lambda^8/128)\},\end{aligned} \tag{33}$$

for the linear detector, and for the quadratic device we have simply

$$\sigma_1(p, \lambda; 2) = 1, \text{ and } \sigma_2(p, \lambda; 2) = p(2 + \lambda^2),\ p \geqq 0. \tag{34}$$

Figs. 4 and 5 illustrate σ_1 and σ_2 when $\nu = 1$, and Fig. 5 also shows $\sigma_2(p, \lambda, 2)$. In the linear rectifier the $(n \times n)$ noise is suppressed as the signal strength is increased, the more so for the lesser percentages of modulation.

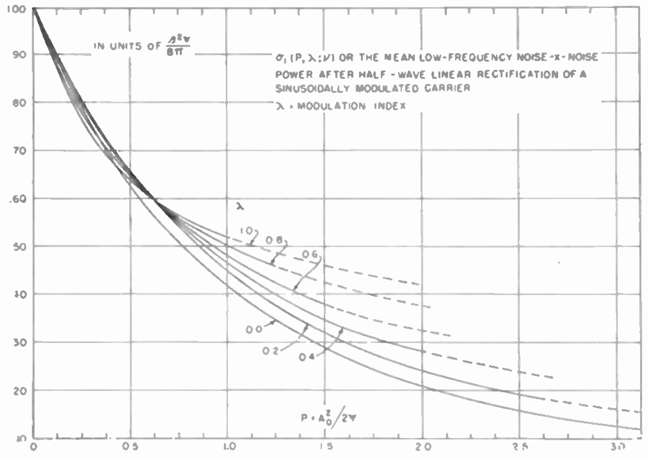

Fig. 4—Mean low-frequency output $(n \times n)$ noise power, for the linear rectifier.

On the other hand, the $(s \times n)$ contribution increases with the larger signal powers, and the less is the increase for greater values of λ. Physically, this behavior is explained by the fact that, when the signal is large, the noise "rides" on top of it, so that for little modulation both positive and negative noise peaks are passed by the detector without much clipping due to cutoff. Greater percentages of modulation mean less noise voltage passed near the bottom of a modulation cycle; hence the decrease in $(s \times n)$ noise for $\lambda \sim 1$, or $\lambda = 0$. When $p \to \infty$, however, there is not much difference between noise outputs with 0 or 100 per cent modulation, as relatively little noise per (modulation) cycle is then blocked in either instance.

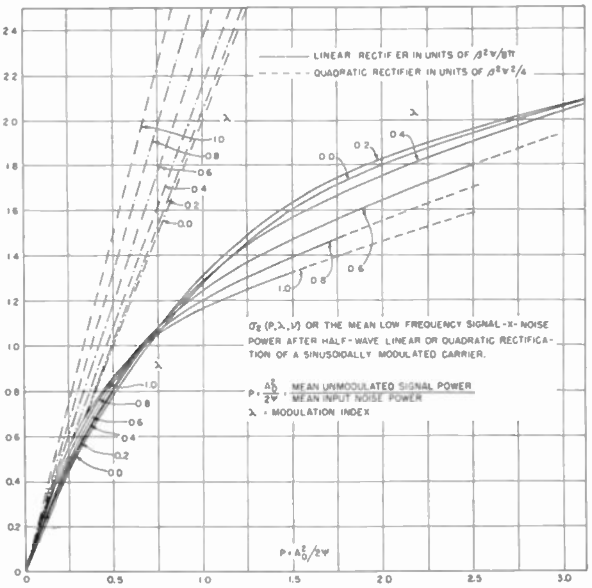

Fig. 5—Mean low-frequency output $(n \times n)$ noise power, for the linear and quadratic rectifiers.

Figs. 6 and 7 show the total mean low-frequency noise power, $\sigma_1 + \sigma_2$, when $\nu = 1$ and $\nu = 2$ for all degrees of modulation. Here again it is evident $(\nu = 1)$ that even in the extreme case of $\lambda = 1$ modulation does not radically change the output noise power. This is important experimentally, because it allows us to measure the output noise separately from the signal, by injecting an *unmodulated* carrier into the if. The output then contains no signal contribution, the dc being filtered out. The noise power so determined may be substituted for the exact expression in the presence of modulation with an

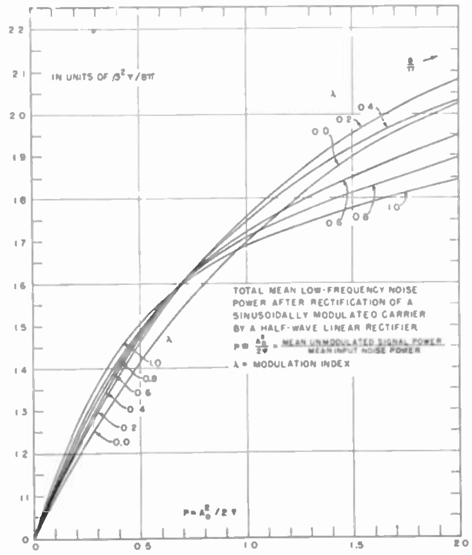

Fig. 6—Total mean low-frequency noise power output, $\nu = 1$.

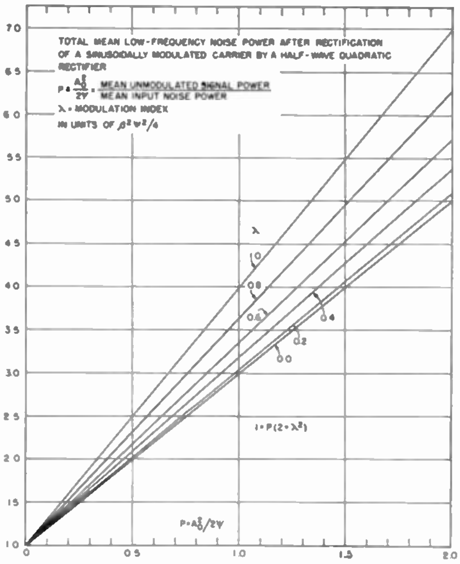

Fig. 7—Same as Fig. 6, but with $\nu=2$.

error of less than 10 per cent, an error which decreases as $p\to\infty$.

The full low-frequency power spectrum follows from (11), (30), (31), and (33), (34), namely,

$$W_0(f) \doteq C_\nu^2\left\{\sigma_1(p, \lambda; \nu)\int_0^\infty r_0(t)^2 \cos \omega t dt + \sigma_2(p, \lambda; \nu)\int_0^\infty r_0(t) \cos \omega t dt\right\}, \quad (35)$$

the first term representing $(n\times n)$ products, the second, $(s\times n)$ products; (35) is exact for $\nu=2$. The mean audio noise power may then be obtained from (35) in (8).

V. AUDIO NOISE POWER ($\nu>0$)

The mean noise power passed by our audio filter, of width Δf, is

$$n_a^2 = \frac{\beta^2\Delta f\Gamma(\nu+1)^2}{\Gamma(\nu/2)^2}\left(\frac{\psi}{2}\right)^\nu\left\{\sigma_1\int_0^\infty r_0(t)^2dt + \sigma_2\int_0^\infty r_0(t)dt\right\}, \quad (36)$$

with different ranges of applicability for p and λ, depending on the value of ν; for $\nu=1$, $p(1+\lambda)^2$ is required to be less than 4, while for $\nu=2$, all values of p and λ are allowed. Now the physical significance of this quantity is conveniently brought out through an examination of its behavior for limiting values of signal strength, i.e., $p=0$, $p\to\infty$, and its dependence on if filter width at these values.

Let us consider first the linear rectifier and restrict ourselves to the case of no modulation ($\lambda=0$); the latter assumption embodies no real loss of generality, as the preceding section has shown that modulation has small effect (~10 per cent or less) on the output noise power when $\nu=1$. To obtain n_a^2 when p becomes very large we must use asymptotic forms for R_0 instead of (36); the result is

$$n_a^2\Big|_{p\to\infty} \cong \frac{4\beta^2\Delta f\psi\Gamma^{(1)}}{\pi^2\omega_F} = \frac{4\beta^2\Gamma^{(0)}\Gamma^{(1)}}{\pi^2} W_0\Delta f, \quad (37)$$

where $\Gamma^{(0)}$, $\Gamma^{(1)}$, and $\Gamma^{(2)}$ are numerical factors depending on the shape of the if. Specifically,

$$\left.\begin{aligned}\Gamma^{(0)} &\equiv \psi/W_0\omega_F = 1/\pi;\ 1/2\sqrt{\pi};\ 1/2\\ \Gamma^{(1)} &\equiv \omega_F\int_0^\infty r_0(t)dt = \pi/2;\ \sqrt{\pi};\ 1\\ \Gamma^{(2)} &\equiv \omega_F\int_0^\infty r_0(t)^2dt = \pi/2,\ \sqrt{\pi/2};\ 1/2\end{aligned}\right\}, \quad (38)$$

for the rectangular, gaussian, and optical distributions, in that order (see Table I).

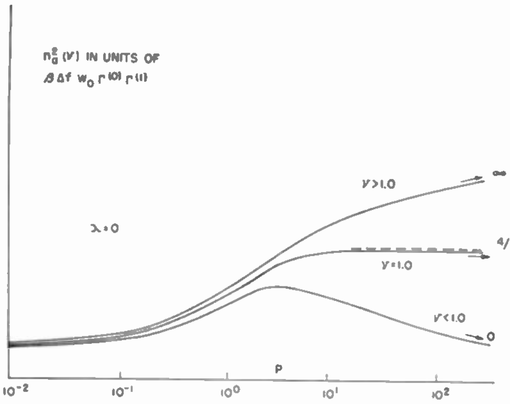

Fig. 8—Mean audio noise power output in the absence of modulation, for three different detector characteristics.

The expression (37) is approached monotonically and is the maximum value of the audio noise power, while the minimum value occurs for no signal at all and is, without the higher terms,

$$n_a^2\Big|_{p=0} = \frac{\beta^2\psi\Delta f\Gamma^{(2)}}{2\pi\omega_F} = \frac{\beta^2W_0\Delta f\Gamma^{(0)}\Gamma^{(2)}}{2\pi}, \quad (39)$$

from (36). The physical explanation is as follows:

When the signal is negligible, the *envelope*, i.e., the low-frequency part of the rectified output, in which we are interested, is formed essentially of the positive peaks of the noise wave. When the signal is very large, the noise "rides" on it, and although only the positive portion of this mixture is passed by the detector, the noise can now fluctuate about its average, on either side, thus effectively doubling the deviation in amplitude of the resulting envelope. It is this deviation that forms the low-frequency voltage, after the dc has been removed. The effect becomes more pronounced as the signal is increased, until for huge signals there is no further gain, since the noise is now effectively lifted above the rectification level. The process is clearly monotonic. Thus the presence of the carrier adds progressively more $(s\times n)$ noise to the low-frequency, at the expense of power in the higher spectral zones centered about harmonics of the carrier, until the limit $p\to\infty$ is attained.

A similar argument holds when $\nu\neq1$. Here we have the more general results, from (36), (24), and (25), analogous to (37) and (39), that

$$p \to 0: \quad n_a^2(\nu) \doteq \frac{\beta^2\Gamma(\nu+1)^2\Delta f W_0^{\nu}\Gamma^{(0)}\omega_F^{\nu-1}\Gamma^{(2)}}{2^\nu\Gamma(\nu/2)^2}, \qquad (n \times n). \tag{40}$$

$$p \to \infty: \quad n_a^2(\nu) \simeq \frac{\beta^2\Gamma(\nu+1)^2\Delta f W_0\Gamma^{(0)}\Gamma^{(1)}s_F^{2\nu-2}}{2^{\nu-1}\Gamma(\nu/2)^2\Gamma(\nu/2+1)^2}, \qquad (s \times n). \tag{41}$$

Now when $\nu<1$, the $(s\times n)$ noise is completely suppressed[12] $(s_F^2 \to \infty)$. The detector characteristic tends to "squash down" the higher peaks of the input wave, and consequently, as the signal is increased, the noise that rides on top is ironed out. When $\nu>1$, the reverse is true: the peaks are enhanced, the greater they and the signal amplitude are, and the $(s\times n)$ grows indefinitely. Equation (41) illustrates this and explains why, when $\nu>1$, one does not have a maximum. Fig. 8 shows the three cases, including $\nu=1$, discussed at the beginning of the section.

We observe further from (41) that for all values of ν the amount of noise is independent of the if spectral width $(\sim\omega_F)$, since it is only the immediate spectral vicinity of the carrier that contributes to the $(s\times n)$ noise which appears near $f=0$ in the output.

We have now to explain the behavior of the audio noise as a function of ω_F. Let us again consider the linear rectifier. Then, spreading the if spectrum, but keeping its intensity W_0 constant, does *not* increase the amount of noise near $f=0$, although at first glance one increases the number of closely adjacent noise products out on the added portions of the spectrum. However, we have neglected to take into account the "crowding effect," which is a consequence of the nonlinear action of the device, and which states that the products of adjacent frequencies near the center of the band outweigh in *intensity* those near its limits, although the number of such products is independent of frequency within the bandwidth. The result is that an increase in the if width simply increases the width of the low-frequency spectrum, but does not change the intensity near $f=0$. This is somewhat easier to understand on the basis of our earlier picture of the noise envelope. Doubling the spectral width merely increases the average and rms heights of the envelope by $\sqrt{2}$. The low-frequency (current or voltage) output is increased by the same factor and the low-frequency output power is doubled. The area of the power-spectrum triangle[13] is doubled, but since the envelope now varies twice as rapidly, the low-frequency power-spectrum triangle must now expand to twice its original value. Thus the area is doubled and consequently the intensity *near* $f=0$ must remain unchanged.[14]

We see also, from (40), that when $\nu<1$ the spectral ordinate near $f=0$ is actually decreased as ω_F is made larger, while the reverse is true when $\nu>1$, the latter a familiar result in the instance of the quadratic detector.

VI. AUDIO SIGNAL-TO-NOISE RATIOS: DISCUSSION

The results of Sections III and IV now enable us to obtain a general expression of the rms audio signal-to-noise ratios, when the lth harmonic of the modulation is chosen as the signal to be observed. In the first case of chief interest, namely, the first harmonic after linear detection, we may write the ratio as

$$(\nu=1): \quad \frac{s_a}{n_a} = p\lambda\left[\frac{(\omega_F)_{\text{opt.}}}{2\Delta f}\right]^{1/2}(1 - 0.2289p^{1/2}[1+\lambda^2/4])$$

$$+ \cdot\left[\begin{Bmatrix}1\\1/\sqrt{2}\\1/2\end{Bmatrix}\sigma_1(p,\lambda;1) + \begin{Bmatrix}1\\1\\1\end{Bmatrix}\sigma_2(p,\lambda;1)\right]^{-1/2}, \qquad p(1+\lambda)^2 \leqq 4 \tag{42}$$

where the quantities in the braces { } apply downward accordingly as the if has a rectangular, gaussian, or optical shape (see Table I and footnote). For the other important special case, quadratic rectification, our more general results are readily shown to reduce to (14) valid for all p and $\lambda \leqq 1$. Curves illustrating s_a/n_a as a function of p are given in Figs. 9–11 for the three types

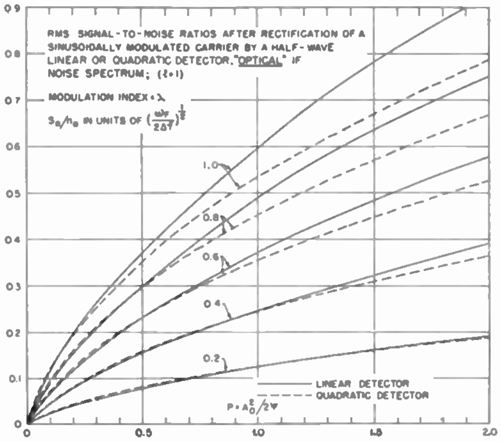

Fig. 9—Rms audio signal-to-noise ratios after half-wave linear or quadratic detection. Optical if filter.

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[12] As has been shown in Part IV, $(n\times n)$ noise is *always* suppressed more rapidly than the $(s\times n)$ products, by a factor p^{-1}, at least.

[13] We have assumed for simplicity that the if is rectangular and that $p=0$. A finite signal introduces no change in the essentials of the argument.

[14] Rice, in footnote reference 4, part IV, comes to the same conclusion, and by much the same train of thought.

of spectra mentioned above. Observe that the rectangular if gives the smallest, and the optical the largest value of s_a/n_a for a given p.

When the signal is small, we find for the general νth-law device that

$$\frac{s_a}{n_a} = \frac{p\lambda}{(2\Delta f)^{1/2}}\left(\int_0^\infty r_0(t)^2 dt\right)^{-1/2}$$

$$= \frac{s_F{}^2\lambda}{W_0(2\Delta f\omega_F\Gamma^{(0)}\Gamma^{(2)})^{1/2}}, \quad \begin{matrix} p \ll 1, \\ l = 1, \end{matrix} \tag{43}$$

which shows that in all cases the behavior is that of a quadratic rectifier, and is quite independent of ν. The reason for this lies in the aforementioned phenomenon of modulation suppression. Now, for small signals the background noise is hardly affected by the signal, and consists almost exclusively of $(n\times n)$ terms. However, the signal is drastically affected by the noise, being reduced as the square of the input signal strength. Further, following precisely the same argument as that given in parts II, III, of footnote reference 2 for radar or pulsed signals, we can show that *the best if filter*, which makes s_a/n_a a maximum for a given input signal, *is one which is the modulus of the Fourier transform of the signal* —in this instance, three "delta-functions" or infinitely narrow pass-bands centered about the carrier and its two sidebands. Again, it is modulation suppression, arising from the nonlinearity of the detector, which dictates this kind of optimum if filter response, for without suppression one would need only an indefinitely narrow filter centered about one of the components.

For very large signals, we find with the aid of the asymptotic form for ${}_1F_1$ applied to (20), (24), and (25),

$$\frac{s_a}{n_a} \cong \frac{\lambda p^{1/2}}{\sqrt{2}(2\Delta f)^{1/2}} \frac{{}_2F_1(1/2-\nu/2, 1-\nu/2; 2; \lambda^2)}{{}_2F_1(1-\nu, 3/2-\nu; 1; \lambda^2)^{1/2}}\left(\int_0^\infty r_0(t)dt\right)^{-1/2}$$

$$= \frac{\lambda s_F\,{}_2F_1(1/2-\nu/2; 1-\nu/2; 2; \lambda^2)}{\sqrt{2}(2\Delta f)^{1/2}[\Gamma^{(0)}\Gamma^{(1)}W_0\,{}_2F_1(1-\nu, 3/2-\nu; 1; \lambda^2)]^{1/2}}. \tag{44}$$

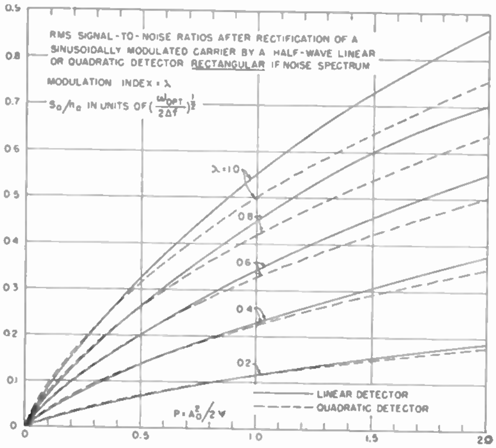

Fig. 10—Same as Fig 9, rectangular if filter.

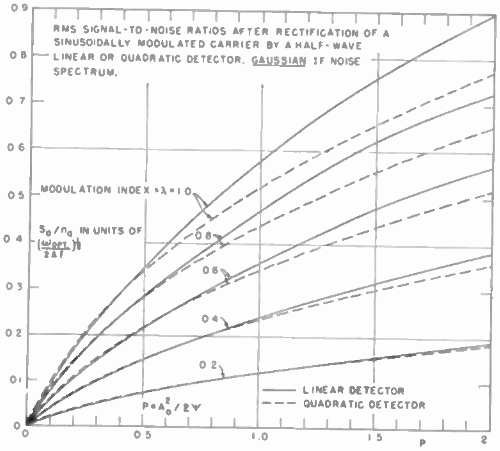

Fig. 11—Same as Fig. 9, gaussian if filter.

Here only $(s\times n)$ products are significant, the if filter shape is unimportant, just as in the corresponding radar case, except for minor numerical terms ($\Gamma^{(0)}$, $\Gamma^{(1)}$, etc.), and so also is the law (ν) of the detector only secondarily significant. Note, however, for the continuous signals considered here, that the audio signal-to-noise ratio is proportional to $p^{1/2}$, while for pulsed signals this ratio becomes proportional to p. The explanation follows from the presence of $(s\times n)$ noise terms, which become dominant ($\sim p^{\nu-1}$ as $p\to\infty$) in the former instance, whereas in the radar example the signal is on such a small fraction of the repetition period that $(s\times n)$ products are ignorable and $(n\times n)$ components ($\sim p^{\nu-2}$ as $p\to\infty$ alone are the obscuring factor. Thus, as we would expect, the chief difference between the detection of pulsed and continuous signals only becomes apparent when $p>1$.

Special forms of (44) are

$$(\nu = 1): \quad \frac{s_a}{n_a} \cong \frac{\lambda p^{1/2}}{\sqrt{2}}\left(\frac{\omega_F}{2\Delta f\Gamma^{(1)}}\right)^{1/2};$$

$$(\nu = 2): \quad \frac{s_a}{n_a} \cong \frac{\lambda p^{1/2}}{\sqrt{2}}\left[\frac{\omega_F}{(2+\lambda^2)\Delta f\Gamma^{(1)}}\right]^{1/2}, \tag{45}$$

and so we find that

$$\frac{(s_a/n_a)_{\nu=2}}{(s_a/n_a)_{\nu=1}} \cong \frac{1}{(1+\lambda^2/2)^{1/2}}, \tag{46}$$

which can be observed in Figs. 9–11: the ratio for the quadratic detector is always reduced relative to the linear, due to a generation of a greater amount of $(s\times n)$ noise when $\lambda>0$ in the latter instance.

The results of this theory give a general picture of the relation between the signal at the output of a receiver as a function of the signal at the input of the if amplifier. For the purposes of making tests on the correlation between theory and experiment, it appears clear that the most useful type of detector, and therefore the one to be investigated, is the linear type.

In the companion paper[15] by Fubini and Johnson, the results of the theory for sinusoidal modulation and linear detection have been tested. The curves given in their paper are a replot of those presented here with a change in scale from linear to logarithmic co-ordinates. It appears in their work that, despite all the approximations made as to shape of if filter, linearity of rectification, and the like, the correspondence between theory and experiment is satisfactory for most practical purposes.

[15] E. G. Fubini and D. C. Johnson, "Signal-to-noise ratio in AM modulated receivers," Proc. I.R.E., vol. 36, pp. 1461–1467; this issue.

An Approximate Solution of the Problem of Path and Absorption of a Radio Wave in a Deviating Ionosphere Layer*

JAMES E. HACKE, JR.†, AND JOHN M. KELSO†

Summary—**A previously obtained double-parabola approximation to a Chapman distribution is used to obtain approximate solutions for ray path and absorption of a radio wave incident obliquely on a deviating plane ionospheric layer from which the wave is "reflected." The solutions, expressed analytically and graphically, are valid when the earth's magnetic field and second-order absorption effects can be neglected.**

INTRODUCTION

A PREVIOUS PAPER[1] proposed a double-parabola approximation to the Chapman distribution[2] which is generally accepted as representing variation of ion density with height in the E and the F_1 layers of the ionosphere. The paper also proposed a parabolic approximation for the product of ion density by electron-collision frequency as a function of height. By using these approximations, analytical solutions were obtained for "true" and apparent reflection height, and for absorption of a radio wave incident normally on the ionosphere. These solutions were also expressed in graphical form, and were compared with numerical results obtained by Pierce[3] and by Jaeger.[4]

The present paper extends the results of the previous paper to oblique incidence while retaining the restrictions that the wave frequency in the ionosphere be everywhere greater than the collisional frequency, that the effects of the earth's magnetic field be neglected, and that absorption per wavelength be small. These restrictions are met in practice for waves above about 1 Mc and correspond to consideration of the "ordinary" ray in quasi-transverse propagations.[5] Attention is confined to refraction and absorption in the deviating region (wave frequency insufficiently high to penetrate the layer) because an exact solution can be made of the problem when the ray penetrates the layer, at least if the ray can be considered as passing on essentially to infinity, and if the wave frequency is much greater than the critical frequency.

SUMMARY OF PREVIOUS RESULTS

In the previous paper[1] it was shown that the ion density in a Chapman region is given by

$$N = N_m Ch(x) \tag{1}$$

where

$$N_m = \sqrt{(\beta S_\infty \sec \chi)/\kappa H \alpha \epsilon}$$
$$Ch(x) = \exp\,(1 - x - \epsilon^{-x})/2 \tag{2}$$
$$x = (h - h_0)/H - \ln \sec \chi$$

α = the recombination coefficient in the ionosphere

ϵ = the base of Naperian logarithms

β = the ionization density produced by unit surface density of incident radiation

S_∞ = the surface density of radiation incident on the ionosphere

H = the "scale height" of the atmosphere in the region where the ionization is produced; division by H yields distances in "scale units"

h_0 = the height at which N is a maximum when $x = 0$

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* Decimal classification: R112.62. Original manuscript received by the Institute, March 29, 1948.

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[1] J. E. Hacke, Jr., "An approach to the approximate solution of the ionosphere absorption problem," Proc. I.R.E., vol. 36, pp. 724–727; June, 1948.

[2] S. Chapman, "The absorption and dissociative or ionizing effect of monochromatic radiation in an atmosphere on a rotating earth," *Proc. Phys. Soc.*, vol. 43, pp. 26–45; January, 1931.

[3] J. A. Pierce, "The true height of an ionosphere layer," *Phys. Rev.*, vol. 71, pp. 698–706; May, 1947.

[4] J. C. Jaeger, "Equivalent path and absorption in an ionosphere region," *Proc. Phys. Soc.*, vol. 59, pp. 87–96; January, 1947.

[5] H. G. Booker, "Some general properties of the formula of the magneto-ionic theory," *Proc. Phys. Soc.*, vol. 147, pp. 352–382; November, 1934.

χ = the sun's angular distance from the zenith.

Equation (2) was approximated by the two parabolas,

$$P_1(x) = 1 - x^2/T^2, \quad x_1 \angle x \angle 0; \tag{3}$$

$$P_2(x) = A^2(x - x_2)^2, \quad x_2 \angle x \angle x_1. \tag{4}$$

In these equations T, A, and x_2 are parameters adjusted to fit (2), and x_1 is the negative point of inflection of (2):

$$x_1 = -\ln(2 + \sqrt{3}) = -1.317$$

$$x_2 = x_1 - 4/(\epsilon^{-x_1} - 1) = -2.781$$

$$T = -x_1/\sqrt{1 - Ch(x_1)} = 1.848$$

$$A = (\epsilon^{-x_1} - 1)\sqrt{Ch(x_1)}/4 = 0.4792.$$

The index of refraction in an ionized layer is given by

$$\mu^2 = 1 - Ch(x)/R^2 \tag{5}$$

where R is the ratio of wave frequency to vertical-incidence critical frequency; and the absorption per unit length is given by

$$K = K_m Ch(x)\epsilon^{-x}/(\mu R^2) \tag{6}$$

where

$K_m = \nu_m/(2c)$

ν_m = the value of the collisional frequency ν at $x = 0$

c = the velocity of light in vacuum.

It is shown in the previous paper that

$$Ch(x)\epsilon^{-x} = P_3(x) \doteq a_0 + a_1 x + a_2 x^2, \quad x_3 \angle x \angle 0 \tag{7}$$

where

$$a_0 = 0.7055; \quad a_1 = -2.382; \quad a_2 = -1.1696;$$

and

$$x_3 = -2.299 \text{ is the negative root of } P_3(x) = 0.$$

The ratios

$$N/N_m = Ch(x), \text{ and } N\nu/N_m\nu_m) = Ch(x)\epsilon^{-x},$$

are plotted in Fig. 1 as functions of x. Shown dotted are the parabolic approximations P_1, P_2, and P_3.

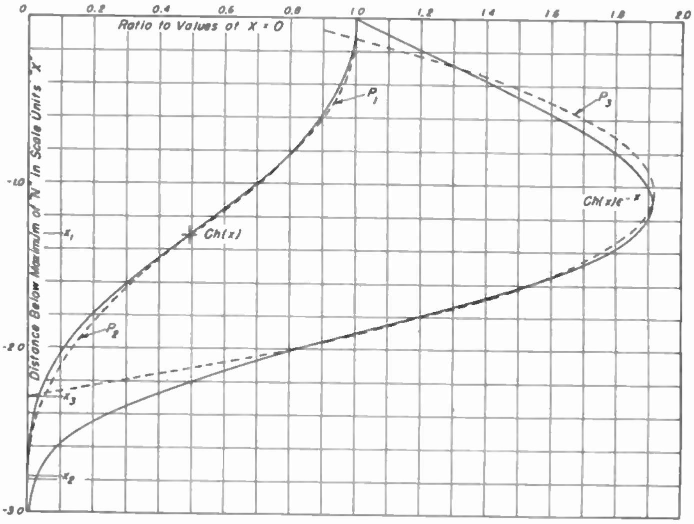

Fig. 1—Ion density and its product with collision frequency as functions of height in scale units below point of maximum ion density. Solid curves, Chapman distributions; dashed curves, parabolic approximations.

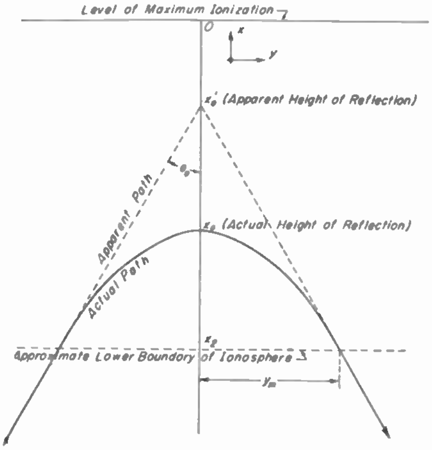

Fig. 2—Diagram illustrating ray path of radio wave through plane ionosphere.

RAY PATH

Fig. 2 depicts an upgoing radio wave entering a plane ionized layer at an initial angle θ_0 with the vertical. The decrease of μ in the region deflects it away from the vertical; at a point x_0 its path becomes horizontal and the wave is said to be "reflected" at this point. The ray leaves the ionosphere at the same angle θ_0 with the vertical as that at which it enters.

By Snell's law,

$$(\sin\theta)/(\sin\theta_0) = 1/\mu \tag{8}$$

where θ = the angle the ray makes with the vertical at any point on its path. This equation gives the angular direction of the ray path at any point in the region which the ray reaches, irrespective of the value of μ in intervening strata (on the downcoming half of the path, of course, θ is greater than $\pi/2$). From this equation,

$$dy/dx = \tan\theta = (\sin\theta_0)/\sqrt{\mu^2 - \sin^2\theta_0} \tag{9}$$

where dy is horizontal displacement in scale units. From equation (5),

$$\begin{aligned} dy/dx &= (\sin\theta_0)/\sqrt{1 - Ch(x)/R^2 - \sin^2\theta_0}, \\ &= (R\sin\theta_0)/\sqrt{R^2\cos^2\theta_0 - Ch(x)}, \\ &= (R\sin\theta_0)/\sqrt{R'^2 - Ch(x)} \end{aligned} \tag{10}$$

where $R' = R\cos\theta_0$ is the ratio of wave frequency to the critical frequency at angle of incidence θ_0.

When (3) and (4) are substituted in (10), the problem divides into three parts:

(i) Behavior in the upper region; the ray penetrates to the region $(x_1 \angle x \angle 0)$ when R' is equal to or greater than 0.7015.

(ii) Behavior in the lower region $(x_2 \angle x \angle x_1; R' \angle 0.7015)$ when reflection occurs in the lower region.

(iii) Behavior in the lower region ($x_2\angle x\angle x_1; R'>0.7015$) when reflection occurs in the upper region.

In Case (i), (9) becomes

$$dy=(R\sin\theta_0)dx/\sqrt{R'^2-(1-x^2/T^2)},$$
$$\doteq(TR\sin\theta_0)dx/\sqrt{T^2(R'^2-1)+x^2},\quad x_1\angle x\angle x_0; \tag{11}$$
$$y-y_0=(TR\sin\theta_0)\int_{x_1}^{x} dx/\sqrt{T^2(R'^2-1)+x^2},$$
$$=(TR\sin\theta_0)\ln\frac{x+\sqrt{T^2(R'^2-1)+x^2}}{x_1+\sqrt{T^2(R'^2-1)+x_1^2}},$$
$$x_1\angle x\angle x_0 \tag{12}$$

where y_0 is defined later (see (15)).

The height of reflection x_0 can be found from (8), for the ray path must be horizontal at $x=x_0$, and therefore $\sin\theta$ must equal one. From (5) and (8),

$$\mu^2 = 1 - Ch(x_0)/R^2 = \sin^2\theta_0;$$

substituting (3),

$$1-(1-x_0^2/T^2)/R^2 = 1-\cos^2\theta_0,\quad x_1\angle x_0;$$
$$x_0 = -T\sqrt{1-R'^2},\quad x_1\angle x_0. \tag{13}$$

If (13) be substituted in (12), one obtains

$$y-y_0 = TR\sin\theta_0\ln\left[\frac{x+\sqrt{x^2-x_0^2}}{x_1+\sqrt{x_1^2-x_0^2}}\right],$$
$$x_1\angle x\angle x_0. \tag{14}$$

Choose y_0 so $y(x_0)=0$:

$$-y_0 = TR\sin\theta_0\ln\left[x_0/(x_1+\sqrt{x_1^2-x_0^2})\right],$$
$$x_1\angle x\angle x_0; \tag{15}$$

substituting in (14),

$$y = TR\sin\theta_0\ln\left[(x+\sqrt{x^2-x_0^2})/x_0\right],\quad x_1\angle x\angle x_0. \tag{16}$$

In numerical calculations the negative value of the square root of $x^2-x_0^2$ is chosen in order that y increase with decreasing x_0^2. This same choice is also made in (21), (23), (25), and (30) for $\sqrt{x_1^2-x_0^2}$.

In Cases (ii) and (iii), where it is the region ($x_2\angle x\angle x_1$) that is being studied, (9) becomes

$$dy=R\sin\theta_0 dx/\sqrt{R'^2-A^2(x-x_2)^2},\quad x_2\angle x\angle x_1; \tag{17}$$
$$y-y_0=R\sin\theta_0\int_{x_2}^{x} dx/\sqrt{R'^2-A^2(x-x_2)^2},$$
$$=(R/A)\sin\theta_0\sin^{-1}\left[A(x-x_2)/R'\right],\quad x_2\angle x\angle x_1. \tag{18}$$

In Case (ii) reflection occurs in the region ($x_2\angle x_0\angle x_1$), and the height of reflection is given by

$$\sin\theta(x_0) = 1 = \sin\theta_0/\mu(x_0);$$

from (5),

$$Ch(x_0)/R^2 = \cos^2\theta_0;$$

substituting (4),

$$A^2(x_0-x_2)^2 = R^2\cos^2\theta_0,\quad x_2\angle x_0\angle x_1;$$
$$x_0 = x_2 + R'/A,\quad x_2\angle x_0\angle x_1. \tag{19}$$

In this case we again define y_0 so $y(x_0)=0$, and obtain from (18)

$$-y_0 = (R/A)\sin\theta_0\sin^{-1}1$$
$$= (R/A)(\pi/2)\sin\theta_0,\quad x_2\angle x_0\angle x_1;$$

and since

$$\pi/2 - \sin^{-1}u = \cos^{-1}u,$$

(18) becomes

$$y=(R/A)\sin\theta_0\cos^{-1}\left[A(x-x_2)/R'\right],\quad x_2\angle x\angle x_0\angle x_1. \tag{20}$$

In Case (iii), when reflection occurs in the upper region, then the value of $y(x_1)$ given by (18) must equal the value of $y(x_1)$ given by (16):

$$y_0 + (R/A)\sin\theta_0\sin^{-1}\left[A(x_1-x_2)/R'\right]$$
$$= -TR\sin\theta_0\ln\left[(x_1+\sqrt{x_1^2-x_0^2})/x_0\right],\quad x_1\angle x_0;$$
$$y_0 = R\sin\theta_0\left[T\ln\frac{x_1+\sqrt{x_1^2-x_0^2}}{x_0} - \frac{1}{A}\sin^{-1}\frac{A(x_1-x_2)}{R'}\right],\quad x_1\angle x_0;$$

and (18) becomes, for this case,

$$y = (R/A)\sin\theta_0\big[\sin^{-1}A(x-x_2)/R'$$
$$-\sin^{-1}A(x_1-x_2)/R'$$
$$+AT\ln(x_1+\sqrt{x_1^2-x_0^2})/x_0\big],\quad x_2\angle x\angle x_1\angle x_0. \tag{21}$$

The quantity $y/(R\sin\theta_0)$ is a function only of R' and x, under all three cases

(i) $x_1\angle x\angle x_0$—see (13) and (16);
(ii) $x_2\angle x\angle x_0\angle x_1$—see (20);
(iii) $x_2\angle x\angle x_1\angle x_0$—see (21).

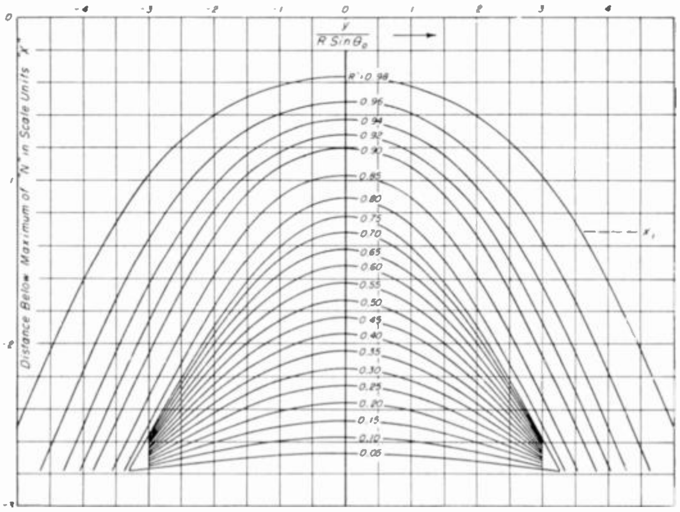

Fig. 3—Ray paths in the ionosphere with R' as a parameter. Ordinates are in units of $y/(R\sin\theta_0)$. See text for notation.

Fig. 3 shows the paths of radio waves through the ionosphere with R' as a parameter. Horizontal distances on the graph must be multiplied by $HR\sin\theta_0$ to obtain physical distances; vertical distances by H. The origin of the graph is at the level of maximum ionization directly above the point of reflection.

The horizontal range in the ionosphere $2y_m$ can be approximated by

$$y_m = y(x_2);$$

from (16) and (21),

$$y_m = (\pi/2)(R/A)\sin\theta_0,\quad x_0 \angle x_1; \tag{22}$$

$$= (R/A)\sin\theta_0\left[AT\ln\left(x_1+\sqrt{x_1^2-x_0^2}\right)/x_0 - \sin^{-1}(x_1-x_2)/R'\right],\quad x_1\angle x_0. \tag{23}$$

The apparent height of reflection x' (see Fig. 2) is given by extrapolating the path directions (upgoing and downcoming) at (x_2, $\pm y_m$) until they meet at x', 0). From geometry, then,

$$x' = x_2 + y_m\cot\theta_0,$$

$$= x_2 + (\pi/2)(R'/A),\ H_0 \subset H_1: \tag{24}$$

$$= x_2 + (R'/A)\left[AT\ln\left(x_1+\sqrt{x_1^2-x_0^2}\right)/x_0 - \sin^{-1}(x_1-x_2)/R'\right],\quad x_1\angle x_0. \tag{25}$$

The actual and the apparent heights of reflection are plotted in Fig. 4 as a function of R'.

When Martyn's equivalent path theorem[6] is put in the notation of this paper, it takes the form

$$x'(R', \theta_0) = x'(R, 0).$$

In words, the apparent height of reflection at oblique incidence is the same function of R' as the apparent height of reflection at vertical incidence is of R. This also applies to the actual height of reflection:

$$x_0(R', \theta_0) = x_0(R, 0).$$

These two relations can be verified by comparing (24) and (25) of this paper with (17) of footnote reference 1.

ABSORPTION

The reflection coefficient ρ is given by

$$\rho = \exp\left(-2\int_{-\infty}^{x_0} K ds\right). \tag{26}$$

Let us define the absorption

$$S = \int K ds; \tag{27}$$

the value of K is given by (6), and that of ds by

$$ds = H\sqrt{(dx)^2+(dy)^2}.$$

Substituting from (9),

$$ds = H\sqrt{(dx)^2+\sin^2\theta_0(dx)^2/(\mu^2-\sin^2\theta_0)},$$

$$= H\mu dx/\sqrt{\mu^2-\sin^2\theta_0};$$

and (27) becomes, substituting from (6),

$$S = \int \left[K_m Ch(x)\epsilon^{-x}/(\mu R^2)\right]\left[H\mu dx/\sqrt{\mu^2-\sin^2\theta_0}\right],$$

$$= (K_m H/R^2)\int Ch(x)\epsilon^{-x}dx/\sqrt{\mu^2-\sin^2\theta_0}. \tag{28}$$

The product of the numerator, $Ch(x)\epsilon^{-x}$, is approximated by (7) in the region $x_3\angle x\angle 0$. As in the determination of path, three cases must be distinguished in the absorption problem:

(i) Absorption in the region ($x_1\angle x\angle 0$) which occurs when R' is greater than 0.7015.

(ii) Absorption in the region below x_1 when R' is less than 0.7015.

(iii) Absorption in the region below x_1 when R' is greater than 0.7015.

In Case (i), μ^2 is approximated by

$$\mu^2 \doteq 1 - P_1(x)/R^2,\quad x_1\angle x\angle x_0;$$

and (28) becomes, for the absorption in the P_1 region,

$$S^{(i)} = \frac{K_m HT}{R}\int_{x_1}^{x_0}\frac{(a_0+a_1x+a_2x^2)dx}{\sqrt{x^2-x_0^2}},\quad x_1\angle x_0. \tag{29}$$

(To this must be added the absorption $S^{(iii)}$ in the P_2 region (see (32)). Equation (29) yields three integrals of the form

$$\int dx/\sqrt{x^2-x_0^2};\qquad \int xdx/\sqrt{x^2-x_0^2};$$

$$\int x^2dx/\sqrt{x^2-x_0^2}.$$

Integration of these is straightforward; the result is

$$S^{(i)} = -(K_m HT/R)\left[(a_0+a_2x_0^2/2)\ln x_0/\left(x_1+\sqrt{x_1^2-x_0^2}\right) - (a_1+a_2x_1/2)\sqrt{x_1^2-x_0^2}\right],\quad x_1\angle x_0. \tag{30}$$

The negative sign in front of the right-hand member of this equation appears because of the choice of the

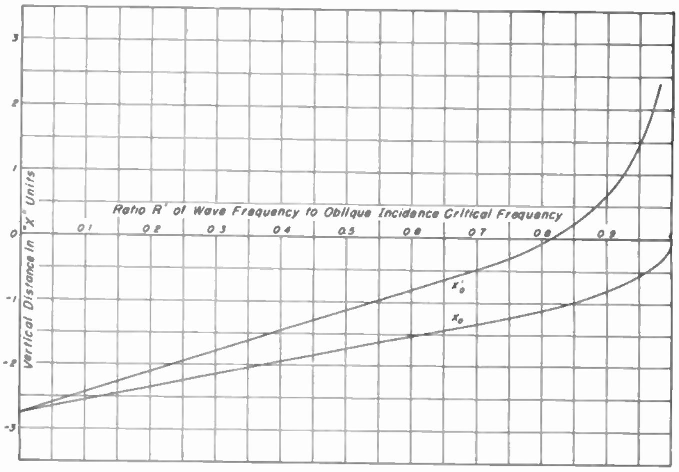

Fig. 4—Apparent and actual height of reflection in scale units below height of maximum ionization of a radio wave as a function of $R' = R\cos\theta_0$ = ratio of wave frequency to oblique-incidence critical frequency.

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[6] D. F. Martyn, "The propagation of medium radio waves in the ionosphere," *Proc. Phys. Soc.*, vol. 47, pp. 323–329; March, 1935.

negative square root of $x_1^2-x_0^2$ as mentioned previously.

In Case (iii) (which is being considered before Case (ii) to complete the absorption $S^{(i)}+S^{(iii)}$ when R' is greater than 0.7015), the absorption $S^{(iii)}$ is given by

$$S^{(iii)}=(HK_m/R)\int_{x_3}^{x_1} P_3(x)dx/\sqrt{R'^2-A^2(x-x_2)^2}, \quad x_1\angle x_0. \quad (31)$$

This can be transformed by a change of variable to three integrals of the form

$$\int du/\sqrt{R'^2-u^2}; \qquad \int udu/\sqrt{R'^2-u^2};$$

$$\int u^2du/\sqrt{R'^2-u^2};$$

and the absorption in this region for this case is given by

$$\begin{aligned}S^{(iii)} = {} & [HK_m/(A^3R)]\{[A^2P_3(x_2)\\ & + a_2R'^2/2][\sin^{-1} A(x_1-x_2)/R'\\ & - \sin^{-1} A(x_3-x_2)/R']\\ & + A[P_3'(x_2)+a_2(x_3-x_2)/2]\\ & \cdot\sqrt{R'^2-A^2(x_3-x_2)^2}\\ & - A[P_3'(x_2)+a_2(x_1-x_2)/2]\\ & \cdot\sqrt{R'^2-A^2(x_1-x_2)^2}\}, \quad x_1\angle x_0 \qquad (32)\end{aligned}$$

where

$$P_3(x_2) = a_0+a_1x_2+a_2x_2^2;$$
$$P_3'(x_2) = a_1+2a_2x_2.$$

When R' is greater than 0.7015, the total absorption is given by

$$S = S^{(i)}+S^{(iii)}, \quad x_1\angle x_0. \qquad (33)$$

In Case (ii), when R' is less than 0.7015, (31) with the upper limit changed to x_0 is the expression for $S^{(ii)}$. From (19),

$$x_0-x_2 = R'/A, \quad x_2\angle x_0\angle x_1;$$

hence,

$$\sqrt{R'^2-A^2(x_0^2-x_2^2)} = \sqrt{R'^2-A^2R'^2/A^2} = 0, \quad x_2\angle x_0\angle x_1;$$

and

$$\sin^{-1} A(x_0-x_2)/R' = \sin^{-1} AR'/R'A = \pi/2, \quad x_2\angle x_0\angle x_1.$$

Also,

$$\pi/2-\sin^{-1} u = \cos^{-1} u.$$

When the upper limit x_0 is substituted in (32) in place of x_1 and the above simplifications made, we have for reflection in the lower region

$$\begin{aligned}S^{(ii)} = {} & [HK_m/(A^3R)]\{[A^2P_3(x_2)\\ & + a_2R'^2/2]\cos^{-1}[A(x_3-x_2)/R']\\ & + A[P_3'(x_2)+a_2(x_3-x_2)/2]\\ & \cdot\sqrt{R'^2-A^2(x_3-x_2)^2}\}, \quad x_3\angle x_0\angle x_1. \qquad (34)\end{aligned}$$

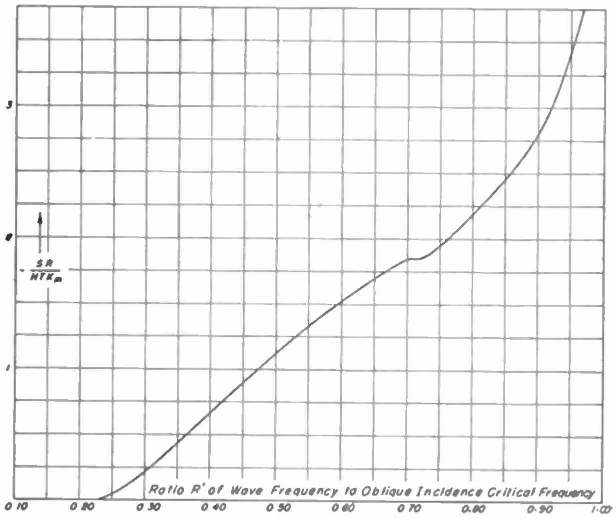

Fig. 5—Absorption as a function of R'. The quantity plotted is $-SR/(HTK_m)$. See text for notation.

Fig. 5 shows the value of $(SR)/(HTK_m)$ as a function of R'. The value of S can be found from the graph in a specific instance, knowing R, H, T, and K_m, and substituted for $\int Kds$ in (26).

Martyn's absorption theorem[6] states, in the present notation,

$$S(R',\theta_0) = (\cos\theta_0)S(R,0);$$

this can be verified by comparison of (30), (32), and (34) of this paper with (22) and (23) of footnote reference 1.

CONCLUSIONS

Approximate ray path and absorption of a radio wave obliquely incident on the ionosphere can be calculated using the parabolic approximations obtained in footnote reference 1. The accuracy of these results are comparable to those obtained at vertical incidence.

In addition to the analytic approximations to the Chapman distribution, the following approximations have been made throughout this paper: (1) plane ionosphere; (2) collisional frequency in ionosphere much less than wave frequency; (3) no magnetic field; and (4) a quasi-homogeneous ionosphere, i.e., $(du/dx \ll \omega/c)$.

ACKNOWLEDGMENTS

This analysis was carried out as a part of research being conducted at The Pennsylvania State College for the Watson Laboratories, Air Material Command, Red Bank, N. J., under Contract W28-099-ac-143. Direction of this research is by A. H. Waynick; his suggestions and criticisms were of material help in obtaining the results given here.

The Negative-Ion Blemish in a Cathode-Ray Tube and Its Elimination*

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Summary—This paper is a critical review of the widely scattered and somewhat conflicting data regarding negative ions in cathode-ray tubes and blemish formations.

The form of the ion blemish is a function of the form of focusing and scanning used, being a small central spot with electric focus and magnetic scan, and being a circle about an inch in diameter with magnetic focus and scan.

The blemish is due to chemical poisoning of fluorescence by negative ions formed on or near the cathode. The paths of such ions are substantially unaffected by the magnetic fields normally used with cathode-ray tubes. The path followed by an ion through an electric field is, however, the same as that of an electron starting from the same point of rest. Thus the ion blemish has the same shape as would the electron spot were the magnetic fields only removed from the crt. The use of a backing layer such as aluminum reduces the blemish but does not eliminate it, apparently owing to porosity of the backing layer. Such a layer has, however, other beneficial properties not related to the blemish problem.

The ion blemish can be eliminated by the use of an ion trap, which is a device usually located in or near the gun.

On the screens of certain types of cathode-ray tubes there appears, after some operation, a blemish which is particularly annoying when the tube is employed for television viewing. The form of the blemish depends upon the type of tube, but is usually a darkened area or spot with a rather well-defined boundary. In Fig. 1, the blemish spot is above the girl's left eyebrow. In the still picture it is rather unobtrusive, but when motion is present it becomes a source of considerable dissatisfaction among television set owners.

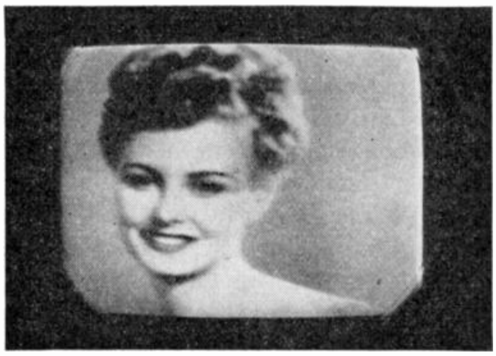

Fig. 1—Picture on the face of an electrically focused, magnetically scanned crt, showing ion blemish over the girl's left eyebrow.

This blemish is due to the localized reduction in the efficiency of light production, but is not visible under external illumination even when the bombarded side of the screen is examined. The blemish can take several forms, the form being determined by the focusing and scanning means employed. It will be shown later that most ions originate on or near the cathode, traverse the same paths as electrons through electrostatic fields, but are substantially unaffected by the magnetic fields used with cathode-ray tubes. Thus a tube employing an electrostatically focused gun and magnetic deflection usually develops, frequently in less than an hour of operation, a small blemish of the general shape and size of an undeflected spot. This is the most objectionable case. If, however, the tube is magnetostatically focused and employs magnetic scan, the resulting blemish has the same size and location as the undeflected "spot" seen with the focusing magnetic lens removed. In this case, the "spot" is a "shadow" of the limiting aperture in the gun, and is usually about an inch in diameter. This form of blemish develops more slowly because the ions are spread over a greater area. It appears after perhaps a hundred hours of operation and is particularly objectionable because of its well-defined boundary.

In a tube employing electrostatic focus and electric deflection, the electrons and ions are both focused into spots which are then scanned about the screen in the same way. The very-slowly forming ion blemish is therefore indistinguishable from the gradual loss in efficiency of the phosphor due to electron bombardment. It is possible to conceive of various other combinations of focusing and scanning means. The form of blemish which might develop can be predicted for each, but because of the lack of commercial importance, this will not be done here.

There are certain generally observed characteristics of the ion blemish. Its prominence decreases with increasing beam voltage. The potential at which it is no longer regarded as objectionable varies with the screen material, being about 12 to 15 kv for the sulfides. The rate of formation for a fixed viewing potential appears to be a decreasing function of beam voltage, also. Thus a tube operated for a short time at 1000 volts will have developed more blemish when viewed at 6 kv than it would have had it been operated for the same period at 6 kv.

The various screen materials have widely differing sensitivities to ion deterioration. Willemite is relatively insensitive to this effect, while zinc sulfide and zinc cadmium sulfide are highly sensitive. This fact points to the chemical poisoning of fluorescence by the ions as the cause of the ion blemish. In general, the susceptibility to ion blemishing appears to be related to the susceptibility of the phosphor to poisoning by impurities during manufacture.

Willemite, manganese-activated zinc orthosilicate, is

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* Decimal classification: R371.5. Original manuscript received by the Institute, May 4, 1948; revised manuscript received, July 16, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 23, 1948.

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a compound requiring a relatively large concentration of activator (~1 per cent) and being relatively insensitive to poisoning of fluorescence by impurities such as nickel. The sulfides, which are usually activated with silver or copper, require activator concentrations of 0.01 per cent or less and are correspondingly more susceptible to poisoning of fluorescence by impurities.

Although the ion blemish has been ascribed by Sharpe[1] to the deposition of material on the surface of the phosphor crystals, which material slows the impinging electrons, the poisoning theory is much more tenable. To reduce the fluorescence the layer would have to have appreciable thickness. Data[2] reported for aluminum indicate that at 2 kv approximately 1000 molecular layers are required to reduce the beam power 20 per cent. If, as will be shown later, the ions are chiefly oxygen, it is difficult to see how a sufficient thickness could be retained. Furthermore, Bachman and Carnahan[3] report that the negative ion blemish can be "developed" by immersing the face removed from a cathode-ray tube in photographer's "hypo." This points to some sort of chemical change rather than to a film formation. It appears, however, that no explanation has been advanced in terms of chemistry of the solid state.

Nature and Origin of the Ion Beam

Several notable papers[3–5] have appeared dealing with the nature and origin of the negative-ion beams. The investigations reported involved mass-spectrographic analyses of crt beams. In two cases,[3,4] standard or substantially standard cathode-ray tubes were employed as mass spectrographs by replacing the customary scanning coil with a strong electromagnet, and using the poisoning effect of the negative ions upon the screen to produce the records. The combined findings appear in Table I. In properly processed tubes, the predominant ion is O_2^-. The two chlorine ions appear to be abundant also during early life, but substantially disappear in a few hours of operation.

Both the origin of the ionizable material and the mechanism of ion formation are subject to some uncertainty. Three mechanisms appear to be involved, although several others have been considered and ruled out. These three are: (1) emission of ions as such from the cathode; (2) ejection of ions from the cathode and grid by impact of positive ions formed by the electron beams; and (3) ion formation by attachment of electrons from the beam to gas molecules.

I. EMISSION OF NEGATIVE IONS

With no exceptions, in a well-processed tube all negative ions appear to have their origins at or very near the cathode (or grid),[6–8] as the ion blemishes are well focused in electrically focused tubes. In certain magnetically focused tubes it is possible, with the magnetic lens removed, to form an electron image of the cathode surface on the screen. In such a tube Liebman[7] showed that this electron image agrees exactly in detail with the ion blemish, except for a few additional features.

Schaefer and Walcher[5] conclude from such observations as the independence of the O_2^- ion current on gas pressure that oxygen is emitted as ions. Broadway and Pearce[4] observed, however, that O^- ion spot formation is reduced by reduction of oxygen pressure. This, and the relatively poorer focus of the O^- spot, they took as evidence that oxygen is not emitted as ions. Both may be right. The former had cathodes operated under adverse conditions with high electrolytic conduction, apparently conducive to a surface reaction leading to O_2^- emission. The well-formed cathodes of the latter apparently emitted uncharged oxygen which contributed to the gas eventually ionized by electron attachment. Note that the oxygen ions reported are different in the two cases. Both papers agree, however, that chlorine is emitted as an ion.

Regardless of the exact mechanism of ionization, however, oxygen appears to originate from the well-known slow decomposition of the oxides of the cathode during operation. Chlorine may originate as an impurity in the cathode materials, although recently Hamaker, Bruining, and Aten[9] have reported evidence that it is due to the following chemical reaction in the glass during baking on exhaust:

$$2NaCl + H_2O + SiO_2 \rightarrow Na_2SiO_3 + 2HCl.$$

The NaCl is a common impurity, while H_2O is always present in commercial glass. The HCl reacts with BaO to yield $BaCl_2$ and H_2O. The former decomposes to yield Cl^- ions.

II. NEGATIVE-ION FORMATION BY POSITIVE-ION BOMBARDMENT

It has been shown that positive-ion bombardment of a surface usually yields spectra of negative ions much like those observed in cathode-ray tubes.[10–14] Bachman[6]

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[1] J. Sharpe, "The ion trap in cr tubes," *Electronic Eng.*, (London) pp. 385–386, December, 1946.

[2] D. W. Epstein and L. Pensak, "Improved cathode ray tubes with metal-backed luminescent screens," *RCA Rev.* Vol. 7, pp. 5–8; March, 1946.

[3] C. H. Bachman, and C. W. Carnahan, "Negative ion components in the cathode-ray beam," PROC. I.R.E., vol. 26, pp. 529–539; May, 1938.

[4] L. F. Broadway, and A. F. Pearce, "Emission of negative ions from oxide cathodes," *Proc. Phys. Soc.*, (London), vol. 51, pp. 335–348; 1939.

[5] Von H. Schaefer, and W. Walcher, "Negative ions in braun tubes and their relation to the oxide cathode mechanism," *Zeit. Phys.*, vol. 121, pp. 679–701; 1943.

[6] C. H. Bachman, "Ring focusing of negative ions in a cathode-ray beam," *Jour. Appl. Phys.*, vol. 11, pp. 83–85; January, 1940.

[7] G. Liebman, "Origin of ion burn in cathode-ray tubes," *Nature*, vol. 157, p. 228; February 23, 1946. Also, *Elec. Eng.*, vol. 18, pp. 289–290; September, 1946.

[8] H. A. Barton, "Negative ion emission from oxide coated filaments," *Phys. Rev.*, vol. 26, pp. 360–363; 1925.

[9] H. C. Hamaker, H. Bruining, and A. H. W. Aten, Jr., "On the activation of oxide-coated cathodes." *Philips Res. Rep.*, vol. 2, pp. 171–176; 1947.

[10] J. S. Thompson, "A new method of producing negative ions," *Phys. Rev.*, vol. 38, p. 1389; October, 1931.

[11] K. S. Woodcock, "The emission of negative ions under the bombardment of positive ions," *Phys. Rev.*, vol. 38, pp. 1696–1703; November 1, 1931.

was able to explain the ring-shaped spots previously observed[3] in certain purposely gassy tubes in terms of positive-ion bombardment of the grid aperture. The resulting negative-ion emission from the grid produced the ring-shaped ion blemishes observed for the oxygen. This theory was confirmed by notching the grid aperture and observing the "notch" in the ion-blemish circle, and also by noting grid-aperture erosion in dissected tubes. The heavy ions such as CaO^-, Ni, and BaO_2^- very possibly are due to positive-ion bombardment of the cathode surface.

III. Negative-Ion Formation by Electron Attachment

From gaseous-discharge research it is well known that electron attachment takes place only to electronegative molecules such as oxygen, and becomes relatively improbable at electron velocities above those corresponding to a few volts. Hence, attachment must occur very near the cathode. Approximate calculations of ion current by this process[4] yield low values, but do not rule out the method. As attachment to molecular oxygen may result in dissociation into O^- and O; the O^- found by Broadway and Pearce[4] very probably was produced by this attachment process.

The organic-vapor ions may be attributed to the breaking down of the cellulose-nitrate binder used in cathode preparation,[3] or to the presence of grease vapor as evidenced by the great number of such ions found by Schaefer and Walcher,[5] whose tube could not be baked out and had greased joints.

Blemish Elimination

It can be seen that the negative ions responsible for the blemish are chiefly O_2^- and Cl^- originating at or near the cathode. They are undeflected by magnetic fields employed with cathode-ray tubes, and reduce the low-voltage (<10,000 v) fluorescent efficiency of the screen material. To eliminate this defect, three approaches have been used. The first is to attempt to eliminate substances likely to form negative ions from the important parts of the tube by special care in parts preparation, followed by rigorous exhaust. In some cases, "getters" activated during operation have been employed to take up gas which might form ions. A hot tantalum filament in parallel with the cathode heater, and zirconium on gun parts, have been employed. It has been the author's experience, however, that such methods merely shift the blemish from an item of manufacturing shrinkage to a field complaint.

Backing Layer As Ion Filter

An interesting approach is to protect the screen with a filter relatively pervious to electrons, but impervious to ions. It is known[15] that the depth of penetration of a particle into a substance increases with particle velocity and decreases in proportion to particle mass. This indicates a strong discrimination by particle mass.

The application of a thin metallic backing layer has received considerable attention, primarily for three reasons other than the elimination of negative-ion blemishes.[16–20] First, an unbacked screen is brighter on the bombarded side because the light-producing interactions occur nearer that side. If a very thin metallic film is "stretched" relatively smoothly over the phosphor grains on the bombarded side of the screen, the brightness of the viewed side may be increased, provided that the loss of energy by the electrons in passing through the film is not too great. This is because the metal film reflects the back-directed light. In practice, the anode potential at which the brightness of a backed screen just equals that of a similar but unbacked screen is a function of backing-layer thickness and is of the order of a few kilovolts. Bachman[18] reports a value of 5 kv, which is presumed to apply for a tube intended for operation at 10 kv.

The second reason for the backing is the improvement of contrast by the elimination of the stray light from the bombarded side of the screen reflected from the inside walls of the tube. The third reason has to do with the provision of a conductive return path for the beam current; an unbacked screen relies upon secondary emission from the screen material to provide the path for the electron current in the beam to the second anode. The screen will not operate at a potential above which the secondary-to-primary current ratio drops below unity, which frequently occurs below 15 kv.

Means for obtaining backing layers are described by Law,[19,22] Schaefer,[20] and Bramley.[21] The aluminum backing currently in commercial use, while providing the advantages mentioned above, appears only to reduce the rate of ion-blemish formation.

Negative-Ion Trap

The third approach to the elimination of the negative-ion blemish is the use of the negative-ion trap, which separates the negative ions from the electrons in the

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[12] R. H. Sloane, and R. Press, "Formation of negative ions by positive ion impact on a surface," *Proc. Roy. Soc.*, vol. A168, pp. 284–300; 1938.

[13] F. L. Arnot, and Clark Beckett, "A new process of negative-ion formation IV," *Proc. Roy. Soc.*, vol. A168, pp. 103–122; 1938.

[14] R. H. Sloane, and Eliza Cathcart, "Formation of negative ions by negative-ion bombardment of surfaces, a new process," *Nature*, vol. 143, pp. 474–475; March 18, 1939.

[15] Beth, *Ann. der Phys.* vol. 5, p. 374; 1930.

[16] M. Von Ardenne, British Patent No. 402,411, accepted November 21, 1933.

[17] Kurt Schlesinger, U. S. Patent No. 2,209,639, issued February 4, 1936.

[18] C. H. Bachman, "Image contrast in television," *Gen. Elec. Rev.*, vol. 48, pp. 13–19; September, 1945.

[19] R. R. Law, U. S. Patent No. 2,233,786, issued March 4, 1941.

[20] Vincent J. Schaefer, U. S. Patent No. 2,374,311, issued April 24, 1945.

[21] Art Bramley, "Aluminum backed phosphor screen in cathode ray tubes," The Electrochemical Society Preprint 91–30; Meeting, April 9–11, 1947.

[22] R. R. Law, U. S. Patent No. 2,303,563, issued December 1, 1942.

beam. As has been pointed out by several authors,[3,4,23] in several different ways, the path followed by any charged particle through any purely electrostatic field is independent of mass-to-charge ratio, provided that all such particles start from rest at the same point. If a magnetostatic field is present, the path of a particle is not independent of its mass-to-charge ratio. This latter principle is employed in the ion trap.

There are three features associated with the trap or its operation. First, the charged particles are at least partially formed into a beam before reaching the trap. By this it is meant that by the time the ions and electrons reach the vicinity of the trap, their trajectories should form a bundle, the cross section of which does not change greatly in a length equal to its diameter. Second, the beam is subjected to a magnetic field having a component perpendicular to the beam length. Third, the ions are disposed of in a manner not to obstruct the passage of the electrons to the useful part of the screen.

The first form of ion trap with which the author had experience[24] is that shown in Fig. 2 and Fig. 3, in which the gun is aimed at the edge of the screen and the vertical scanning coil is provided with strong, steady bias. It is obvious that the beam must be bent through a rather large angle by the steady component of the magnetic field in order that the ion blemish will fall outside of the scanning raster. This results in considerable spot distortion. The tube[24] of Fig. 4 proved much more satisfactory from that standpoint, but presents manufacturing difficulties because of the nonaxial symmetry of the envelope. However, quite a number of such tubes have been made. This manufacturing disadvantage can be overcome by bending the gun in such a way that it will fit in a straight neck, as shown in Fig. 5. In this case, bending occurs before the beam electrons reach final velocity. A version of this arrangement, employing a magnetic lens, was used by Philco Radio and Television Corporation before the war.

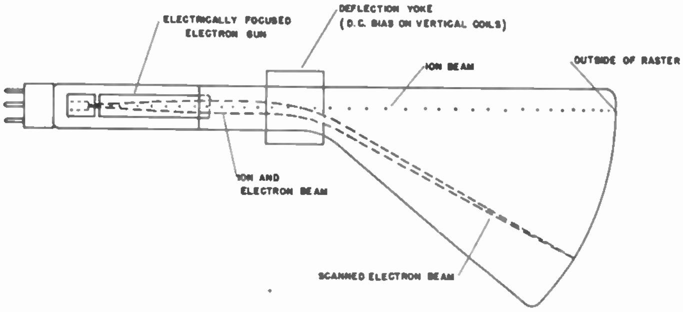

Fig. 2—An early ion-trap tube of the bent-beam variety, requiring excessive beam bending.

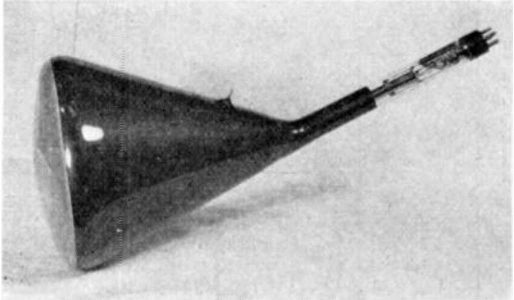

Fig. 3—Photograph of a tube of the type shown in Fig. 2.

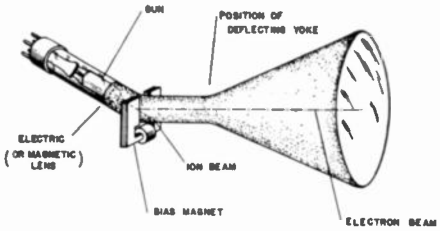

Fig. 4—Bent-neck variety of ion-trap tube employing the bent-beam principle.

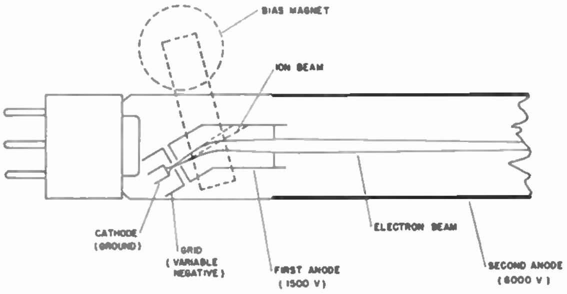

Fig. 5—Bent-gun variety of ion trap employing the bent-beam principle.

H. Branson[25] described the gun shown in Fig. 6, in which the beam is focused and magnetically deflected so as to pass through a slightly off-center hole in the end of the extended second-anode cylinder.

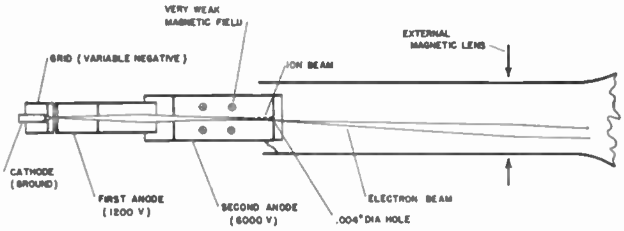

Fig. 6—The Branson[25] ion-trap gun.

In England, a type of crt employing an ion-trap gun due to Woodbridge[26] is manufactured by Electronic Tubes Ltd. The trap and gun are shown schematically in Fig. 7. Note that the cathode, the grid aperture, and the anode aperture are eccentric. The resulting field causes the entire beam to be deflected in the opposite direction to that which one might at first expect. At the

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[23] H. Busch, and E. Bruche, "Beitrage zur Elektronenoptik," Johann Ambrosius Barth; Leipzig, p. 34; 1937.

[24] R. M. Bowie, U. S. Patent Nos. 2,211,613 and 2,211,614.

[25] H. Branson, U. S. Patent No. 2,274,586, issued February 24, 1942.

[26] Leonard A. Woodbridge, British Patent Pending.

place where the beam crosses the gun axis, a magnetic field is provided which directs the electron beam along the axis, while the ions proceed to the stop in the anode.

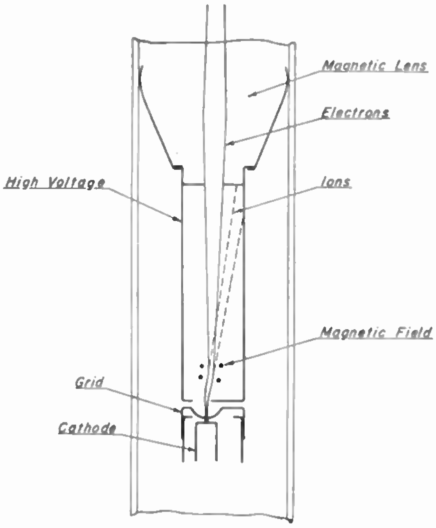

Fig. 7—The Woodbridge ion-trap gun.

The ion trap described above may be classed as of the bent-beam variety. Another class is that in which the electron beam is substantially unbent. This is accomplished by applying a transverse electric field with the magnetic field and of such a strength as to compensate the tangential force exerted by the magnetic field on the electrons. Such a trap was described by Bowie,[24] and is shown in Fig. 8. The transverse electric field bends the ion beam to one side, causing it to impinge upon the baffle, while the magnetic field counteracts the bending force on the electrons. A modification[27] of this unbent-beam trap is currently used in the type 10PB4[28] crt shown schematically in Fig. 9. The trap is so incorporated in the gun that no extra potential need be applied to the tube itself. By tilting the slot between grid No. 2 and the adjacent end of the anode, a component of electric field transverse to the gun axis is obtained. This component substantially compensates the action of the magnetic field upon the electrons. However, as the two fields do not compensate point-by-point along the axis, the electron beam deviates somewhat from the axis and is brought back by slightly overcompensating the electric field with one magnetic field, and then correcting the direction subsequently by means of a weak magnetic field in the opposite direction.

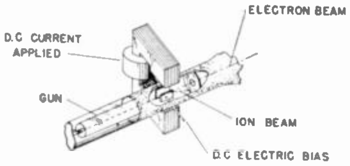

Fig. 8—Ion trap employing the undeflected-beam principle.

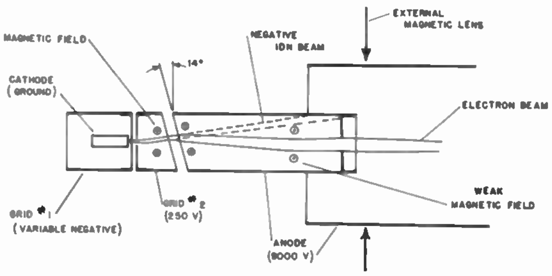

Fig. 9—Ion-trap gun of the type employed in the 10BP4, using the undeflected-beam principle.

CONCLUSIONS

The widespread use of magnetically deflected, directly viewed cathode-ray tubes operating below 10 kv in high-quality receivers has again pointed up the negative-ion-blemish problem which had received considerable attention before the war. The art has now progressed to the point at which the mechanisms of formation of the negative ions producing the spot are reasonably well understood, while means for reducing and for preventing the formation of the blemish are known.

ACKNOWLEDGMENT

The author wishes to thank G. D. O'Neill for his helpful suggestions in the preparation of this manuscript.

TABLE I

NEGATIVE IONS IN CRT BEAMS AS REPORTED BY BACHMAN AND CARNAHAN,[3] BROADWAY AND PEARCE,[4] AND SCHAEFER AND WALCHER[5]

Ion Mass	Bachman and Carnahan	Broadway and Pearce	Schaefer and Walcher, oxide cathode	Schaefer and Walcher, tungsten cathode
1		H	H strong	H strong
12	C	C		
13			CH_1 strong	
14			CH_2 strong	
16	O or CH_4	O		
17			OH strong	
18	OH_2		OH_2 weak	
19	F			
23			Na weak	
24			C_2 strong	
25			C_2H medium	
26	C_2H_2	CN-C_2H_2	C_2H_2 medium	C_2H_2 weak
30	NO			
32	O_2	O_2	O_2 strong	O_2 weak
33			?	
34			?	
35	Cl	Cl	Cl medium	
37	Cl	Cl	Cl medium	
40	Ca		Ca weak	
42		CNO-C_3H_6	? weak	
43			? weak	
48			? weak	
56	CaO		CaO weak	
58			Ni weak	
60			Ni weak	
62			? weak	
68			? weak	
74	$Ca(OH)_2$			
80		Br		
101	$CaCO_3$			
120	SrO_2			
127		I		
169	BaO_2			
261	$Ba(NO_3)_2$			

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[27] "Television receivers in mass production," *Electronics*, vol. 20, pp. 86–91; June, 1947.

[28] See RCA registration of crt type 10BP4, RMA Data Bureau, release no. 482, April 15, 1946; and of focusing coil, deflection yoke, and ion-trap magnet, release no. 661, May 13, 1948.

The Patterns of Slotted-Cylinder Antennas*

GEORGE SINCLAIR†, SENIOR MEMBER, IRE

***Summary*—A method is described for calculating the patterns of arrays of axial-slot antennas mounted on the surface of a metallic cylinder. The method of calculation gives information on the relative phase of the radiated field in addition to the amplitude. Measurements have been made for certain of the antennas to verify the accuracy of the calculations. A number of calculated patterns are shown to indicate the extent of the control of the pattern which can be obtained by using arrays of slots on a cylinder.**

INTRODUCTION

ANTENNAS WHICH USE the radiating properties of slots in the surface of a metal cylinder are being used to an increasing extent for vhf and uhf applications. Slotted-cylinder antennas are particularly useful when it is desired to produce a horizontally polarized field with a horizontal pattern which is essentially circular. If the diameter of the cylinder is not too large, a single slot parallel to the axis produces a pattern which is usually sufficiently near a circle for most purposes. As the diameter of the cylinder is increased, the antenna becomes directional with a pattern which may have pronounced minima. While these directional properties can sometimes be utilized to secure more advantageous coverage of a given area, they are generally undesirable. It is, therefore, important to be able to predict the amount by which the pattern departs from a true circle.

As the number of stations using the higher frequencies increases, it is to be expected that there will be an increased demand for antennas with controlled directional patterns. Alford[1] has shown that a certain amount of control of the pattern of a slotted-cylinder antenna can be achieved by putting wings on the slots, but the amount of control which can be obtained is limited. Another method for modifying the pattern consists of using a number of slots spaced around the periphery of the cylinder. If the slots are equispaced around the periphery and fed in phase with equal amounts of power, the pattern can be made more nearly circular.[2] By feeding power of different amounts and different phases to the slots, a wide variety of patterns can be obtained.

Some data have been published on the patterns of slot antennas.[1,3] However, in order to design antennas consisting of arrays of slots, it is necessary to have available information on the variations in phase of the radiated field, in addition to information on the amplitude of the field. While it is possible to obtain this information from either full-scale or model measurements using well-known techniques, such a procedure is generally impractical because the necessary measuring equipment is not available to the antenna designer. Calculation of the pattern is a much more satisfactory procedure, and can be carried out quite readily.

CALCULATION OF THE PATTERNS OF SLOT ANTENNAS

The exact calculation of the field radiated by a slotted-cylinder antenna is a difficult boundary-value problem. However, the antenna designer is generally most interested in the horizontal pattern of a vertical slotted-cylinder antenna, and fortunately the calculation of this pattern is a much simpler problem. It can be shown that, when the breadth of the slot is small in comparison with the wavelength, the horizontal pattern is independent of the axial distribution of the field along the slot. By assuming the simplest possible axial distribution, namely, one in which the field is uniformly distributed in a slot which is infinitely long, an expression for the pattern can be obtained in the form of a Fourier series.[4]

In the following, only the horizontal patterns of vertical slotted-cylinder antennas will be considered. However, it is also possible to make approximate calculations of the vertical patterns, if certain simplifying assumptions are made.[1,3] Jordan and Miller[3] have shown that the vertical pattern of a vertical slotted-cylinder antenna is approximately the same as the corresponding pattern for a similar slot in a perfectly conducting plane sheet of infinite extent. This pattern can be approximated by calculating the H-plane pattern of the field radiated from the open end of a rectangular waveguide.[5]

PATTERN OF A SINGLE SLOT IN A CYLINDER

Fig. 1 shows the co-ordinate system employed in the calculation of the field, and also illustrates the notation used. The axis of the cylindrical surface coincides with the z axis of a cylindrical co-ordinate system. The slot is assumed to be parallel to the z axis, and located at $\phi=0°$ on the surface of a cylinder of diameter $D=2a$, a

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* Decimal classification: R326.7. Original manuscript received by the Institute, May 10, 1948; revised manuscript received, July 19, 1948. Presented, National Electronics Conference, Chicago, Ill., November, 1947; and IRE Canadian Section Conference, Toronto, Canada, May 1, 1948.

A co-operative research contribution of the United Broadcasting Company, Cleveland, Ohio, and The Ohio State University Research Foundation, Columbus, Ohio. The research described was conducted in the Antenna Laboratory of the department of electrical engineering, the Ohio State University.

† Formerly, The Ohio State University Research Foundation; now, University of Toronto, Toronto, Ontario, Canada.

[1] A. Alford, "Long Slot Antennas," *Proc. Nat. Electronics Conf.*, Chicago, Ill., pp. 143–155; 1946.

[2] H. J. Riblet, "Microwave omnidirectional antennas," PROC. I.R.E., vol. 35, pp. 474–478; May, 1947.

[3] E. C. Jordan and W. E. Miller, "Slotted-cylinder antennas," *Electronics*, vol. 20, pp. 90–93; February, 1947.

[4] G. Sinclair, E. C. Jordan, and E. W. Vaughan, "Measurement of aircraft-antenna patterns using models," PROC. I.R.E., vol. 35, pp. 1451–1462; December, 1947.

[5] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., 1943; p. 359.

being the radius. In the slot the field is assumed to be uniformly distributed axially, and polarized so that there is only an E_ϕ component. Thus, at the surface $\rho = a$, the field is assumed to be of the form

$$E_\phi = E_0 \epsilon^{j\omega t} \quad \text{for } -\delta < \phi < \delta \tag{1a}$$

$$E_\phi = 0 \quad \text{for other values of } \phi \tag{1b}$$

where 2δ is the angular width of the slot in radians. It is assumed that the width of the slot is small, in comparison with the wavelength and with the diameter of the cylinder.

For these assumptions, it can be shown[4] that the field produced at a fixed large distance from the cylinder is a Fourier series

$$E_\phi \approx \frac{A}{j\pi}\left[a_0 + \sum_{n=1}^{\infty} a_n \cos n\phi\right] \tag{2}$$

where

$$a_0 = \frac{1}{2H_0^{(2)\prime}(ka)}$$

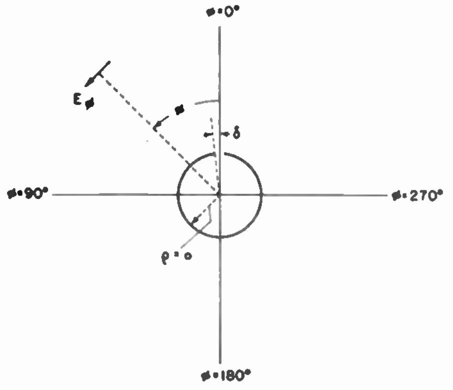
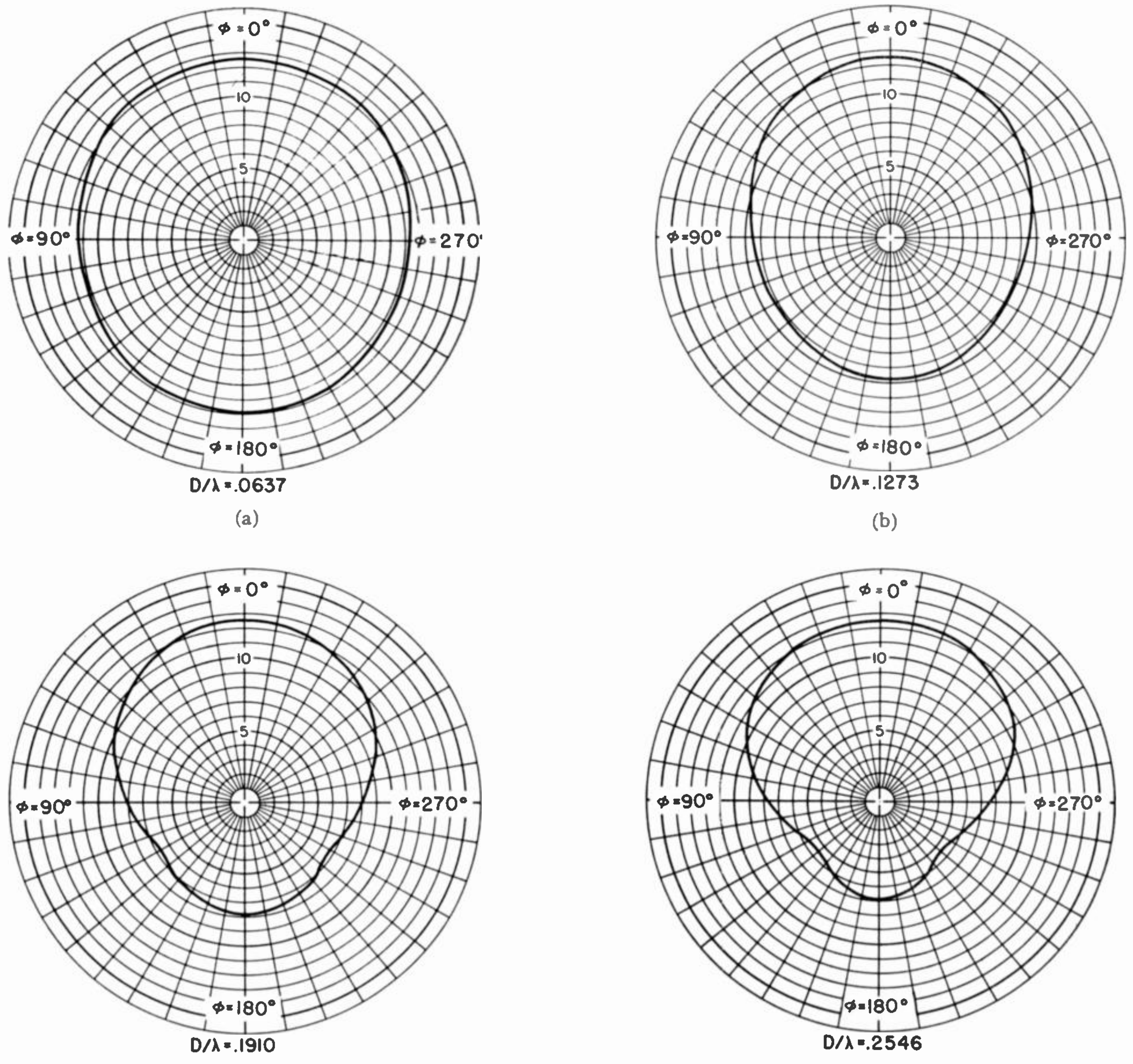


Fig. 1—Diagram showing the co-ordinate system.



D/λ=.0637

(a)

D/λ=.1273

(b)

D/λ=.1910

(c)

D/λ=.2546

(d)

Fig. 2—Patterns for a single slot in a cylinder. The slot is located at $\phi = 0°$.

$$a_n = \frac{j^n}{H_n^{(2)\prime}(ka)}$$

$$ka = \frac{2\pi a}{\lambda}$$

λ = wavelength measured in same units as a.

The parameter A depends on the exciting voltage, and the distance from the antenna, but does not depend on the azimuth angle.[4] For calculating relative patterns, the value of A need not be known.

In (2), the coefficients a_n are complex, to take account of the phase variations in the field. The reference for phase in (2) is at the axis of the cylinder. The azimuth angle ϕ is measured from a reference line through the center of the slot (see Fig. 1).

Amplitude and Phase Patterns for a Single Slot

Amplitude and phase patterns have been calculated, using (2), for a series of slotted-cylinder antennas of various diameters. Some of the amplitude patterns are shown in Fig. 2, and relative phase patterns in Fig. 3.

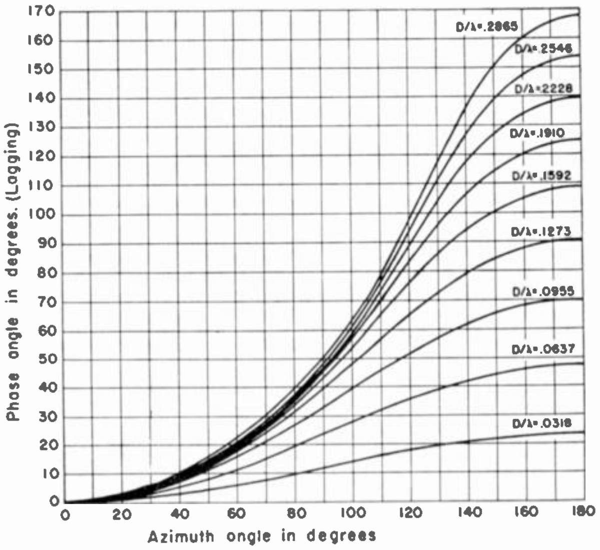
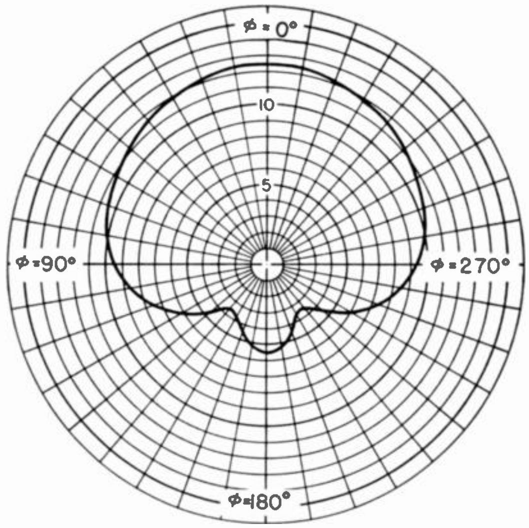


Fig. 3—Curves showing the variation in relative phase of the field with azimuth for a single slot in a cylinder. The slot is located at $\phi = 0°$ in each case.

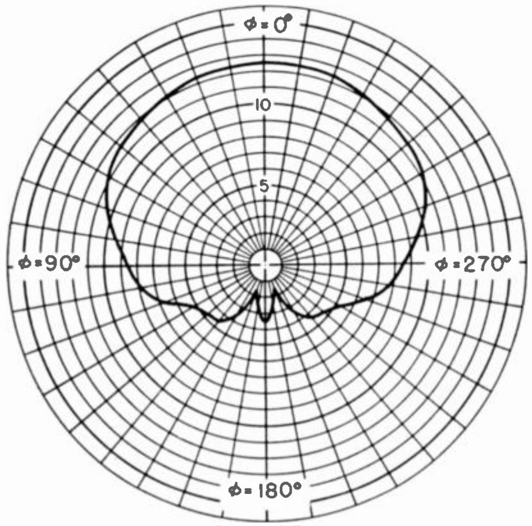
The patterns in Fig. 2 cannot be used to compare the gains obtained with the various antennas, since (2) does not give any information on gain, due to the assumptions made in deriving the equation. Hence, the patterns in Fig. 2 have been plotted to the same maximum value of relative field intensity in each case. The phase patterns in Fig. 3 show the phase of the field in a given direction, relative to the field at the same distance in the direction $\phi = 0°$. The absolute phase computed from (2) has little significance for practical antennas, because of the large phase shifts which are unavoidably introduced by most feeding systems.

In order to verify the accuracy of the calculations of relative phase patterns, measurements were made on a slotted-cylinder antenna having a diameter of 0.1910



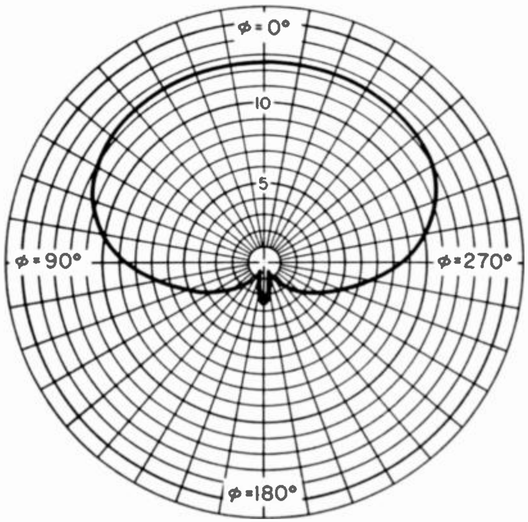
D/λ = 0.5

(a)



D/λ = 1.25

(b)



D/λ = 8.0

(c)

Fig. 4—The patterns of slotted-cylinder antennas having diameters large in terms of the wavelength. Slot is located at $\phi = 0°$.

wavelength. The measurements were made for every 20° of azimuth. The maximum deviation of the measured values of phase from the calculated curve was ±5°.

SLOTTED CYLINDERS OF LARGE DIAMETER

Three patterns have been calculated for cylinders whose diameters are comparatively large in terms of wavelength. The patterns are shown in Fig. 4, and are for diameters of one-half, one and one-quarter, and eight wavelengths. It is apparent from these patterns that the effect of increasing the diameter is to reduce the signal in the region of the shadow cast by the cylinder. For cylinders of large diameter, the pattern is always approximately a cardioid.

THE PATTERNS OF ARRAYS OF SLOTS

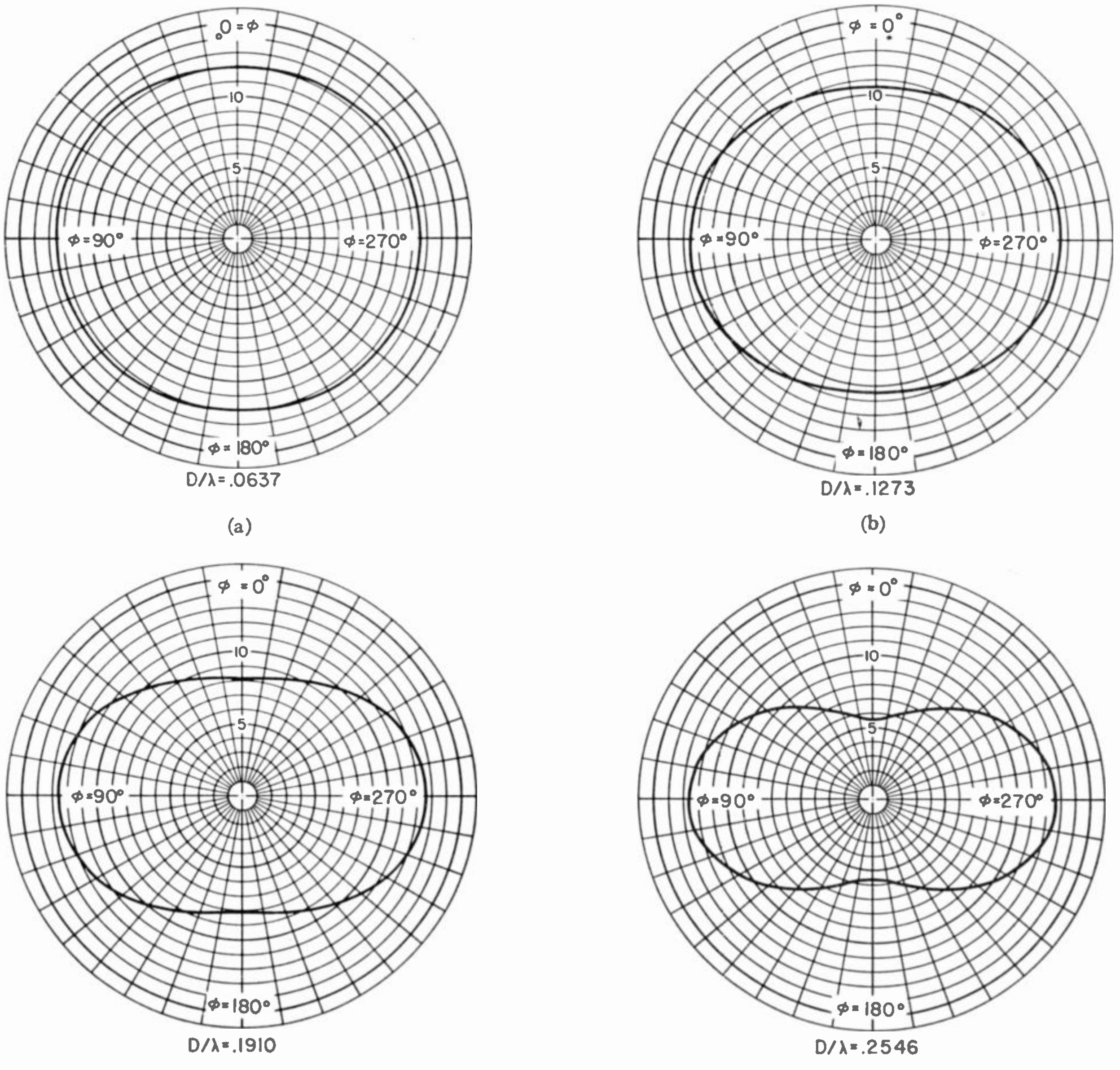
The patterns of arrays of slots can be computed by suitably combining, in the proper phases, the fields from the various slots, each calculated using (2). Since the reference for phase for each slot is at the axis of the cylinder, it is not necessary to introduce any phase factor involving the distance of each slot from the axis. Hence, the pattern of the array is found by simply superimposing the fields of each of the slots, taking due account of the relative amplitudes and phases of the power being fed to each slot.

Consider the situation when two slots are used, slot No. 1 being located at $\phi=0°$, and slot No. 2 at $\phi=\phi_2$. The field radiated by slot No. 1 is given by

$$E_{\phi_1} = \frac{A}{j\pi}\left[a_0 + \sum_{n=1}^{\infty} a_n \cos n\phi\right]. \qquad (3)$$

Assume that slot No. 2 is fed in such a way that the field it produces is given by

$$E_{\phi_2} = \frac{MA}{j\pi}\left[a_0 + \sum_{n=1}^{\infty} a_n \cos n(\phi_2 - \phi)\right] e^{j\psi} \qquad (4)$$



(a) (b)

(c) (d)

Fig. 5—Patterns for arrays of two diametrically opposed slots, fed equally and in phase. Slots are located at $\phi_1=0°$ and $\phi_2=180°$.

where M is the ratio of the amplitude of the field radiated in a given direction by slot No. 2 to the amplitude radiated by slot No. 1 in a similar direction from slot No. 1, and ψ is the time phase difference between these fields. If V_1 is the voltage fed to slot No. 1, and V_2 the voltage fed to slot No. 2, then

$$V_2 = Me^{j\psi}V_1. \tag{5}$$

The field produced by the array of the two slots is the sum of the fields in (3) and (4). The patterns of arrays of several slots can be obtained in a similar fashion.

AMPLITUDE AND PHASE PATTERNS FOR ARRAYS OF SLOTS

To illustrate some of the patterns which can be obtained from arrays, patterns have been calculated for arrays of two slots, the slots being located at opposite ends of a diameter and fed equally and in phase. Such arrays are of interest in that they are sometimes suggested as a means for obtaining patterns which are more nearly circular than the patterns for a single slot.

The patterns for the two-slot arrays are shown in Fig. 5, for the same diameters of cylinders used incomputing the patterns in Fig. 2. The relative phase patterns are shown in Fig. 6.

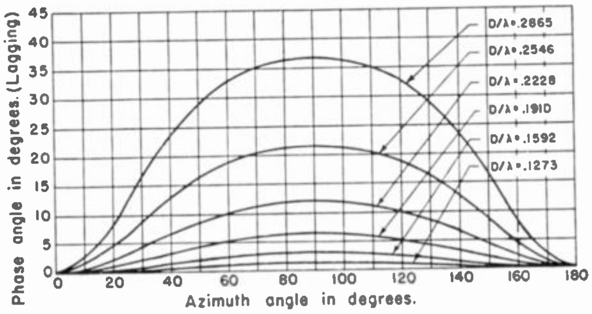


Fig. 6—Curves showing the variations in relative phase of the fields of arrays of two slots.

The ratios of the maximum to minimum field intensities in the patterns of Fig. 5 have been plotted in Fig. 7 as a function of cylinder diameter, for comparison with the corresponding ratios for a single slot. It is apparent that, for most diameters, two slots give some improvement in pattern over one slot. The improvement in phase is quite marked, as can be seen from Fig. 6. Fig. 7 also shows the maximum-to-minimum ratio for arrays of three slots, equispaced and fed equally and in phase. When four slots spaced 90° apart around the cylinder, are employed, the improvement in the maximum-to-minimum ratio is substantial for the range of diameters in Fig. 7.

Fig. 9 shows some patterns of arrays of two slots on a cylinder one-half wavelength in diameter. The pattern in Fig. 9(a) is for two slots diametrically opposed and fed equally and in phase, so that $\phi_2=180°$, $M=1$, and $\psi=0°$. Fig. 9(b) shows the pattern obtained when the phase of one of these slots is reversed ($M=1$ and $\psi=180°$).

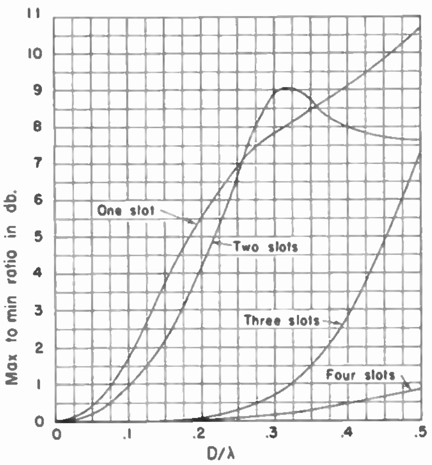


Fig. 7—Maximum-to-minimum ratio for the patterns of arrays of slots in cylinders.

It is apparent from the patterns in Fig. 9(a) and (b), and also from an examination of the manner in which (3) and (4) depend on ϕ, that when two diametrically opposed slots are used, the pattern must be symmetrical about the diameter through the slots. However, when the angle ϕ_2 between the slots is other than 180°, there will be, in general, no line of symmetry to the pattern. For example, with the arrangement of slots shown in Fig. 8, where $\phi_2=90°$, $M=0.5$, and $\psi=90°$, the pattern of Fig. 9(c) is obtained. Hence, in designing an antenna to produce a given asymmetrical pattern, it may be possible to achieve the desired pattern with only two slots.

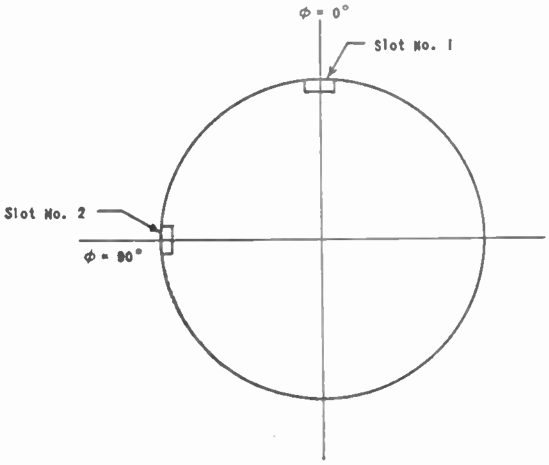


Fig. 8—Diagram showing the location of the slots for the pattern in Fig. 9(c).

ADJUSTMENT OF SLOTTED-CYLINDER ARRAYS

It is necessary to use a trial-and-error method in adjusting antenna arrays of this type to produce a given calculated pattern. In certain special cases it is easy to design a feed system to produce the calculated pattern, as, for example, when the slots are fed in phase with equal amounts of power.[4] Measurements have been made to prove it is possible to obtain the calculated patterns with properly designed feed systems, as shown in Figs. 9(a) and (b).

There seems to be very little information available on the impedance properties of slotted-cylinder antennas. Jordan and Miller have published some data on the impedances of single slots.[3] When two or more slots are used, there will be mutual impedances between the slots. No data on mutual impedances between slots on cylinders has appeared in the literature as yet.

CONCLUSIONS

It is apparent that considerable control over the patterns of slotted-cylinder antennas can be obtained by using arrays of slots. By using the expression in (2) for the field radiated from a single slot, it is possible to calculate, with adequate accuracy, the horizontal patterns of such arrays.

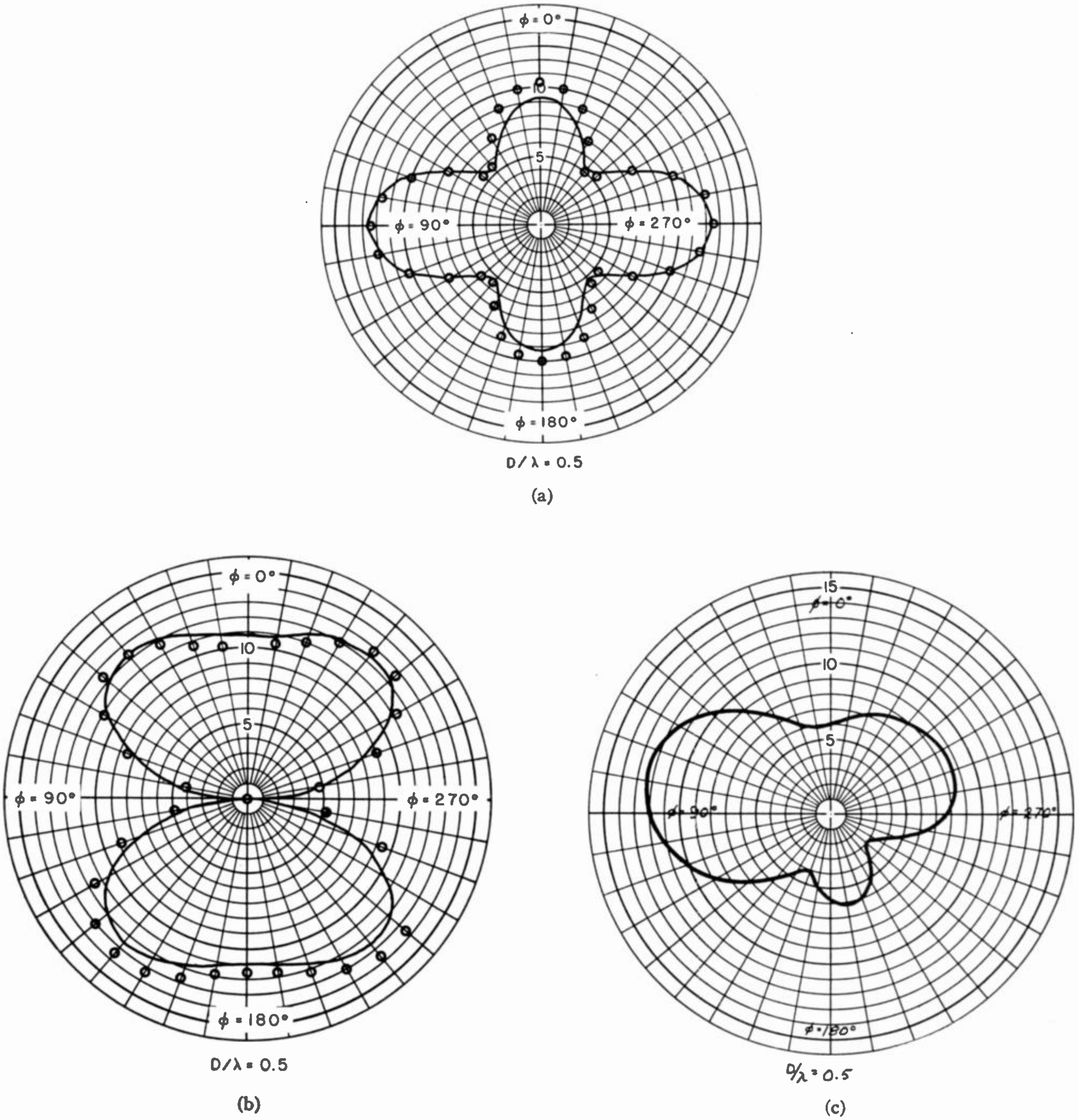


Fig. 9—Patterns of two slots in a cylinder one-half wavelength in diameter.
(a) Diametrically opposed slots fed equally and in phase.
(b) Diametrically opposed slots fed equally and out of phase.
(c) Slots at $\phi_1 = 0°$ and $\phi_2 = 90°$ with $M = 0.5$ and $\psi = 90°$.
The measured points shown in (a) and (b) were obtained at 1500 Mc.

A Swept-Frequency 3-Centimeter Impedance Indicator*

HENRY J. RIBLET†, ASSOCIATE, IRE

Summary—**An item of test equipment is described which is capable of presenting on a cathode-ray tube sufficient information to determine the magnitude and phase of the impedance of a load at a number of closely spaced frequencies over a 12 per cent frequency range centered in the 3-cm band. The first model of the equipment measures reflection coefficients with an accuracy, for low standing-wave ratios, of ±4° in phase and ±8 per cent in magnitude. A novel rf circuit, called a wave sampler, makes the accuracy of this system independent of frequency.**

I. INTRODUCTION

THE MEASUREMENT of impedance over a broad band of frequencies is one of the most tedious and time-consuming chores in the design and testing of radio-frequency components. Time is not the only loss, however, in using conventional standing-wave equipment. Inability to obtain quickly a picture of the over-all effect of small dimensional changes in the sample under test limits the performance obtainable from certain critical components. It is the object of this paper to describe a system which will present rapidly on a cathode-ray tube information from which may be readily deduced the magnitude and phase of a load at a number of closely spaced points in the 3-cm band.

Two essentially distinct systems for obtaining somewhat similar results have been discussed previously. Gaffney[1] has described a reflectometer which uses a "magic-tee" hybrid junction[2] to separate the power incident on a load from the power reflected by it. One may then compare these powers and determine the standing-wave ratio. For a complete discussion of the limitations of this system, the reader is referred to the original paper. In brief they are:

1. No measurement of phase is convenient. This restricts the usefulness of this piece of test equipment severely, since, for most experimental work, phase information is fully as important as amplitude information.
2. The accuracy, for standing-wave ratios near unity, is limited (for varying frequencies) by the performance of the hybrid circuit, since in this limit the measured reflected power depends on the phase of the power actually reflected in a manner not easily taken into account.
3. The need for comparing power levels differing by as much as 40 db in the case of well-matched components—actually the situation of principal interest for many applications—requires the additional complication of a more sensitive receiver than is ordinarily used in standing-wave measurements.

On the basis of these objections, work was begun early in 1946 on the arrangement to be described and has been continuing on a low-priority basis at the Submarine Signal Company ever since. We believe that it meets all of the objections listed above; however, it pays for this by giving correct impedance information only at a set of discrete frequencies. As a consequence, the present equipment appears to be useful only in measuring components known to have reasonable bandwidths.

The arrangement suggested by Samuel[3] and Korman[4] gives the magnitude and phase of a waveguide termination for a continuous range of frequencies. Its accuracy depends on a rather narrow-band waveguide circuit Calculations made on the basis of the analysis given by Samuel indicate serious errors for this waveguide circuit when it is used over a 12 per cent frequency band. Unfortunately, what amounts to crosstalk between the amplitude and the phase of the reflected wave would appear to make it rather difficult to remove these errors by calibration. Accordingly, it is felt that this system is a natural complement for the system here described.

II. THE SYSTEM

For an explanation of the operation of the swept-frequency impedance indicator, reference is made to the schematic diagram of Fig. 1. The rf energy, originating in a variable-frequency signal generator having reasonably constant output, is fed through a special waveguide circuit devised for this purpose and called a wave sampler, for want of a better name, and then through a known length of transmission line to the load whose impedance is to be measured. The wave sampler is a four-terminal-pair waveguide circuit. At two of its arms appear rf signals proportional to $A^2+B^2+2AB\cos\phi$ and $A^2+B^2-2AB\cos\phi$, regardless of frequency. Here A is the incident voltage in the main guide, B is the reflected voltage, and ϕ is their relative phase. It will be seen that the wave sampler, to be described in detail later on, is

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* Decimal classification: R244.3. Original manuscript received by the Institute, March 3, 1948; revised manuscript received, May 11, 1948. Presented, 1948 IRE National Convention, March 24, 1948, New York, N. Y.

The work discussed in this paper was carried out at the Submarine Signal Company in connection with its Section "T" activity under contract NOrd-9788 with the Bureau of Ordnance, Navy Department, Washington, D. C.

† Submarine Signal Company, Boston, Mass.

[1] F. J. Gaffney, "Microwave measurements and test equipments," PROC. I.R.E., vol. 34, pp. 778–780; October, 1946.

[2] W. A. Tyrrell, "Hybrid circuits for microwaves," PROC. I.R.E., vol. 35, pp. 1294–1307; November, 1947.

[3] A. L. Samuel, "An oscillographic method of presenting impedances on the reflection-coefficient plane," PROC. I.R.E., vol. 35, pp. 1279–1283; November, 1947.

[4] N. A. Korman, "Theory and design of several types of wave selectors," *Proc. Nat. Electronics Conference*, vol. 2, pp. 418–422; 1946.

equivalent to a section of waveguide having on it two identical, uncoupled, short probes one-quarter wavelength apart at all frequencies. Two voltages proportional to $A^2+B^2+2AB\cos\phi$ and $A^2+B^2-2AB\cos\phi$ are developed in suitable identical detectors and fed after amplification to a continuous computer which has an output proportional to the ratio of these voltages.

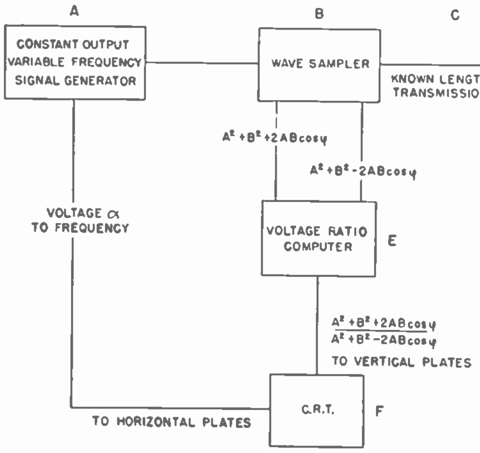

Fig. 1—Schematic of system.

This output voltage is applied to the vertical plates of a crt while the voltage on the horizontal plates is synchronized with the frequency of the signal generator, so that on the oscilloscope face we have a graph of $R=(A^2+B^2+2AB\cos\phi)/(A^2+B^2-2AB\cos\phi)$ against frequency.

It is fundamental to an understanding of the operation of this device to appreciate that ϕ, the phase of the reflected voltage B relative to the incident voltage A (measured at the wave sampler) is a function of the frequency, the length of line separating the load from the slots of the wave sampler, and the phase of the impedance of the load. At 3 cm, for a transmission line (C) whose length approximates 3 feet, ϕ changes sufficiently rapidly with λ, for reasonably broad-band loads, so that R takes on alternately the values $(A+B)^2/(A-B^2)$ and $(A-B)^2/(A+B)^2$ about once for every 1 per cent change in wavelength. These, of course, are true values of the square of the standing-wave ratio and its reciprocal, respectively. All other values of R lie between these extremes. Thus, on the face of the crt we see traces as shown in Figs. 6 and 7. True values of standing-wave ratio are given only at those wavelengths $\lambda_1, \lambda_2 \cdots \lambda_n$ which correspond to maxima and minima of the trace. Since the phase of the reflected wave at the wave sampler, $\lambda_1, \lambda_2, \cdots, \lambda_n$ and the length of the transmission are known, we are in a position to determine the phase as well as the magnitude of the impedance of the load at these wavelengths.

It is clear that the principles by which the impedance of a termination is determined with this arrangement are identical to those employed in slotted-line measurements. The use of the wave sampler and rather wide frequency variations combine to make *any mechanical motion unnecessary*. As a consequence, the speed with which the data may be presented on a crt is limited only by the time constants of the electronic circuits and the characteristics of the frequency-modulated signal generator. It seems reasonable to believe that, with a suitable signal generator, it should be possible with this arrangement to present the data containing the impedance of the load on a crt at a speed sufficient to give a continuous trace, as is now done at this frequency with conventional spectrum analyzers. Lacking such a signal generator, we have employed a tunable magnetron and have had no difficulty in covering a 6 per cent frequency band in 3.5 seconds.

SYSTEM COMPONENTS

Fig. 2 is a photograph of the waveguide and mechanical components used. A 2J51 tunable magnetron pulsed at a repetition rate of 1000 cps is used as the variable-frequency signal source. A very suitable reversing mechanism for tuning the magnetron back and forth across a 6 per cent frequency band was available out of war surplus, at a nominal cost, in the form of an aircraft control cable motor and clutch system. The magnetron, blower, reversing mechanism, and the potentiometer pick-off for moving the crt spot horizontally are shown at the left-hand side of Fig. 2. To avoid interaction between the load under test and the magnetron, about 30 db of padding is provided between them by a cross-type directional coupler.[5] The physical appearance of the wave sampler is clear from the photograph. At the output arms of the wave sampler may be seen the two Sperry barretters which detect the rf signals. The electronic circuits are not shown. Two chassis contain the power supplies, crt, and associated amplifiers. Another chassis contains the computer. It is a two-channel amplifier arranged so that a simultaneous automatic gain control holds the output of one of the channels constant. In this way, the output of the other channel gives the ratio of the input signals. To minimize the effects of tube unbalances, the same tubes are used in both channels. The numerator channel operates at 1 kc and the denominator at 3 kc. These frequencies are obtained directly from

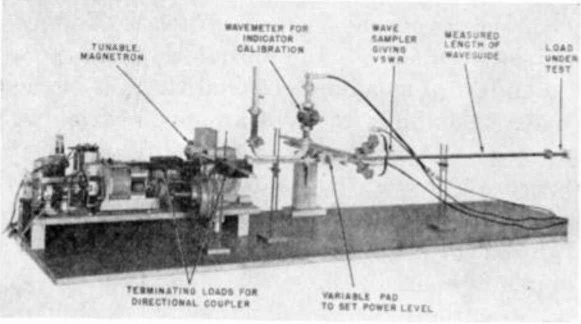

Fig. 2—Rf components.

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[5] M. J. Surdin, "Directive couplers in wave guides," *Jour. I.E.E.* (London), Pt. IIIA, vol. 93, p. 728; March-May, 1946.

the repetition rate of the magnetron by filtering in two sharply tuned amplifiers.

The wave sampler, which is fundamental to the accuracy and bandwidth of this scheme, is a special waveguide circuit designed specifically for this project. Although it is the outgrowth of considerable general discussion, its present form was suggested by Saad. The possibility that its rather surprising performance might be traced to basic symmetry considerations[6] was suggested by Lippmann and then worked out in detail by the writer. These arguments are presented in detail in the Appendix. Fig. 3 shows a cutaway view of the wave sampler. Terminals (1) and (2), as shown, are respectively connected to the signal generator and load. When terminals (4) and (6) are well matched, signals proportional to $A^2+B^2-2AB\cos\phi$ and $A^2+B^2+2AB\cos\phi$ are obtained from terminals (3) and (5). This is accomplished by cutting slots (a) and (b) in the main guide so that their centers fall in a plane perpendicular to the axis of this guide. Since slot (*b*) is excited by the total longitudinal current flow in the main guide, while slot (*a*) is excited by the total transverse current flowing in the main guide, it is easily argued that the excitation of slot (*b*) is proportional to the vector difference of the incident and reflected voltages, while the excitation of slot (*a*) is proportional to their vector sum (i.e., the equivalent probes are one-quarter wavelength apart at all frequencies). Fairly careful analysis, however, is required to show that the factors of proportionality are identical at all frequencies. As will be pointed out in the next section, this theoretical fact has been checked experimentally to within ±0.1 db over the 12 per cent 3-cm band.

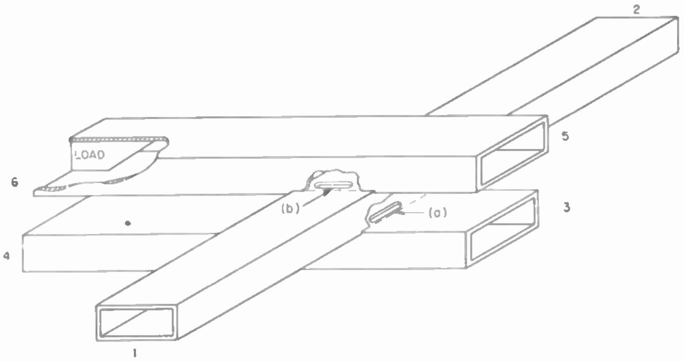

Fig. 3—Wave sampler.

PERFORMANCE

The most accurate way of using this equipment is to tune the magnetron manually across its band and record the maximum and minimum values of the output of the computer as indicated by a voltmeter, together with the frequencies at which they occur. Normalization of these values to the values obtained with a matched load at the same frequencies then allows one to determine the standing-wave ratio over the frequency band. The phase of the impedance of the load is determined from the length of the wave sampler and the frequencies at which the maximum and minimum values occur. Of course, the most rapid procedure consists in mechanically tuning the magnetron across the band and photographing the trace as it appears on the crt. The impedance data may then readily be obtained by comparison of the photograph obtained from the load under test with that obtained when the wave sampler is terminated with a matched load and with a short circuit.

Of these two methods, the first is the more accurate and gives the ultimate accuracy of the present apparatus, since it places no strain on the speed of the electronic circuits or on the linearity of the crt. Accordingly, three loads have been measured in this way, and the results compared with those obtained from slotted-line measurements. The results as concerns the absolute value of the standing-wave ratio are shown in Fig. 4. On the basis of these data, it is felt that the system will measure reflection coefficients of less than 0.20 (standing-wave ratios of magnitude less than 1.50) with an accuracy of better than ±8 per cent. An idea of the phase accuracy is obtainable from Fig. 5, which shows the impedance in both magnitude and phase at five selected points in the band for the three loads of Fig. 4. It is clear

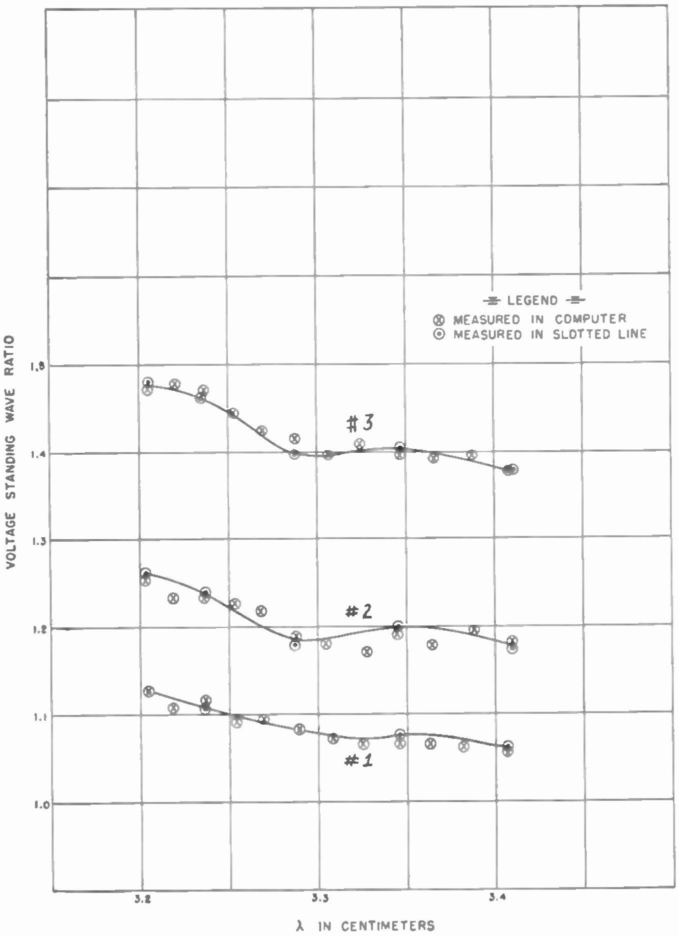

Fig. 4—Comparison of slotted-line and computer data.

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[6] The thinking that led Saad to this result was based on formulas due to H. Bethe (see footnote reference 5), and is thus rigorous only in the limit of small coupling slots.

that the phase accuracy is inversely related to the length of the wave sampler. It is our experience that the frequencies of the maximum and minimum values of the computer output can be determined to about ±1 part in 3000. If the length of the wave sampler centers around 20 guide wavelengths, as in this case, it is readily determined that the phase error will be in the neighborhood of $\pm 0.01\lambda$ g. It will be observed that this is just about the accuracy obtained in Fig. 5. Absolute phase information would appear to depend on a precise knowledge of the frequency and the dimensions of the wave sampler. Actually, the wave sampler and wavemeter may be calibrated by shorting the output terminal of the wave sampler and determining the wavemeter readings at which the slots of the wavesampler are an integral number of quarter guide wavelengths from the short. By linearly interpolating between these readings, the absolute phase information plotted in Fig. 5 was obtained.

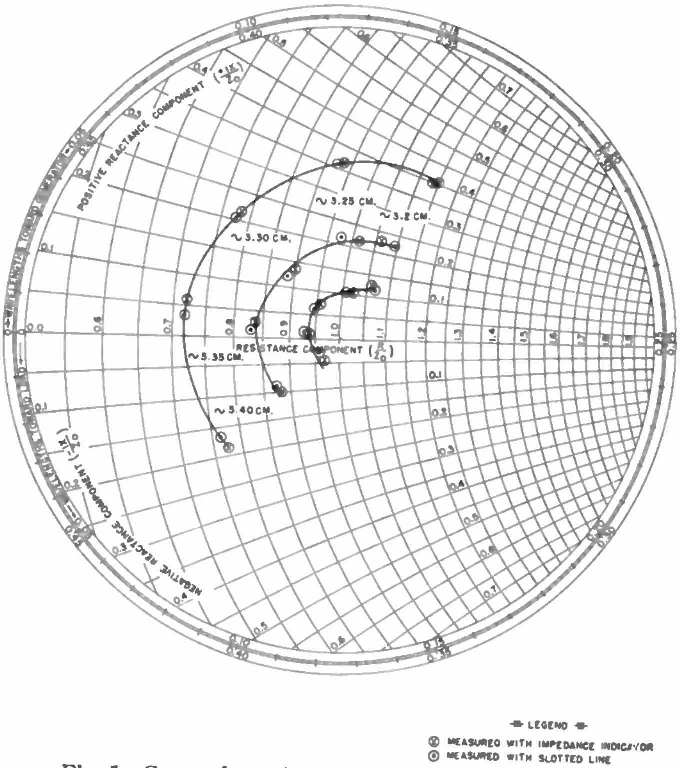

Fig. 5—Comparison of data on an impedance diagram.

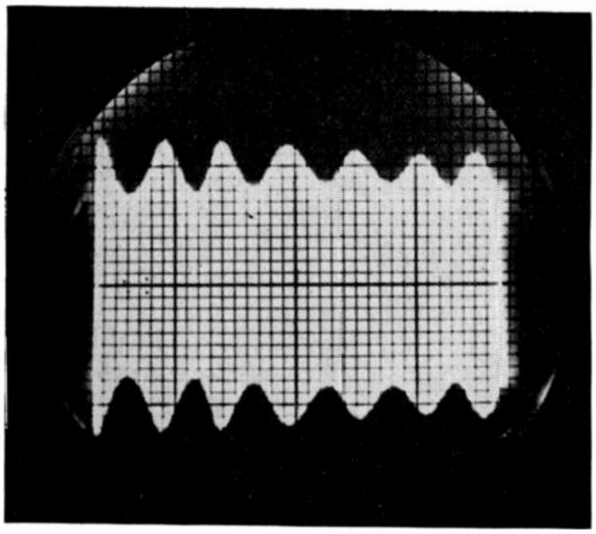

Fig. 6(a)—Trace of load No. 1.

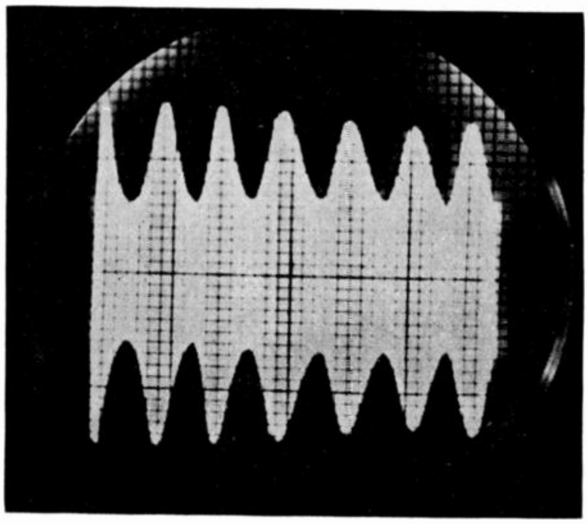

Fig. 6(b)—Trace of load No. 2.

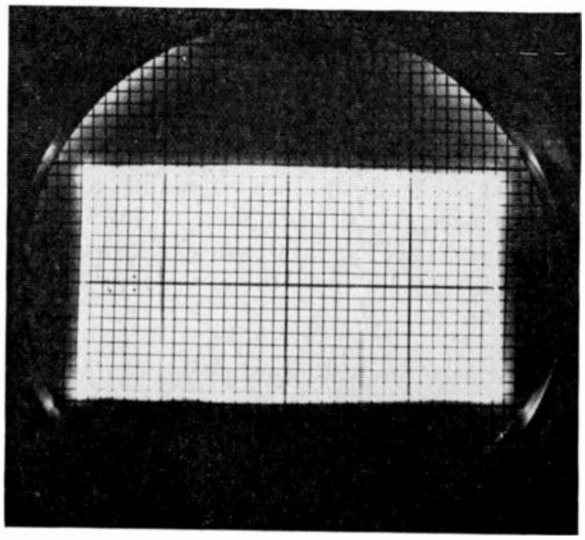

Fig. 6(c)—Trace of matched load.

Data obtained photographically are shown in Figs. 6(a), (b), (c) and 7(a), (b), (c). Figs. 6(a) and (b) show the traces for loads No. 1 and No. 2. These are to be compared with the trace obtained with a matched load in Fig. 6(c). Fig. 7(a) gives the data for load No. 3. Figs. 7(b) and (c) are the comparable data for a terminating short and matched load, respectively. Fig. 8 compares the values of the standing-wave ratio obtained from the photographs with those obtained from a slotted line. It should be recalled that the scope values were normalized to the values obtained from a matched load, so that the data are inherently more accurate for low standing-wave ratios. It is clear from Fig. 8 that, allowing for a certain amount of smoothing, which appears to be very permissible, the accuracy of the photographic procedure is more than adequate for most purposes. The phase characteristics of the load No. 3, say, may be readily determined by comparing Figs. 7(a) and (b). A lit-

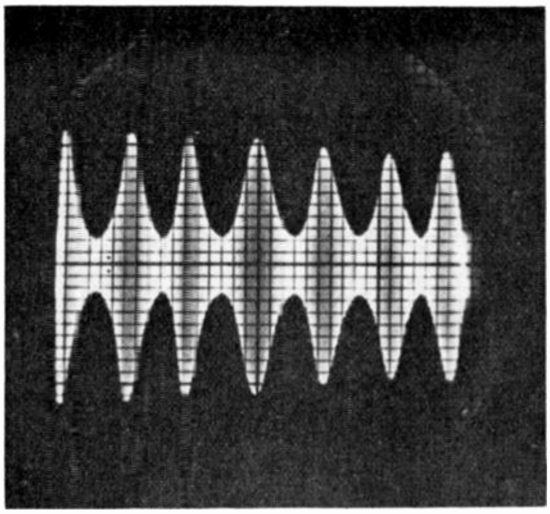

Fig. 7(a)—Trace of load No. 3.

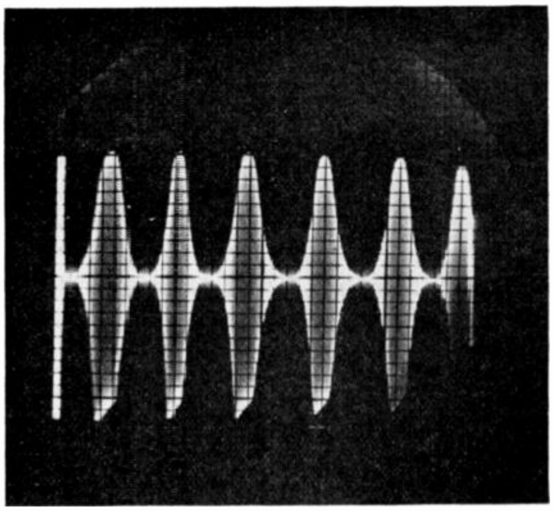

Fig. 7(b)—Trace of short circuit.

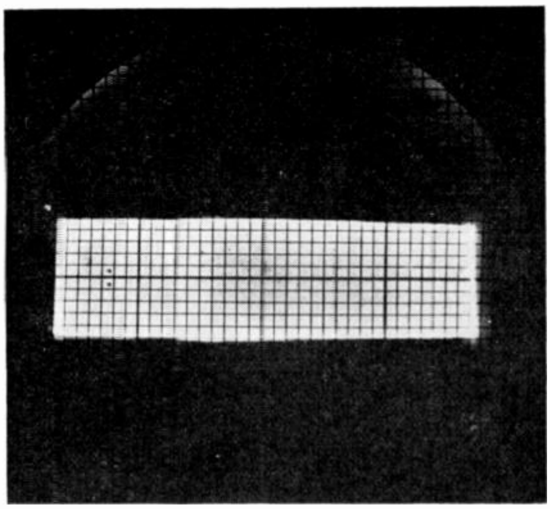

Fig. 7(c)—Trace of matched load.

tle consideration will show that the impedance of this load moves about half way around the circle diagram over the frequency band involved.

It will be observed that the data so far have been limited to a 6 per cent frequency band. The reason for this is simply that this band is adequate for the purposes for which the equipment was constructed, while a wider band was rather awkward to obtain with the mechanical reversing mechanism which was available. The only components of the system which in any way involve the frequency have, however, been tested over the 12 per cent band. The output of the tunable magnetron is, of course, constant over the band within 1 db. The items about which there remain some doubt are, then, the wave sampler and the barretters. The performance of these is shown in Fig. 9. The lower graph of this figure shows that the barretters are identical to within 0.1 db over the band. The signals out of the wave sampler, when terminated in a matched load, are the same within 0.2 db, as shown in the middle graph. The upper graph indicates that the combined unbalance of wave sampler and barretters taken together is less than 0.2 db over the 12 per cent band. Incidentally, the trace of Fig. 6(c) shows about this amount of variation for a matched load.

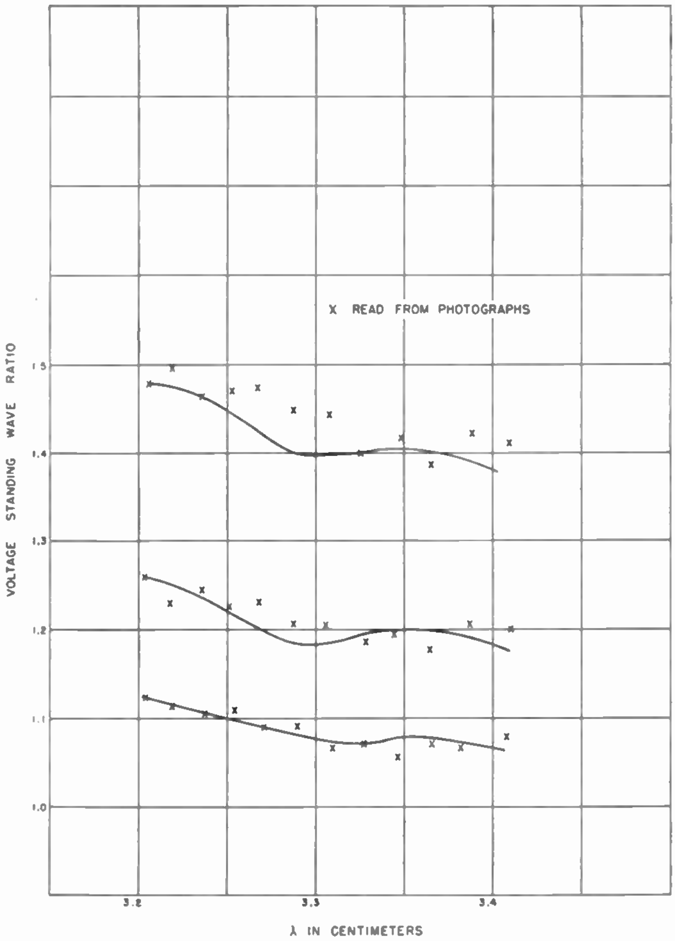

Fig. 8—Comparison of slotted line and photographed data.

ACKNOWLEDGMENT

This development has required the co-operation of a number of persons, several of whom are no longer at the Submarine Signal Company. John Daspit designed and tested the ratio computer; Chester W. Young designed and tested the amplifier and indicator circuits; Charles Aker arranged the mechanical drive for the magnetron;

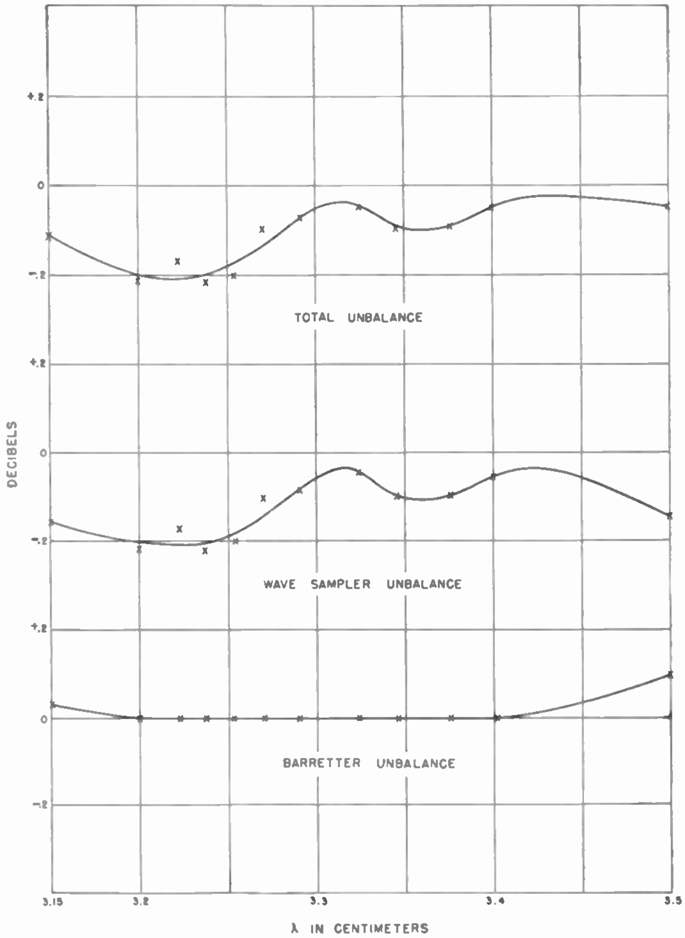

Fig. 9—Wave sampler unbalances.

and Miss Eileen Quigley has made the measurements reported in the paper. The contributions made by Saad and Lippmann have already been described.

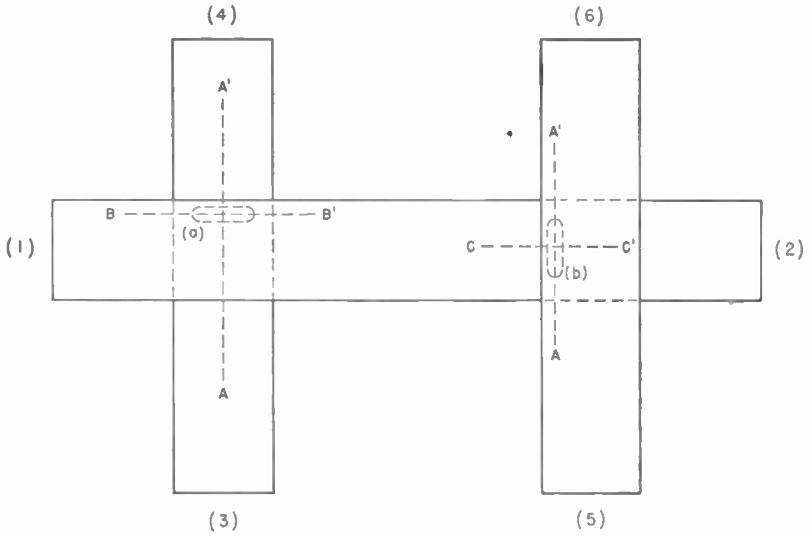

Fig. 10—Schematic of wave sampler.

Appendix

Fig. 10 gives a plan view of the wave sampler. For the sake of clarity, the wave guides 3–4 and 5–6 are shown displaced laterally from each other. The two lines *AA'* in the actual device are identical. Waveguide 5–6 is above waveguide 1–2, just as is shown in Fig. 5. Terminals (1) and (2), (3) and (4), and (5) and (6) are taken to be the same distance from the lines *AA'*, *BB'*, and *CC'*, respectively.

On the assumption that the identical slots (*a*) and (*b*) are infinitesimally wide, the following statements follow from the mechanical and electrical symmetry of the device:

I. A field in slot (*a*) does not excite a field in slot (*b*), and vice versa.

II. Slot (*a*) is not excited by incident voltages antisymmetric about *AA'*, and similarly for slot (*b*) as seen from guide 5–6.

III. Slot (*b*) is not excited by voltages symmetric about *AA'*, and similarly for slot (*a*) as seen from 3–4.

IV. A field in slot (*a*) generates antisymmetric outgoing voltages in guide 3–4, and similarly for (*b*) in 1–2.

V. A field in slot (*b*) generates symmetric outgoing voltages in guide 5–6, and similarly for (*a*) in 1–2.

Consider in guide 1–2 the electromagnetic field which is established by two symmetric voltages of amplitude Φ_e incident on (1) and (2). Then, by III and IV, two outgoing fields antisymmetric about *BB'* will be generated in 3–4. The voltages at (3) and (4) will be denoted respectively by Ω_0 and $-\Omega_0$. We should now like to invoke the reciprocity theorem to show that antisymmetric voltages incident on 3–4 of magnitude Φ_e and $-\Phi_e$ respectively, will generate symmetric outgoing voltages at (1) and (2) of magnitude Ω_0. Since, by I and III, slot (*b*) is not excited by the assumed field configuration, there is no current flow across it, and we may assume that it is shorted out or closed over without altering the field in any of the guides. During this part of the argument, slot (*b*) need not exist, so we can consider the network whose terminals are (1), (2), (3), and (4). We write the equations relating reflected to incident voltages:

$$\begin{aligned} B_1 &= S_{11}A_1 + S_{12}A_2 + S_{13}A_3 + S_{14}A_4 \\ B_2 &= S_{12}A_1 + S_{22}A_2 + S_{23}A_3 + S_{24}A_4 \\ B_3 &= S_{13}A_1 + S_{23}A_2 + S_{33}A_3 + S_{34}A_4 \\ B_4 &= S_{14}A_1 + S_{24}A_2 + S_{34}A_3 + S_{44}A_4 \end{aligned} \tag{1}$$

where B_i and A_i are the reflected and incident voltages at the i^{th} terminal. The matrix (Sij) of this system of equations is known as the scattering matrix of the network and may easily be shown to be symmetric, a fact already assumed in (1). Since the network is symmetric with respect to an interchange of terminals (1) and (2), $S_{11}=S_{22}$, $S_{13}=S_{23}$, and $S_{14}=S_{24}$. Moreover, it is clear that $S_{33}=S_{44}$ and, by IV, $S_{13}=-S_{14}$ and $S_{23}=-S_{24}$. Thus (1) becomes

$$\begin{aligned} B_1 &= S_{11}A_1 + S_{12}A_2 + S_{13}A_3 - S_{13}A_4 \\ B_2 &= S_{12}A_1 + S_{11}A_2 + S_{13}A_3 - S_{13}A_4 \\ B_3 &= S_{13}A_1 + S_{13}A_2 + S_{33}A_3 + S_{34}A_4 \\ B_4 &= -S_{13}A_1 - S_{13}A_2 + S_{34}A_3 + S_{33}A_4. \end{aligned} \tag{2}$$

In the absence of energy incident on terminals (3) and (4), incident voltages Φ_e at (1) and (2) generate outgoing waves at (3) and (4) which are $2\ S_{13}\ \Phi_e$ and $-2\ S_{13}\ \Phi_e$, respectively. Thus $\Omega_0=2\ S_{13}\ \Phi_e$.

We then see from (2) that voltages Φ_e and $-\Phi_e$ incident on (3) and (4) respectively, in the absence of energy incident on (1) and (2), generate outgoing voltages at (1) and (2) which are both Ω_0. This is what we wished to prove.

Now the geometry seen by incident antisymmetric voltages at (3) and (4) is identical to that seen by antisymmetric voltages at (1) and (2), since slot (*a*) may be assumed to be closed for this case. Thus, if symmetric incident voltages at (1) and (2) of magnitude Φ_e establish voltages Ω_0 and $-\Omega_0$ at (3) and (4), antisymmetric voltages Φ_e and $-\Phi_e$ incident at (1) and (2) will establish symmetric voltages of magnitude Ω_0 at (5) and (6). The sum of two solutions of Maxwell's equations is again a solution and we conclude that, if the only voltage incident on the network is $2\ \Phi_e$ at terminal (1), the outgoing voltages at (3), (4), (5), and (6) will be Ω_0, $-\Omega_0$, Ω_0, and Ω_0, respectively. Thus the device has the property, under the condition of perfect match at terminal (2), that for power incident at (1), equal powers appear at the other terminals. Clearly, then, if terminals (4) and (6) are perfectly terminated, terminals (3) and (5) are equivalent to identical uncoupled probes at all frequencies. That they are one-quarter wavelength apart, as has already been indicated, follows from the fact that slot (*a*) is in parallel with guide 1–2, while slot (*b*) is in series with it.

Of course, the slots, unless small themselves, will interact with the quantities to be measured and give rise to appreciable errors. We may evaluate this type of error as follows. The scattering matrix for the complete six-terminal network is, on the basis of remarks of the previous paragraph,

$$\begin{aligned} B_1 &= S_{11}A_1+S_{12}A_2+S_{13}A_3-S_{13}A_4+S_{13}A_5+S_{13}A_6 \\ B_2 &= S_{12}A_1+S_{11}A_2+S_{13}A_3-S_{13}A_4-S_{13}A_5-S_{13}A_6 \\ B_3 &= S_{13}A_1+S_{13}A_2+S_{33}A_3+S_{34}A_4 \quad 0 \quad 0 \\ B_4 &= -S_{13}A_1-S_{13}A_2+S_{34}A_3+S_{33}A_4 \quad 0 \quad 0 \\ B_5 &= S_{13}A_1-S_{13}A_2 \quad 0 \quad 0 \quad S_{55}A_5+S_{56}A_6 \\ B_6 &= S_{13}A_1-S_{13}A_2 \quad 0 \quad 0 \quad S_{56}A_5+S_{55}A_6. \end{aligned} \tag{3}$$

For voltages A_1 and A_2 incident on terminals (1) and (2), $B_3=S_{13}\ (A_1+A_3)$ and $B_5=S_{13}\ (A_1-A_2)$. Thus the ratio of the power out of terminals (3) and (5) is

$$\frac{|A_1+A_2|^2}{|A_1-A_2|^2} = \frac{\left|1+\frac{A^2}{A_1}\right|^2}{\left|1-\frac{A^2}{A_1}\right|^2} \tag{4}$$

independently of frequency. If the voltage incident on the wave sampler A_1 were the same as the voltage B_2 incident on load under test, this device would be exact, regardless of the size of the coupling slots. Of course, this is not the case. Actually, by (3), $B_2=S_{12}A_1+S_{11}A_2$. Thus the ratio of the power at terminals (5) and (3) equals

$$\frac{|B_2+(S_{12}-S_{11})A_2|^2}{|B_2-(S_{12}+S_{11})A_2|^2}. \tag{5}$$

In the limit of small slots $S_{13}\rightarrow S_{11}\rightarrow 0$ and $|S_{12}|\rightarrow 1$, so that (5) approaches

$$\frac{A^2+B^2+2AB\cos\phi}{A^2+B^2-2AB\cos\phi}.$$

For the wave sampler pictured in Fig. 2, $|S_{13}|=0.01$ and $|S_{11}|$ was approximately 0.005. Since the matrix of (3) is known to be unitary, we have that

$$|S_{11}|^2+|S_{12}|^2+4|S_{13}|^2=1. \tag{6}$$

Then $|S_{12}|^2=1-4.10^{-4}-0.25\times10^{-4}\approx 0.9995$. For small standing-wave ratios, then, this gives an error in the reflection coefficient of, at most, about one-half of one per cent. This is a truly insignificant error in the standing-wave ratio.

It might be interesting to point out that by making guides 3–4 and 5–6 coincide, and by superimposing slots (*a*) and (*b*) in the form of a cross, one is led to a type of directional coupler whose directivity is, in theory at least, independent of the frequency. The proof given here, however, is more general than that given by Surdin,[7] since no restrictions need be placed on the size of the slots other than those required for the validity of I–V. Furthermore, it is rather easy to imagine analogous coaxial-line circuits by making use of the notion of series and parallel coupling essential to the symmetry of the wave sampler.

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[7] See footnote reference 5. Surdin's discussion of the properties of the cross-type directional coupler depends on formulas due to Bethe, which may readily be shown to be incorrect for slots of finite size.

The Ultrasonic Interferometer with Resonant Liquid Column*

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Summary—The ultrasonic interferometer is simple in construction and operation, and yields accurate and consistent data. From these data, one can readily determine the velocity of sound in a liquid with high accuracy. Formerly, the absorption of sound in the liquid and the coefficient of reflection at the reflector surface has been obtained through a complicated analysis of the electrical and equivalent-electrical circuits of the quartz crystal and the associated fluid column. The simplified analysis of the equivalent circuit given here is made possible by limiting the discussion to the conditions that exist when all parts of the system (electrical, mechanical, and acoustical) are adjusted to resonance. Under these conditions, the analysis of the complete electrical and equivalent-electrical circuit is greatly simplified. One does not need to analyze the shape of a reaction dip in order to obtain coefficients of absorption and reflection.

VARIOUS AUTHORS, such as Cady,[1] Van Dyke,[2] and Dye,[3] have developed the analysis of the equivalent electrical circuit of the quartz-crystal resonator. Hubbard[4] has extended the analysis to include the modification introduced by a fluid column coupled to the resonator. He assumed the sound source to be an infinite plane executing simple harmonic motion. The plane progressive waves thus generated in the fluid are reflected from an infinite plane surface at a distance r from the source. Hubbard expressed the particle velocity (pressure) and displacement at each point in the fluid and calculated the effect of the multiply reflected waves on the sound source. These assumptions are valid if one uses a source with a diameter large compared to the wavelength of the sound in the fluid.

Hubbard measured the current variation produced as the reflector-to-source distance r is varied and from these changes calculated the coefficient of absorption in several gases and the coefficient of reflection at the reflector surface. Fox[5] adapted Hubbard's theory to the corresponding measurements for liquids and determined the absorption coefficient for water. Using the interferometer, Hunter[6] measured the absorption coefficients for several highly viscous liquids. Previous to these interferometric measurements the values obtained by other methods had varied over several orders of magnitude. Recent results using other methods (e.g., sound pulse) have converged to the values first measured with the ultrasonic interferometer.

Hubbard's treatment of the interferometer necessitates extensive analytical and graphical analysis in order to obtain coefficients of absorption and reflection from experimental data. It is the purpose of this paper to propose a simplified theory for the ultrasonic interferometer at resonance, i.e., electrical resonance for the LC circuit, mechanical resonance for the crystal, and acoustical resonance for the fluid column.

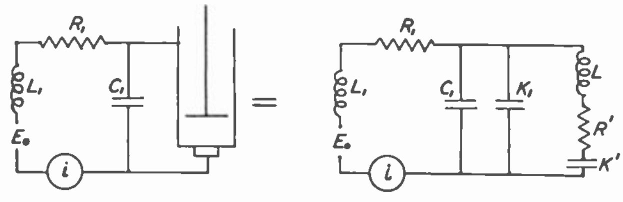

Fig. 1—The equivalent electrical circuit of the ultrasonic interferometer and associated electrical pickup circuit.

Hubbard's theory gives a general analysis of the current in a pickup circuit which has an ultrasonic interferometer connected across a tuning capacitor as shown as shown in Fig. 1. We give his results, using the following notation. L_1, C_1, and R_1 are the purely electrical inductance, capacitance, and resistance of the pickup circuit. L, K, and R are the corresponding equivalent electrical constants of the quartz resonator vibrating in a vacuum close to its natural mechanical resonance frequency, and K_1 is its clamped[7] dielectric capacitance. A voltage $E_0 e^{i\omega t}$ is induced in L_1 by a loosely coupled oscillator. When a fluid column is in contact with the resonator, the equivalent resistance and capacitance of the resonator is changed, and these modified values R' and K' are defined by

$$R' = R + AB\rho vP \qquad (1)$$

$$1/K' = 1/K + AB\rho v\omega Q \qquad (2)$$

where A is the effective area of the resonator face exposed to the fluid column, B is a piezoelectric constant of the quartz, ρ is the density, and v is the velocity of sound in the liquid, P and Q are expressions[8] determining the

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* Decimal classification: R214.2×R141.2. Original manuscript received by the Institute, February 11, 1948. Presented, 1948 National IRE Convention, New York, N.Y., March 25, 1948.

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[4] J. C. Hubbard, "Acoustic resonator interferometer: I. The acoustic system and its equivalent electric network," *Phys. Rev.*, vol. 38, pp. 1011–1019; September, 1931; also, "Acoustic resonator interferometer: II. Ultrasonic velocity and absorption in gases," vol. 41, pp. 523–535; August, 1932; and errata, "Acoustic resonator interferometer," vol. 46, p. 525; September, 1944.

[5] F. E. Fox, "Ultrasonic interferometry for liquid media," *Phys. Rev.*, vol. 52, pp. 973–981; November, 1937.

[6] J. L. Hunter, "The absorption of ultrasonic waves in highly viscous liquids," *Jour. Acous. Soc. Amer.*, vol. 13, pp. 36–40; July, 1941.

[7] K_1 is the dielectric capacitance at constant strain; or the capacitance the resonator would have if, other things being the same, the quartz were not piezoelectric. See P. Vigoreux, "Quartz Oscillators," p. 11, London, 1939; W. G. Cady, "Piezoelectricity," p. 311, McGraw-Hill Book Co., New York, N. Y., 1946.

[8] In the Hubbard paper, P and Q are written for any position x between the source and the reflector. Only the values at the source interest us here, and these are obtained by placing $x=0$ in Hubbard's general equations.

acoustic resistance and reactance of the fluid column at the resonator surface. The current in the L_1C_1 branch is i, and I_0 is used to denote (E_0/R_1), the resonance current in the L_1C_1 branch when the interferometer is disconnected and C_1 readjusted to resonance. The current ratio (i/I_0) is designated by σ. Then

$$\sigma^2 = \frac{\phi_1^2\left[\phi_2^2 + \left(q + \frac{K'}{C_1 + K_1}\right)^2\right]}{\left[pq - \phi_1\phi_2 - (1 - p)\frac{K'}{C_1 + K_1}\right]^2 + \left[\phi_2 p + \phi_1\left(q + \frac{K'}{C_1 + K_1}\right)\right]^2} \tag{3}$$

where

$$p = 1 - (C_1 + K_1)L_1\omega^2$$

$$q = 1 - LK'\omega^2; \quad q_0 = 1 - LK\omega^2$$

$$\phi_1 = R_1(C_1 + K_1)\omega$$

$$\phi_2 = R'K'\omega.$$

In interferometry the usual practice is to make certain experimental adjustments that permit one to use a simplified form of (3). Thus, if the frequency of the oscillator is adjusted, as described by Hubbard, to the response frequency of the quartz before it is modified by the loading of the liquid column, $[q_0+K'/(C_1+K_1)]$ is zero; if the L_1C_1 circuit is adjusted to resonance[9] at this same frequency, p is zero, and we have

$$\sigma^2 = \frac{(1 + SP)^2 + S^2Q^2}{(1 + SP + C)^2 + S^2Q^2} \tag{4}$$

where[10] $S=AB\rho v/R$, and C is written for $[R\omega\phi_1(C_1+K_1)]^{-1}$, an expression containing only electrical and equivalent electrical constants unmodified by the fluid column. This is the fundamental equation of ultrasonic interferometry. It has been discussed at length by Hubbard and Fox. In it P and Q are functions of the reflector positions. It is the variations of P and Q that produce the characteristic dips at odd quarter-wavelength settings of the reflector when liquids are used, or peaks at half-wavelength settings in gases. In order to obtain the coefficients of absorption and reflection, one must determine the shape of a dip when the reflector is near the source, the successive dip minima, and the value of C. The calculations are involved, and graphical interpolation is used in the evaluation. A more serious difficulty comes from the fact that with many resonators the shape of the dip is distorted by the presence of "satellite" dips. Even if these are too small to appear separately, they may nevertheless distort the shape of the dip.

It will be shown here that it is possible to avoid these difficulties and obtain the acoustic coefficient quite simply when the absorption is large, and this can always be realized by working at sufficiently high frequencies. The method may be outlined as follows. At a dip minimum the liquid column is at resonance, and the equation for σ at these positions assumes a very simple form. From a set of these σ at dip minima together with values of σ found when the reflector is far enough away for the dips to disappear, and when the liquid is removed from the interferometer, one can plot a set of points that should lie on straight line with slope equal to the coefficient of absorption and a y intercept at $r=0$ from which the coefficient of reflection can be determined.

In (4), P and Q are defined:

$$P = \frac{1 - \gamma^2 e^{-4r\alpha}}{1 + \gamma^2 e^{-4r\alpha} - 2\gamma e^{-2r\alpha}\cos\left(\frac{2r\omega}{v}\right)} \tag{5}$$

$$Q = \frac{2\gamma e^{-2r\alpha}\sin\left(\frac{2r\omega}{v}\right)}{1 + \gamma^2 e^{-4r\alpha} - 2\gamma e^{-2r\alpha}\cos\left(\frac{2r\omega}{v}\right)} \tag{6}$$

where $\omega=2\pi$ times the frequency of the driving voltage; α is the amplitude coefficient of absorption of the sound in the liquid; and γ is the amplitude coefficient of reflection.[11] We write

$$\gamma = e^{-\beta}$$

$$2r\alpha + \beta = X$$

$$\frac{2r\omega}{v} = \frac{4\pi r}{\lambda} = \delta$$

and then[12] have

$$P = \frac{\sinh X}{\cosh X - \cos\delta} \tag{7}$$

$$Q = \frac{\sin\delta}{\cosh X - \cos\delta}. \tag{8}$$

For $r=\infty$, $X=\infty$ and

$$P = 1 \tag{9}$$

$$Q = 0. \tag{10}$$

Whenever $r=m\,\lambda/4$ where m is an integer, Q is zero and the liquid column is at resonance, having no acoustic reactance. These are the critical points in acoustic inter-

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[9] These are the adjustments that insure a "symmetrical crevasse" when σ is plotted as a function of frequency.

[10] S is the ratio of the equivalent resistance of an infinitely long fluid column to the equivalent resistance R of the unloaded quartz resonator. For columns of finite length, SP is the ratio of the resistive (real) part of the equivalent impedance due to the column to the equivalent resistance R, and SQ is the ratio of the equivalent reactance of the finite column to the equivalent resistance R.

[11] The theoretical treatment of the coefficient of reflection has been developed by K. F. Herzfeld, "Reflection of sound," *Phys. Rev.*, vol. 53, pp. 899–906; June, 1938.

[12] The interpretation of the coefficients of reflection as measured in acoustic interferometry, following Herzfeld's theory, has been given by R. S. Alleman, "Dissipative acoustic reflection coefficients in gases by ultrasonic interferometry," *Phys. Rev.*, vol. 55, pp. 87–93; January, 1939.

ferometry, and the variations of σ with r near these points[13] depends on the SP term in (4). Where m is even, P is a maximum; where m is odd, P is a minimum. Since S is large compared to C for liquids, σ remains near unity except where P is a minimum as Q goes through zero. This occurs at the positions of the reflector where $r=(2n-1)\lambda/4$ and is the reason for the "dip" in σ that is characteristic of acoustic interferometry with liquids. Here n is the dip number and the value of P at these points is P_0, so we have

$$\sigma_{\text{on}} = \frac{1+SP_{\text{on}}}{1+SP_{\text{on}}+C} \tag{11}$$

where

$$P_{\text{on}} = \frac{\sinh X_{\text{on}}}{\cosh X_{\text{on}}+1} = \tanh\left(\frac{X_{\text{on}}}{2}\right). \tag{12}$$

At this point it is useful to rewrite (4) in the simplified form that applies when the interferometer is at resonance electrically, mechanically, and acoustically, as described above. From the definition of S, and (1), $SP=(R'-R)/R$, so we have at resonance

$$\sigma_{\text{on}} = \frac{1}{1+\dfrac{R}{R_{\text{on}}'}C} = \frac{1}{1+\dfrac{1}{R_{\text{on}}'R[\omega(C_1+K_1)]^2}}, \tag{13}$$

recalling that, since $p=0$, $[\omega(C_1+K_1)]^{-1}=\omega L_1$,

$$\sigma_{\text{on}} = \frac{1}{1+\dfrac{\omega^2L_1{}^2}{R_1R_{\text{on}}'}} \tag{14}$$

or

$$i_{\text{on}} = \frac{E_0}{R_1+\dfrac{\omega^2L_1{}^2}{R_{\text{on}}'}} = \frac{E_0}{R_1+\left(\dfrac{L_1}{C_1+K_1}\right)\dfrac{1}{R_{\text{on}}'}}. \tag{15}$$

Thus at resonance the total impedance of the combined interferometer and electrical pickup circuit can be simply expressed as the resistance Z in the equivalent forms

$$Z = R_1+\left(\frac{L_1}{C_1+K_1}\right)\frac{1}{R_{\text{on}}'} = R_1+\frac{\omega^2L_1{}^2}{R_{\text{on}}'} \tag{16}$$

where R_1, L_1, and (C_1+K_1) are purely electrical constants of the pickup circuit and the clamped dielectric capacitance of the quartz, and R_{on}' is the equivalent resistance of the resonator loaded by the resonant liquid column. This equation is valid only at dip minima. To indicate this we have used R_{on}' for the value of R' at the nth dip minimum, and i_{on} for the corresponding current.

Equations (13) and (14) show clearly and simply how the current in the pickup circuit increases as R' grows larger with longer liquid columns. They show, too, how the difference in i_{on} from dip to dip depends upon the square of the surge impedance $L_1/(C_1+K_1)$ or $\omega^2L_1{}^2$ of the pickup circuit, and indicate the advantage of keeping ωL_1 large. Also, from (16) we see that the equivalent circuit of the interferometer with the associated pickup circuit and resonant liquid column, adjusted as described above, may be represented as shown in Fig. 2, where the notation is identical to that used in Fig. 1.

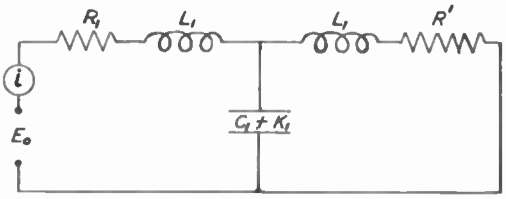

Fig. 2—The equivalent electrical circuit of the ultrasonic interferometer at resonance.

In Fig. 3 we have plotted σ_{on} as a continuous function of R_{on}', although of course (14) is valid only for discrete values of the independent variable R_{on}'. The minimum

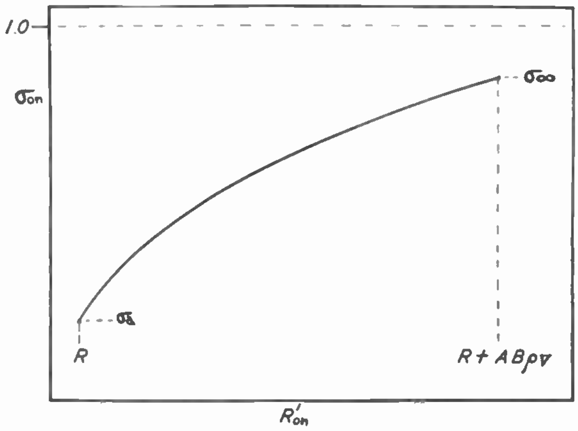

Fig. 3—σ_{on} as a function of R_{on}' as given by equation (14). The minimum value of R_{on}' is simply R, the equivalent resistance of the unloaded resonator. The maximum value of R_{on}' is R plus the equivalent resistance of an infinitely long liquid column.

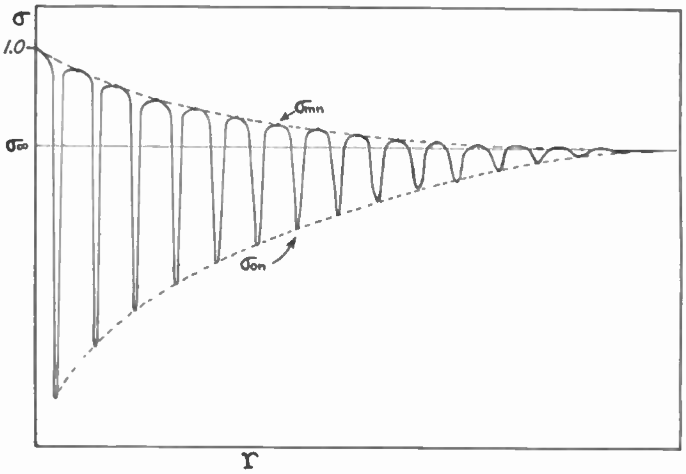

Fig. 4—σ as a function of the column length r as given by equation (4)

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[13] See footnote reference 5 for detailed discussion of the behavior of σ in liquids and gases, and for graphs of P and Q as functions of r.

value of R_{on}', (1), is simply the equivalent resistance R of the unloaded resonator, and the maximum value is $R+AB\rho v$ where $AB\rho v$ is the equivalent resistance of an infinitely long column of liquid of cross-sectional area A and acoustic radiation resistance ρv.

In Fig. 4 we have sketched σ as function of r as given by (4). The envelope of the dip minima is to be compared to the curve of Fig. 3.

We now discuss (9) to show how the coefficients of absorption and reflection may be evaluated. We first determine C, which is independent of the fluid column. This can be done by measuring σ with the liquid removed from the interferometer. Ideally, this measurement should be made with the quartz resonator vibrating in a vacuum, but this is unnecessary since S is so small in air compared with S in liquids (because of the difference in densities). However, one can make the SP term entirely negligible by setting the reflector in air so that P is a minimum, i.e., at $r=\lambda/4$ in air. Calling σ thus measured σ_a, we have

$$\sigma_a = \frac{1}{1+C} \tag{17}$$

or

$$C = \frac{1-\sigma_a}{\sigma_a}. \tag{18}$$

The liquid is placed in the interferometer and S is determined in the following manner. At dip minima where $\delta=(2n-1)\pi$,

$$P_{\text{on}} = \tanh\left(\frac{X_{\text{on}}}{2}\right),$$

while for points midway between minima where $\delta=2n\pi$,

$$P_{mn} = \left[\tanh\,(X_{mn}/2)\right]^{-1}. \tag{19}$$

Obviously, for large values of X

$$P_m \rightarrow P_0 \rightarrow P_\infty = 1 \tag{20}$$

where P_∞ is the value of P for $r=\infty$.

By measuring σ at very large values of r, one can determine σ_∞. The experimental criterion of how large r must be is supplied by the disappearance of the dips (see Fig. 4) as the reflector is withdrawn from the crystal.[14] It is, however, unnecessary to measure σ_∞ directly, since one can easily and accurately determine it by plotting maximum and minimum values of σ and estimating the value σ_∞ that both approach asymptotically from opposite sides. From equation (11),

$$SP_{\text{on}} = \frac{\sigma_{\text{on}}(1+C)-1}{1-\sigma_{\text{on}}} \tag{21}$$

and

$$SP_\infty = S = \frac{\sigma_\infty(1+C)-1}{1-\sigma_\infty} = \frac{\sigma_\infty - \sigma_a}{\sigma_a(1-\sigma_\infty)}. \tag{22}$$

Therefore S can be found from experimentally[15] determined σ values.

It is now simple to determine α and β. From the current at the nth dip minimum one finds σ_{on} and calculates P_{on} for r_n:

$$\begin{aligned} P_{\text{on}} &= \frac{\sigma_{\text{on}}(1+C)-1}{S(1-\sigma_{\text{on}})} = \frac{\sigma_{\text{on}}-\sigma_a}{S\sigma_a(1-\sigma_{\text{on}})} \\ &= \frac{(\sigma_{\text{on}}-\sigma_a)(1-\sigma_\infty)}{(1-\sigma_{\text{on}})(\sigma_\infty-\sigma_a)}. \end{aligned} \tag{23}$$

From (10),

$$P_{\text{on}} = \tanh\left(\frac{X_{\text{on}}}{2}\right) = \tanh\left(\frac{2r_n\alpha+\beta}{2}\right) \tag{24}$$

$$2r_n\alpha + \beta = 2\tanh^{-1}(P_{\text{on}}). \tag{25}$$

Thus we plot (Fig. 5) 2 $\tanh^{-1}(P_{\text{on}})$ (ordinate) against r_n (abscissa) and obtain a straight line of slope 2α, and with a y intercept ($r=0$) equal to $\beta=-\ln\gamma$.

It should be emphasized that the simplified interferometric theory presented here is not an approximation of the exact general theory as developed by Hubbard and Fox, but is the exact theory of the current minima; that is, of the interferometer at resonance as defined above. In order to evaluate the acoustic coefficients one need only measure σ_a, σ_∞, and a series of σ_{on}. Formerly, one was obliged to determine the shape of a dip, and one of the great difficulties of ultrasonic interferometry was the necessity of obtaining quartz-crystal resonators for which the dips are entirely free of secondary or satellite responses.

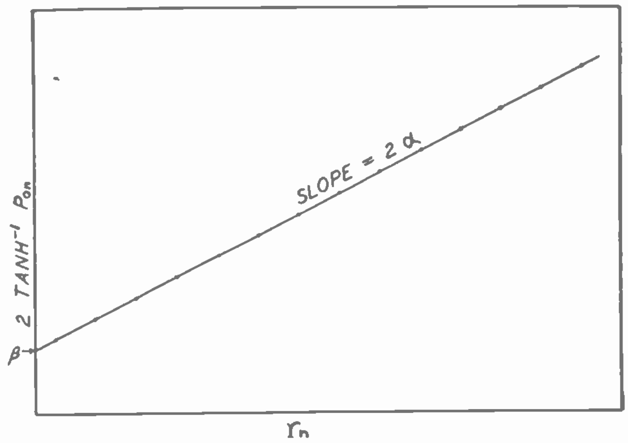

Fig. 5—2 $\tanh^{-1}(P_{\text{on}})$ as a function of the column length r. The value of P_{on} for any dip n is obtained from σ_a, σ_∞, and σ_{on}.

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[14] Whether one could determine σ_∞ by skewing the reflector so as to make the effective β and thus X large would have to be determined by measurements in a highly absorbing liquid. One would determine σ_∞ directly, then skew the reflector, and see how close to the source one could advance the reflector without changing σ_∞.

[15] It is possible, although slightly less simple, to eliminate S by measurements of the maximum values σ_{mn} using (17). In this case it is not necessary to determine σ_∞ (or, if one chooses, σ_a, instead of σ_∞). Hubbard, in an unpublished treatment, has pointed out how this can be done. In either case, the labor both of observation and calculation is greatly reduced. See E. E. Swomley, "Dispersion of the velocity and anomalous absorption of sound in hydrogen," p. 4, doctoral dissertation, Johns Hopkins University, 1946.

Contributors to Proceedings of the I.R.E.

John E. DeTurk

John E. Deturk was born in Urbana, Ill., on February 5, 1921. From 1938 to 1942 he attended the University of Illinois, where he pursued studies leading to the B.S. degree in physics. From 1943 to 1946 he was a member of the staff of the Harvard University Underwater Sound Laboratory, a war research organization sponsored by the Office of Scientific Research and Development.

Since 1946 Mr. DeTurk has been employed by the Raytheon Manufacturing Company, as a senior electronics engineer. In June, 1947, he became associated with the computer equipment section of the Raytheon Company, and has since been engaged in the design and development of automatic digital-computing equipment.

◆

Francis E. Fox

Francis E. Fox was born on May 22, 1909, in Wilmington, Del. In 1932, he received the A.B. degree at the Catholic University of America, in Washington, D. C. In 1934, this was followed by the M.Sc. in physics, and in 1937 by the Ph.D. from the same University.

Since 1936 Father Fox has been associated with the physics department at the Catholic University in various capacities, beginning as a research assistant. He became an instructor in physics in 1939, and at present holds the rank of associate professor of physics at that institution.

Since 1942, Father Fox has been associated with the naval research program of the David Taylor Model Basin. His research has been chiefly in the field of underwater sound and ultrasonics.

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For a biography and photograph of R. M. Bowie, see page 1049 of the August, 1948, issue of the Proceedings of the I.R.E.

Eugene G. Fubini

Eugene G. Fubini (A'36–SM'46) was born in Italy on April 19, 1913, and was educated at the Universities of Turin and Rome, Italy. He received the Ph.D. degree in 1933, Dr. Fubini did work on ultrahigh frequencies at the National Electrochemical Institute from 1936 to 1938. He joined the Columbia Broadcasting System staff in New York, N. Y., in 1939 as acting engineer-in-charge of the short-wave division. He later worked on design and installation of vhf links.

In December, 1942, Dr. Fubini joined the Radio Research Laboratory at Harvard University to undertake theoretical and experimental work. At the end of 1943, he was assigned as technical observer to the Mediterranean Army Air Force headquarters, and from September, 1944, to April, 1945, he was in charge of the RCM Division of the Operational Analysis Section of the Eighth Air Force.

Upon returning to the United States, Dr. Fubini became expert consultant to General McClelland, Air Chief of Operations. He joined Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y., in September, 1945, where he now heads the special devices section of the Laboratory.

◆

Donald C. Johnson

Donald C. Johnson attended Cornell and Columbia Universities and received the B.S. degree from the latter in 1941. He joined the staff of Sperry Gyroscope Company as an engineer, and served until the end of 1945. Early in 1946, he transferred to Reynolds Research, Glen Cove, N. Y., as senior chemical engineer. From December, 1946, to July, 1948, he worked at Airborne Instruments Laboratory, Inc., as an engineer in the special devices section. He is now affiliated with the Hazeltine Electronics Corporation.

◆

For a biography and photograph of James E. Hacke, see page 742 of the June, 1948, issue of the Proceedings of the I.R.E.

Joseph L. Hunter

Joseph L. Hunter (M'45) was born in New York, N. Y., on May 16, 1913. He received the B.S. degree in civil engineering from the Manhattan College in New York, N. Y., in 1934, the M.S. degree in physics from the Catholic University of America in 1936, and in 1940 was awarded the Ph.D. degree in physics, also from Catholic University.

Dr. Hunter was an instructor in physics at the John Carroll University in Cleveland, Ohio, from 1940 to 1941. During the war he served as a civilian with the signal corps at Camp Evans Signal Laboratory until 1943; thereafter, he was a member of the technical staff of the Bell Telephone Laboratories, New York, N. Y., until the end of the war. In 1945 he joined the staff of the Link Radio Company, and returned in 1946 to John Carroll University as a professor of physics. Since that time he has been chiefly engaged in supersonic research.

Dr. Hunter is a member of the American Physical Society, the American Institute of Electrical Engineers, and the Acoustical Society of America.

◆

John M. Kelso

John M. Kelso was born in Punxsutawney, Pa., on March 12, 1922. He received the A.B. degree from Gettysburg College in 1943, and the M.S. degree from The Pennsylvania State College in 1943, both in physics.

Since 1943, Mr. Kelso has been a part-time graduate student and a full-time employee at The Pennsylvania State College. Two years of this time were spent in teaching physics, and three years were spent with the Wind Tunnel Laboratory of the physics department. He is now employed by the ionosphere studies project in the electrical engineering department.

Mr. Kelso is a member of the American Physical Society, Sigma Pi Sigma, Pi Mu Epsilon, and an associate member of Sigma Xi.

◆

For a biography and photograph of George Sinclair, see page 1389, of the November, 1948, issue of the Proceedings of the I.R.E.

Contributors to Proceedings of the I.R.E.

Henry J. Riblet

Henry J. Riblet (A'45) was born at Calgary, Canada, on July 21, 1913. He received the B.S. degree in 1935, the M.S. degree in 1937, and the Ph.D. degree in mathematics in 1939, all from Yale University.

From 1939 to 1941 Dr. Riblet was instructor in mathematics at Adelphi College, and from 1941 to 1942 he was assistant professor of mathematics at Hofstra College. From 1942 to 1945 he was at the Radiation Laboratory, where he was in charge of that section of the antenna group which devoted its time to the design of omnidirectional, linear-array, broadband, and circularly polarized microwave antennas. He is now employed in the radar engineering group at the Submarine Signal Company.

He is a member of Sigma Xi, the American Mathematical Society, and the American Physical Society.

❖

David Middleton

David Middleton (S'42–A'44–M'45) was born on April 19, 1920, in New York, N. Y. In 1942 he received the A.B. degree in physics from Harvard University, followed by the A.M. degree in 1945, and the Ph.D. degree in 1947, obtained as a National Research Council predoctoral fellow in physics. During the last half of 1942 he was employed as a teaching fellow in electronics in the preradar training course carried on at that institution during the war. From December, 1942, through 1945, he was engaged as a research associate in the theory group of the Harvard Radio Research Laboratory, where much of his work dealt with noise and problems encountered in its use as radar and communication countermeasures.

Dr. Middleton is at present a research fellow in electronics at Harvard University, engaged in theoretical research in noise and electromagnetic theory problems. He is a member of Phi Beta Kappa, Sigma Xi, and the American Physical Society.

❖

Charles F. West

Charles F. West (M'48) was born in Oakland, Calif., on March 3, 1921. In the spring of 1942 he received the degree of B.S. in electrical engineering from the University of California at Berkeley. From 1942 to 1945 he was a staff member of the Radiation Laboratory at the Massachusetts Institute of Technology. During the early part of the war, and until the establishment of the British and French branches of the Radiation Laboratory, he served as a special overseas representative of the Office of Scientific Research and Development in London, prior to the British and French branches of the Radiation Laboratory.

Following the war, Mr. West was attached to the Dunham Laboratory of Electrical Engineering at Yale University. In 1946 he joined the Columbia Broadcasting System's color-television research group in New York.

Since June, 1947, Mr. West has been with the Raytheon Manufacturing Company at Waltham, Mass., where he has been engaged in the development of high-speed electronic digital-computing equipment. Since February, 1948, he has headed the computer equipment section.

Correspondence

Multifrequency Bunching in Reflex Klystrons*

W. H. Huggins has made an interesting theoretical study of mode competition in reflex klystrons.[1] The importance of the mode-competition problem in microwave oscillators in general gives his conclusions a particular significance. The simultaneous presence of oscillations of different frequencies on a transient or steady-state basis, the build-up of one oscillation as another dies out, etc., are phenomena associated with the character of the nonlinear circuit element in an oscillator and the frequency spectrum of its resonator. It should consequently be possible to analyze the behavior of such an oscillator in these respects on a more general basis; for instance, in terms of an expansion of the nonlinear conductance (or resistance) in powers of the voltage or current variable. In other words, I am suggesting that an extension of van der Pol's analysis of the nonlinear oscillator would contribute substantially to the qualitative understanding of the phenomena under discussion.[2,3] Van der Pol's treatment of the coupled-circuit triode oscillator illustrates the feasibility of this approach.

It is surprising (and regrettable) that radio engineers in general, and authors of handbooks and text books in particular, so far have made little use of the insight into the properties of oscillators that the nonlinear circuit analysis has given in such respects as initiation of oscillations, amplitude stability, synchronization with a small applied voltage, pulling in coupled systems, etc. The more complicated phenomena encountered in the microwave field due to the multiple resonances of distributed-constant systems make it even more desirable to take advantage of the powerful methods of the nonlinear circuit analysis.

Gunnar Hok
University of Michigan
Ann Arbor, Mich.

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* Received by the Institute, August 26, 1948.

[1] W. H. Huggins, "Multifrequency bunching in reflex klystrons," Proc. I.R.E. vol. 36, p. 624; May, 1948.

[2] B. van der Pol, "The nonlinear theory of electrica oscillations," Proc. I.R.E., vol. 22, p. 1051; September, 1934.

[3] N. Minorsky, "Introduction to Nonlinear Mechanics," Edward Brothers, Ann Arbor, Mich., 1947.

Modern Single-Sideband Equipment*

In regard to the paper "Modern Single-Sideband Equipment," which appeared in the August, 1948, issue of the Proceedings,[1] in which is treated the schematic of an oscillator developed by Mr. Prins, I beg herewith to inform you that it has just appeared to us that this same oscillator is already included in the U. S. patent number 2,321,354, 1943. Although the mentioned oscillator had been developed here independently, it appears that the idea of it was already known at that moment in the United States.

We should be very much obliged to you if you would publish this letter as soon as possible.

C. T. F. van der Wyck
Telegrafie en Telefonie
Centraal Laboratorium
Kortenaerkade II, Holland

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* Received by the Institute, August 30, 1948.

[1] C. T. F. van der Wyck, "Modern single-sideband equipment of the Netherlands postals telephone and telegraph," Proc. I.R.E., vol. 36, pp. 970–980; August, 1948.

Institute News and Radio Notes

Board of Directors

At the Board of Directors' Meeting held at the Institute on September 9, Ralph Bown was elected winner of the 1949 Medal of Honor. The following 31 members will also receive fellow awards, effective January 1, 1949:

H. A. Affel
K. C. Black
J. E. Brown
Cledo Brunetti
W. L. Carlson
P. S. Carter
F. F. d'Humy
J. N. Dyer
L. A. Gebhard
T. T. Goldsmith, Jr.
F. W. Grover
E. A. Guillemin
Ross Gunn
A. V. Haeff
L. C. Holmes
J. K. Johnson
S. R. Kantebet
W. B. Lodge
K. A. Mackinnon
H. F. Olson
L. S. Payne
L. M. Price
H. J. Reich
J. D. Reid
Karl Spangenberg
George Sterling
O. E. Strong
Franz Tank
Norris Tuttle
I. R. Weir
George O'Neill

The Danish Academy of Technical Sciences requested in a letter to President Shackelford, dated May 18, 1948, that the Institute recommend a candidate to receive the Valdemar Poulsen Gold Medal for 1948. The medal is awarded once every year in principle—although in practice it may be awarded every two years—to a radio engineer who has contributed significantly to the science or art of radio communication. This person is selected from recommendations made by "competent institutions in Denmark and abroad." Accordingly, the Institute Awards Committee suggested the name of G. W. Pierce.

The Institute was invited to send a representative to the inauguration reception for Dwight D. Eisenhower, new President of Columbia University. Dr. Goldsmith was asked to represent the Institute and accepted.

Executive Committee

A regular meeting of the Executive Committee was held at the Institute on September 8. C. G. Mayer, RCA European Technical Representative in London, was appointed IRE Representative on the IRE-AIEE International Liaison Committee.

EMPORIUM SECTION CONVENES WITH CHEMISTS

The third joint dinner meeting of the Pennsylvania-New York Western Border Section of the American Chemical Society and the Emporium Section of the IRE was held in Emporium, Pa., on September 16. An audience of 120 heard Albert C. Walker present his paper on "The Development of Piezoelectric Crystals," after which they saw a motion picture "Crystal Clear," prepared by J. J. Harley to illustrate Dr. Walker's paper.

The success of this meeting, as well as of the previous joint meetings, indicates the potential value to small or geographically isolated sections of the Institute of combined meetings with related scientific groups in order to facilitate the presentation of important developments by recognized authorities.

AGREEMENT ON SCREW THREADS STANDARDS

A joint conference of representatives of government committees and ranking industrial standardization groups from Great Britain, Canada, and the United States is scheduled to meet within the next three months in order to make final agreements on common standards for screw threads used on most types of threaded fasteners, including bolts and nuts.

It is planned to hold this conference at the National Bureau of Standards, which has actively co-operated in the attainment of this objective over a period of many years. The purposes of the joint conference are to assure complete agreement on all fundamentals and to celebrate the reaching of an agreement after years of negotiation.

Calendar of COMING EVENTS

1948 Southwestern I.R.E. Conference Dallas, Tex., Dec. 10–11, 1948

AIEE Symposium on High-Frequency Measurements, Washington, D. C., Jan. 10–12, 1949

American Physical Society Meeting, New York City, Jan. 27–29, 1949

1949 IRE National Convention, New York City, March 7–10, 1949

RMA—IRE Spring Meeting, Philadelphia, Pa., Apr., 25–27, 1949

1949 IRE Convention Plans Under Way

Under the guidance of George W. Bailey, Chairman, and the 1949 Convention Committee, arrangements for the 1949 IRE National Convention are moving ahead at top speed. All over the United States, Canada, and points even further distant, members are setting aside the second week in March of the new year for what promises to be one of the biggest and best conventions in the Institute's history. Indications show that the attendance should equal if not exceed that at the record 1948 convention. Again members are urged to make their hotel reservations well in advance because, although the hotel situation in New York City is less serious than it was, there is still a seasonal shortage of rooms. Desirable accommodations will be at a premium, and this is a convention no member will want to miss.

In response to the December 1 deadline set by D. B. Sinclair, Chairman of the Technical Program Committee, proposals for papers have been pouring in at a gratifying rate, and Dr. Sinclair assures members of a stimulating and varied papers program that will cover the entire gamut of subjects interesting to workers in the radio and electronics and allied fields. The complete technical program, including summaries of the papers scheduled and dates of presentation, will be printed in the February, 1949, issue of the PROCEEDINGS.

Dates have already been confirmed for most of the social activities. The annual meeting of the Institute will start the Convention on Monday, March 7, at 10:30 A.M. Later in the day a cocktail party from six to eight P.M. arranged by Walter Knoop, Chairman of the Cocktail Party Committee, will give the members and their families an opportunity to become better acquainted with each other. A sections meeting will follow the party.

The president's luncheon at 12:45 P.M. on Tuesday planned by R. D. Chipp, Chairman of that activity, will honor the incoming president, Stuart L. Bailey, who is himself on the Convention Committee. Wednesday's highlight will be the annual banquet at 6:45 P.M. Raymond F. Guy, toastmaster, will introduce Frank Stanton, president of the Columbia Broadcasting System. Dr. Stanton's topic will be "Television Today." Karl Spangenberg has been invited to deliver the speech of acceptance on behalf of the newly elected Fellows.

Although plans for a program of Women's Activities have not yet been consolidated, ladies who enjoyed themselves at previous Conventions can look forward to a program which bids fair to outdistance all the Institute's past successes. Ladies who have never before attended an IRE Convention have a treat in store for them.

Exhibit Manager William C. Copp has reported that 140 exhibitors have already taken 80 per cent of the available space at Grand Central Palace in which to display the latest in radio and electronic devices. By the time arrangements are completed, the total number of exhibitors is expected to reach a figure close to two hundred. Among the special features contemplated for this year's exhibits is a center devoted entirely to nuclear instrumentation.

Other members of the Convention Committee who are working along with Mr. Bailey in order to make this Convention an unforgettable one are Trevor H. Clark, Vice-Chairman; Austin Bailey, in charge of Finance; L. G. Cumming, Institute Activities; Clinton B. DeSoto, Printed Program; E. K. Gannett, Facilities; Virgil M. Graham, Publicity; Roscoe Kent, Hospitality; J. Harold Moore, Registration; and Emily Sirjane, Secretary.

The IRE Professional Group System—A Status Report

In the May, 1948, issue of the Proceedings of the I.R.E., on page 570, plans for the inauguration of a Professional Group System within the broad structure of The Institute of Radio Engineers were announced. Now, nine months later, many of those plans have been fulfilled, and the IRE Professional Group System is substantially an accomplished fact. Indeed, insofar as its acceptance by the membership and the degree to which the mechanics for its future operation have been charted are concerned it is an accomplished fact.

This seems an appropriate time, therefore, for a progress report directed to the Institute membership at large.

At this writing the following steps, envisioned by the Board of Directors when it first formally adopted the Group principle of organization at its March 25. 1948, meeting, have been realized: (1) Necessary amendments to the Bylaws of the Institute to effectuate the Group system have been adopted. (2) A permanent Professional Groups Committee, replacing the original ad hoc committee which did the initial planning and promotion of the plan, has been established. (3) A model constitution and bylaws for individual Groups has been prepared, incorporating the requisite tenets with sufficient latitude in construction to allow of alteration to satisfy the specialized requirements of individual groups. (4) A detailed and highly informative Manual for the guidance of would-be Group promoters has been prepared and published, and is available upon request. (Requests should be directed to the Technical Secretary of IRE.) (5) Most importantly of all, two Groups have been actively organized, and more are in the process of formation.

In accordance with the thoroughly democratic principles which were decreed for the Professional Group System from the time of its earliest inception in the minds of several IRE officials in 1947, the greatest part of this progress—and in particular all of the steps leading to the formation of the Groups so far established or in process—has sprung from the enthusiasm with which the membership at large greeted the original announcement. In this connection, it should be emphasized that, under the system as adopted, the promotion and organization of Groups is entirely a membership matter. Groups are organized by members; they are conducted by members; and their programs and activities are entirely guided by members. Once a group is established, any number of members within a section or locality can hold Group meetings and carry on authorized activities—even if the number is as small as two!

Principle of Group Organization

The principle of Group organization should be clear to all. It is not an idea novel with IRE. Some of the older engineering organizations have already adopted similar arrangements, as their fields of interest have enlarged with a rapidly expanding technology, and for them it has proved a successful mode of operation. The same problem that faced these societies has in recent years increasingly faced The Institute of Radio Engineers. Time was when our scope was fully indicated by our name, and all members of the organization had an essentially equal interest in the activities of all others. But in the past decade or so, and particularly in the postwar epoch, the amoeba-like multiplication of applications of radio and electronics has led to a divergence of fields of interests so great that the broadcast engineer has in common with the computer engineer mainly only the fundamental phenomena of the electron tube—and even that interest is an indirect association maintained largely through the person of yet another specialist, the designer of electron tubes. Similarly, the analyst of wave propagation phenomena can apprehend the problems of the man working with industrial heating equipment only to the extent that one wants radiation to escape and the other does not, while the audio man and the microwave man are a billion cycles removed from each other. These are, admittedly, only arbitrary and singular examples; but they can be elaborated in manifold fashion, and they illustrate the problems of community of interest in our recently incredibly expanded art.

The fundamental consideration, however, is that underneath it all each of these individual specialists is linked with each by his use of the same basic phenomena—electronics and radiation. That underlying binding force is the thing that is The Institute of Radio Engineers. It is essential to the self-interest of all that this force be maintained, that unity of purpose and of seeking shall be preserved, that research findings and design applications and, above all, professional zeal be sustained and disseminated and utilized to strengthen the whole, both in terms of organization and in extension of common knowledge. The question is how to accomplish this, and yet allow full exchange of mutual interest and information and enthusiasm among those most intimately connected with specific phases of the specialized directions in which this broad art of ours is growing.

To answer that question is the objective of the Professional Groups System.

Progress in Group Organization

As was stated above, two Groups already have been approved. The Audio Group was established on June 2, 1948, its purview being the "technology of communication at audio frequencies and with the audio-frequency portion of radio systems." Those interested in this Group should communicate with O. L. Angevine, Jr., c/o Stromberg-Carlson Co., Rochester, N. Y.

The second Group to be approved on July 7, 1948, was the Broadcast Engineers Group, encompassing not only "engineers associated with individual radio stations, broadcast networks, operating companies and broadcast consultants," but also including "government engineers associated with broadcasting, and engineers in commercial and other laboratories and manufacturing companies whose prime interest is in the development, design, production, operation and maintenance of broadcast equipment used by the various classes of broadcast stations." This professional group is intended to encompass the engineers of "any FM, AM, Facsimile, TV, International, Relay, Experimental, Emergency or any other associated class of broadcast stations." Those interested in this Group should communicate with Orrin W. Towner, c/o Radio Station WHAS, Louisville, Ky.

Other fields of activity in which interest in the formation of Groups has been expressed include wave propagation, antennas, telemetering, electronic computers, quality control, and standardization.

How to Organize a Group

For the benefit of those who are interested in organizing Groups in their particular fields of interest, the following step-by-step procedure has been outlined by the Professional Groups Committee. (Page numbers refer to the Group Manual.)

1. Write to the Executive Secretary of the Institute, for general information on the formation of Groups. (See page 1, Initial Inquiry, in the Manual.)
2. Communicate with others interested in the proposed field to get sufficient members willing to promote the Group and over 25 prepared to sign the petition. (See page 1, Promoting Petitioners.)
3. Prepare the petition. (See pages 1, 2, Preparing Petition, and pages 13, 22.)
4. Forward the petition to the Chairman, Committee on Professional Groups, through Institute Headquarters. (See page 2.)
5. Foward suggested names for the initial Administrative Committee to the Chairman, Committee on Professional Groups, through Institute Headquarters. (See pages 2, 14.)
6. After notification by Institute Headquarters of the appointment of the Administrative Committee by the Committee on Professional Groups, the Administrative Committee elects a chairman, notifies Institute Headquarters of the result of the election, and draws up a constitution. (See pages 2A, 15, and 19.)
7. Send the constitution to the Chairman, Committee on Professional Groups, through Institute Headquarters, for approval by that Committee and by the Executive Committee of the Institute. (See page 2A.)
8. The Administrative Committee draws up bylaws and appoints committees. (See pages 2A, 15, 16, and 21.)
9. Forward a copy of the bylaws and names of officers and committee chairmen to the Executive Secretary for his information. (See page 21.)

Note particularly that, once a petition bearing 25 signatures has been filed and a national Group approved in any field of interest, any number of members of any section, however small, may join the Group and hold meetings. As few as two members might conceivably hold local meetings. The only restriction is that all meetings of all Groups must be open to all members of IRE.

Industrial Engineering Notes[1]

Continued Use of Loran in Northeast Atlantic Asked

The State Department has notified the International Telecommunication Union that aids to radio navigation which are suitable for the Northeast Atlantic area and designed for operation in the allotted frequency bands under international regulations cannot be made available by July 1, 1949. The Department said that due to the great use now being made of loran in the area by both air and surface craft, it is the government's opinion that standard loran should be continued in the Northeast Atlantic area until an acceptable substitute which can be agreed upon internationally is in operation.

New Transducer Development

A highly sensitive mechano-electrical transducer, which transforms slight displacements into large changes of resistance, current, or voltage, is being developed by the Bureau of Standards. The active element of the device is a helical or conical spring wound in such a way that the initial tension varies slightly along its length. Thus, when the ends of the spring are pulled apart, the turns separate one by one, rather than simultaneously.

The October issue of the Technical New Bulletin, Standards Bureau publication, has printed a description of the new development. Copies of this monthly magazine may be obtained at 10 cents each from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

FCC Issues Life-Saving Revisions

The FCC has issued a public notice (No. 27112) containing an outline of the revisions of the International Convention for the Safety of Life at Sea, 1929, which were adopted at London last June, and which become effective on January 1, 1951. Copies of the outline and further information on the radio aspects of the revised Convention may be obtained from the Secretary of the FCC, Washington 25, D. C.

Television and Radio Production and Sales Rise as Excise Tax Collections Fall

RMA member-companies manufactured 64,953 television receivers in August, thus setting up a new monthly record which represents an increase of almost 10,000 over the July output. FM-AM set production totalled 110,879 in August, the largest output of this type of receiver since last March. Radio receiver production of all types totalled 870,044 in August as compared with 627,349 in July. Production of automobile and portable radios aggregated 256,594 and 178,323, respectively.

Indicative of the sharp increase in television receiver production during the first half of 1948, sales of cathode-ray tubes to set manufacturers rose more than 68 per cent during the second quarter over sales in the first quarter. During the first six months of 1948 cathode-ray tube sales to manufacturers totalled 426,469, with a value of $10,250,218, as compared with sales during the entire year of 1947 of 253,035 units, valued at $7,218,358.

With U. S. government purchases accounting for 71 per cent of the total, sales of radio and television transmitting and communications equipment by RMA member-companies aggregated $50,318,006 during the second quarter of 1948, and brought sales of this type to $80,346,321 for the first six months of this year.

August collections of the 10 per cent excise tax on radios and phonographs and certain of their components dropped below the collections of both July, 1948, and August, 1947, the U. S. Bureau of Internal Revenue reported. Collections of the radio excise tax in August amounted to $3,927,009.08, as compared with $4,060,785.34 last July and $5,084,018 in August a year ago.

FM Developments

The FCC has denied petitions presented by Major Edwin H. Armstrong and the FM Association, seeking to reopen the record in proceedings concerning the sharing of television channels. Major Armstrong's petition requested that the FCC order of May 5, 1948, be reconsidered "insofar as it denies requests made at the hearing that a portion of the 44- to 50-Mc band be allocated to FM broadcast relay purposes, and insofar as it required FM broadcast stations operating in the 44- to 50-Mc band to discontinue operation by December 31, 1948." The FM petition sought to extend the FCC proposed discontinuance date of FM operation in the 44- to 50-Mc band until December 31, 1950.

Since the last issue of the Proceedings, 30 FM stations have gone on the air, bringing the total to 673 stations, which includes 24 noncommercial outlets. New stations have begun operation in the following states: *Ala.*, Mobile (WMOB-FM), Montgomery (WMGY-FM); *Cal.*, San Diego (KFSD-FM); *D. C.*, Washington (WOL-FM and WCFM); *Ga.*, Atlanta (WSB-FM) and La Grange (WLAG-FM); *Iowa*, Des Moines (KCBC-FM); *Ill.*, Chicago (WMAQ-FM) and Woodstock (WILA); *Ind.*, Warsaw (WRSW); *Ky.*, Ashland (WCMI-FM); *Md.*, Cumberland (WTBO-FM); *Mass.*, Boston (WNAC-FM), Lowell (WLLH-FM), Lynn (WLYN-FM); *N. Y.*, Endicott (WENE-FM), Massena (WMSA-FM), New Rochelle (WGNR), Troy (WEVR); *Ohio*, Canton (WAND-FM), Columbus (WHKC-FM); *Pa.*, Uniontown (WNIQ); *Tex.*, Longview (KLTI-FM); *Wash.*, Tacoma (KTNT); *W. Va.*, Parkersburgh (WPAR-FM); and *Wis.*, Madison (WISC-FM), St. Cloud (KFAM-FM), and Wisconsin Rapids (WFHR-FM).

Canadian Radio News

The Canadian RMA Service Committee has published four bulletins for service technicians: "The Treatment of Hum in Radio Receivers," "FM Servicing" (in two parts), "Capacitor Colour Coding," and "Antennae." The earlier bulletin, "Trouble Shooting in Radio Receivers," has been reprinted, and the Committee is now preparing bulletins on "Alignment" and "An Open Letter to Radio Service Technicians," the latter of which covers basic theory . . . . R. M. Brophy, Chairman of the Canadian RMA Industrial Mobilization Committee, has been requested by the Industrial Defence Board and Canadian Ordnance Association to organize and convene an Industrial Defence Preparedness Committee on Communications and Electronics. Its principal task will be to effect liaison among the Armed Services, the arsenals, and industry. . . . Radio set sales in Canada have been dropping. Sales for the first seven months of this year totalled 224,762 units worth $20,069,485, as against 429,234 units worth $27,707,066 for the first seven months of 1947 . . . . Imports and exports also have been decreasing. During the first six months of this year, 17,800 sets valued at $206,988 were imported, as compared with 44,775 sets valued at $1,559,517 imported during the first half of 1947. Canada exported 441 sets worth $15,146 for the first six months of 1948; whereas she exported 21,597 sets worth $781,672 in the first half of 1947 . . . . One hundred and twenty-five members and guests attended a joint IRE-Canadian RMA Golf Tournament and Dinner at the Cutten Fields Golf Club in Guelph, Ont., on September 17, J. R. Longstaffe, a member of both the RMA and the Toronto Section of the IRE, acted as master of ceremonies and presided at the dinner and evening program.

RMA Holds Conference; Publishes Radio Booklet

Aggressive action to develop television and to expedite the adoption by the government of a mobilization plan of the radio and electronics industry highlighted a three-day RMA fall conference from October 6 to 8, at the Roosevelt Hotel in New York City.

Among the other matters on its agenda, the Board of Directors authorized legal action to contest the validity of a Pennsylvania state license tax on taverns equipped with television receivers, established a special committee to confer with the FCC regarding pending study of future expansion of television services into the UHF band, authorized President Max F. Balcolm to set up a committee representing the Set, Tube, and Transmitter Divisions, and authorized the Engineering Department to work with the Export Committee in the promotion of American television standards and equipment in foreign countries.

The RMA has published a booklet, "Classroom Radio Receivers," which is being distributed to 45,000 educators throughout the country. The work of the U. S. Office of Education and the RMA Joint Committee on Specifications for School Audio Equipment, the pamphlet is the third in a series of such studies, the other two being "School Sound Systems" and "School Sound Recording and Playback Equipment."

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[1] The data on which these Notes are based were selected, by permission, from "Industry Reports," issues of September 17 and 24, and October 1, 11, and 15, published by the Radio Manufacturers' Association of America; and from the September 14 issue of "News," published by the Radio Manufacturers' Association of Canada. The helpful attitude of both of these organizations is hereby gladly acknowledged.

Sections

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W. A. Edson Georgia School of Tech. Atlanta, Ga.	Atlanta December 17	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
	Baltimore	J. W. Hammond 4 Alabama Ct. Baltimore 28, Md.
John Petkovsek 1015 Ave. E Beaumont, Texas	Beaumont— Port Arthur	C. E. Laughlin 1292 Liberty Beaumont, Texas
R. W. Hickman Cruft Laboratory Harvard University Cambridge, Mass.	Boston	A. F. Coleman Mass. Inst. of Technology 77 Massachusetts Ave. Cambridge, Mass.
G. E. Van Spankeren San Martin 379 Buenos Aires, Arg.	Buenos Aires	A. C. Cambre San Martin 379 Buenos Aires, Arg.
J. F. Myers 249 Linwood Ave. Buffalo 9, N. Y.	Buffalo-Niagara December 15	R. F. Blinzler 558 Crescent Ave. Buffalo 14, N. Y.
G. P. Hixenbaugh Radio Station WMT Cedar Rapids, Iowa	Cedar Rapids	V. R. Hudek Collins Radio Co. Cedar Rapids, Iowa
K. W. Jarvis 6058 W. Fullerton Ave. Chicago 39, Ill.	Chicago December 17	Kipling Adams General Radio Co. 920 S. Michigan Ave. Chicago 5, Ill.
C. K. Gieringer 3016 Lischer Ave. Cincinnati, Ohio	Cincinnati December 14	F. W. King RR 9 Box 263 College Hill Cincinnati 24, Ohio
F. B. Schramm 2403 Channing Way Cleveland 18, Ohio	Cleveland December 23	J. B. Epperson Box 228 Berea, Ohio
Warren Bauer 376 Crestview Rd. Columbus 2, Ohio	Columbus January 14	George Mueller Electrical Eng. Dept. Ohio State University Columbus, Ohio
S. E. Warner Aircraft Electronics Assoc. 1031 New Britain Ave. Hartford 10, Conn.	Connecticut Valley December 16	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
J. G. Rountree 4333 South Western Blvd. Dallas 5, Texas	Dallas-Ft. Worth	J. H. Homsy Box 5238 Dallas, Texas
George Rappaport 132 East Court Harshman Homes Dayton 3, Ohio	Dayton December 16	C. J. Marshall 1 Twain Place Dayton 10, Ohio
C. F. Quentin Radio Station KRNT Des Moines 4, Iowa	Des Moines- Ames	F. E. Bartlett Radio Station KSO Old Colony Bldg. Des Moines 9, Iowa
A. Friedenthal 5396 Oregon Detroit 4, Mich.	Detroit December 17	N. C. Fisk 3005 W. Chicago Ave. Detroit 6, Mich.
E. F. Kahl Sylvania Electric Products Emporium, Pa.	Emporium	R. W. Slinkman Sylvania Electric Products Emporium, Pa.
W. H. Carter 1309 Marshall Ave. Houston 6, Texas	Houston	J. C. Robinson 1422 San Jacinto St. Houston 2, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	Indianapolis	Eugene Pulliam 931 N. Parker Ave. Indianapolis, Ind.
Karl Troeglen KCMO Broadcasting Co. Commerce Bldg. Kansas City 6, Mo.	Kansas City	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
R. W. Wilton 71 Carling St. London, Ont., Canada	London, Ontario	G. H. Hadden 35 Becher St. London, Ont., Canada
Walter Kenworth 1427 Lafayette St. San Gabriel, Calif.	Los Angeles October 19	R. A. Monfort L. A. Times 202 W. First St. Los Angeles 12, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	Louisville	D. C. Summerford Radio Station WKLO Henry Clay Hotel Louisville, Ky.
F. J. Van Zeeland Milwaukee School of Eng. 1020 N. Broadway Milwaukee, Wis.	Milwaukee	H. F. Loeffler Wisconsin Telephone Co. 722 N. Broadway Milwaukee 1, Wis.
K. R. Patrick RCA Victor Div. 1001 Lenoir St. Montreal, Canada	Montreal, Quebec January 12	S. F. Knights Canadian Marconi Co. P.O. Box 1690 Montreal, P. Q., Canada
L. A. Hopkins, Jr. 1711 17th Loop Sandia Base Branch Albuquerque, N. M.	New Mexico	T. S. Church 637 La Vega Rd. Albuquerque, N. M.
J. W. McRae Bell Telephone Labs. Murray Hill, N. J.	New York January 5	R. D. Chipp DuMont Telev. Lab. 515 Madison Ave. New York, N. Y.
C. G. Brennecke Dept. of Electrical Eng. North Carolina State College Raleigh, N. C.	North Carolina- Virginia	C. M. Smith Radio Station WMIT Winston-Salem, N. C.
W. L. Haney 117 Bourque St. Hull, P. Q.	Ottawa, Ontario November 16	G. A. Davis 78 Holland Ave. Ottawa, Canada
A. N. Curtiss Radio Corp. of America Camden, N. J.	Philadelphia January 6	C. A. Gunther Radio Corp. of America Front & Cooper Sts. Camden, N. J.
M. A. Schultz 635 Cascade Rd. Forest Hills Borough Pittsburgh, Pa.	Pittsburgh January 10	E. W. Marlowe Union Switch & Sig. Co. Swissvale P.O. Pittsburgh 18, Pa.
O. A. Steele 1506 S.W. Montgomery St. Portland 1, Ore.	Portland	F. E. Miller 3122 S.E. 73 Ave. Portland 6, Ore.
A. V. Bedford RCA Laboratories Princeton, N. J.	Princeton	L. J. Giacoletto 9 Villa Pl. Eatontown, N. J.
K. J. Gardner 111 East Ave. Rochester 4, N. Y.	Rochester December 16	Gerrard Mountjoy Stromberg-Carlson Co. 100 Carlton Rd. Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	Sacramento	W. F. Koch 1340 33rd St. Sacramento 14, Calif.
G. M. Cummings 7200 Delta Ave. Richmond Height 17, Mo.	St. Louis	C. E. Harrison 818 S. Kings Highway Blvd. St. Louis 10, Mo.
C. L. Jeffers Radio Station WOAI 1031 Navarro St. San Antonio, Texas	San Antonio	H. G. Campbell 233 Lotus Ave. San Antonio 3, Texas
C. N. Tirrell U. S. Navy Electronics Lab. San Diego 52, Calif.	San Diego January 4	S. H. Sessions U. S. Navy Electronics Lab. San Diego 52, Calif.
F. R. Brace 955 Jones St. San Francisco 9, Calif.	San Francisco	R. A. Isberg Radio Station KRON 901 Mission St. San Francisco 19, Calif.
W. R. Hill University of Washington Seattle 5, Wash.	Seattle January 13	W. R. Triplett 3840—44 Ave. S.W. Seattle 6, Wash.
F. M. Deerhake 600 Oakwood St. Fayetteville, N. Y.	Syracuse	S. E. Clements Dept. of Electrical Eng. Syracuse University Syracuse 10, N. Y.
A. R. Bitter 4292 Monroe St. Toledo 6, Ohio	Toledo	J. K. Beins 435 Kenilworth Ave. Toledo 10, Ohio

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G. P. Adair 1833 "M" St. N.W. Washington, D. C.	WASHINGTON	H. W. Wells Dept. of Terrestrial Magnetism Carnegie Inst. of Washington Washington, D. C.
J. C. Starks Box 307 Sunbury, Pa.	WILLIAMSPORT	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

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A. M. Glover RCA Victor Div. Lancaster, Pa.	LANCASTER (Philadelphia Subsection)	C. E. Burnett RCA Victor Div. Lancaster, Pa.
H. A. Wheeler Wheeler Laboratories 259–09 Northern Blvd. Great Neck, L. I., N. Y.	LONG ISLAND (New York Subsection)	M. Lebenbaum Airborne Inst. Lab. 160 Old Country Rd. Box 111 Mineola, L. I., N. Y.
L. E. Hunt Bell Telephone Labs. Deal, N. J.	MONMOUTH (New York Subsection)	G. E. Reynolds, Jr. Electronics Associates, Inc. Long Branch, N. J.
J. B. Minter Box 1 Boonton, N. J.	NORTHERN N. J. (New York Subsection)	A. W. Parkes, Jr. 47 Cobb Rd. Mountain Lakes, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (Chicago Subsection) December 16	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
R. M. Wainwright Elec. Eng. Department University of Illinois Urbana, Illinois	URBANA (Chicago Subsection)	M. H. Crothers Elec. Eng. Department University of Illinois Urbana, Illinois
S. S. Stevens Trans Canada Airlines Box 2973 Winnipeg, Manit., Canada	WINNIPEG (Toronto Subsection)	S. G. L. Horner Hudson's Bay Co. Brandon Ave. Winnipeg, Manit., Canada

Books

Microwave Transmission Design Data, by Theodore Moreno

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y., 241 pages, 6-page index, x pages, 92 figures. 6¼×9¼. $4.00.

This volume, intended as a reference handbook for radio engineers working with microwave transmission lines and associated components, is a greatly revised version of a wartime confidential publication of the same title, put out in 1944 by the Sperry Gyroscope Co. Such matters as generation, reception, propagation, and measurement techniques are not included, but several chapters pertaining to elementary basic theory have been added, as well as considerable new technical information to bring the book up-to-date. On the other hand, some data have been left out, and all graphs have been reduced in size and redrawn with fewer co-ordinate lines. Perhaps the publisher saved a few pages in this way, but he made the book a less useful tool.

The principal subjects covered are transmission-line relations and charts; general formulas for coaxial lines and waveguides; coaxial-line and waveguide obstacles, junctions, and other structures; and cavity resonators. The extensive tables listing the dimensions and properties of standard coaxial lines and waveguides, as well as the table giving the microwave properties of over a hundred dielectric materials, should prove very valuable to the practicing microwave engineer.

The author has covered his topics well and produced a text that is clearly written and easily understandable. Although there are some typographical errors, very few should result in confusion. More confusing is the inconsistent use of units, which is particularly troublesome in the chapter on cavity resonators. The consistent use of mks units throughout the volume would have made it an even more useful guide.

SEYMOUR B. COHN
Sperry Gyroscope Co.
Great Neck, N. Y.

Microwave Duplexers, by Louis D. Smullin and Carol G. Montgomery

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 430 pages, 7-page index, xiv pages, 392 figures. 6¼×9¼. $6.50.

This is another of the MIT Radiation Laboratory series of books and, like the others, it attempts to summarize within the confines of a single volume the information accumulated at the Radiation Laboratory during the war years. The rapid progress made during this period, together with the emphasis on results rather than on understanding, inevitably lead to the accumulation of engineering information rather than of scientific knowledge. This trend was particularly pronounced in the field of microwave duplexers, where the needs for operating devices were acute and where the poorly understood gaseous-discharge TR switch was found to be surprisingly effective. The authors are to be commended for the excellent job which they have done in compiling this information. It would be manifestly unfair to criticize them personally for the rather unsatisfactory state of knowledge which is thereby revealed. As a matter of fact, the advances made during the war years in the engineering applications of gas discharges were so substantial that a well-written summary such as this is a real aid to anyone wishing to conduct a more leisurely inquiry into the fundamental reasons for the observed effects.

After an introductory chapter, the book considers in turn the linear theory of high-Q TR tubes, the band-pass TR tubes, and the characteristics of ATR switches at low levels. This is followed by a rather long chapter on microwave gas discharges which covers those scientific aspects of the TR switch which are characteristic of this device and are not common to other microwave devices. Chapter six discusses the operation of TR and ATR tubes at high powers, and considers this subject largely from the user's point of view. The principles of branched duplex circuits are then discussed, using matrix notation. This chapter constitutes a worth-while contribution to the general solution of complex branching circuits. The design details of practical branched duplexers are next described, and some consideration given to balanced duplexers. The concluding chapter on measurement techniques is a good summary of some of the more important aspects of this problem.

The joint effort of six contributing authors, the volume inevitably is inconsistent in spots. Considering the fact that a whole chapter has already been devoted to the linear theory, the analysis starting on page 71 is surprisingly elementary. Furthermore, the authors of chapter two and chapter three do not agree on the equivalence or lack of equivalence of different methods of defining Q, as may be seen by comparing the material on pages 10 and 72. Such inconsistencies are more apparent in the earlier chapters of the book, where the chapter order makes it necessary to assume some knowledge of material which is not defined until a later chapter. There is also a certain amount of looseness in terminology to be noted, such as the frequent use of the word "susceptance" to mean "admittance."

On the whole, however, the book is very well written, and a high standard of excellence in style, format, and typography is maintained throughout. The authors have been most fair in their assignment of credit to persons outside of the Radiation Laboratory. Nevertheless, there is a certain lack of historical perspective evident in the book, and a persistent implication that the Radiation Laboratory was the fountainhead of all knowledge. But, with the few defects already mentioned, the book is a "must" for anyone interested in the field.

A. L. Samuel
University of Illinois
Urbana, Ill.

Microwave Transmission Circuits, edited by George L. Ragan

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 624 figures, 45 tables, 716 pages, 9-page index, xvii pages. 6×9 inches. $8.50.

This book, number nine in the Radiation Laboratory Series of 28 volumes, deals with the problems of power transmission from one place to another at microwave frequencies, and its contents are applicable to almost all of the problems that come up in the design of microwave circuits. Seven authors have combined to produce a well-written and interesting volume, of value to both the scientist and the layman. Copious illustrations and tables supplement the text. Taking into consideration the amount of time between the preparation of a book and its publication (about a year), the mathematical treatment is up-to-date.

In the first two chapters, Mr. Ragan defines the microwave region referred to in this volume as extending from 2 to 25 kM, and offers a discussion of elementary line theory. The second chapter is presented with especial clarity: after each step of new theory a short interpretation of the terms of the equations developed are given. Divided into four parts, the chapter includes conventional line theory; transmission lines as guides for electromagnetic waves; transmission-line charts and impedance-matching, and design procedure.

R. M. Walker covers materials and construction techniques in chapter three. He discusses metallic materials, finishes and electroplating, dielectric materials, and pressurization problems. As far as this reviewer knows, a great deal of this material is completely new, in that it has not been published before.

In chapter four, Mr. Ragan and Mr. Walker both discuss rigid transmission lines, covering coaxial lines; waveguide couplings; corners, circular bends and twists; impedance matching; pressurizing windows; and voltage breakdown.

Chapter five is again presented by two authors, F. E. Ehlers and F. P. Worrell, who give an up-to-date review of flexible coupling units and lines.

In chapter six Mr. Ragan covers transition units, including transitions for one coaxial line to another, coaxial to waveguide, one waveguide to another, and one waveguide mode to another. Mr. Ehlers and F. L. Neimann discuss approximately twelve different kinds of motional joints in chapter seven; and in chapter eight Neimann and Ragan cover tuners, power dividers, and switches, describing about thirteen different kinds of tuners.

The concluding two chapters by R. M. Fano and A. W. Lawson discuss the theory of microwave filters and design of microwave filters, each of the chapters being followed by a bibliography. Using the information contained in these last chapters, the reader should be able to design microwave filters.

The one significant defect in the book is its inadequate index, in that the table of contents is the more complete reference. Otherwise, it is an excellent and useful presentation of the subject.

Allen F. Pomeroy
Bell Telephone Laboratories
New York 14, N. Y.

Antenna Manual, by Woodrow Smith

Published (1948) by Editors and Engineers, 1300 Kenwood Rd., Santa Barbara, Calif. 301 pages, 5-page index, iii pages, 155 figures. 6½×9½.

A clearly written, well-illustrated book, the "Antenna Manual" is couched in simple language manifestly addressed to the amateur experimenter.

The first quarter of the book is devoted to an elementary discussion of propagation, ionosphere reflection, and related subjects. Only about 40 per cent of the manual is actually devoted to antennas. This portion is a compilation of descriptive material on the construction and salient characteristics of antennas, most of which are varieties appealing to the amateur. The antennas are classified into low, medium, high, very-high, and ultra-high-frequency types. No information is included on microwave antennas or waveguides.

The remaining thirty-five per cent of the book has chapters on transmission lines, coupling methods, and measurements. Throughout, the discussion is general, with the accent on the practical. Occasionally the explanations are over-simplified, and even such basic aids as transmission-line impedance charts are omitted.

The book is nonmathematical, including only four very simple equations, and contains only two references. It is regrettable that the author has used considerable material without indicating its source. As a result, much of the data lacks authoritativeness, and its value is depreciated in many instances.

In spite of this minor deficiency, the book fills a definite need, and is recommended for those desiring a readable, elementary, and concise discussion of some of the more practical aspects of radio propagation antennas.

John D. Kraus
Ohio State University
Columbus, O

IRE People

William L. Everitt (A'25–M'29–F'38), one of America's foremost authorities on electronics, has been appointed Dean of the University of Illinois' College of Engineering. The appointment will be effective on September 1, 1949.

Born in Baltimore, Md., on April 14, 1900, Professor Everitt received the E.E. degree from Cornell University in 1922, acting meanwhile as instructor of electrical engineering during the last two years of his studies there. In 1922 he joined the North Electric Manufacturing Co. of Galion, Ohio, where he took charge of the design and development of their relay automatic public switchboard exchanges. Two years later he became an instructor in electrical engineering at the University of Michigan, remaining there until he received the M.S. degree in 1926 when he went to Ohio State University to teach communication engineering. Ohio State awarded him the Ph.D. degree in 1933 and the following year promoted him to a full professorship. Meanwhile, during the summer vacations, from 1925 to 1930, he had served with the department of development and research of the American Telephone and Telegraph Co. and also acted as consultant for a number of broadcast stations and radio manufacturing companies.

In 1942 Dr. Everitt was granted a leave of absence from the University in order to act as Director of Operational Research with the U. S. Army Signal Corps. Three years later, while he was still working with the Army, he was appointed head of Ohio State's electrical engineering department in absentia. Upon the conclusion of the war he resumed his work with the University.

Dr. Everitt is the author of a number of texts and articles on electrical engineering. A former director of the Institute, and a member of numerous Committees, he was President in 1945. He is a Fellow of the AIEE and a member of the National Council of Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

◆

C. Russell Cox (A'29–SM'48), formerly sales manager and chief engineer of the Andrew Corp. in Chicago, has been chosen for the newly created office of director of sales and engineering.

Born in Chicago, Ill., in 1916, Mr. Cox attended the University of Chicago, from which he received the B.S. degree in 1937 and the M.S. degree in physics in 1939. The following year he joined the Andrew Corp.

Mr. Cox is the author of numerous technical papers on coaxial transmission lines. He is a member of Phi Beta Kappa, and has served on the RMA subcommittees to develop standards on coaxial transmission lines.

A. Hoyt Taylor (M'16–F'30), one of the U. S. Government's most distinguished scientists, has retired after more than 31 years of continuous service to the Navy. In stepping down from the Naval Research Laboratory's top civilian position—Chief Consultant for Electronics—Dr. Taylor brings to a close a half-century devoted to research in radio.

Dr.Taylor was born in Chicago on New Year's Day in 1879. While in high school at Evanston, Ill., he became interested in electrical engineering and decided to specialize in that field at Northwestern University, from which he received the B.S. degree in 1902. After graduation, he became an instructor in physics, first at Michigan State College and later at the University of Wisconsin.

In 1908 he returned to his studies, qualifying for the doctorate at Goettingen in Germany. The following year he joined the University of North Dakota as head of the physics department. He remained there for eight years, concentrating on experimental work in radio.

When the entry of the United States in the first World War seemed imminent, good radio engineers were at a premium. The Navy persuaded Dr. Taylor to apply for a Naval Reserve Commission and he was sworn in as a lieutenant shortly before war was declared. For almost five years he remained on active duty, advancing to the rank of commander. After the war he assumed direction of the U. S. Naval Aircraft-Radio Laboratory.

When the Naval Research Laboratory was established in 1923, the staff of its original radio division was formed by combining the Naval Aircraft Laboratory with two other experimental units then located in Washington. Dr. Taylor was appointed superintendent, and in this position significantly influenced the entire development of U. S. naval radio. It was largely as a result of successful experiments and demonstrations carried out at the laboratory that the U. S. Navy led the world in adopting high frequencies for improving its communication system. Dr. Taylor's own contributions to radio include the development and the application of the propagation of high-frequency energy, the quartz crystal oscillator, and high- and super-frequency radio communication systems.

It is, however, for his share in the discovery and development of radio-detection techniques now called radar that Dr. Taylor will probably be best remembered. Beginning as early as 1922, he and an assistant conducted a series of experiments which disclosed the amazing possibility of detecting and tracking targets by radio waves. Later the Naval Research Laboratory's radio engineers under Dr. Taylor's guidance perfected the weapon which was to have such a profound effect on U. S. Naval tactics in World War II. For this work he was awarded the U. S. Medal for Merit in 1944.

Aside from his achievements for the Navy, Dr. Taylor has won wide recognition in the world of science. In 1927 he was awarded the IRE's Morris Liebmann Memorial Prize; in 1929 he was elected President of the Institute; in 1942 he received the IRE's Medal of Honor, and in the same year was selected to receive the Franklin Institute's John Scott Medal. He is a fellow of the American Physical Society, the American Association for the Advancement of Science, and the AIEE; and a member of the Geophysical Society.

❖

L. Grant Hector (A'26–SM'39) has been elected vice-president in charge of technical operations of the Sonotone Corp., where he has been employed as director of research since the end of the war.

Dr. Hector was born on December 15, 1894, at Clarendon, Pa. In 1920 he received the A.B. degree from Oberlin College. While he was doing graduate work at Oberlin, Dr. Hector worked as a private research assistant to A. P. Wills, leaving in 1922, when he received the M.A. degree to become an instructor of physics at Oberlin. The following year he became Tyndall Fellow at Columbia and, upon being awarded the doctorate in 1924, he became an assistant professor of physics at the University of Buffalo. Three years later he was promoted to full professor.

After engaging in the development of the proximity fuze for the U. S. Government's Office of Scientific Research and Development from 1941 to 1943, Dr. Hector jointed the National Union Radio Corp. as director of engineering. At the same time he worked as a dollar-a-year consultant on the production of subminiature radio tubes used in the radio proximity fuze for the War Production Board from 1943 to 1945. When the war was over, Dr. Hector joined the Sonotone Laboratories in Elmsford, N. Y. to work on the research and development of tiny vacuum tubes for peacetime use in hearing aids.

Dr. Hector has written a number of papers, articles, and books dealing with magnetic, dielectric, and acoustical measurements by electronic techniques. Text books by Dr. Hector include "Modern Radio Receiving" (1927), "Introductory Physics" (1933), and "Electronic Physics" (1943). His papers and articles have been delivered before the American Physical Society and the International Scientific Union and have been printed in the PROCEEDINGS OF THE I.R.E., *Physical Review*, and the *Review of Scientific Instruments*. He is a fellow of the American Physical Society and the American Association for the Advancement of Science, a member of the Acoustical Society of America and of Sigma Xi.

❖

Joseph P. Maxfield (SM'47) formerly associated with the Bell Telephone Laboratories for thirty-one years, has accepted the position of Superintending Scientist at the U. S. Navy Electronics Laboratory in San Diego, Calif., where he will be in charge of the laboratory's scientific and technical research in the field of radio, radar, and sonar.

A distinguished pioneer in the field of electronics, Mr. Maxfield has played a leading role in the development of radio broadcasting, motion-picture sound systems, and record reproducing equipment. After receiving the Bachelor of Science degree from the Massachusetts Institute of Technology in 1910, Mr. Maxfield taught there for four years. He then joined the engineering department of the Western Electric Co. to engage in research work on the physical and electrical properties of microphone contacts. An outgrowth of this work was the "Maxfield Transmitter," one of the first microphones used in radio broadcasting.

During the first World War, Mr. Maxwell helped develop methods for the acoustic detection of aircraft and the sound ranging of artillery, and resumed work on this problem with a staff of twenty-five leading scientists and engineers at Duke University during World War II. From 1919 to 1926 he made significant contributions to the art of transmitting, recording, and reproducing high-quality sound. The techniques which he evolved as a result of this work have been applied directly to public-address systems, the design of broadcasting studios, the development of microphone techniques, and the adaptation of sound recording to motion pictures. In the field of home musical entertainment, Mr. Maxfield directed engineering and research on the orthophonic phonograph for the Victor Talking Machine Co.

Upon his return to Bell in 1929, Mr. Maxfield joined the Electrical Research Products, Inc., and made many contributions which advanced the quality of sound motion pictures during the ensuing years. From 1936 to 1943 Mr. Maxfield was director of commercial engineering of ERPI and was responsible for developing equipment to measure, analyze, and record sound and vibration. He also was responsible for a study of airplane vibration and flutter, work requested for the Civil Aeronautics Authority, and he designed the acoustic treatment of several buildings at the New York World's Fair.

In 1942 Mr. Maxfield returned to the Bell Laboratories, but six months later was given a leave of absence to become director of the National Defense Research Committee's Division of Physical Research at Duke University. There he and his staff made significant contributions to the problem of sound ranging of artillery. Returning again to Bell Laboratories in 1944, this time as a member of the Acoustic Products Development Group, he worked with liveness as related to sound-pickup techniques until his retirement in the fall of 1947. In 1948 he became associated with the Altec-Lansing Corp. as a consulting engineer.

Mr. Maxfield is a Fellow of the American Physical Society, and a member of the AIEE, the Acoustic Society of America, and the Society of Motion Picture Engineers. He is the author of numerous papers and articles on acoustics, electrochemistry, and physics, and the inventor of numerous devices used in telephony, radio broadcasting, and the electrical recording and reproduction of sound on film.

❖

John S. Brown (A'44–M'45), former assistant chief engineer at the Andrew Corp., has been promoted to chief engineer. Associated with the company since 1943, Mr. Brown has been active in development and manufacturing design work on transmission lines and antenna equipment.

Louis Cohen, (F'15), distinguished engineer, consultant, and inventor of many devices in radio and cable telegraphy, died recently of a heart attack.

Born in Kiev, Russia, in 1876, Dr. Cohen was brought to this country as a boy. He studied at the Armour Institute of Technology in Chicago, receiving the B.Sc. degree in 1901, and subsequently attended the University of Chicago and Columbia University, the latter of which conferred the doctor's degree on him in 1905. That year he joined the staff of the U. S. Bureau of Standards and remained there until 1909 when he joined the Electrical Signaling Co., leaving four years later to open his own consulting practice. From 1916 on he also taught at George Washington University.

Over the years Dr. Cohen became internationally known for his researches into radio and telegraphy. During the first World War he developed for the Navy an instrument that became known as the Cohen receiver. From 1920 to 1924 he acted as consulting engineer for the War Department. Later he served on several international communications commissions, and wrote technical books and papers in the general field of electricity.

While practicing privately as a consulting engineer, he accepted many public assignments. In 1921 he was a United States delegate to the International Conference on Electrical Communications in Paris. Later he served on an advisory board at the Washington Conference on Limitation of Armaments. From 1929 to 1931 he was a technical expert with the German-Austrian Claim Commission.

Emanuel R. Piore (A'37–M'42–SM'43), director of the Physical Sciences Division of the Office of Naval Research, has been granted a year's leave of absence in order to do research work at the Massachusetts Institute of Technology's Electronics Research Laboratory.

An expert in electronics, composite surface physics, and color television, Dr. Piore received the Ph.D. degree in physics from the University of Wisconsin in 1936. After working as a physicist in the RCA Electronics Laboratory, he was placed in charge of the Columbia Broadcasting Co.'s Television Laboratories. From 1942 to 1946 he was a senior physicist for the Bureau of Ships, and served the Deputy Chief of Operations for Air as a specialist on guided missiles. Since 1946 Dr. Piore has directed the fundamental research program in the physical sciences sponsored by the Office of Naval Research.

◆

Walter F. Kean (S'41–A'43), who has headed the Broadcast Consulting Division of the Andrew Corp. since its formation in 1944, has becomes sales manager of the same company. Prior to joining Andrew, he was with the Western Electric Co. in Chicago.

◆

George F. Callahan (A'39) has been appointed division engineer of cathode-ray tubes in the General Electric Co.'s tube division. There he will assume the responsibility for all design and application engineering and standardizing activities relating to cathode-ray-tube product lines.

A native of Elk Point, S. D., Mr. Callahan was graduated from the University of Nebraska with the B.S. degree in electrical engineering. Since 1924, he has been engaged in the design, development, and manufacture of lamps and electronic tubes. Joining the Ken-Rad Tube and Lamp Corp. at Owensboro, Ky., in 1939, he worked on the design of receiving tubes and served for four years as the works manager of that company's Bowling Green plant. When General Electric acquired Ken-Rad, it acquired Mr. Callahan along with it. Since that time he has been active in the development of new miniature and cathode-ray receiving tubes.

◆

William Roth Work (A'42), Assistant Director of the Carnegie Institute of Technology's College of Engineering and Science, and one of the school's first faculty members, died recently.

Dr. Work, who had been affiliated with Carnegie since 1905, was born in Steelton, Pa., on May 4, 1881. After attending Wittenberg Academy in Springfield, Ohio, he received the A.B. degree from Wittenberg College in 1902, and the Master of Engineering degree from Ohio State in 1905. Wittenberg awarded him the honorary degree of Doctor of Science in 1920.

Associated with the Westinghouse Electric Corp. from 1905 to 1906, Dr. Work simultaneously served as a part-time instructor of mathematics at Carnegie. In 1906 he was appointed an instructor in electrical engineering on the CIT faculty, and three years later he was named an assistant professor. While holding an associate professorship at Carnegie from 1915 to 1920, Dr. Work was also a member of the Committee on Education and Special Training of the War Department General Staff during World War I. Subsequently he served as veterans' adviser for the school.

Vitally interested in scientific progress, Dr. Work was a Fellow of the AIEE, and a member of the American Society for Engineering Education, the Engineers' Society of Western Pennsylvania, and the American Association of University Professors, as well as belonging to a number of honorary fraternities.

Arnold Everett Bowen (SM'46), research engineer who contributed significantly to the development of radar, died recently following a brief illness.

Mr. Bowen was born on October 21, 1900, in Lowell, Mass. Educated at the Sheffield Scientific School, Yale University, he received the Ph.B. degree, summa cum laude, in 1921. Later he attended the Yale Graduate School, where he served as an assistant in the Department of Physics. In 1923 he joined the Development and Research Department of the American Telephone and Telegraph Co., and, with the entire department, transferred to Bell Laboratories in 1934, remaining with them until his death.

Mr. Bowen did much of the pioneer work in developing a system for transmitting microwaves through hollow guides. This technique made possible new forms of radar, which were used extensively in World War II. His research work and his inventions also furthered the development of microwave devices now used in radio, telephone, and television transmission.

In 1942 Mr. Bowen was commissioned as a major in the Army and worked first on antisubmarine equipment. For a short while he was stationed in Trinidad, and later served as a consultant at Langley Field. He then returned to Bell Laboratories, but was commissioned again, and, as a lieutenant-colonel, served in Washington as officer-in-charge of the Air Forces' Airborne Radar Equipment Board.

Mr. Bowen was a member of the American Physical Society and of Sigma Xi, national science honorary fraternity. Currently he was serving on the IRE Board of Editors.

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Philips B. Patton (A'46) has joined the Lenkurt Electric Co. in San Carlos, Calif., as a field engineer in the carrier division.

A native of California, Mr. Patton attended the University of Maryland and San Francisco State College. He began his communications career with the Western Union Telegraph Co. in 1931. After eleven years, he joined Pan American Airways as flight radio officer. In 1942 he became associated with the Federal Communications Commission, serving for the next two years and again from 1945 to 1946. In the intermediary period he returned to Pan American.

While with the FCC, Mr. Patton served with the Board of War Communications on telegraph and telephone operations, the San Francisco field office of the common-carrier division, and in 1946 was appointed acting chief of the radiotelephone and telegraph section of the division's engineering department. Subsequently he joined the Farnsworth Corp. as technical co-ordinator for the Fort Wayne, Ind., division.

G. E. Van Spankernen

Chairman, Buenos Aires Section

Gerardo Evan van Spankernen was born in Amsterdam, Holland, in 1907, and was educated in that country, receiving his degree as an electrical engineer in 1929. Immediately after his graduation he joined the Philips Co. in Eindhoven, Holland, starting as an assistant in the engineering laboratory, where he worked on electronic measuring instruments. From 1932 until 1933 he was stationed with the Philips Co.'s British Division in London and with their French division in Paris, where he installed electrical measuring equipment and acted as an instructor to explain its usage.

In 1933 he was assigned to the engineering laboratory to develop special radio receivers designed for Swiss usage. Upon completing their development, he was sent to Le Chaux de Fonds in Switzerland to help install the Philips factory there for manufacturing the receivers. Returning to Holland at the end of the year, he was appointed to do a similar job for Philips in Buenos Aires, Argentina.

The beginning of 1934 found Mr. van Spankernen in Argentina, starting the Philips Co.'s new factory there. Until 1938 he took charge of the entire development and manufacture of radio receivers, but, as the plant grew, he found that the engineering laboratory demanded the whole of his attention. After that, as chief engineer, he headed the laboratory which develops radio receivers and transmitters and also includes a chemistry laboratory to test materials and develop manufacturing procedures.

From 1938 on Mr. van Spankernen made a series of short trips to Europe and the United States in order to familiarize himself with the latest developments and procedures. In 1947 he was given the title of "Gerente Tecnico," which is technical manager with power of attorney.

Mr. van Spankernen became an Associate Member of the Institute in 1939 and a Senior Member in 1945.

K. R. Patrick

Chairman, Montreal Section

K. R. Patrick was born in St. John, New Brunswick, in 1915, and received his early education there. When his family moved to New Haven, Conn., he continued his studies there, afterward attending various colleges.

Immediately after graduating from school Mr. Patrick formed and operated an industrial sound company of his own in Massachusetts, returning to Canada when war broke out. He was commissioned a Flying Officer in the RCAF, serving until his discharge in October, 1945, with the rank of Wing Commander. During the war he served as Commanding Officer of the Canadian Air Force's Number One Wireless School in Montreal, and also was in command of the Number Five Radar and Communications Station, the joint Canadian-American-British development and training organization at Clinton, Ont. He received the O.B.E. for his work in radar in 1943, and in 1945 the American Government presented him with the Legion of Merit in recognition of his services in the research and development of guided missiles.

Since the war Mr. Patrick has been connected with the RCA Victor Co., Ltd., of Canada, of which he now manages the engineering products division. In 1947 he was appointed industrial member of the Electronics Advisory Committee of the Defense Research Board and this spring was re-appointed for an additional three years. He is also a member of the executive committee of the National United Services Institute. Becoming a Member of the IRE in 1943, Mr. Patrick was advanced to Senior Membership later the same year. He is a member of the Electronics Advisory Committee for the Ontario Government School of Electronics, the Board of Governors of Sir George Williams College in Montreal, and the Society for the Promotion of Engineering Education.

JTAC Requests Technical Co-operation in Connection with FCC Television Hearings

THE JTAC (Joint Technical Advisory Committee) was formed jointly by The Institute of Radio Engineers and the Radio Manufacturers Association in May, 1948, for the purpose of "rendering additional public service in their fields of activities."

The JTAC objectives are quoted below from the Charter:

"To obtain and evaluate information of a technical or engineering nature relating to the radio art for the purpose of advising Government bodies and other professional and industrial groups. In obtaining and evaluating such information, the JTAC shall maintain an objective point of view. It is recognized that the advice given may involve integrated professional judgments on many interrelated factors, including economic forces and public policy."

The duties of the JTAC are:

"To consult with Government bodies and with other professional and industrial groups to determine what technical information is required to insure the wise use and regulation of radio facilities.

"To establish a program of activity and determine priority among the problems selected by it or presented to it in view of the needs of the profession and the public.

"To sift and evaluate information thus obtained so as to resolve conflicts of fact, to separate matters of fact from matters of opinion, and to relate the detailed findings to the broad problems presented to it.

"To present its findings in a clear and understandable manner to the agencies originally requesting the assistance of the Committee.

"To make its finding available to the profession and the public.

"To appear as necessary before Government or other parties to interpret the findings of the Committee in the light of other information presented."

The first task assumed by JTAC was at the request of the Federal Communications Commission, and was to assist the FCC in its deliberations concerning the future use of frequencies from 475 to 890 Mc for television broadcasting.

With the assistance of IRE and RMA Technical Committees and others, a preliminary study was made and a report was presented to the FCC at the hearing, docket 8976, held on September 20, 1948. As a result of this hearing and in order to arrive at a decision, the FCC has called an engineering conference to be held on November 30, December 1 and 2, 1948. In the correspondence which is reproduced on the following pages, Chairman Wayne Coy of the Commission has asked the JTAC to make preliminary studies in preparation for this conference.

It will be observed that the date of this conference anticipates the date of publication of this issue of the PROCEEDINGS. Members are advised, however, that this promises to be a continuing study and that all pertinent additional information will be of value beyond the established deadline. **It is urged, therefore, that the membership of the Institute co-operate in supplying the Secretary of the JTAC, L. G. Cumming, at the Institute Headquarters, 1 East 79 Street, New York 21, N. Y., with all available data. The attention of all Technical Committees is invited to the importance of the contributions to the preliminary report made by the IRE and RMA Technical Committees.**

Following presentation of the September 20, 1948, report, the chairman of the JTAC offered the further services of the committee to the chairman of FCC. The following reply is reproduced as received:

FEDERAL COMMUNICATIONS
COMMISSION
WASHINGTON 25, D. C.

October 28, 1948

Mr. Philip F. Siling, Chairman
The Joint Technical Advisory Committee,
IRE-RMA
1 East 79th Street
New York 21, New York

Dear Mr. Siling:

I have your letter of October 1, 1948, offering the continued assistance of the JTAC in providing information which will be helpful to the Commission. I quite agree that the major tasks which JTAC undertakes for the Commission should be in response to requests through the Commission rather than through staff contacts. In this way we can assure that emphasis is placed upon the problems which the Commission considers to be the more important.

While the question of JTAC assistance may appear to have lain dormant since the September 20th hearing, I assure you that such is not the case. Your participation in the current proceedings relative to very-high-frequency and ultra-high-frequency television has been one of the principal topics for discussion in the several Commission and staff meetings which we have held on these subjects since that date. I have deferred answering your letter, however, until our plans for proceeding in the television matters had crystallized and a definite proposal could be made.

The more urgent of the two proceedings is of course the one regarding very-high-frequency, involving as it does a freezing of assignments for an indefinite, but we hope a limited, period. It would appear, therefore, that any activity in regard to the ultra-high-frequency situation which will result in a delay in the very-high-frequency considerations should be postponed until a later date.

Although the time is short and it is appreciated that only a small amount of new information can be developed within its limitations, nevertheless, it should be sufficient for collecting and processing certain pertinent information which has been accumulating since the last major revision of the standards.

I am forwarding herewith a copy of FCC 48-2256, "Notice of Further Proposed Rule Making" in Dockets 8975, 8976, and 9175, outlining the procedure to be followed for VHF television. There is also enclosed a copy of FCC 48-1966, the notice of issuance of four reports in preparation for the engineering conferences which are part of the procedure.

Ten additional copies of the notices and ten copies of each of the reports are being sent under separate cover for use by JTAC. Copies of the channel studies, referred to in paragraph IVB of FCC 48-2256, will be forwarded when issued.

A reading of the procedure outlined in FCC 48-2256 will indicate that the agenda for the engineering conferences is a rather heavy one, although it is being limited to factors which have an appreciable effect on the station allocation plan. A considerable preparation will be involved, both by the Commission and by the Industry.

While the Commission will be able to make contributions in some degree to a substantial number of the items listed in the agenda, personnel who will be available for this proceeding will devote their time principally to the following:

1. Further study of tropospheric effects, particularly the effect of transmitting antenna heights, Item IVD(1) (a).
2. Further study of terrain effects, IVD (2).
3. TV and FM channel studies, IVB.
4. Preparation of a film showing the effects on the received picture of various signal/noise ratios and desired/undesired ratios of co-channel and adjacent-channel interference, affecting items IVE(3) and (4).

It is not desired to circumscribe the activities of JTAC in its efforts to assist the Commission in this proceeding, but I would recommend that the following matters be given particular attention, the procedure indicated in each case, of course, being merely by way of suggestion.

1. That JTAC submit the accompanying reports to the proper committees of the IRE and the RMA for comment, recommending that any similar studies which are known or can be prepared by persons on the committees be made available for consideration at the first conference.
2. That JTAC and the committees of the IRE and the RMA study the agenda announced in FCC 48-2256 critically with a view to (a) detecting any omissions or any items which are believed to be unnecessary to a resolution of the channel allocation problems, (b) determining the items upon which information can be furnished by JTAC and its associated committees and (c) assessing the adequacy of the time schedule in permitting the formulation of answers to the various items.

Some of the items on which the JTAC

and its committees can be of particular assistance are the following:

IVD(3) Antennas. In addition to transmitting antennas and the practicability of assuming directional operation in the allocation plan, practical receiving antennas and their directional effects should be considered.

IVE(3) Re-examination of co-channel and adjacent-channel ratios, on the basis of test data to be furnished by the manufacturers of various commercial receivers.

IVE(4) Re-examination of contours involving:

- a. Noise and interference levels in urban and suburban areas at various frequencies.
- b. Noise figures for commercial receivers.
- c. Acceptable signal/noise ratios.
- d. Typical receiving antennas and transmission lines.

IVE(6) & (7) Present capabilities of power generation for frequencies between 54 and 216 megacycles.

The JTAC may desire to furnish answers to items other than those which I have listed and should, of course, feel perfectly free to do so. Complete or partial answers to some of the items may be found in previous information furnished by JTAC, in which case the reply should so indicate, with appropriate references.

Do not hesitate to call upon the Commission if further details or clarification of our request is desired.

Sincerely yours,
Wayne Coy,
Chairman

Enclosures

The FCC Notice 48-2256 referred to above is here reproduced:

Received October 18, 1948
FCC 48-2256
27297

FEDERAL COMMUNICATIONS
COMMISSION
Washington, D. C.

In the Matter of Amendment of Section 3.606 of the Commission's Rules and Regulations	Docket Nos. 8975 and 8736
In the Matter of Amendment of the Commission's Rules, Regulations and Standards concerning the Television and Frequency Modulation Broadcasting Services.	Docket No. 9175

Notice of Further
Proposed Rule Making

I. Notice is hereby given of further proposed rule making in the above-entitled matter.

II. During the hearing held by the Commission in the above-entitled proceeding (Docket Nos. 8975 and 8736) to consider proposed revisions of the Commission's table of television channel allocations, evidence was presented concerning (A) tropospheric interference to existing and proposed television stations, (B) the use of directional antennas, (C) the use of increased power, and (D) conflicting proposals for closer spacing and wider spacing between television stations than is presently provided for by the Commission. In order to assure that the Commission's national television allocation plan should be based on the soundest engineering foundation, an Industry-Commission Conference was held on September 13 and 14, 1948. The issues for decision at the Conference were:

"1. Whether the Commission should initiate proceedings to revise the television allocation rules and standards prior to final decision in Dockets 8975 and 8736.

2. If the standards are to be revised, what policy should be adopted with respect to applications now pending before the Commission.

3. What procedures should be adopted in order that the revised standards can be based on the best available engineering information."

III. At the conclusion of said Conference it was announced that the Commission would call an engineering conference to consider questions regarding revisions of the Commission's Rules, Regulations and Standards with respect to the technical phases of television allocations. This Notice deals with issues "1" and "3" set forth above. On September 30, 1948, the Commission issued its Report and Order herein concerning issue number "2." Further, since the Frequency Modulation Broadcasting Service is directly affected by any action taken with respect to propagation in the VHF band, revisions of the Rules, Regulations and Standards of that service is made a part of this proceeding.

IV. In order to facilitate and expedite the promulgation of rules, regulations and standards herein, the following schedule will be followed:

(A) On or about October 20, 1948, the Commission will make public:

(1) A report containing (a) a summary of available measurements of tropospheric fields, (b) empirical method of treating measurements to formulate field intensity vs distance curves for various frequencies for various percentages of the time and (c) representative tropospheric field intensity curves for antenna heights of 500 feet and 30 feet for various frequencies and percentages of time derived by the foregoing method.

(2) A study of the effects on service of the simultaneous fading of both the desired and undesired fields from tropospheric causes.

(3) A report on measurements made at Princeton, Southampton and Laurel on frequencies of 47.1, 106.5 and 700 Mc. radiated from transmitters in New York City.

(4) A study of the effects of terrain upon average signal levels as compared to smooth earth values and upon the variability of signal levels within limited areas.

(B) On or about November 15, 1948, the Commission will make public:

(1) A TV channel study showing the effects of ground wave and tropospheric interference on representative service areas of stations allocated in accordance with the Commission's Notice herein of May 5, 1948, as amended in the Commission's Supplemental Notice of July 15, 1948.

(2) A TV channel study in a representative area showing the effects of ground wave and tropospheric interference on the service areas of presently operating stations and CP's, but with other allocations spaced so as to protect the 500 u/m contours 90% of the time. (All allocations to be based on 50 kw/500 ft. in the center of the principal city).

(3) Channel study for FM showing the effects of ground wave and of tropospheric interference for 1% and 10% of the time on representative channels.

(C) On or about November 30, 1948, December 1, 1948 and December 2, 1948, the Commission will hold a series of engineering conferences in Washington, D. C.[1] All interested persons are invited to attend said conferences, participate fully therein, and to submit written data, views, or arguments with respect thereto. To assist the Commission in the expeditious conduct of said conferences, it is requested that persons who plan to participate therein file (by letter) notice of their intention to do so at least one week prior to the date of commencement of said conferences. Written statements may be filed on or before the dates of the respective conferences.

(D) The first conference to be held on or about November 30, 1948, will be on VHF propagation standards to arrive at standard methods of evaluating the effects upon propagation of the following factors:

(1) Tropospheric effects—
  (a) Variations with time in the field intensities to be expected at various distances from the transmitter, as functions of transmitting antenna height and of frequency.
  (b) Range of diurnal variations.
  (c) Range of seasonal variations.
  (d) Effects on service of the simultaneous fading of both the desired and undesired signals.

(2) Terrain effects—
  (a) Shadows—relation of the average field intensity in a limited area or limited section of a radial to calculated values as a function of the profile between the area and the transmitter.
  (b) Urban field intensities—validity of the FCC standards on Ground Wave Signal Range charts for predicting near-in fields in city areas.
  (c) Local terrain effects—variability of field intensities as compared to the average over a limited area or distance.
  (d) Receiving antenna height-gain factor–validity of assuming a

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[1] The exact date and place of each conference will be announced at a later date.

uniform variation of field intensity with receiving antenna height for relating mobile measurements made at low antenna height to the standard receiving antenna height of 30 feet. Consideration of the alternative method of spot measurements at 30 feet height.

    (e) Apparent transmitting antenna height—validity of the 2–10 mile rule for estimating the apparent height of the transmitting antenna.
    (f) Validity of the method presently prescribed in the Commission's Standards for equalizing coverage obtained by transmitters of varying antenna heights and power.

(3) Antennas—
    (a) Practical limitations on vertical and horizontal directivity of transmitting antennas.
    (b) Methods for establishing and maintaining the performance of directional antennas.
    (c) The engineering basis for utilizing horizontal directivity in allocation problems.

(E) The second conference to be held on or about December 1, 1948, will consider the following items with respect to VHF television broadcasting:

(1) Tropospheric effects:
    (a) Specification of grade or grades of service resulting from variations in the intensities of desired and undesired fields.
    (b) Discussions of the effects of the specification of various grades of service on particular channel allocation plans.
    (c) The development of standard tropospheric curves for various frequencies and antenna heights, calculated in accordance with methods approved at the propagation conference.
(2) Examination of current standards for the prediction of service areas to determine whether any modifications are dictated by the terrain effects considered in the propagation conference.
(3) Re-examination of co-channel and adjacent channel ratios at the receiver terminals in the light of more recent information; and a determination whether a terrain factor should be included in the field intensity ratios.
(4) Re-examination of the contours specified for protection and for recognized service levels at various frequencies.
(5) Re-examination of assumptions as to typical receiving antenna heights for urban and rural areas and of methods of proving station performance by measurement of received fields at such heights.
(6) Examination of the effects of horizontal increases in power upon protected contours in the channel allocation plans.
(7) Examination of the effects of differential increases in power on the protected contours and on the allocation plans.
(8) Examination of the effects of directional antennas on allocation plans.

(F) The third conference to be held on or about December 2, 1948, will consider the following items with respect to FM broadcasting:

(1) Tropospheric effects:
    (a) Specification of grade or grades of service resulting from variations in the intensities of desired and undesired fields.
    (b) Study of the areas provided with various grades of service under the present channel assignments and under the tentative allocation plan.
    (c) The development of standard tropospheric curves for various antenna heights, calculated in accordance with methods approved at the propagation conference.
(2) Examination of current standards for the prediction of service areas to determine whether any modifications are dictated by the terrain effects considered in the propagation conference.
(3) Re-examination of assumptions as to typical receiving antenna heights for urban and rural areas and of methods of proving station performance by measurement of received fields at such heights.

V. Authority to issue amendments of the Commission's Rules, Regulations and Standards with respect to the matters to be discussed at the conferences listed above is vested in the Commission by Sections 301, 303(b), (c), (d), (f), (h) and (r), and 4(i) of the Communications Act of 1934, as amended.

VI. In accordance with the provisions of Section 1.764 of the Commission's Rules and Regulations, an original and 14 copies of all written data, views, or arguments filed shall be furnished the Commission.

FEDERAL COMMUNICATIONS
COMMISSION

T. J. Slowie, *Secretary*

Adopted: October 14, 1948
Released: October 15, 1948

THE JOINT TECHNICAL ADVISORY COMMITTEE

THE INSTITUTE OF RADIO
ENGINEERS, INC.
RADIO MANUFACTURERS ASSOCIATION

Pursuant to this letter, the chairman of JTAC addressed the following letter to the chairmen of all Technical Committees and other interested groups:

October 29, 1948

Dear Sir:

As you doubtless know, the FCC has scheduled tentatively a three-day engineering conference for November 30, December 1 and 2. The schedule for this engineering conference is included in the attached FCC release and requests information on propagation and allied subjects in the vhf portion of the spectrum, particularly as it affects standards and allocation problems of the television and FM services.

The Commission, through the letter copy attached, has requested the JTAC to assist in the collection of information for this conference. Accordingly, JTAC is requesting the assistance of engineering groups within the RMA and IRE, as well as other industry groups. As Chairman of one of these groups, we urgently request that you read carefully the attached material from the FCC.

The JTAC will greatly appreciate any information, engineering data, and comments regarding these subjects that you may be able to submit by November fifteenth to the Office of the Secretary.

Sincerely,
PHILIP F. SILING

Enclosures 2

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Electronics in Nuclear Physics*

W. E. SHOUPP†

The publication in the PROCEEDINGS OF THE I.R.E. of a series of papers on nuclear particles and phenomena, and on electronic instrumentation and control of nuclear processes, continues through the presentation of the following paper. This paper clearly sets forth certain fundamental nucleonic facts, their relationship to electronics, and some of the many and intriguing opportunities offered to the electronics expert who enters the nuclear field.—*The Editor.*

INTRODUCTION

IN AN EFFORT to determine where the science of electronics crosses the path of nucleonics and just what are the results when they meet, it will be necessary to define reasonably well the two words "electronics" and "nucleonics." The word "electronics" requires little discussion here; however, it may be interesting to try to put a definition into words.

Generally speaking, electronics embraces all phenomena associated with the passage of electric currents through gases and high vacua.[1] Such phenomena may be utilized for many purposes, such as the generation, amplification, and detection of high-frequency electrical oscillations, and the production of light from electric current, as well as the converse process. The propagation processes and the study of interstellar radiations are likewise included. Studies concerning the acceleration, generation, and use of atomic particles and radiations may also be included in electronics.

Electronics may then be defined as that science, art, and industry concerned with electrical phenomena involving electrically charged atomic particles outside of liquids and solids. Under this definition we have included phenomena other than those due to the action of electrons.

Now, if we define nucleonics, we may see where the two fields fit together or possibly overlap, and it may be possible to draw some conclusions as to their interaction. The nucleus is supposed to be made up of particles called "nucleons." Until recently it was supposed that protons and neutrons in combination make up all nuclei. In line with recent meson theory, this does not necessarily follow; however, the term nucleon will usually refer to a nuclear proton or neutron. Nucleonics, then, is the science dealing with the rearrangement, transformation, and utilization of the nucleons. This includes nuclear synthesis, radioactivity, nuclear fission, and their applications. We will not expect interactions between electromagnetic waves of the microwave and radio spectrum and the fundamental nuclear particles and radiations because of the vast difference in wavelength. Electronics will, therefore, be of chief importance in observing the physical processes that occur in ordinary and induced nuclear phenomena. In fact, if electronics is to influence nucleonics, it must be in the course of the discovery, production, measurement, control, use, or disposal of nuclear substances. The better to see just how this may come about, a brief discussion of some of the general properties of nuclei is necessary.

Most radioactive substances that are generally available are generated in the nuclear reactor commonly called a "pile." After looking into the operation principles of the pile, the characteristics of its radiations, and its radioactive products, it will be possible to point out where electronics may contribute to the technology of nuclear processes through the fabrication and location of the proper materials for pile construction, the control of the pile during operation, or the utilization of the products manufactured within the pile.

THE URANIUM PILE OR NUCLEAR REACTOR

The active ingredient of the pile is uranium, usually in the metallic form. The largest known deposit was discovered in Northern Canada, where it is found to exist in the form of the uranium-oxide complex $U(UO_4)_2$. Already we have met electronics head-on, because our modern uranium prospector who made the discovery used a Geiger-counter radiation detector to find the ore, which is a radioactive material. The detector used here is not so complex as some other electronic equipment, but is a model of compactness, reliability, and clever electronic engineering. Detectors of this type have been made that weigh less than four pounds and are sometimes powered by four flashlight cells.

The uranium oxide, after being discovered and mined, is reduced to pure metallic uranium and is found to have a specific gravity of 18.7, an atomic weight of 238.07, and a melting point of 1150°C. It is also known that 99.3 per cent of the metallic material is composed of the heavy isotope ${}_{92}U^{238}$ and 0.7 per cent is ${}_{92}U^{235}$. Suppose, now, that after considerable mining and chemical reduction we have accumulated sufficient uranium metal to set up a self-sustaining fission reaction in an assembly called a nuclear reactor or pile.

The essential thing about the process of uranium fission is that the uranium atom falls apart in such a way as to produce two more or less equal fragments—and to liberate several more free neutrons. It is this neu-

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* Decimal classification: 621.375×539.7. Original manuscript received by the Institute, August 26, 1948.

† Westinghouse Research Laboratories, East Pittsburgh, Pa.

[1] E. U. Condon, *Bull. Nat. Electronics Con.*; 1946.

tron liberation that makes a self-maintaining process possible. The splitting requires a neutron to make it go; the splitting process itself acts as a source of neutrons which can cause more uranium atoms to split. Here is the basis for a chain reaction.[2]

Why, then, does not ordinary uranium explode, or at least "burn," in a nuclear sense? Since several neutrons are released at every fission, a chain reaction is possible. But one of the several neutrons released must actually produce another fission to keep the process going.

If all the neutrons released produced more fissions, the material would explode violently. Since the neutrons move rather freely through matter (like X rays) many are lost by escaping through the surface. Consequently, it is necessary to use a big enough lump of fissionable material to get a smaller surface-to-volume ratio. In other words, unless the lump exceeds a certain critical size, the chain reaction cannot proceed.

Another complication is that impurities in the uranium have a powerful effect on the neutron loss through absorption. This loss of neutrons is very difficult to obviate, for appreciable losses result from the presence of only one part per million of some materials, and it is no easy matter to produce anything of that purity on an industrial scale.

The worst complication of all is that uranium itself absorbs neutrons in other ways than those that produce fission. This phenomenon is both a blessing and a curse. It turns out, fortunately, that the over-all effect of this nonfission absorption of neutrons by uranium is sufficiently great to prevent the explosion of perfectly pure uranium, in normal isotopic abundance, even in so large a lump that escape of neutrons through the surface is negligible.

Neutrons given out in the fission process are "fast"; that is, they possess speeds corresponding to several million electron volts of kinetic energy. Such fast neutrons colliding with uranium atoms have a rather great chance of losing energy without being caught and without producing fission. On the other hand, neutrons of intermediate speed are unable to produce fission in U^{238}. They can do so only in U^{235}, which forms only 1/140 part of natural uranium.

Neutrons of a particularly low energy (about 10 electron volts) are very like to be captured by U^{238} to form U^{239}. This is very important! In fact, this happens so readily that so many neutrons are used up in this process that a chain reaction cannot be maintained in ordinary uranium alone.

An uncaptured neutron continually loses energy by colliding with atoms as it diffuses throughout any material, until its average energy is that of the heat motion of the atoms of the material. Neutrons of certain extremely low energies are strongly captured by U^{235} to produce fission.

To cause and maintain a chain reaction with ordinary pure metallic uranium, which contains all kinds of uranium atoms but is predominantly U^{238}, the uranium is arranged in a lattice of small lumps so that many of the fast-moving neutrons diffuse out of the uranium into some surrounding material. Here many of them are slowed down before diffusing back into the uranium. Most neutrons thus escape being caught by U^{238} until they have lost so much energy that capture by U^{238} is unlikely. Ultimately, unless the neutrons are lost in the surrounding medium, they return to the uranium lumps and are sufficiently reduced in speed to cause fission in U^{235}, but have energies too low to be captured by U^{238}.

In the technical vocabulary of nuclear engineering, the material that temporarily traps the neutrons and helps them lose energy until they are safe from capture by U^{238} is called the "moderator." Evidently the moderator material must not absorb too many neutrons itself, or the reaction will be stopped. Besides not absorbing neutrons, a desirable moderator has a low atomic weight. For, to be slowed most efficiently, the neutrons are allowed to collide elastically with the nuclei of the moderator material. More energy is given up at each impact if the two partners of the collision have nearly the same mass.

The hydrogen content of ordinary water (H_2O) would be ideal with respect to mass, but ordinary hydrogen absorbs too many neutrons. Heavy water (D_2O) is satisfactory from a neutron-absorption standpoint, but is not generally available in sufficient quantity. Metallic beryllium is a possibility, but is quite expensive. Therefore, specially purified graphite was finally adopted in the original piles constructed. In a typical graphite-moderated pile a neutron that has escaped from the uranium into the graphite travels roughly about 1 inch between collisions and makes, on an average, 200 elastic collisions before returning to the uranium, losing about $\frac{1}{6}$ of its energy in each collision.

A chain reaction is then possible, but since the nuclear fission process takes place in very short time and with considerable evolution of energy, how can the pile be kept from blowing up? If a pile is so arranged that, on the average, more than one fission results from the neutrons produced by each fission, then clearly the number of neutrons present, and the amount of heat generated, increase by the compound-interest law. If a great multiplication happens rapidly—say, in a small fraction of a second—then the phenomenon becomes an explosion. In short, we have an atomic bomb. Even if the reaction occurs slowly, the pile would soon be destroyed by melting if the multiplication were allowed to proceed.

One way to control the pile[3] is to provide passageways through it into which rods of material that strongly

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[2] E. U. Condon, *Westinghouse Eng.*, vol. 5, pp. 3, 9; November, 1945.

[3] H. D. Smyth, "Atomic Energy," Princeton University Press, 1945; p. 135.

absorb neutrons can be placed. When these rods are all the way in, they absorb so many neutrons that the chain reaction is stopped. As they are slowly withdrawn, a point is reached at which the reaction is just able to proceed. If pulled out farther, the neutrons are able to multiply more rapidly and the pile operates at a higher power level. To stop the pile the absorbing rods are simply pushed back in farther. Cadmium and boron-containing steel are suitable materials for the control rods.

We have so far assumed that the build-up time for the chain reaction used in the pile is long enough for an operator to maintain control by manual operation of the rods or by use of a similar slow-acting control mechanism. That is in fact the case, due to another fortunate phenomenon in the fundamental physics of fission—delayed neutrons.

It was discovered in May, 1942, that not all neutrons emitted in the fission process come out instantly. The uranium nuclei in splitting apart spill out most neutrons immediately. But the atomic fragments formed are also in a highly unstable condition and some of them throw out additional neutrons after a short time delay, amounting on the average to half a minute. It is the delayed ones that set the time scale on which the neutron multiplication in the pile builds up. They set it for such a long time that slow-acting controls are easily able to regulate the activity of the pile. In fact, 1 per cent of the neutrons are delayed by at least 0.01 second and 0.07 per cent are delayed by as much as a minute.[3]

Here is an ideal chance for electronics to contribute. The delayed neutrons do provide a bit of time to activate controls, but sensing and control-activating methods are necessary. It is presumably possible to measure the temperature within a pile by appropriate sensing elements such as bolometers or thermocouples, and to use the temperature level so measured properly to position the control rods. It would also seem feasible to measure the neutron intensity at some arbitrary point, and to use the signal thus obtained to operate the controls. In any event, electronic controls for the drive motors operating the control rods seems to be indicated. This is particularly true since operations of this sort probably require antihunting methods and accurate control.

Since a nuclear reactor is a very valuable and expensive machine, it is likewise important that elaborate precautions be taken against operating the pile at such a high level that internal damage to the reactor might occur. Here, again, electronics has an ideal application. More or less standard methods are indicated for this job.

There is, however, one missing link. How do we measure the level of neutron intensity so as to operate our controls and precautionary devices? We have seen that neutrons are nonionizing particles and for that reason we cannot use exactly the same techniques that we apply to β rays, α particles, and the like. What is done is to cause the neutrons to produce an ionizing particle (usually an α particle) and to infer the neutron intensity by counting the number of particles produced. One of the best ways of doing this is to allow the neutron flux to fall upon a chamber, filled with boron-trifluoride gas (BF_3) and surrounded by considerable hydrogeneous material, usually about 10-cm thickness of paraffin. The neutrons entering the paraffin will be slowed down by making elastic collisions with the hydrogen in the paraffin and then will become very active in causing the following nuclear reaction to occur:

$$_5B^{10} + {}_0n^1 \rightarrow {}_3Li^7 + {}_2He^4. \tag{1}$$

In this reaction an α particle ($_2He^4$) of 2.7 Mev is emitted and will produce either a pulse in an ionization chamber or will trigger a Geiger or proportional counter. We have here a means for generating electric pulses from neutrons, but electronics must supply all the technology between the sensing element and the control activator. Various types of amplifiers and electronic circuitry are used for this service. It will be the task of succeeding papers by other authors in this series to describe this technology in detail. Suffice it to say that there is need for a vast array of fast and slow pulse amplifiers, dc amplifiers, coincidence circuits, discriminators, scaling circuits, and others too numerous to mention. Their use for this and similar applications forms one of the most interesting electronic fields of endeavor today.

Now let us return to the nuclear reactor and see what else is going on. The first pile was built on the University of Chicago campus during the fall of 1942. It contained 12,400 pounds of uranium together with a graphite moderator. It was intended to be spherical in shape, but since the critical dimensions proved to be smaller than the original calculations indicated, the sphere was left incomplete, giving the actual pile the shape of a large inverted doorknob.

It was first operated on December 2, 1942, at a power level of ½ watt, and on December 12 the power level was stepped up to 200 watts but the pile was not allowed to go higher because of inadequate provision for shielding the personnel. Further studies on piles were made by the construction of one in Tennessee designed for 1000-kw level of operation. Later a pile using heavy water instead of graphite as moderator was built.

It should be remembered that, although a pile is built with ordinary uranium, it is only the 0.7 per cent that represents active U^{235} metal. Most of the metal is U^{238} and actually tends to stop the process. Only by an ingenious lattice arrangement for slowing neutrons in a moderator is the pile able to operate in spite of the presence of the more prevalent U^{238}.

This means that, regarded as a fuel, only 1/140 of the total weight of uranium is being directly used; the rest is an inert material that remains largely untransformed by the pile.

How does the bomb chain reaction differ from that in the pile? The atomic bomb explodes, whereas the reaction in the pile proceeds at a slow rate, easily controlled by manual operation of absorbing rods. The big, fundamental distinction is that the bomb (one type) is made of essentially pure U^{235} and without the use of moderator. The chain reaction in the bomb is carried on by fast neutrons directly released by fission. As already remarked, this cannot happen with ordinary uranium; the U^{238} slows the neutrons to the point where they cannot produce fisssion in U^{235}. In addition, U^{238} absorbs many of them. With essentially pure U^{235} these competing absorption processes do not occur and the reaction is carried by the fast neutrons directly emitted from a U^{235} fission. These are utilized at once to produce fission in other U^{235} atoms. Here the main factors tending to stop the reaction are the loss of neutrons through the surface (which sets a minimum size to the bomb) and losses by absorption due to impurities, including any remaining U^{238}.

What is plutonium? This is a newly discovered chemical element not known to exist in nature, but which is made from uranium by atomic transmutation. Plutonium is important because it, like U^{235}, is a material from which atomic bombs can be made.

That U^{238} can capture neutrons has already been mentioned as a phenomenon detrimental to the operation of a pile. When U^{238} captures a neutron it becomes U^{239} and emits gamma radiation, as does radium. This U^{239} is not stable but emits high-speed electrons by a process of spontaneous radioactivity. The mean life of the U^{239} atoms is only about 20 minutes. By this activity they are transformed into atoms having essentially the same mass but one greater positive charge, 93, on the nucleus, and hence a new chemical element. It is called neptunium and written as Np^{239}. Neptunium 239 is also spontaneously radioactive and emits another high-speed electron, becoming thereby an atom having 94 positive charges on the nucleus but still essentially of mass 239. This process is slower; the mean life of the neptunium atoms is about two days. The resulting atom of charge 94 and mass 239 is another new element that does not occur in nature. It is called plutonium and is written Pu^{239}.

Actually, the purpose of piles in the military project was not to get atomic power but to produce the new element plutonium, which provides a second bomb material. It is, in short, a competitor to U^{235}. Since the uranium lumps in the pile are exposed to a dense atmosphere of neutrons, the means is at hand for changing a part of the U^{238} into Pu^{239}.

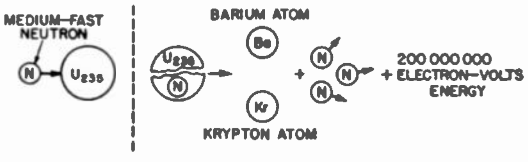
The several large piles thus put in operation generated many hundreds of thousands of kilowatts as heat. This heat was, however, not utilized, as the main purpose of the operation was production of plutonium for use in the atomic bomb. To utilize the heat efficiently it would have been necessary to operate the pile at high temperatures, requiring extensive modification of pile construction.

The pile, when run at a high power level, also generates an enormous amount of radioactive material, far more potent than all the radium ever mined. This greatly complicates the problem of operation of the large piles by requiring a high standard of reliable operation that must depend entirely on remote controls.

The plutonium is formed in the blocks of uranium in the pile. These have to be removed from the pile and the plutonium extracted by fairly simple chemical methods, because plutonium and uranium, being completely different elements, are dissimilar chemically. The process, however, is greatly complicated by the intense radioactivity of the materials.

RADIOACTIVE BYPRODUCTS OF THE PILE

In all this involved process, what has happened to the uranium atoms we were watching? We have seen that some of the U^{235} has been caused to fission. For example, the U^{235} nuclei have split into roughly two equal parts, as indicated in Fig. 1. We now have a barium and a krypton atom, and two or three neutrons have been emitted. The nuclei of the barium and krypton are very different from the ordinary types found in nature. In fact, in this special case the krypton formed is deficient by three positive nuclear charges, and the barium nucleus is deficient by four, from a stable condition. Consequently, these nuclei are radioactive and will emit β particles and γ rays until stability is attained. Barium and krypton and their intermediate decay products are not the only radioactive isotopes formed as a result of the fissioning. In fact, possibly a hundred different radioactive isotopes are produced in this process.



$$_0n^1 + {}_{92}U^{236} \longrightarrow {}_{36}Kr^{89} + {}_{56}Ba^{144} + 3\,{}_0n^1$$

Fig. 1—Typical neutron-induced fission of U^{235} atom.

Fission into precisely equal masses seldom occurs, the most abundant fragments being one of a mass number between 127 and 154 and the other between 83 and 115. The half-lives of the unstable isotopes thus formed vary from a small fraction of a second to over a year. Radiations from these radioactive elements present one of the major health problems to the personnel working in this field. The isotopes are, of course, chemically recoverable and are valuable byproducts from pile operation. In addition to the radioactive fission products, heavy nuclei, U^{239} and Np^{239}, are present in the pile. The concentration of these materials builds up within the pile during normal operation until eventually their decay is equal to their rate of formation; their concentration then becomes constant, thus presenting interesting limitations of the economics of pile operation.

Within the chain-reacting pile, a very high flux density of neutrons is maintained. If certain elements are inserted within the pile, intense radioactivity may be induced in most isotopes by neutron capture. As an example, radioactive phosphorus (P^{32}) is made by inserting ordinary phosphorus within a pile and inducing the following reaction:

$$P^{31} + n^1 \rightarrow P^{32} + \gamma \qquad (2)$$

where the superscripts indicate the total number of protons and neutrons in the nucleus (mass number), and the letter is the symbol for the element or radiation involved. At the present time seventy-five radioactive isotopes made by this method may be obtained from the Atomic Energy Commission, and several hundred more may be made within piles or by bombardment by the beam of nuclear accelerators. These radioactive materials are useful for a vast variety of applications, a few of which will be mentioned. Radioactive sulphur may be used to determine the amount of sulphur remaining in coke; radioactive phosphorus may be used to analyze the phosphorus content of steel; while solid or liquid and other radioactive materials may be used to examine the finished steel for flaws or inclusions.

In medicine, clinical use in made of radioactive materials in the diagnosis and treatment of certain thyroid disorders. The motion of the blood, heart action, and digestive processes may also be investigated by means of radioactive techniques. In fact, radioactive tracers may be followed quantitatively throughout the various physical and chemical processes to which they may be subjected. All applications of these materials depend upon the detection and accurate measurement of the position and quantity of disintegrating radioactive nuclei. It is in this application that electronics finds its greatest possibilities. In fact, the real importance of analysis by radioactive isotopes lies in the fantastic sensitivity of the electronic detection methods that are applicable. Here we reach the ultimate in detection sensitivity, where individual exploding atoms are detected and accurately counted. This leads to applications where tremendous dilutions are encountered. Ionizing radiations are emitted by radioactive materials and the ions formed within gases, liquids, and solids may be used to indicate the presence of radioactive atoms. A vast variety of electronic devices is available for these applications, and a series of comprehensive articles written by well-known authorities on the various types of instruments will be published in succeeding issues.

Nuclear Radiations from Radioactive Materials

The radioactive material may emit β radiation (high-speed electrons), positrons (β_+), α particles (helium nuclei), γ radiation, neutrinos, and antineutrinos. The latter two are nonionizing particles of low mass, more or less hypothetical, and so far have not been directly observable. Nearly all of the detection methods for α, β, β_+, and γ radiations are based upon the electrical response generated when these radiations cause primary and secondary ionization processes as they pass through matter. For α and β radiations, the ionization is produced in the slowing down of the particles themselves. On the other hand, γ radiations are detected by the ionization produced by the Compton recoil electrons, photoelectrons, and electron-positron pairs that are produced as they pass through matter. Since the electronics specialist plans to use ionization to indicate the presence of such radiation, it is of great importance that the energy, penetrating power, and ionization density of the radiation be known. In general, ordinary radioactive materials emit radiations having energies that vary from several kilovolts to a few million volts. When dealing with these radiations, it is convenient to remember the following rough characteristics of these radiations.

Properties of Alpha Radiation

The range of α particles in air at ntp (normal temperature and pressure) is about 1 cm for 2 Mev, 2 cm for 3.4 Mev, and 5 cm for 6.3 Mev. On the average the energy loss is about 1 Mev per cm of air forming about 40,000 ion pairs. A better insight into the ionization by α radiation may be obtained from examining Fig. 2, which shows the specific ionization along the path of the particle. Similarly, Fig. 3 shows the range of α particles in air at ntp.

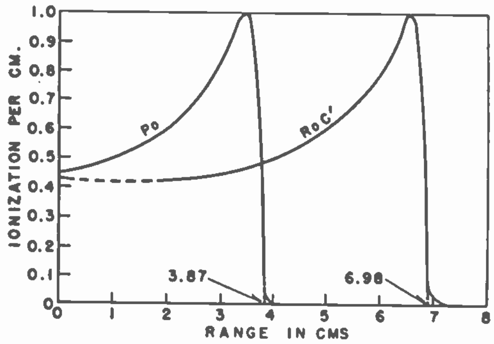


Fig. 2—Relative specific ionization of α particles from polonium and radium C'.

Properties of Electrons (β) and Positrons (β_+)

Electrons and positrons (β and β_+ particles) have a mass that is only 1/1838 that of the α particle and are scattered considerably as they pass through matter, so that their paths are generally far from being straight lines. The "straightened-out" length of the electrons increases with the energy and is about 1 cm at 25 kev, 10 cm for 90 kev, and 100 cm for 0.4 Mev in air at ntp. Since it is frequently necessary to shoot these particles through foils into detection equipment, their penetrating ability for aluminum is of some importance. Roughly, their penetration in aluminum is about 0.0005 that in air, but at high energies the range varies almost linearly with energy, while at low energies it is more nearly proportional to the square of the energy. This is shown graphically in Fig. 4. It is seen that in the region of tens of kilovolts the range-energy relation for electrons having an initial energy E_0 follows roughly the quadratic law:

$$R_{1\mathrm{cm}} = 7.5 \times 10^{-7} E_{\mathrm{kv}}^2.$$

The energy E in kv possessed by the electrons after penetrating t grams per square centimeter of material is given by

$$E^2 = E_0^2\left(1 - \frac{13 \cdot 10^5}{E_0^2} t\right).$$

The previous discussion is for high-speed electrons of monokinetic energy. When using these data it should be remembered that radioactive materials emit electrons or positrons with a continuous distribution in energy from zero up to a certain maximum energy ($E_{\max}$) that is characteristic of the particular radioactive isotope.

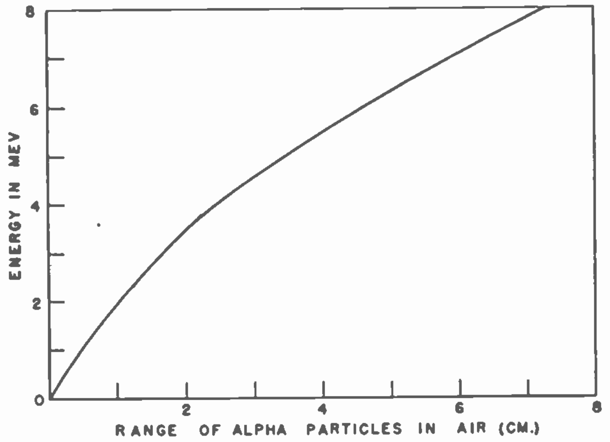


Fig. 3—Range of α particles in air at ntp.

Properties of Gamma Rays

Gamma rays are similar to X rays of high energy. The distinction between the two lies only in their method of production. As γ rays pass through matter they lose energy by interaction with electrons through the Compton effect, the photoelectric effect, and, when energetically possible, through formation of positive-negative electron pairs. The intensity of a monoenergetic beam of γ radiation passing through an absorber that does *not* interact with the absorber will decrease exponentially as

$$I = I_0 e^{-\mu t}$$

where I_0 is the incident intensity and I is the intensity after having penetrated a thickness t of material (in grams per cm^2) of absorption coefficient μ (in cm^2 per gram). The absorption coefficient is about 0.06, cm^2/gram for γ radiation of a few Mev and increases rapidly below 1 Mev. The contribution of the various absorption processes is shown in Fig. 4. The total absorption coefficient has a minimum at about 3.5 Mev and above 8 Mev is nearly entirely due to pair production.

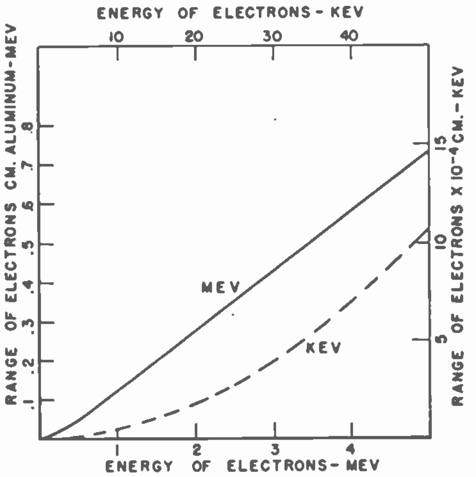


Fig. 4—Range energy of electrons in aluminum.

Radiation Detectors

The various properties of α, β, β_+, and γ radiation as discussed are used in the many types of electronic radiation detectors. A brief summary of the operation of the various electronic devices used for the detection of these radiations follows.

Scintillation Detectors

Certain materials have the property of emitting visible light when bombarded by ionizing materials. Among the more useful materials are zinc sulphide, aluminum oxide, calcium tungstate, anthracene, and napthalene. These materials may be used in the form of powdered material deposited on a backing material, although usually their use as a clear, transparent crystal is preferred. The minute flash of light produced by a single α particle is visible to the eye, provided it is "dark-adapted." The flashes are also capable of being picked up by photomultiplier tubes such as the 1P28 or 931A. With appropriate amplifiers, pulses due to single α, β, β_+, or γ radiations may be detected; and if fissionable material, hydrogeneous material, boron 10, or other materials that react with neutrons be mixed with the fluorescent materials through fission, scattering or nuclear transformations may also be indicated. The electronic circuitry for these detectors is not particularly complex since considerable amplification is attained within the photomultiplier, and it is comparatively simple to obtain pulses 10^{-7} seconds wide from the scintillation detector.

Ionization Chamber Detectors

An ionization chamber is a gas-filled volume into which electrodes are inserted and maintained at a potential difference sufficiently low that gas breakdown or gas amplification does not occur, yet sufficient to collect the primary ionization due to the passage of the ionizing radiation. In integrating types of ionization chambers, a given charge is applied to the electrodes of the ionization chamber, acting as a capacitor. The ionizing radiation then discharges this capacitor and the change in potential difference across the electrodes is read by an electroscope or an electrometer, usually of the filar type. Special electrometer tubes sensitive to lower than 10^{-15} amperes have also been developed for measuring the ionization current produced within the chamber. All types of ionizing radiations are detected by these devices. Generally speaking, instruments of this type have long time constants and are chiefly used where simplicity of operation and low sensitivity (2 to 5000 milliroentgens per hour) are measured.

Pulse Ion Chambers

The alternating-current pulse generated by the ionizing event in an ionization chamber may be amplified and the rate of occurrence of the pulses thus generated may also be used to measure a flux of ionizing radiations. Until recently, the amplified pulse due to the displacement current generated by the motion of the positive ions has been used. At customary strengths of the collecting field, this results in pulse widths which require amplifiers to be peaked at about 1 kc. The motion of the electrons generated by the ionizing event cause a much more rapid pulse to be generated in addition to that due to the positive ions. The mobility of the electrons is much greater than that of the positive ions; therefore, "faster" amplifiers are required, resulting in amplifiers having bandwidths of from 0.2 to 2.0 Mc. In order that pulse heights from the amplifier may be used as a measure of particle energy, pulse amplifiers are usually made linear by means of inverse feedback. The "high-speed" amplifiers have the advantages over the "low-speed" type of comparatively low microphonic noise and higher resolving time.

Proportional Counters

If the field strength within the ionizing volume is increased until "gas amplification" takes place, the size of the pulse, due to the traverse of the ionizing particle, may be considerably increased. When coaxial proportional counters are used, this increased ionization, which may be several thousandfold, is due to high field about the central wire. For this reason amplifiers used with proportional counters need to have much less gain than those used with ionization chambers, and this indeed is the reason proportional counters have replaced ionization chambers for many applications. However, the pulse size obtained from proportional counters is quite sensitive to the voltage across the counter, and the "simpler amplifier" advantage is partially nullified by the additional requirement of a very stable high-voltage power supply.

Geiger Counters

If the voltage across a coaxial proportional counter is increased, the pulse size of the discharge due to an ionizing particle will at first increase and then level off at what is called the "Geiger region." In the "Geiger region" electron avalanches are set up and a continuous gas discharge will result, unless either the voltage is removed from the counter after the ionizing event takes place, or a "quenching gas" is used to extinguish the discharge. Either or both methods are frequently used and offer interesting electronic design problems so as to minimize recovery time and double counting.

Other Detectors of Nuclear Radiations

Photographic films, Wilson cloud chambers, chemical effects, and crystal counters offer other means of radiation detection, of which the crystal counter is of greatest interest to the electronic nuclear instrumentation field. The crystal counters utilize a slab of crystalline material mounted between two high-voltage electrodes. When an ionizing particle strikes crystals of certain types, ionization pulses are produced. The amplifiers and associated electronic equipment are very much the same as that used with the proportional counter; however, since the crystal is of high density, considerably more of the radiation energy is converted to electrical pulses for a unit volume of material than for the case of the gas-filled counters.

Use of Radiation Detectors

The α, β, β_+, and γ radiations emitted from radioactive sources may be measured quantitatively by the various radiation detectors previously discussed. Generally speaking, the various radiation instruments are used as described in Table I. In addition to the applications listed, the α-particle-sensitive devices may be used to measure neutrons, which are nonionizing, by inserting certain materials that react with the neutrons to emit α or fission particles. A typical example of this is the use of a boron-trifluoride atmosphere in a proportional counter, thus permitting the neutrons to interact with the boron to emit an α particle which is then recorded in the customary manner. The nuclear reaction used for this process is given in (1). From this reaction it is seen that the boron 10 (${}_5B^{10}$) isotope is mainly responsible. In ordinary boron only 18 per cent is ${}_5B^{10}$, the remainder being the isotope of mass 11, ${}_5B^{11}$, which

TABLE I

Apparatus	*Particles Detected*	*Range of Intensity*	*Uses and Remarks*
Integrating Ion Chambers	α, β, γ (usually only γ)	0.1 to 100 r	Used mainly for health dosimeter. Advantages: compact, simple, rugged.
Ion Chamber with DC Amplification	α, β, γ (usually only γ)	0.001 to 10 r/hr	Used mainly for reading radiation flux (r rate)
Pulse Ionization Chambers	α, protons, deuterons	Counts single particles up to 10^5 per second	Research-laboratory purposes and monitoring of α radiations.
Proportional Counters	α, protons, sometimes β, γ	Counts single particles up to 10^4 per second	Monitoring of α-particle intensity and measurement of heavy particles in presence of β and γ background.
Geiger Counters	β, γ, sometimes α	Counts single particles up to 10^3 per second	Most-used instrument—counting chiefly β, γ from radioactive materials. May be used for α if necessary. Location of contamination and range dosimetry.
Scintillation Counters	α, β, γ	10 per second up to 10^6 per second	Research purposes requiring high resolution. Probably more extensive uses later.
Crystal Counters	α, β, γ	Single events up to 10^5 per second	Largely research in character at present. Brilliant possibilities.

has one more neutron in its nucleus. It is possible to obtain from the Atomic Energy Commission certain boron compounds enriched in the isotope ${}_5B^{10}$ produced through the use of isotope separators. Use of this "enriched" material as an atmosphere in proportional counters and ionization chambers increases the sensitivity of neutron detectors of this type by roughly a factor of 5.

Electronic Requirements of Radiation Detectors

The characteristics of the various nuclear radiations and instruments capable of detecting and measuring them have been briefly discussed. It remains to consider in greater detail just where electronics may be expected to contribute to the design and operation of these instruments. If one examines any of the various detectors listed in Table I, it becomes obvious that the main content is a maze of electronic tubes and electronic circuitry of ordinary type, but of unusual application. In fact, in most radiation detectors electronics is the "tail that wags the dog." Succeeding papers will discuss in detail the electronic and physical details of these detectors. However, for illustrative purposes, consider the Geiger counter as used in the laboratory for analytical radioactive purposes. For γ-ray counting the "Geiger tube" used is a fundamentally very simple device composed of a copper cylinder 10 cm long and roughly a centimeter in diameter, having an insulated 3- to 10-mil tungsten wire stretched down the cylindrical axis. The tube is surrounded by an envelope enclosing a reduced-pressure atmosphere that, as an example, may be composed of 10 cm of argon and an organic vapor which is usually about 1 cm of ethyl alcohol. Across this tube is imposed about 1000 volts dc. This voltage may be obtained from simple rectification of the output of a 60-cps transformer or, to reduce weight, volume, and filtering required, the rectified output of a 100-kc oscillator is frequently used. When an ionizing event takes place in the tube, a gas discharge takes place. This discharge may be extinguished by the organic vapor present. If, however, long counter-tube life is desired, an electronic extinguishing circuit is used to remove the voltage from the counter after a discharge has been triggered.

Several circuits[4] have been developed for this purpose, each having its own peculiar difficulties and advantages.

The generation and extinction of the gas discharge has lowered and raised the voltage across the Geiger counter tube, thereby indicating the presence within the tube of a single ionizing particle. This pulse, which is a few volts in amplitude and some 10^{-5} seconds wide, is amplified in the conventional manner and fed into an electronic pulse-equalizer circuit and then to either an electronic integrating circuit, to indicate the counting rate, or into a power amplifier that may operate a number register to totalize the pulses. Since electromechanical registers require at least 0.01 second to operate, they place an undesirable limit upon the speed of counting. Consequently, electronic scaling circuits, which permit only a known fraction of the pulses to reach the register, are frequently necessary. It is obvious that, even though the input Geiger counter tube is very simple, a complex array of electronic circuitry is necessary to utilize the device properly and in the most efficacious manner.

The preceding remarks apply only to the Geiger counter. However, if we examine working models of other nuclear radiation detectors, it is apparent that the major portion of their useful operation depends upon appropriate utilization of electronic circuits. This is to be expected, since nuclear detection usually requires fast, reliable indication of phenomena that occur at a

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[4] J. Strong, "Proceedings in Experimental Physics," Chap. VII, p. 259, Prentice-Hall, Inc., New York, N. Y., 1938.

very low voltage level. The application of electronic methods and circuits naturally offers the more obvious solutions to these problems. In fact, electronic applications to the detection, indication, and measurement of nuclear radiations represent the most apparent, but by no means the only, field of application to nuclear science.

ELECTRONICS AND NUCLEAR PARTICLE ACCELERATORS

The application of electronics to the vast number of accelerators has been no less spectacular. In fact, the world's most powerful oscillators are being developed for use in the larger cyclotrons. Whether the nuclear accelerator be a synchrocyclotron or a cavity linear accelerator, the electronics expert finds countless electronic problems that have been previously unsolved and, consequently, tax his ingenuity to the utmost.

ELECTRONICS AND THE FUTURE OF NUCLEONICS

As time passes, nuclear electronic applications and problems seem to increase in number, thus offering to the engineer a bright new field of the future, fraught with difficulty, but great in interest.

On the other hand, the electronics part of the nuclear sciences presents a challenge to the electronics engineer. In this borderline or dual subject, it is not sufficient for the electronics expert to be capable in his own field. He must also be an expert in nucleonics. Since the science of nuclear phenomena is a new field, the nuclear expert is not always able to put his electronics problems to the electronics expert in the proper language so that he may obtain the benefit of his thinking and experience. It is usually necessary for the electronics scientist to follow the nuclear problems involved, and to find and point out the defects in existing techniques and to offer new solutions to these difficulties. To do this, he must have a sound understanding of the basic phenomena in nuclear physics, in addition to his electronics training. While this may appear to be too much to expect, there is no substitute for this combination of knowledge. To those that qualify, new horizons are opening up, second in importance to none, and with tremendous capabilities for the good or the destruction of mankind.

Considerations in the Design of a Universal Beacon System*

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Summary—**Airborne beacons are an essential element in aircraft navigation and traffic-control systems since they provide a means of (a) radar range extension and (b) intelligence transmission. However it is important that a universal beacon be provided which will, in effect, operate with all ground and airborne radar equipments, regardless of the operating frequency of the primary radar equipment. Also, the airborne beacon must provide the maximum of facilities for intelligence transmission. The paper outlines the specifications for a proposed universal beacon system satisfying the above basic requirements, and discusses certain design criteria for the several components of the proposed system.**

I. INTRODUCTION

AIRBORNE BEACONS or transponders have, during the past few years, received considerable attention by both military and civil agencies because of their universal application to aircraft navigation and traffic-control systems.[1,2] It has been found, in particular, that airborne radar beacons show considerable promise of contributing to the solution of two problems commonly associated with all such systems; i.e., (1) radar range extension, or "radar assist," and (2) automatic intelligence transmission. In fact, it has been shown that airborne transponder beacons are capable of performing so many essential functions that the problem is no longer one of deciding if a beacon is required; rather, the problem has become one of deciding how many functions can be accomplished by a simple beacon, or of deciding how many beacons must be carried in a single aircraft to accomplish essential functions. It is, of course, desirable to carry only one small, light-weight beacon in the aircraft. This situation received careful consideration during the recent deliberations of Special Committee No. 31, which was established by the Radio Technical Commission for Aeronautics to determine the basic system considerations and equipment types required for a nationwide air navigation and traffic-control system.[3] The material presented in this paper is believed to be essentially in

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* Decimal classification: R526.1. Original manuscript received by the Institute, March 22, 1948; revised manuscript received, July 6, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 22, 1948.

† Communication and Navigation Laboratory, Air Materiel Command, Wright Field, Dayton, Ohio.

[1] Ralph D. Hultgren and Ludlow B. Hallman, Jr., "The theory and application of the radar beacon," PROC. I.R.E., vol. 35, pp. 716–730; July, 1947.

[2] Arthur Roberts, "Radar Beacons," vol. 3, MIT Rad. Lab. Series, McGraw-Hill Book Co., Inc., New York, N. Y., 1947.

[3] "Air Traffic Control," prepared by RTCA SC-31, Paper 27-48/DO-10, May 12, 1948, Radio Technical Commission for Aeronautics, Rm. 597, Department of State Bldg., 17 St. and Pennsylvania Ave., Washington 25, D. C.

accord with the applicable guiding principles established by SC-31. The purpose of the paper is to outline basic requirements. A discussion of detail techniques will not be attempted.

II. THE PROBLEM OF RADAR RANGE EXTENSION

Small aircraft, particularly those with fabric-covered airfoil and fuselage surfaces, make poor radar targets, especially under adverse weather conditions, and are often difficult to detect beyond a few miles from the radar site. These aircraft become hazards when the ground radar equipment is used to control the traffic pattern around the airport. If all such aircraft were equipped with radar beacons, this problem would, of course, be much less serious. To be universally effective, the airborne beacon must, in effect, respond to all ground and airborne navigational radar frequencies, and any universal beacon system must satisfy this requirement. It also follows that the airborne beacon must be so designed that it is effective out to the maximum operating range of the associated primary radar equipment.

III. THE PROBLEM OF AUTOMATIC INTELLIGENCE TRANSMISSION

It is convenient to divide the intelligence to be transmitted, both air-to-ground and ground-to-air, into two fundamental types, as follows: (a) intelligence which is fundamental to the air traffic-control system; and (b) intelligence which is fundamental to the navigation of the aircraft.

In establishing the principles upon which the universal beacon system is based we propose that the universal airborne beacon be required to handle only that form of intelligence transmission which is fundamental to the air traffic-control system.

The following types of intelligence, listed in order of operational importance, are fundamental to the air traffic-control system and should, therefore, be capable of automatic transmission to the ground traffic-control agency through the universal beacon system: (a) aircraft range and azimuth (automatically provided in connection with the radar-assist or range-extension function described above); (b) aircraft altitude; and (c) aircraft identity.

In addition to the above, the proper control of the air traffic requires a private-line communication channel between ground control and each aircraft, for the automatic transmission and reception of routine traffic-control intelligence such as "clear to land," "increase altitude to———," "decrease altitude to———," "hold," "ahead of schedule," "behind schedule," "turn right —— degrees," "turn left —— degrees," etc. Also, the private-line communication channel may be used to indicate the failure of the system at any point. It is probable that all such information required could be transmitted using a four- or five-pulse binary code.

IV. PROPOSED SPECIFICATIONS FOR THE UNIVERSAL BEACON SYSTEM

a. General

Consideration of the basic problem, as outlined in the foregoing, leads to the following conclusions regarding the over-all system:

(1) The airborne transponder should be so designed that its reply signal is capable of being seen on the display of any pulse-type ground or airborne radar equipment operating in the approximate frequency range of 100 to 20,000 Mc. The use of an interrogator-responser equipment operating in a common frequency band provides a convenient means of satisfying this requirement, and a brief description of such equipment will be given later in this paper. Assignments have been provided in the bands 960–1215 and 1300–1660 Mc for aeronautical radio aids. It is proposed that the common interrogator-responser channel be assigned in one of these bands.

(2) The characteristics of the airborne beacon (secondary radar) system should be such that its operating range, azimuth, and range definition is at least as great as that of the highest-power (greatest-range) primary radar equipment with which it will be used, when the primary radar equipment is operating at maximum range against an effective reflecting area equivalent to, say, a close formation of six aircraft of the DC-6 type. The beacon system should also provide satisfactory operation down to a minimum of at least 1/2 mile range at an altitude of 1000 feet.

(3) When the azimuth and range definition of the primary radar equipment is greater than that which is practicable at the selected common beacon interrogator-responser frequency, then simultaneous or dual interrogation at the primary radar frequency may be required. Even when this is not necessary, however, the L-band common interrogator-responser channel is extremely useful in that it provides a means for preventing overinterrogation, or saturation, of the transponder by spurious radar signals. Also, the L-band channel may be used to provide a coded signal for adjusting the sensitivity of the transponder receiver to provide improved azimuth resolution when the aircraft is less than, say, ten miles from the ground radar equipment.

(4) The capability of supplying the types of data indicated below should be provided by the universal beacon system.

(i) *Radar Assistance.* (Both range and azimuth.) This constitutes the basic radar-beacon function.

(ii) *Altitude.* It appears technically feasible to provide aircraft altitude data by suitable interrogation or reply codes, frequency channeling, or by a combination of pulse codes and frequency channeling. Further study will be necessary before a decision can be made as to just how this facility may be best provided. System requirements indicate a need for 20 altitude layers, although a continuous indication of altitude may ultimately prove the more desirable.

(iii) *Identity*. The simultaneous display of the identity of all aircraft is not required. Rather, provision should be made for the display of the identity of particular aircraft upon request. The period of time between the selection of the target and the display of the identifying information required should be as short as practicable and preferably less than ten seconds.

There are two fundamental types of identity to be considered; namely, identity of the "Who are you?" type, and identity of the "Where are you?" type.

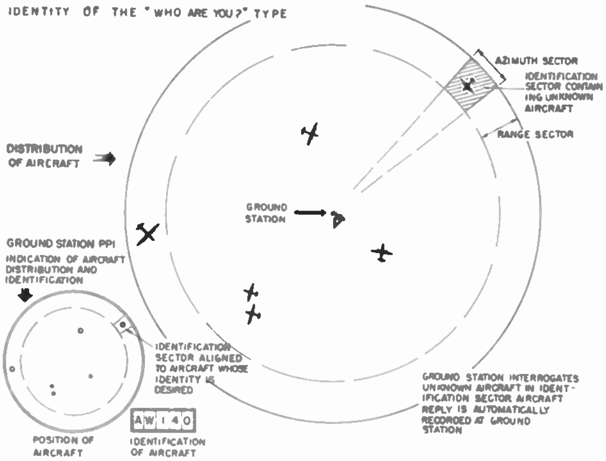

Fig. 1—Display of identity of the "Who Are You?" type as applied to the universal beacon system.

Identity of the "Who are you?" type is illustrated in Fig. 1. Here the interrogation is limited to the particular air-space segment in which the unknown aircraft is located. Azimuth definition is obtained from the directivity of the L-band ground antenna. Range identity may be obtained by a pulse code which is controlled by the range position of the unknown aircraft as viewed on the ground-equipment plan-position indicator (PPI). Both the azimuth positioning of the interrogator-responser L-band antenna and the range identity pulse-code adjustment are automatically accomplished when the ground traffic controller adjusts the cross hairs of his range and azimuth cursors so that they cross over the spot on the PPI displaying the unknown aircraft. In the aircraft, provision is made to utilize the airborne distance-measuring equipment (DME) to establish a decoding circuit, such as a double-pulse gate, which will pass the identity interrogating pulse only if it is properly coded to correspond to the aircraft's distance from the ground station. Use of the airborne DME to establish range identity requires that the ground beacon used with airborne equipment be suitably located with respect to the ground radar and interrogator-responser equipments. Also, the same ground beacon must be used by all aircraft in a given control area. Where this is not feasible, identity of the "Who are you?" type must be accomplished using azimuth discrimination only. Ultimately, it may be found advantageous to further limit the air space interrogated for identity of the "Who are you?" type by interrogating only that altitude sector occupied by the unknown aircraft.

Identity of the "Where are you?" type is illustrated in Fig. 2. This type of identity is provided when the aircraft equipment automatically, or the aircraft pilot or navigator manually, operates a control which causes a special identifying signal to be transmitted when the ground controller broadcasts a request for this type of identity either by radiotelephonic or coded interrogation means; or by means of a special interrogation over the private-line circuit.

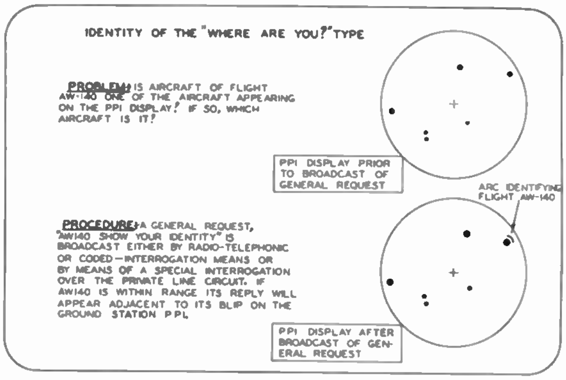

Fig. 2—Display of identity of the "Who Are You?" type, showing a problem, and the procedure to be followed.

(iv) *Private-Line Communication*. It is proposed that a separate communication channel, consisting of one transmitting and one receiving frequency, be set aside in the 960- to 1660-Mc band, and designated specifically for automatic intelligence transmission of the type which is required in connection with the universal beacon system.

Since it is required to handle all aircraft within range on this single communication channel, some method of time division is required. We may, for example, choose a 6-second cycle. If each aircraft is assigned a 3000-microsecond interval in each 6-second cycle, two thousand aircraft can be handled in a given area on the same pair of frequencies. This requires synchronization of all ground stations and all aircraft equipments in the area. This requirement could be satisfied by utilizing the rotational rate of the navigational omnirange pattern as the source of synchronization voltage. Intelligence would be transmitted by means of suitable binary codes.

b. Airborne Components

The airborne transponder components of the universal beacon system are illustrated diagrammatically in Fig. 3. L-band transmitter No. 1 with its associated decoder and modulator provide the radar assist (range and azimuth), aircraft altitude, and aircraft identity functions. L-band transmitter No. 2 and its associated decoder and modulator provide the private-line communication facility.

Only S- and X-band coincident-channel receivers are provided, since it does not appear at present that coincident interrogation at other frequencies will be a requirement within the next ten years.

It is proposed that the inherent delay of all airborne transponders in replying be standardized to remain constant under all service conditions to within ±0.2 microsecond, so that resulting range inaccuracies from this source shall not exceed ±100 feet.

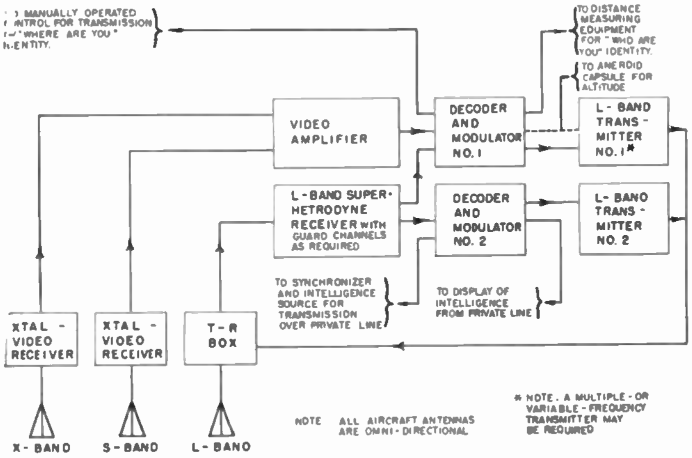

Fig. 3—Airborne transponder components of the universal beacon system.

In addition to the airborne transponder components illustrated in Fig. 3, an L-band interrogator-responser unit is proposed for aircraft carrying navigational-type radar equipment. The unit should have facilities provided for interconnection with any standard airborne radar system. However, it should also be capable of satisfactory operation independently of the airborne radar equipment.

The receiving system should be capable of receiving a selectable range of altitude-indicating signals.

It is suggested that the beacon reply be displayed on the indicator of the associated airborne radar equipment when such is available. For use in nonradar-equipped aircraft, a suitable display unit should be provided. The identity information could be displayed separately, and it is not necessary that the airborne equipment display the identity of all beacon replies.

c. Ground Components

The proposed ground components of the universal beacon system are diagrammatically illustrated in Fig. 4. Although only one L-band directional antenna is shown in Fig. 4, it is likely that two such antenna systems may be found operationally desirable. One would be used for continuous automatic azimuth tracking, the other would be rotated or controlled manually for obtaining aircraft identity without interference with the automatic-tracking function.

Facilities should be provided for interconnection with any standard ground radar system, though the interrogator-responser equipment should also be capable of operation independent of the ground radar equipment.

The receiving system should be capable of receiving and displaying a selectable range or the entire range of altitude-data signals.

It is suggested that the beacon reply be displayed on the same type of indicator or the same indicator as the radar. The "radar assist" function and altitude information should be available continuously. The identity information may be displayed separately.

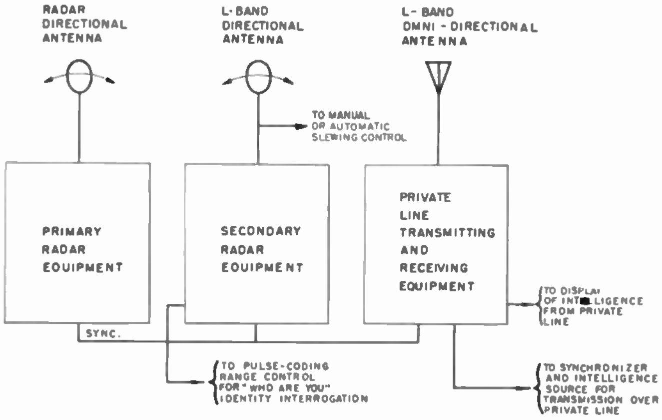

Fig. 4—Ground components of the universal beacon system.

ACKNOWLEDGMENT

The author wishes to acknowledge the contribution by personnel of the Communication and Navigation Laboratory, Wright-Patterson Air Force Base, in the general study of the universal beacon system. It is especially desired to acknowledge the assistance of S. A. Mundell, formerly Chief, Communication and Navigation Laboratory, and N. Braverman, Assistant Chief Engineer, Communication and Navigation Laboratory, in reviewing and criticizing the manuscript, particularly from the standpoint of their experience in working with RTCA Special Committee No. 31.

Three-Dimensional Representation on Cathode-Ray Tubes*

CARL BERKLEY†, ASSOCIATE, IRE

Summary—**A procedure for the representation of functions of a number of variables on the screen of a cathode-ray tube is developed. The procedure may be applied to any regular system of coordinates or any number of variables. The representation may take the form of an oblique perspective picture. The procedure consists of the following steps:**

1. Setting the form of the representation.

2. Deriving the position of the spot on the cathode-ray tube as a function of its true position in space.

3. Making the indicated corrections electrically.

Applications in the fields of mathematics, radar, electromagnetic theory, mechanical measurements, topographic surveying, and meteorology are described.

INTRODUCTION

PATTERNS WHICH give the illusion of a three-dimensional object are frequently encountered in work with cathode-ray tubes. These patterns are sometimes obtained intentionally, but more usually occur as a result of some cross coupling, oscillation, or defect in the circuits used. An analysis of such patterns was undertaken to find if they could be reproduced at will, and whether use could be made of the results. It was found that analysis is simple in many cases, and it is expected that the results will find wide application.

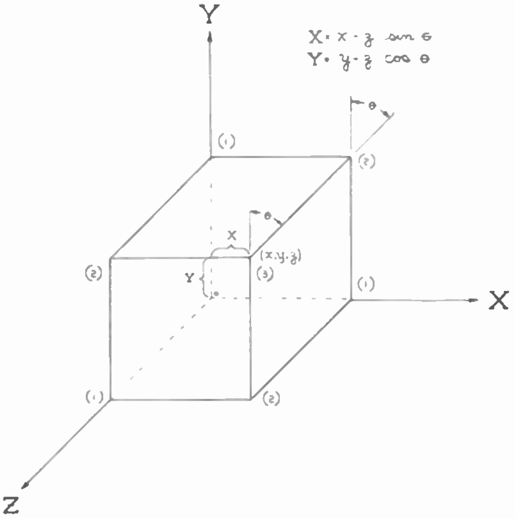

Fig. 1—Figure for the derivation of translating voltages for oblique representation.

Examples of such patterns appear in Figs. 2–5 and 8, which are photographs directly from the face of a cathode-ray tube. Since the same cathode-ray-tube pattern is viewed by both eyes in the majority of cases, it will be obvious that these pictures can only be representations or perspective projections which, at the most, can give an illusion of three dimensions. Two separate and distinct patterns, each viewed by one eye, are necessary to give the true effect of depth for the observer. Schmitt[1] has independently arrived at the electrical and optical conditions which must be satisfied for true stereographic representation, in addition to those to be described for the metrical projections.

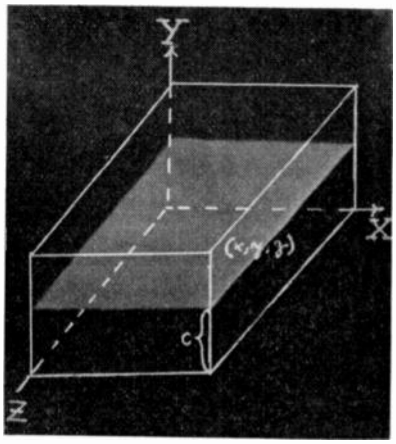

Fig. 2—Representation of the plane: $y-c=0$
V signal = 15-cps sawtooth
H signal = 3-kc sawtooth + 15-cps sawtooth.

An analysis of the geometry of perspective drawings shows that the procedures used in conventional representations may be expressed by mathematical notation. As an example, this may be very readily done as follows for the so-called "oblique" perspective representa-

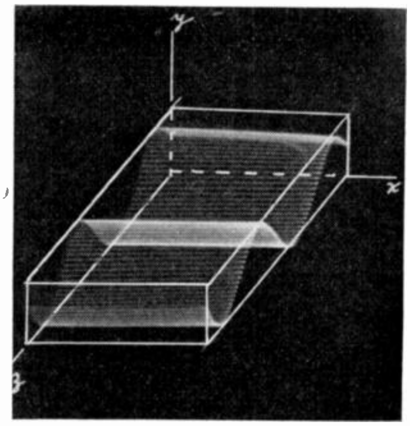

Fig. 3—The surface: $A \sin z-y=0$
V signal = 30-cps sawtooth − 60-cps sine wave
H signal = 1.5-kc sawtooth + 30-cps sawtooth.

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* Decimal classification: R201.7×621.375.9. Original manuscript received by the Institute, April 14, 1947; revised manuscript received, June 7, 1948. Presented, 1947 IRE National Convention, March 5, 1947, New York, N. Y.

† Allen B. DuMont Laboratories, Inc., Clifton, N. J.

[1] O. H. Schmitt, "Cathode ray presentation of three dimensional data," *Jour. Appl. Phy.*, vol. 18, pp. 819–829; September, 1947.

tion most used in mathematical and engineering work. Let x, y, and z be the co-ordinates of a point in three-dimensional space represented by an oblique projection in Fig. 1. X and Y are the co-ordinates in the two-dimensional representation on the cathode-ray-tube face. If the origin is the same in both spaces and θ is the angle the Z axis makes with the Y axis in the projection, then it is evident from the geometry of the figure that

$$X = x - z \sin \theta \quad (1)$$

$$Y = y - z \cos \theta. \quad (2)$$

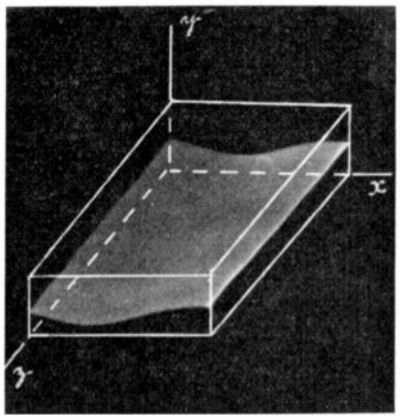

Fig. 4—The surface: $A \sin x + y = 0$
V signal = 60-cps sawtooth + 3-kc sine wave
H signal = 3-kc sawtooth + 60-cps sawtooth.

These formulas show that, for a proper oblique representation, to the x value of the function in space must be added a certain fraction of the z value at that point to get the proper crt co-ordinate in the X or horizontal direction. Similarly for the y value. If z is the independent variable, z may be given a form such as a sawtooth function, in which case it is only necessary to add to the x and y values sawtooths whose value is multiplied by $\sin \theta$ and $\cos \theta$. This has been done in Figs. 2–5 in the text, using the circuit shown schematically in Fig. 6. The results show that it is possible, at least for the oblique representation, to represent these functions by the simple addition of properly obtained electrical values to the deflection plates. Any empirically desired values can readily be obtained by electronic means; for instance, by scanning, photoelectrically, a printed image of the function by a slit.[2] Consider the usefulness of this method. In any mathematical analysis of a complicated function, expressions are arrived at which are extremely difficult to visualize merely from the formulas. By translating these functions into electrical values, these otherwise unimaginable properties can be readily made visible, and the correspondence in shape between functions which appear unrelated when expressed as formulas can be easily recognized.

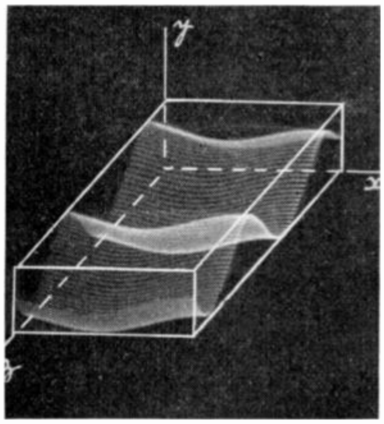

Fig. 5—The plane: $A \sin x - y + B \sin z = 0$
V signal = 30-cps sawtooth + 1.5-kc sine wave + 60-cps sine wave
H signal = 1.5-kc sawtooth + 30-cps sawtooth.

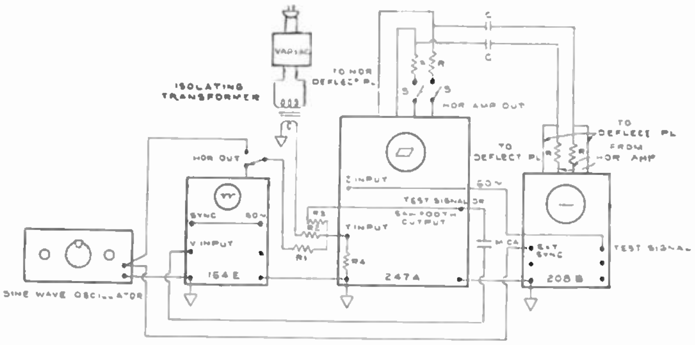

Fig. 6—Schematic for production of three-dimensional patterns.

PROCEDURE

While the procedure which is used here is described for up to three variables, it will likewise be useful for the two-dimensional representation of any number of variables, once the form of the projection is agreed upon. While the illustration is also limited to the case of the oblique projection, it will be evident that other projections can be similarly treated.

An equation in three variables evidently represents a surface. This surface may be an abstract mathematical function or, as shown later, may be that of a real object, such as the surface of a cylindrical piston, which can be expressed analytically, or a topographic map, which cannot. When an analytically expressed surface is to be plotted, the usual procedure is to allow the independent variables to assume all values between the limits desired and then plot the values of the function. This generally entails a tremendous amount of work. However, in many cases the effort is justified by the importance of the results.[3]

It is possible to represent a variable on a cathode-ray tube and to make the variable assume all values within limits by very simple electrical means. This can be done, for example, by applying a sweep voltage as the independent variable or variables. The peak-to-peak

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[2] R. C. Walker, "Electronic Equipment," Chemical Publishing Company, 1945, p. 143.

[3] E. Jahnke and F. Emde, "Funktionentafeln," Dover Publications, 1943.

values of the sweep, which need not necessarily be a sawtooth, are chosen to correspond to the limits within which it is desired to plot the variable. The corrections of (1) and (2) are then necessary to produce the particular projection desired for an oblique perspective representation.

In order to assume all the values within limits in an area, it is necessary, of course, to use two sweeps at right angles to one another which scan this area. To cover a solid, it is necessary to use three sweeps essentially at right angles. If sweeps are chosen for the x and z values, the mathematical corrections required in (1) and (2) can be seen to represent the addition of a portion of the z sweep to the x and y signals; that is, to the horizontal and vertical deflection plates. The application of this method will now be described for a number of cases.

Fig. 2 evidently represents a plane whose formula is

$$y - c = 0 \qquad 0 < x < x_1,\ 0 < z < z_1$$

in which x and z are independent variables.

In this case the two sweeps which were chosen to cover the area were synchronized 15-cps and 3-kc sawtooth sweeps. These may be expressed by

$$x = nkt \tag{3}$$

for the actual value of one sweep in three-dimensional space in the x direction, and

$$z = kt \tag{4}$$

for another sweep in the x, z plane

where t = time

k = a constant

n = frequency ratio between sweeps.

Then, applying the operation of (1) to the X coordinate on the oscillograph by substituting the value of X from (3) in (1) to find the horizontal deflection for proper representation, we obtain

$$X = nkt - kt \cos\theta$$

Since $\cos\theta$ is constant, $k \cos\theta = k_1$, which changes the previous formula into

$$X = nkt - k_1t. \tag{5}$$

This can be seen to be the algebraic sum of the high-frequency sweep and a fraction, $\cos\theta$, of the low-frequency sweep. This is the signal applied to the horizontal deflection plates in Fig. 2. To find the Y deflection, we apply the indicated correction to the function

$$y - c = 0,$$

obtaining for the Y value, or vertical deflection voltage,

$$\begin{aligned} Y &= c - kt \sin\theta \\ &= c - k_2t, \quad \text{where} \quad k \sin\theta = k_2. \end{aligned} \tag{6}$$

The c in this equation represents the vertical positioning voltage, and k_2t is proportional to the Z signal, which value can be obtained by adjusting the Z-sweep gain.

A useful accessory for setting up the proper amplitudes is a DuMont type 264A voltage calibrator. If, for example, 1 volt is selected as representing one unit of the independent variable z, and $\cos\theta = 0.707$, then the vertical sweep signal amplitude is set at 0.707 volt peak-to-peak as read on the 264A to represent $0 > z > 1$. The value of C may be read with a voltmeter on the positioning voltage or by having the positioning control directly calibrated.

A similar analysis can obviously be carried through for other functions. Consider the surface $A \sin z - y = 0$.

An analysis of this formula shows that it can be represented as follows:

On the vertical plates is produced the algebraic sum of

$$y = k_3t,$$

the sweep, and

$$y = A \sin z.$$

On the horizontal plates is produced the sum of

$$x = nk_3t - k_4t,$$

which is the algebraic sum of the high- and low-frequency sweeps. This surface is shown in Fig. 3.

Consider Fig. 4, which is similar to Fig. 3. The equation of this surface is

$$A \sin x + y = 0.$$

This is similar to Fig. 2 except that the wave is traveling down the X axis. The figure is seen (after analysis) to be obtained by putting on the vertical plates the sum of the functions:

$$y = -\sin x, \quad \text{and} \quad y = kt, \quad \text{or,}$$

expressed as a single waveshape,

$$y = kt - \sin x,$$

and on the horizontal plates, the sum of

$$x = -kt \quad \text{and} \quad x = nkt, \quad \text{or}$$

$$X = nkt - kt.$$

Putting this into electrical terms, we again see that on the vertical plates there is the algebraic sum of a sawtooth and sine-wave function, and on the horizontal plates the sum of two sawtooth functions.

It will be seen, upon study of these last three diagrams, that the illusion of a perspective drawing is given essentially by scanning the area with the two sawtooth voltages and by adding a portion of the vertical-deflection sawtooth to the horizontal plates to give the illusion of a Z plane in space. Once this is realized, it will be fairly easy to convert from the formulas to the desired pattern and back merely by inspection.

The equation for the surface in Fig. 5 is $A \sin x - y + B \sin z = 0$. This can be readily visualized from Fig. 5 as a surface resulting from the combination of two sine waves traveling along the x and z axes.

Figs. 3 and 4 may represent the magnetic field intensity between the walls of a waveguide, neglecting boundary conditions (that is, assuming the side walls have no effect), if the waves are traveling at right angles to each other.[4]

Fig. 5 shows the sum of the magnetic field intensities produced by these waves.

REPRESENTATION OF FUNCTIONS OF THREE VARIABLES

A function of one variable, $f(x)=y$, requires a two-dimensional surface (a plane) for its representation. A function of two variables requires a three-dimensional surface as in Figs. 2–5. Reasoning by extension requires a four-dimensional figure to represent a function of three variables $f(x, y, z)=u$. For a generalized representation, however, and with the restriction that the result is an arbitrary convention, and not a four-dimensional figure, any representation on a plane surface (such as a cathode-ray-tube face) may be used from which the values of x, y, z, and u can be read at any point. To produce such a representation, it is necessary to find one additional parameter which can be easily recognized on a cathode-ray tube. The one additional variable already available on most oscillographs is the brightness. This may be chosen to represent the value of u. This value can then be measured by reference to a source of calibrated brightness or by a comparison of the beam current obtained at any point in the function.

The appearance of many simple functions plotted in this manner can be easily visualized and they can be readily produced. Consider the equation

$$x + y + z = u \qquad \begin{array}{l} 0 < x < 1 \\ 0 < y < 1 \\ 0 < z < 1. \end{array}$$

This is represented on the cathode-ray tube by a three-dimensional cube similar to Fig. 1, in which any point has a brightness equal to the sum of the x, y, and z values at that point. This can easily be produced on an oscillograph by adding the x, y, and z components of the signal and applying the sum to the blanking amplifier. In Fig. 1, the numbers in parenthesis indicate the value of the function, and therefore the brightness of the corresponding corners of the cube. This pattern has actually been produced, but is not shown since the reproduction process does not adequately represent the slight brightness shadings which exist.

EXPERIMENTAL SETUP OF PATTERNS

Fig. 6 shows schematically the experimental setup used to produced Figs. 2–5. The final representation appears on the screen of the DuMont type 247A projection oscillograph. The 164E produces a sweep at some submultiple of 60 cps. The variac produces a variable 60-cps wave which is adjusted to obtain the proper vertical amplitude in Figs. 3 and 5. The 208B produces the 3-kc sweep in Figs. 2–5. The sine-wave oscillator generates the 3-kc sine wave in Figs. 4 and 5. The network consisting of resistors R_1 to R_4 adds the signals from the 247A sawtooth output, the variac, and the 164E or the oscillator, as may be shown from the superimposition theorem. The resistors are chosen of a large enough value so that each generator output is not loaded down by the internal impedances of the others in parallel. For demonstration purposes, the 247A is used as a projection oscillograph. To heighten the illusion of three dimensions, the co-ordinate axes or the limits to which the variables are permitted to explore may be projected separately or by electronic means. The axes have been drawn in the figures for this purpose. A certain percentage of the observers have difficulty in visualizing the shape of a surface or volume unless the axes are so indicated.

To produce Fig. 2, only the 247A and 208B are needed. The high-frequency sweep of the 208 is added to the low-frequency sweep from the horizontal amplifier and applied to the horizontal deflection plates of the 247. The sweep in the 247 is also taken from the front panel output and applied to the vertical amplifier. By varying the gains of the amplifiers, the constants in (5) and (6) may be adjusted so that the angle at which the projection is viewed may be changed at will. By varying the frequencies and adding a third sawtooth from the 164E, either the top or bottom of the figure may be viewed or a series of planes one above the other will be displayed, as will be evident after consideration of the motion of the spot. By using potentiometers ganged so as to control the value of θ, the pattern may be rotated at will to examine various aspects of the figure on the screen.

Fig. 3 may be produced by adding the output of the variac to the vertical deflection. Fig. 4 is produced by turning the variac to 0 and applying the sine-wave oscillator output instead. Fig. 5 is produced by adding both sine waves of the same amplitude.

The function $x+y+z=u$ can be produced by adding to the sweeps of Fig. 2 a third sweep applied as the independent y variable and obtained from the 164E. The sum of these three sweeps is then produced separately and applied to the blanking amplifier of the 247.

OTHER APPLICATIONS

1. *Mathematics:* The use of apparent three-dimensional representation of mathematical functions on a cathode-ray tube has been covered above. The costly and laborious drawings for mathematical and theoretical texts can be greatly facilitated by producing such patterns on a cathode-ray tube and making a photographic record.

2. *Electromagnetic Fields:* This method will be very useful in the correct adjustment and study of uhf

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[4] J. Skilling, "Fundamentals of Electric Waves," John Wiley & Sons, Inc., New York, N. Y., 1942; figs. 57 and 58, p. 33.

resonators, couplings, and transmitters. The wave patterns in Figs. 2–4 can be obtained in an actual case by using a moving pickup probe whose motion is synchronized with that of the spot. Any other spatial field distribution, such as in acoustics, hydrodynamics, aerodynamics, optics, and fluid mechanics, can be plotted with appropriate pickup devices.

3. *Topographic Surveying and Meteorology:* By applying a third sweep to the sum of the two functions on the vertical plates, it is possible to make the locus of the spot appear to scan a volume. For example, in Fig. 2, by adding to the vertical sawtooth signal another voltage

$$y = \frac{kt}{n}$$

where k/n is a submultiple of the other sweeps, the volume shown in the figure will be scanned. This volume may be used to represent a certain section of the earth's surface being mapped by an airplane flying above it, as shown in Fig. 7. By properly arranging to brighten the beam or displace it, depending upon the reflected rf pulses, a three-dimensional picture of the territory will be shown. This may be used similarly for meteorological purposes, since it is possible to so gate the transmitter and receiver as to scan only a particular volume of the aerosphere. This method would make possible the indication of the contour of a cloud or storm formation, in addition to its size and location.

4. *Radar Indication of Azimuth, Range, and Elevation:* The same principles obviously can be applied to radar. By putting in three sets of controls for a movable marker spot, it is possible to blank in a particular spot in the volume scanned and then read off azimuth, range, and elevation from the three controls. In order to heighten the illusion of perspective for these last two applications, it may be desirable to modulate the size of the pattern from one end to the other by means of keystoning or intensifier potential modulation, as is indicated in Fig. 7. A typical example of the type of pattern to be expected from such radar or topographic application is shown in Fig. 8. This shows approximately what can be expected from the shore line of a river with a road over a pontoon bridge crossing the river. This was actually obtained by height-modulating the $y=c$ plane in Fig. 1 by the addition of a sharp vertical pulse synchronized with the sweeps. The exact occurrence of the pulse was varied during the z-sweep cycle.

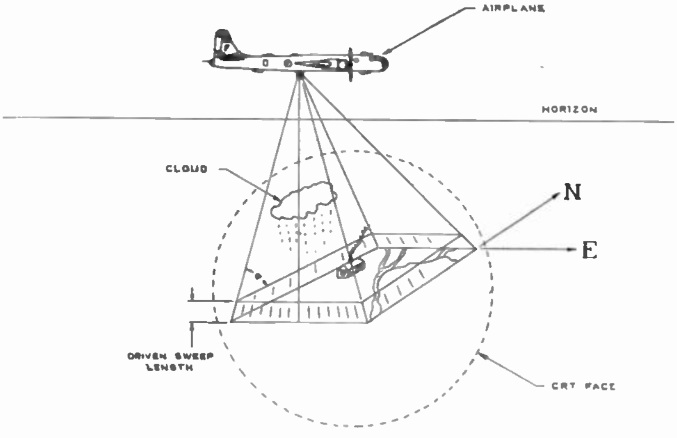

Fig. 7—Oblique perspective representation.

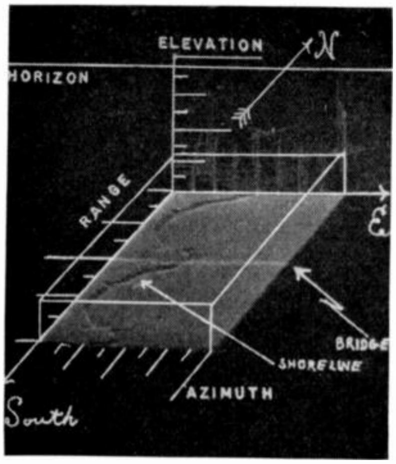

Fig. 8—Simulated radar pattern such as may be obtained from a bridge and the shoreline of a river.
V signal = 30-cps sawtooth + sharp pulse at 3-kc average repetition rate.
H signal = 3-kc sawtooth + 60-cps sawtooth.

Another radar representation that may be improved by the use of these techniques is the PPI. If the vertical gain is reduced, the circular field appears ovoidal as if viewed from a point in space outside the area being scanned and not directly over the transmitter. The video information may then be clipped to eliminate the noise and applied as an addition to the vertical deflection. This gives a picture in which an untrained observer can readily recognize objects which otherwise would be merely dark or bright blobs on the map. If some video is added to the grid, as well, mountains may be seen as bright projections from the flat earth plane.

5. *Mechanical Measurements:* Fig. 9 shows a setup for scanning the surface of an engine piston. This may be done on a lathe, for example, using a Brush surface-analyzer pickup. The helical travel of the spot obtained from the two-phase generator coupled to the piston, and the motion of the positioning potentiometers, can then

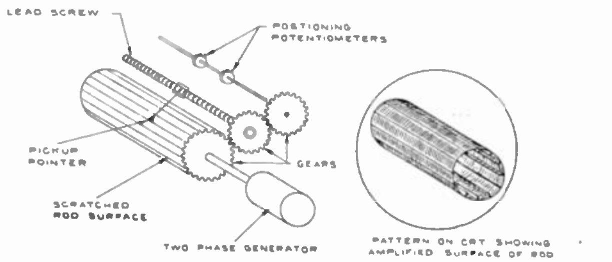

Fig. 9—Application showing the representation of surface defects.

be modulated using a DuMont type 275A cathode-ray oscillograph in order to obtain an accurate but amplified visualization of the surface of the piston as seen in Fig. 9. If a representation in polar form is not desired, the oblique method described in the preceding sections can also be used, in which case a pattern similar to Fig. 8 is obtained.

6. *Complex Time Relations:* This method will also prove useful whenever there are complex time or phase relations to be plotted. These exist, for example, in the representation of electron bunching in velocity modulated tubes,[5] or in a composite sync signal such as is used for color television.

Numerous methods have been developed recently for the construction of equivalent circuits for three-dimensional objects.[6] Such equivalent circuits have been carried so far as to incorporate an equivalent circuit for the universe.[7] The operation of these circuits corresponds exactly to their equivalents. The results can best be shown in the three-dimensional projections proposed. Numerous other applications will doubtless occur for this method. It should be possible to extend this method to include functions of more than three variables.

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[5] Klystron Technical Manual, Sperry Gyroscope Co., Inc., December, 1944.

[6] Gabriel Kron, "Equivalent circuits to represent the electromagnetic field equations," *Phys. Rev.* vol. 64, pp. 126–128; August, 1943.

[7] S. Austen Stigant, "Equivalent circuits for the electromagnetic field," *Beama Jour.*, pp. 412–416; December, 1944.

A Single-Control Variable-Frequency Impedance-Transforming Network*

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Summary—**The basis of the variable-frequency transformer to be described is the familiar transmission-line stub matching section. Although an infinite number of structures may be found to produce the desired transformation at any one frequency, it is generally necessary to vary the lengths of the component lines to maintain matching when the frequency is changed.**

Mechanical complexities and electrical noise and losses in sliding contacts often make such methods undesirable. The particular solution to the stub matching problem described here is a structure which will maintain a perfect impedance match over a wide frequency range with the adjustment of a single reactance shunted across the load at the receiving end.

I. Design Formulas

THE FOLLOWING symbols will be used:

Z_{01} = characteristic impedance of a line between the generator and the load

Z_{02} = characteristic impedance of a short-circuited stub line across the generator

$\theta_1 = 2\pi$ (length of line between the generator and the load)/(wavelength of applied energy)

$\theta_2 = 2\pi$ (length of stub line across the generator)/(wavelength of applied energy)

φ = desired voltage transformation

E_S = sending-end emf

I_S = sending-end current

E_R = receiving-end emf

I_R = receiving-end current

R_R = shunt load resistance

X_R = shunt load reactance

X_0 = value of X_R for perfect matching

R_g = internal resistance of generator

$$Y_{01}=\frac{1}{Z_{01}},\quad Y_{02}=\frac{1}{Z_{02}},\quad g_R=\frac{1}{R_R},\quad g_g=\frac{1}{R_g},$$

$$ib_R = \frac{1}{jX_R}.$$

The formulas required for design are:

$$Z_{01} = \varphi R_g \qquad X_0 = \frac{\varphi^2 R_g}{\varphi - 1}\tan\theta_1$$

$$Z_{02} = \frac{\varphi}{\varphi - 1} R_g \qquad R_R = \varphi^2 R_g,\ \theta_2 = \theta_1.$$

The schematic diagram is shown in Fig. 1. The bandwidth characteristics will be discussed in Section III.

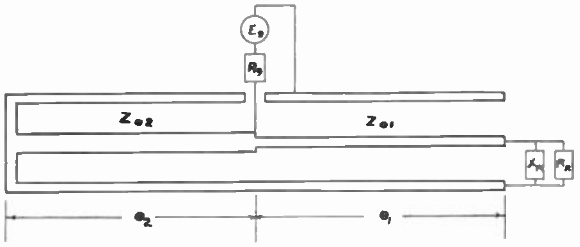

Fig. 1

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* Decimal classification: R117.121. Original manuscript received by the Institute, December 4, 1947; revised manuscript received, May 17, 1948.

† Columbia Broadcasting System, Inc., New York, N. Y.

II. THEORY OF OPERATION

The input conductance of a line is a function of the characteristic impedance of the line, the length of the line, the load conductance, the load susceptance, and the frequency. If the first three parameters are correctly chosen, only the load susceptance need be changed when the frequency is changed to keep the input conductance constant.

The input susceptance with this choice of parameters is always the negative of the susceptance of a stub of fixed dimensions. The derivation of the required relationships between these parameters is shown below.

The input admittance of the line connecting the generator to the load, g_R+jb_R, is given by

$$Y_{01}\frac{Y_R+jY_{01}\tan\theta_1}{Y_{01}+jY_R\tan\theta_1}.$$

The real component of this admittance is

$$g_RY_{01}^2\frac{1+\tan^2\theta_1}{(Y_{01}-b_R\tan\theta_1)^2+g_R^2\tan^2\theta_1}$$

If b_R is selected as $(Y_{01}\pm g_R)\cot\theta_1$, the real component becomes Y_{01}^2/g_R. It may be noted that if $\theta=\pi/2$, then $b_R=0$, and we have as a special case the quarter-wave series transformer.

The imaginary component of the input admittance is

$$Y_{01}\frac{-Y_{01}b_R\tan^2\theta_1+(Y_{01}^2-b_R^2-g_R^2)\tan\theta_1+b_RY_{01}}{(Y_{01}-b_R\tan\theta_1)^2+g_R^2\tan^2\theta_1}.$$

When the value of b_R selected above is substituted in this expression, the imaginary admittance component is

$$-\frac{Y_{01}}{g_R}(g_R\pm Y_{01})\cot\theta_1.$$

This component may be tuned out by placing across the input terminals another susceptance of opposite sign which also varies as $\cot\theta_1$. This requirement is met by a shorted line of equal length. Because the shorted line has a negative susceptance, the sign of the previous expression must be positive. Then,

$$Y_{02}=\frac{Y_{01}}{g_R}(Y_{01}-g_R),\quad Y_{01}>g_R,\quad \theta_2=\theta_1.$$

The total input admittance is, then,

$$\frac{Y_{01}^2}{g_R}=g_g.$$

Letting

$$g_g=\varphi^2g_R,$$

then

$$Y_{01}=\frac{g_g}{\varphi},\qquad Y_{02}=\frac{\varphi-1}{\varphi}g_g$$

and

$$b_0=g_g\frac{\varphi-1}{\varphi^2}\cot\theta_1.$$

III. RESPONSE AT ANY FREQUENCY

The equations

$$E_S=AE_R+BI_R$$

$$I_S=CE_R+DI_R$$

completely characterize the network. The coefficients may be found by writing similar equations for each network element and eliminating the intermediate emf's and currents. This is equivalent to the following matrix multiplication:

$$\begin{Vmatrix}E_S\\I_S\end{Vmatrix}=\begin{Vmatrix}1 & 0\\ \dfrac{-j(\rho-1)}{Z_{01}}\cot\theta_1 & 1\end{Vmatrix}\times\begin{Vmatrix}\cos\theta_1 & jZ_{01}\sin\theta_1\\ \dfrac{j\sin\theta_1}{Z_{01}} & \cos\theta_1\end{Vmatrix}\times\begin{Vmatrix}1 & 0\\ jb_R & 1\end{Vmatrix}\times\begin{Vmatrix}E_R\\I_R\end{Vmatrix}$$

Matrix of Stub Matrix of Series Line Matrix of b_R

$$\begin{Vmatrix}E_S\\I_S\end{Vmatrix}=\begin{Vmatrix}\cos\theta_1-Z_{01}b_R\sin\theta_1 & jZ_{01}\sin\theta_1\\ \dfrac{j}{Z_{01}}\left[Z_{01}\rho b_R\cos\theta_1+\dfrac{1-\rho\cos^2\theta_1}{\sin\theta_1}\right. & \rho\cos\theta_1\end{Vmatrix}\times\begin{Vmatrix}E_R\\I_R\end{Vmatrix},$$

or

$$A=\cos\theta_1-Z_{01}b_R\sin\theta_1$$

$$B=jZ_{01}\sin\theta_1$$

$$C=\frac{j}{Z_{01}}\left[Z_{01}\rho b_R\cos\theta_1+\frac{1-\varphi\cos^2\theta_1}{\sin\theta_1}\right]$$

$$D=\varphi\cos\theta_1.$$

It has been shown that the mismatch ratio η, which is the ratio of the power delivered to the load with perfect matching to the power delivered to the load under actual conditions, is given for any four-terminal network by

$$\eta = \frac{AR_R + B + CR_gR_R + DR_g}{4R_RR_g}.$$

(For the sake of completeness, the four-terminal mismatch ratio formula is derived below:

$$E_S = AE_R + BI_R \qquad\qquad E_S = (AR_R + B)I_R$$
$$\text{or}$$
$$I_S = CE_R + DI_R \qquad\qquad I_S = (CR_R + D)I_R.$$

or the open-circuit generator emf E_g is then

$$E_S + I_SR_g = I_R(AR_R + B + CR_RR_g + DR_g)$$

and the maximum possible power delivered is

$$\frac{E_g^2}{4R_g} = I_R^2\frac{|AR_R + B + CR_RR_g + DR_g|^2}{4R_g}.$$

The actual power delivered is I^2R_R, and η is the ratio defined above.)

For the network considered here,

$$\eta = \frac{\left|R_R(\cos\theta_1 - Z_{01}b_R\sin\theta_1) + R_g\varphi\cos\theta_1 + jZ_{01}\sin\theta_1 + j\frac{R_gR_R}{Z_{01}}\left(Z_{01}\varphi b_R\cos\theta_1 + \frac{1-\varphi\cos^2\theta_1}{\sin\theta_1}\right)\right|}{4R_RR_g}.$$

Since, by initial choice, $R_R = \rho^2R_g$ and $Z_{01} = \rho R_g$, the formula simplifies to

$$\eta = 1 + \frac{R_R^2}{4}\left(b_R - \frac{\varphi - 1}{R_R}\cot\theta_1\right)^2.$$

An equivalent circuit for which the formula holds is shown in Fig. 2. To show the equivalence, consider a potential φE_g impressed across the load in series with $\varphi^2R_g = R_R$. The maximum possible load power will then

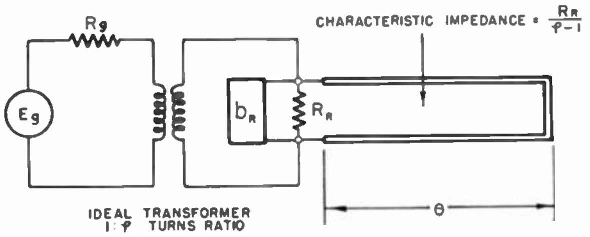

Fig. 2

be $\varphi^2E_g^2/4R_R$. If ξ_R represents the reactance of the parallel circuit composed of the transmission line and X_R, the actual load power will be

$$\frac{\varphi^2E_g^2}{R_R}\left|\frac{\dfrac{j\xi_RR_R}{j\xi_R + R_R}}{R + \dfrac{j\xi_RR_R}{j\xi_R + R_R}}\right|^2$$

or

$$\frac{\varphi^2E_g^2}{R_R}\;\frac{1}{4 + \left(\dfrac{R_R}{\xi_R}\right)^2}$$

$$\eta = \frac{\varphi^2E_g^2}{4R_R}\,\frac{R_R}{\varphi^2E_g^2}\left(4 + \frac{R_R^2}{\xi_R^2}\right) = 1 + \frac{R_R^2}{4}\,\frac{1}{\xi_R^2},$$

and, by definition,

$$\frac{1}{\xi_R^2} = \left(b_R - \frac{\varphi - 1}{R_R}\cot\theta_1\right)^2.$$

From the above discussion it may be seen that the circuit parameters which may be varied to obtain diverse frequency characteristics are the length of the line and the configuration of b_R.

IV. A Practical Application

Transformers of this type were built to match antenna cables to mixers of color television receivers operating over the range of 480 to 920 Mc. A parallel-tuned circuit operating above resonance was used to obtain the tuning susceptance. The capacitor was ganged to the local-oscillator shaft, and varied the susceptance as the frequency changed. Tracking was accomplished by sectionalizing the stator and varying the spacing of each section to the rotor to obtain the correct capacitance curve. Because the desired transformation ratio was 2:1, the structure became simply a uniform transmission line of length 2θ fed at the center. Fig. 3 is a photograph of the transformer detached from the oscillator.

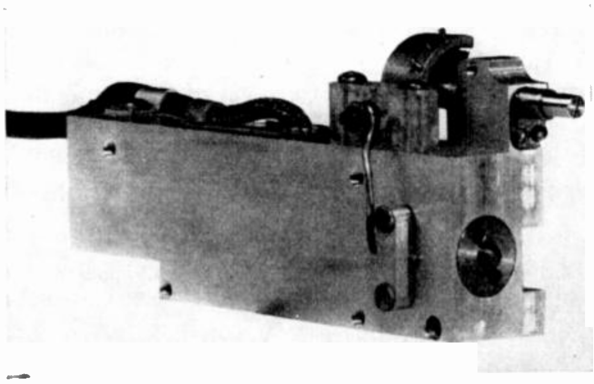

Fig. 3

Phase Difference Between the Fields of Two Vertically Spaced Antennas*

E. W. HAMLIN†, ASSOCIATE, IRE, AND A. W. STRAITON‡, MEMBER, IRE

Summary—**This paper discusses the phase differences between the fields of two vertically spaced antennas as a function of transmitter height, receiver height, and distance between transmitter and receiver. The results are applicable to microwave propagation within the line-of-sight region when a direct wave and a ground-reflected wave are present, and have been useful in interpreting phase-difference measurements made by the Electrical Engineering Research Laboratory of the University of Texas.**[1,2,5]

INTRODUCTION

IN THE PAST two years, the University of Texas, investigating the angle of arrival of radio waves under a contract with the Office of Naval Research, has made many measurements[1,2] of phase difference between signals received at two identical antennas spaced a number of wavelengths apart. One of the primary objects of the study was to determine how fast the angle of arrival might change. It was this consideration which prompted the decision to use the phase-difference method, and equipment[3] was designed and put in operation which could record changes in phase of a few degrees taking place in a second. The work has been done entirely at a wavelength of 3.2 cm.

Microwave angle-of-arrival measurements by the maximum signal strength method have been reported by Sharpless.[4]

To measure horizontal angle, the two antennas are placed at the same height above the ground, but spaced a known distance apart horizontally. For vertical-angle measurements, they are spaced vertically one above the other. In either case, if only one ray is received at the two antennas, the angle of arrival of the ray is directly proportional to the difference in phase of the signal received in the two antennas.

Even in making horizontal measurements,[2] however, the effect of the ground-reflected component is noticeable. In making vertical phase-difference measurements, the effect of the reflected wave, even with low values of reflection coefficient, must be taken into account. A reflected wave only 20 per cent as large as the direct can change the measured phase difference by ±20° from that for the direct wave alone. Since the resolution of the equipment is about ±2°, it is essential to investigate this effect mathematically. The results of the analysis presented in this paper have proved very useful in the study of field-measured phase-difference data.[1,2,5]

The present paper investigates this effect under the simplest conditions. Two identical receiving antennas are spaced vertically in a plane perpendicular to the direction of propagation. Mutual coupling between the antennas is assumed to be negligible. The earth is assumed flat and two rays only are considered, a direct wave and one reflected from the ground. The reflected wave is assumed to suffer a complete phase reversal on reflection. For the horizontal polarization which has been used, and the high frequency at which the measurements were made, this latter assumption is very nearly true. Effects similar to those to be presented here may be expected under practical field conditions.

SYMBOLS AND DEFINITIONS

h_1 and h_2 = height of transmitting and receiving antennas, respectively

h_2^0 = average height of two receiving antennas

d = horizontal distance between transmitter and receiver

d_1 and d_2 = distance from reflecting point to transmitter and receiver, respectively

λ = wavelength

L_D and L_R = length of path of direct and reflected rays, respectively

ϕ, ϕ_D and ϕ_R = phase of resultance, direct and reflected rays with reference to the transmitter

θ, θ_D and θ_R = phase difference between signal received at upper and lower horns for resultant, direct, and reflected waves, respectively

$\Delta\theta_D$ = deviation in phase difference from that of direct wave only $= \theta - \theta_D$

$\Delta\phi$ = deviation in phase from that of direct wave only $= \phi - \phi_D$

K = magnitude of reflected wave with respect to the direct

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* Decimal classification: R248.14×R115.4. Original manuscript received by the Institute, December 19, 1947; revised manuscript received, May 3, 1948.

† Cornell University, Ithaca, N. Y.

‡ University of Texas, Austin, Tex.

[1] E. W. Hamlin, et al., "Preliminary Report on Phase Front Measurements in Arizona During April, 1946," University of Texas, Electrical Engineering Research Laboratory, Report No. 6P, February, 1947.

[2] A. W. Straiton and J. R. Gerhardt, "Results of horizontal microwave angle-of-arrival measurements by phase-difference method," PROC. I.R.E., vol. 36, pp. 916–922; July, 1948.

[3] F. E. Brooks, Jr. and C. W. Tolbert, "Equipment for Measuring Angle-of-Arrival by the Phase Difference Method," University of Texas, Electrical Engineering Research Laboratory, Report No. 2, May, 1946.

[4] W. M. Sharpless, "Measurement of the angle-of-arrival of microwaves," PROC. I.R.E., vol. 34, pp. 837–845; November, 1946.

[5] E. W. Hamlin and W. E. Gordon, "Comparison of calculated and measured phase difference at 3.2 centimeters wave length," PROC. I.R.E., vol. 36, pp. 1218–1224; October, 1948.

V_C and V = field strength of resultant and direct waves, respectively

Δh_2 = spacing between receiving antennas

Δh_1 = difference between two transmitter heights

Δh_2^0 = difference between two receiver heights

α = angle of arrival

$n = \dfrac{2\, h_1\, h_2}{\lambda d}$

Δn = difference between two values of n.

Single-primed values refer to the lower receiving antenna. Double-primed values refer to the upper receiving antenna.

PHASE RELATIONS RELATIVE TO THE TRANSMITTER

The geometry is shown in Fig. 1. The angle of incidence equals the angle of reflection. The distance between the transmitter and receiver is large compared to the heights. Using images, and the usual approximations (neglecting terms higher than the square in the binomial expansion of a square root), the path lengths of the direct and reflected waves become

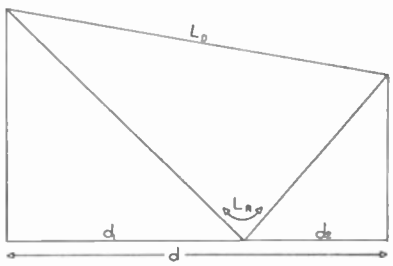

Fig. 1—Distance relationships.

$$L_D = d + \frac{(h_1 - h_2)^2}{2d} \tag{1}$$

and

$$L_R = d + \frac{(h_1 + h_2)^2}{2d}. \tag{2}$$

The path difference will be

$$L_R - L_D = 2h_1h_2/d. \tag{3}$$

The direct wave will lag the transmitter in phase by

$$\phi_D = \frac{2\pi}{\lambda}\left(d + \frac{(h_1 - h_2)^2}{2d}\right). \tag{4}$$

This approximation makes ϕ_D parabolic with respect to the transmitter height h_1, or receiver height h_2. The phase lag decreases with increasing h_1 or h_2 until $h_1 = h_2$, and then increases.

Similarly, the reflected wave lags the transmitter by

$$\phi_R = \frac{2\pi}{\lambda}\left(d + \frac{(h_1 + h_2)^2}{2d}\right) - \pi, \tag{5}$$

assuming that the phase is advanced π radians on reflection. For horizontal polarization, and any ordinary ground constants, this will be true at microwave frequencies. ϕ_R is also parabolic when plotted versus h_1 or h_2. If h_1 or h_2 are plotted vertically, the axis of the parabola is horizontal and below the reflecting surface by h_1 or h_2, whichever is constant. For positive h_1 or h_2 (transmitter and receiver above ground), ϕ_R increases with increasing h_1 or h_2.

The angle by which the reflected wave lags the direct is

$$\phi_R - \phi_D = \frac{4\pi h_1 h_2}{\lambda d} - \pi = (2n - 1)\pi. \tag{6}$$

Thus, with transmitter or receiver at the level of the reflecting plane, the direct and reflected waves are 180° out of phase.

The angle by which the combined signal lags the transmitter depends upon the relative magnitude of the direct and reflected components. If the field strength of the direct wave is V, and that of the reflected wave KV, the vector field strength of the combined wave will be

$$V_C\underline{/\phi} = V\underline{/\phi_D} + KV\underline{/\phi_R}. \tag{7}$$

The space orientation of the two components is the same for horizontal polarization. Equation (7) may be written

$$V_C\underline{/\phi_D + \Delta\phi} = V\underline{/\phi_D}(1 + K\underline{/(2n - 1)\pi})$$

or

$$V_C\underline{/\Delta\phi} = V(1 - K\underline{/2n\pi}) \tag{8}$$

where $\Delta\phi$ is the deviation in phase from that due to the direct wave alone.

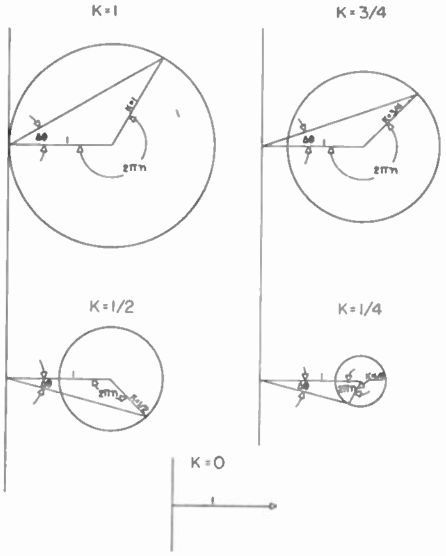

Fig. 2—Loci of $V/\Delta\phi$.

As n varies, the locus of $V_C\underline{/\Delta\phi}$ will be a circle with its center at $(V, 0)$ and radius equal to KV. For $n = 0$, the vector KV falls horizontally to the left. As n increases, the rotating vector KV turns counterclockwise, making a complete revolution as n goes from 0 to 1. The loci of V_C for several values of K are shown in Fig. 2.

From (8),

$$\Delta\phi = \tan^{-1}\left[\frac{-K\sin 2\pi n}{1-K\cos 2\pi n}\right] \tag{9}$$

and the combined wave lags the transmitter by $\phi+\Delta\phi$. The relation (9) is shown graphically in Fig. 3 for several values of K, with ϕ plotted as a function of h_1 or h_2.

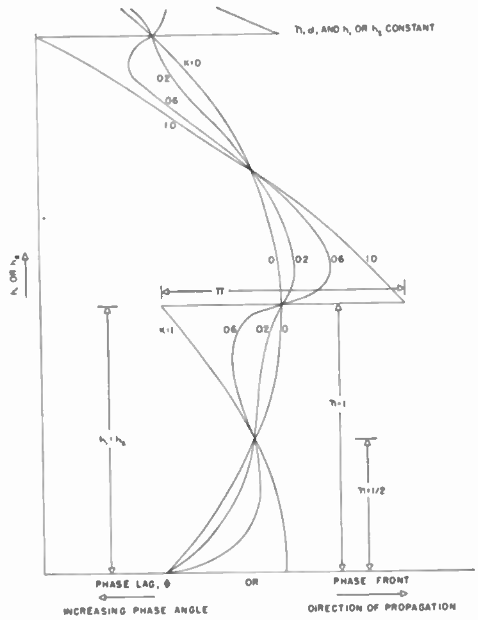

Fig. 3—Phase fronts.

The curvature of the parabola representing the relationship for $K=1$ is greatly exaggerated for large distances, but is used for the sake of illustration. The graph of Fig. 3 also represents the phase front.

$\Delta\phi$ as a function of n is shown in Fig. 4 for several values of K. Maximum and minimum values of $\Delta\phi$ occur when $\cos 2\pi n = K$, and are $\pm\sin^{-1} K$.

PHASE DIFFERENCE BETWEEN TWO ANTENNAS AT DIFFERENT HEIGHTS

The phase difference between two identical receiving antennas, one vertically above the other in a plane perpendicular to the direction of propagation, will be the difference between the phases of the signals at each antenna relative to the transmitter. Each will be of the form $\phi_D+\Delta\phi$, from (4) and (9). Let the height of the upper antenna be h_2'', and of the lower, h_2'; then $h_2''=h_2'+\Delta h_2$ where Δh_2 is the vertical spacing.

For a direct wave alone, the angle by which the lower antenna lags the upper in phase is $\theta_D=\phi_D'-\phi_D''$ where the single prime refers to the lower antenna, and the double prime to the upper. θ_D may be written

$$\theta_D = \frac{2\pi}{\lambda d}(h_1-h_2^0)\Delta h_2 \tag{10}$$

where h_2^0 is the average height of the receiving antennas.

A graph of θ_D versus transmitter height will be a straight line with a slope $2\pi\Delta h_2/\lambda d$. The phase lag of the lower antenna continually increases as the transmitter is raised. If the receiving antennas are raised together, keeping their spacing constant, the slope of the line is negative but of the same magnitude.

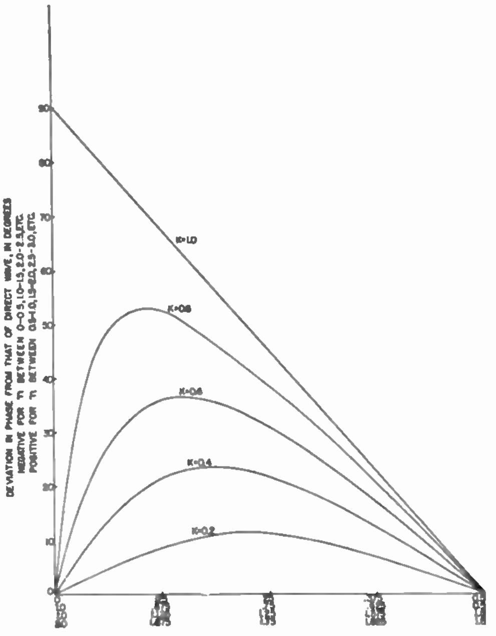

Fig. 4—$\Delta\phi$ as a function of 1.

The angle of arrival of the direct wave alone, at the average height of the receiving antennas, is obtained from (10). This angle is

$$\alpha = \tan^{-1}\frac{h_1-h_2^0}{d} \cong \frac{h_1-h_2^0}{d} \cong \frac{\theta_D\lambda}{2\pi\Delta h_2}$$

$$\text{when } d \gg h_1 - h_2^0. \tag{11}$$

Positive α indicates angles of arrival above horizontal. The resolution of the equipment is determined by the smallest θ_D which may be measured, and the receiving antenna spacing in wavelengths. At 3.2-cm wavelength, with antennas spaced 10 feet apart, the phase angle θ_D is 600 times the space angle of arrival. Since θ_D is ambiguous in intervals of 360°, the space angle is ambiguous in intervals of 0.6°.

For a wave containing both direct and reflected components, the phase lag of the lower antenna behind the upper is $\theta=\theta_D+\Delta\theta$ where $\Delta\theta=\Delta\phi'-\Delta\phi''$. This is

$$\theta = \frac{2\pi}{\lambda d}(h_1-h_2^0)\Delta h_2 + \tan^{-1}\left[\frac{-K\sin 2\pi n'}{1-K\cos 2\pi n'}\right] - \tan^{-1}\left[\frac{-K\sin 2\pi n''}{1-K\cos 2\pi n''}\right] \tag{12}$$

where the primes refer to the respective antennas, as above.

VARIATION OF PHASE DIFFERENCE OF THE RESULTANT WAVE WITH TRANSMITTER HEIGHT

As the transmitter height is increased, the first term in (12), θ_D, increases linearly, and the second two terms which make up $\Delta\theta$ will be alternately plus and minus (Fig. 5). The magnitude of these variations starts at zero with the transmitter at the reflecting surface ($h_1 = 0$), increasing to a maximum and returning to zero when

$$n'' = n' + 1.$$

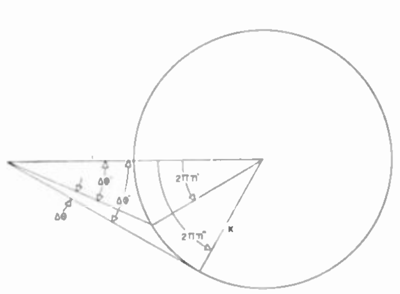

Fig. 5.—Deviation in phase from that of the direct wave determined from loci diagram.

Starting from any value of h_1, a complete set of values of $\Delta\theta$ will be run through when the change in n'', $\Delta n''$ exceeds the change in n', $\Delta n'$ by one, or when h_1 changes by

$$\Delta h_1 = \frac{\lambda d}{2\Delta h_2}. \tag{13}$$

This depends only upon the spacing between receiving antennas, the wavelength, and the distance, and is independent of the height of either the transmitter or the receiver. A plot of θ_D versus h_1 will show a straight line with height, about which θ varies, the variations repeating at intervals Δh_1.

The variation $\Delta\theta$ in phase from that of the direct wave alone will be zero at transmitter heights at which

$$\frac{\sin 2\pi n'}{1 - K\cos 2\pi n'} = \frac{\sin 2\pi n''}{1 - K\cos 2\pi n''}, \tag{14}$$

from which it follows that

$$\sin\frac{2\pi h_1\Delta h_2}{\lambda d}\left(\cos\frac{4\pi h_1 h_2^0}{\lambda d} - K\cos\frac{2\pi h_1\Delta h_2}{\lambda d}\right) = 0. \tag{15}$$

This equation is satisfied if either factor is zero.

If the first factor is zero,

$$n'' - n' = N \tag{16}$$

where $N = 1, 2, 3$, etc., or

$$\frac{2h_1\Delta h_2}{\lambda d} = N, \tag{17}$$

corresponding to (13) above with the first zero at ground level.

The second factor taken equal to zero results in the condition

$$\tan\pi n' \tan\pi n'' = \frac{1-K}{1+K}. \tag{18}$$

Curves showing the variations of θ with h_1 for typical cases are shown in Figs. 6 and 7 for different values of K. In each case only one interval of height corresponding to (17) is shown. The other zeros of $\Delta\theta$ are due to (18).

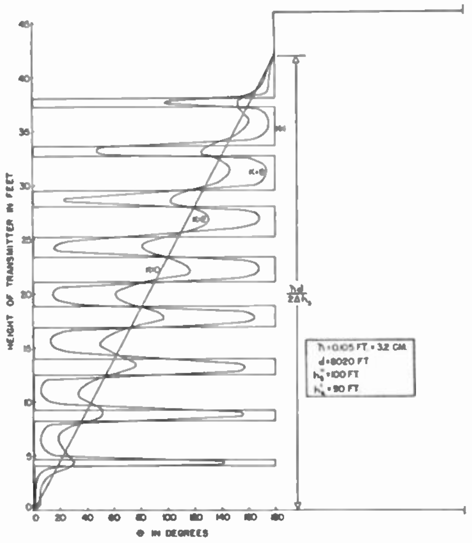

Fig. 6—Variation of phase difference with transmitter height, example No. 1.

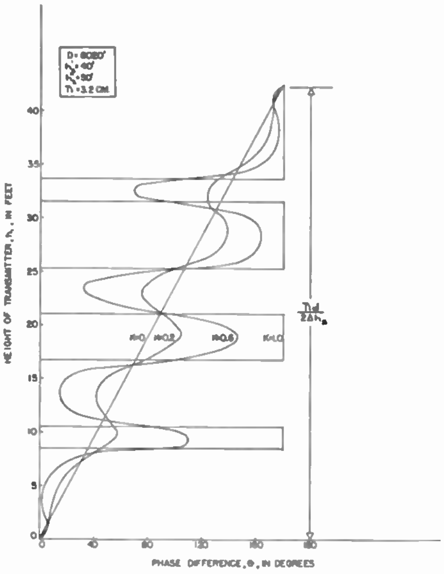

Fig. 7—Variation of phase difference with transmitter height, example No. 2.

VARIATION OF PHASE DIFFERENCE OF THE RESULTANT WAVE WITH RECEIVER HEIGHT

If the receiver height h_2^0, is increased, the general slope of the curves of θ versus h_2^0 will be opposite to that occurring when the transmitter is raised. Also, n'' and n' increase with h_2^0 at exactly the same rate. Hence, phase-difference versus receiver-height curves repeat at intervals

$$\Delta h_2^0 = \frac{\lambda d}{2h_1}, \tag{19}$$

which do not depend upon receiver height or spacing of receiving antennas, but only upon wavelength, distance, and transmitter height.

The zeros of $\Delta\theta$ are still given by (14). The first factor of (15) does not vary with h_2^0, and for an abritrary transmitter height and antenna spacing, it will not be zero. Also, the second term of the second factor does not depend upon h_2^0, so that equating this factor to zero gives only two values of h_2^0 at which $\Delta\theta = 0$ in each repeating interval. An example of the type of curve to be expected is shown in Fig. 8, in which the constants have been chosen so as to make $\Delta\theta$ symmetrical about the θ_D line. h_2^0 must, of course, be greater than half Δh_2.

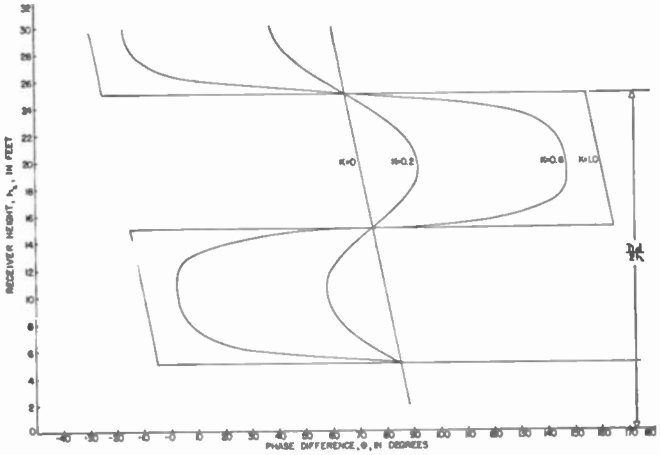

Fig. 8—Variation of phase difference with receiving height.

If transmitter height, or antenna spacing, or both, is chosen so as to make the first factor of (15) zero, the phase difference is always that of the direct wave alone. Thus, when

$$h_1 = \frac{N\lambda d}{2\Delta h_2} \qquad N = 1, 2, 3, \text{etc.}, \tag{20}$$

the last two terms in (12) cancel for all receiver heights. The plot of θ versus h_2^0 is just a straight line. On the other hand, if h_1 is chosen at midway between the values indicated by (20), large variations from θ_D occur, as shown in Fig. 8.

SPECIAL CONDITIONS WHEN $K=1$

For the special case, $K=1$, $\Delta\phi$ in (9) becomes

$$\Delta\phi = \tan^{-1}\frac{-\sin 2\pi n}{1-\cos 2\pi n} \tag{9}$$

$$\Delta\phi = \pi n - \frac{\pi}{2} \quad \text{for} \quad 0 < n < 1$$

$$\Delta\phi = \pi n - \frac{3\pi}{2} \quad \text{for} \quad 1 < n < 2$$

and, in general,

$$\Delta\phi = \pi n - \frac{(2N-1)}{2}\pi \quad \text{for} \quad (N-1) < n < N. \tag{21}$$

Then the deviation in phase difference from that of the direct wave alone $\Delta\theta$ is given by

$$\begin{aligned} \Delta\theta &= -\frac{2\pi h_1 \Delta h_2}{\lambda d} && \text{for} \quad 0 < n'' < 1 \\ &= -\frac{2\pi h_1 \Delta h_2}{\lambda d} + \pi && \text{for} \quad 1 < n'' < 2 \\ & && \text{and} \quad 0 < n' < 1 \\ &= -\frac{2\pi h_1 \Delta h_2}{\lambda d} && \text{for} \quad 1 < n'' < 2 \\ & && \text{and} \quad 1 < n' < 2. \end{aligned} \tag{22}$$

But θ_D is given by (10) as

$$\theta_D = \frac{2\pi \Delta h_2}{\lambda d}[h_1 - h_2^0].$$

Adding (10) and (22), it is found that

$$\begin{aligned} \theta &= -\frac{2\pi h_2^0 \Delta h_2}{\lambda d} && \text{for} \quad 0 < n'' < 1 \\ &= -\frac{2\pi h_2^0 \Delta h_2}{\lambda d} && \text{for} \quad 1 < n'' < 2 \\ & && \text{and} \quad 0 < n' < 1 \\ &= -\frac{2\pi h_2^0 \Delta h_2}{\lambda d} && \text{for} \quad 1 < n'' < 2 \\ & && \text{and} \quad 1 < n' < 2. \end{aligned} \tag{23}$$

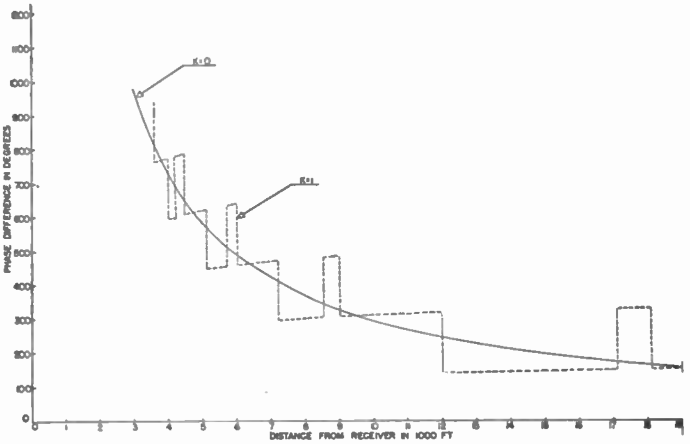

Fig. 9—Variation of phase difference with distance.

This shows θ to be equal to $-2\pi h_2^0\Delta h_2/\lambda d$, or this value increased by π in the range $0>n''>2$. If h_1 is the variable, these values are constants. As h_1 is increased from zero, θ will jump from one of these values to the other when either n'' or n' passes through an integer value, as shown by the $K=1$ lines in Figs. 6 and 7. This continues until n'' exceeds n' by one, corresponding to the condition of (17). From this point, the alternate values of θ will be $-2\pi h_2^0\Delta h_2/\lambda d+\pi$, and $-2\pi h_2^0\Delta h_2/\lambda d+2\pi$ until n'' exceeds n' by 2.

If h_2^c is the variable, $-2\pi h_2^0\Delta h_2/\lambda d$ is not a constant, but a line with the same slope as θ_D. Also n'' is larger than n' at all times by a constant value $2h_1\Delta h_2/\lambda d$. Again, each time n'' or n' passes through an integer value a jump of π occurs, but because of the constant difference the jumps are spaced at regularly repeating intervals. This is shown by the $K=1$ lines in Fig. 8. It should be pointed out that variations about the line θ_D are not necessarily symmetrical, although Fig. 8 is drawn for such a case.

VARIATION WITH DISTANCE AND WAVELENGTH

If the distance or wavelength is variable, the variations are similar to those for variable transmitter height, when plotted against inverse distance or frequency. A typical example of phase-difference variations plotted directly against distance is shown in Fig. 9.

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Andrew Bark (M'47) was born in Seattle, Wash., on December 2, 1919. He was graduated from the University of Detroit in 1942, with the B.E.E. degree, after working for three years at the Detroit Edison Company Research Laboratories under a cooperative system with the University of Detroit. After graduation, he went to the Radio Research Laboratory, at Harvard University, as a research associate. He was a special representative of the Office of Scientific Research and Development in England.

Mr. Bark joined the Columbia Broadcasting System, Inc., in New York, N. Y., in 1944. His work was initially concerned with radar countermeasures; at present he is engaged in research on color television and an instantaneous audience-measurement system.

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CARL BERKLEY

Carl Berkley (A'45) was born in New York, N. Y., on February 19, 1917. He attended the College of the City of New York. He was a motion-picture cameraman at the Browning Studios from 1936 to 1938, a teacher of photography in the New York public schools from 1938 to 1939, and a technician engaged in biological research and exhibit work in the Laboratory of Experimental Biology at the American Museum of Natural History from 1939 to 1941. In 1941, he was employed at the Agfa Ansco Company at Binghamton, N. Y., in development work on multilayer color films.

Since 1942 Mr. Berkley has been engaged at the Allen B. DuMont Laboratories in engineering and application work on cathode-ray tubes and instruments. In 1947, he assumed charge of the applications section of the Instrument Division of that Company.

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L. B. HALLMAN, JR.

Ludlow B. Hallman, Jr., (J'27–A'29–SM'44) was born in Sneads, Fla., on July 21, 1907. He studied electrical engineering at the Alabama Polytechnic Institute in Auburn, Ala., from 1925 to 1929. During his undergraduate career he served part time as an instructor in radio at the Institute, and as an engineer for broadcast station WAPI. He received the B.S. degree in electrical engineering from the Alabama Polytechnic Institute in 1929, and the E.E. degree in 1934.

Shortly after graduation, Mr. Hallman became associated with the Montgomery Broadcasting Co., at Montgomery, Ala., as chief engineer of station WSFA. He continued to work in the field of broadcast station engineering until 1936, when he accepted a position in the Aircraft Radio Laboratories at Wright Field, Ohio. Since then he has been associated with various projects involving the development, procurement, and testing of special electronic equipment for the U. S. Air Forces. At present, he is assistant chief of the Communication and Navigation Laboratory, Electronic Subdivision, at Wright Field.

Mr. Hallman served as Chairman of the Dayton IRE Section during 1945, and is now Chairman of the Headquarters Relations Committee of this Section. He holds membership in the honorary fraternities of Eta Kappa Nu, Tau Beta Pi, and Phi Kappa Phi.

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For a photograph and obituary of the late E. W. HAMLIN, see page 887 of the July, 1948, issue of the PROCEEDINGS OF THE I.R.E.

WILLIAM E. SHOUPP

William E. Shoupp (SM'45) was born in Troy, Ohio. He was graduated from Miami University in Oxford, Ohio, in 1931 with the degree of bachelor of science in physics, and for the next six years served as graduate assistant and instructor in physics in the University of Illinois. Here he was awarded the master of arts degree in 1933 and the doctor of philosophy in physics in 1937. In 1938 he joined the Westinghouse Research Laboratories as a Westinghouse Research Fellow, and three years later was named a research engineer. He was appointed to his present position as manager of the electronics department at the Laboratories in 1943, a responsibility which later was broadened to include all nuclear physics work being carried on in the Westinghouse Research Laboratories.

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Dr. Shoupp is a Fellow of the American Physical Society, the Pittsburgh Physical Society, and a member of the American Institute of Electrical Engineers. He is a member of Sigma Xi and Phi Beta Kappa.

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For a photograph and biography of A. W. STRAITON, see page 931 of the July, 1948, issue of the PROCEEDINGS OF THE I.R.E.

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534.21 2997

Directional Characteristic of a Cylindrical Radiator with a Coaxial Conical Reflector—G. Bacchi. (*Alta Frequenza*, vol. 17, pp. 74–78; April, 1948. In Italian with English, French, and German summaries.) Magnetostriction oscillators of the above type, adapted for radial vibration, have radiation patterns with an absolute maximum along the axis. The wavelength is of the same order as the dimensions of the cylinder. The radiation pattern is calculated. With a 45° cone, the characteristic is represented by the integral of Bessel's function of zero order.

534.242 2998

Formula Involving the Shape for Calculating the Resonance Frequency of Helmholtz Resonators—D. Chervet and J. Henry (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1891–1893; June 7, 1948.) Rayleigh's formula was established on the assumptions that the pressure inside the resonator is constant and that the resonator dimensions are small compared with the resonance wavelength. Measurements show that the pressure varies from point to point according to an exponential law, the maximum pressure being at the bottom of the resonator. From this a correction formula is derived which is not easy to use and can be replaced by an empirical formula involving simply the product of the resonator length L and the resonance frequency f_0 given by Rayleigh's formula. Experimental results for resonators of various shapes are tabulated, together with the frequencies calculated from Rayleigh's formula and the two correction formulas.

534.321.9 2999

Distortion of Progressive Ultrasonic Waves—J. C. Hubbard, J. A. Fitzpatrick, B. T. Kankovsky, and W. J. Thaler. (*Phys. Rev.*, vol. 74, pp. 107–108; July 1, 1948.) Spark shadow photographs of ultrasonic waves in air, water, glycerine, and CCl_4 show definite changes of waveform into a type with steeper fronts and increased harmonic content.

534.321.9:621.391.63 3000

Distortion in Light Modulation by an Ultrasonic Cell—D. Sette. (*Alta Frequenza*, vol. 17, pp. 51–68; April, 1948. In Italian with English, French, and German summaries.) Description of an 8-Mc cell similar to that of Giacomini (1847 of August). Frequency distortion in a modulated light beam transmitted through the cell is small even for comparatively high modulation frequencies. The low distortion and the high modulation factor possible are particular advantages of such a system.

53.321.9.001.8:614.8 3001

Notes on Using High-Power Ultrasonics—(*See* 3195.)

534.75 3002

Audibility of High-Frequency Sounds—V. Gavreau. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 2053–2054; June 21, 1948.) It is found that sounds of frequencies from 17.5 kc to 26 kc are perfectly audible when sufficiently intense. When the intensity of such sounds is steadily increased, one or more sudden apparent changes of pitch, of about an octivae, are observed. These apparent changes of pitch may be due to abrupt changes in the order of the harmonic response of the resonators of the ear. Magnetostriction oscillators were used and gave sound powers up to 5 w.

534.756.1 3003

Phase Memory of the Ear: A Proof of the Resonance Hypothesis—R. J. Pumphrey and T. Gold. (*Nature* (London), vol. 161, p. 640; April 24, 1948.) Two alternative trains of sine-wave pulses are used. Each pulse contains n cycles, the pulse separation being m cycles, where m and n are both integers. In one train, the phase of all pulses is the same, while in the other the phase of alternate pulses is reversed. The ear can differentiate between the two trains. It follows that the value of Q for the resonators of the ear must be at least 100 and is probably considerably higher. See also 8 of February.

534.83 3004

Acoustic Materials—(*Electronics*, Buyers' Guide Issue, vol. 21, pp. M2–M3; June, 1948.) Noise reduction coefficients of various sound-absorbing materials are tabulated and compared with those of ordinary building materials. Mountings used in sound absorption tests are discussed briefly.

534.839 3005

Methods for the Study of Noise—A. Moles. (*Radio Franç.*, pp. 9–15; June, 1948.) Subjective and objective methods are discussed, the principal characteristics of various types of microphone are considered and a description of the General Radio decibel meter is given. Harmonic analyzers are also considered briefly.

53.851:621.395.813 3006

Measuring Wow—U. R. Furst. (*FM and Telev.*, vol. 8, pp. 30, 50; May, 1948.) Discussion of methods of measurement, and description of a commercial type of meter.

534.861.1:621.316.345:621.396.664 3007

Modern Design Features of CBS Studio Audio Facilities—Monroe and Palmquist. (*See* 3241.)

534.861.1:621.396.712.3 3008

Speech-Input Equipment for New Oslo Broadcasting House—E. Julsrud and G. Wieder. (*Elec. Commun.*, vol. 25, pp. 21–29; March, 1948.) An illustrated description, giving details of the facilities available. Easy, flexible, and reliable operation is ensured by extensive use of automatic switching.

621.395.623.7.015.3 3009

Study of Loudspeakers in the Transient Regime—G. Guyot. (*Rev. Gén. Élec.*, vol. 57, pp. 245–253; June, 1948.) Discussion of the response of loudspeakers to steady tones, with methods of obtaining the corresponding curves, and a detailed account of test methods and results for square-shape sine-wave signals and for pulse signals. Preliminary results indicate a correlation between the quality appreciated by the ear and the duration of the transient signal. For a loudspeaker 21 cm in diameter, signal quality was found to fall off if the duration of the pulse signal was below about 2.7 ms. Discussion of the various methods of testing loudspeakers shows that transient tests cannot replace ordinary methods completely, but that they furnish valuable information as to loudspeaker quality.

621.395.625.3 3010

Two-Channel Two-Way-Drive Magnetic Tape Recorder—R. E. Zenner and R. B. Vaile, Jr. (*Audio Eng.*, vol. 32, pp. 11–15; April, 1948.) Designed for good performance at low tape speed.

621.395.625.3 3011

Test Characteristics of Recording Wire—Carter and Koontz. (*See* 3149.)

621.395.625.6:621.317.75.015.3 3012

Sweeping Device for the Display of Transient Phenomena and Nonlinear Distortion—Meyer-Eppler. (*See* 3183.)

ANTENNAS AND TRANSMISSION LINES

621.315 3013

Transmission Line Theory Simplified—W. Redmayne. (*Distrib. Elec.*, vol. 20, pp. 346–348, 350; April, 1948.) The general equations for voltage and current at any point of a dc transmission system are derived without using differential equations. Both loaded and unloaded lines are considered. The equations can be easily adapted to the case of ac transmission.

621.315.2.017.71 3014

Heating of Radio-Frequency Cables—W. W. Macalpine. (*Elec. Commun.*, vol. 25, pp. 84–99; March, 1948.) Discussion with particular reference to solid-dielectric coaxial cables; only the simplest physical or mathematical assumptions are made.

621.315.23:620.197.6 3015

Rubber Thermoplastic Jacket for Buried Cable—C. V. Lundberg. (*Bell Lab. Rec.*, vol. 26, pp. 148–151; April, 1948.) Protection of the lead casing of cables against corrosion is secured by two layers of a material made from reclaimed rubber, clay, resin, paraffin wax, and mineral rubber. A fabric tape, corrugated Cu sheath and bitumen layer are applied outside the rubber.

621.392.029.64 3016

Attenuation of the H_{10} Wave in a Rectangular Waveguide—N. N. Malov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 417–420; April, 1948. In Russian.) Discussion of a method based on an examination of the multiple reflection of the transverse wave from the walls of the waveguide. It is shown that the H_{m0} and H_{0n} types of wave cannot be regarded as particular cases of the H_{mn} type with m or n zero.

621.392.029.64 3017

The Excitation of a Rectangular Waveguide through a Slot—I. I. Vol'man. (*Radiotekhnika* (Moscow), vol. 3, pp. 49–55; May and June, 1948. In Russian.) Methods are indicated for calculating the field produced between two parallel planes by energy supplied through an aperture. The results obtained can be applied in the theory of rectangular waveguides.

621.392.029.64 3018

Effect of Junctions on the Field in a Waveguide—H. Buchholz. (*Arch. Elek.* (Übertragung), vol. 2, pp. 14–22; January, 1948.) Formulas are derived for the effect of a narrow circumferential slot in the wall of a waveguide on the transmission of waves having an axial component of wall current.

621.392.029.64+621.396.611.4]:621.3.015.3 3019

Study of the Transient Regimes in Waveguides and Cavity Resonators—T. Kahan and S. Colombo. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 2060–2061; June 21, 1948.) A method is given which enables formulas applicable to the transient state to be derived from formulas established for the steady state with frequency $\omega/2\pi$.

621.392.029.64:621.317.3 3020

On the Representation and Measurement of Waveguide Discontinuities—N. Marcuvitz. (Proc. I.R.E., vol. 36, pp. 728–735; June, 1948.) The various equivalent circuit representations of a general 2N-terminal waveguide structure, obtained by selection of different terminal planes, are discussed and interrelated. A precision method of measuring the circuit parameters of such structures is described. Weissfloch's tangent relation for the input versus output behavior of a four-terminal waveguide structure is used.

621.392.029.64:621.396.67 3021

The Theory of the Ring Resonant Slot in a Waveguide—M. L. Levin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 639–652; May, 1948. In Russian.) A theoretical analysis is given of a resonant slot antenna having the form of a narrow ring cut in the wall of a semi-infinite waveguide of circular cross section.

621.392.43:621.317.72 3022

A Modified Micromatch—Corfield and Cragg. (*See* 3176.)

621.396.67 3023

Relations between the Transmitting and Receiving Properties of Antennas—A. F. Stevenson. (*Quart. Appl. Math.*, vol. 5, pp. 369–384; January, 1948.) A rigorous mathematical discussion. The parasitic and receiving properties of a perfectly conducting antenna can be derived from its transmitting properties under stated conditions. The particular case of a linear antenna in a homogeneous isotropic medium is considered in detail. A rigorous proof of Thévenin's theorem for antennas is given without any appeal to circuit theory. The polar diagram of an antenna used for transmission is not in general identical with that of the same antenna used for reception, though these polar diagrams are approximately identical in practical cases. Reciprocal relations between two antennas are discussed. The case of an imperfectly conducting antenna is also considered briefly.

621.396.67 3024

Calculation of the Field Strength Produced by a Half-Wave Aerial at a Given Point above a Plane Earth as a Function of the Energy Supplied per Second: Part 1—K. F. Niessen. (*Physica, 's Grav.*, vol. 7, pp. 586–602; July, 1940. In German.) Formulas are derived for the vertical field due to a $\lambda/2$ antenna with its foot at a given distance above a plane earth, taking account of the energy absorbed in the earth. Numerical results are given for various assumed values of earth conductivity and dielectric constant. See also 3025 below.

621.396.67 3025

Electric Field Strength as a Function of the Energy Supplied to the Aerial: Part 2—K. F. Niessen. (*Physical's Grav.*, vol. 7, pp. 897–908; December, 1940. In German.) Calculations are made for the same wavelength and geometrical ratios as those of part 1 (3024 above), but for values of the earth conductivity ρ and dielectric constant ϵ corresponding to dry ground. The results are compared with those of part 1 and discussed. Tables give the vertical field, due to a vertical $\lambda/2$ antenna with its foot at a height of $\lambda/4$ above a plane earth, at a distance of 5λ and a height of $\lambda/4$, for various values of ϵ and ρ, and also the corresponding fields due to a dipole. For high values of ρ the corrected reflection formula should be used, particularly when ϵ is small; for low values of ρ, Sommerfeld's formula is preferable.

621.396.67 3026

The Cylindrical Antenna with Gap—R. King and T. W. Winternitz. (*Quart. Appl. Math.*, vol. 5, pp. 403–416; January, 1948.) The King-Middleton theory (1453 and 3547 of 1946) is generalized to show the effect of a finite gap on the current and the impedance of a cylindrical antenna with a gap of length 2δ between the halves of the antenna. For small gaps, the impedance is not very sensitive to gap length, so that impedances calculated for zero gap are good approximations for antennas for which $2\pi a\delta/\lambda_0 < 0.01$, a being the radius of the conductors. For gap lengths not satisfying this inequality, correction curves are given for use with curves for resistance and reactance.

621.396.67 3027

The Field of a Dipole with a Tuned Parasite at Constant Power—R. King. (Proc. I.R.E., vol. 36, pp. 872–876; vol. 36, July, 1948.) Theoretical curves are given for the electric field, in the forward and backward directions, of a center-driven $\lambda/2$ dipole with a parallel center-tuned reflector of the same dimensions, for various values of antenna spacing and reflector impedance.

621.396.67:517.392 3028

Concerning a New Transcendent, Its Tabulation and Application in Antenna Theory—Bouwkamp. (*See* 3158.)

621.396.67:517.512.2 3029

Fourier Transforms in Aerial Theory: Part 6—Ramsay. (*See* 3159.)

621.396.671 3030

The Radiation Resistance of End-Fire and Collinear Arrays—C.H. Papas and R. King. (Proc. I.R.E., vol. 36, pp. 736–741; June, 1948.) Formulas obtained for this resistance involve circular functions and can conveniently be used for computation. The only mathematical approximation is a Fourier representation of the field of a single $\lambda/2$ dipole, discussed in 3326 of 1941 (King). The new formulas are in satisfactory agreement with the results of Pistolkors and Bontsch-Bruewitsch.

621.396.677 3031

Note on Practical Limitations in the Directivity of Antennas—R. M. Wilmotte. (Proc. I.R.E., vol. 36, p. 878; July, 1948.) Discussion on 2731 of November (Riblet).

621.396.677 3032

Artificial Dielectric Lenses for Microwaves—W. E. Kock. (*Bell. Lab. Rec.*, vol. 26, pp. 145–147; April, 1948.) For a fuller account see 2176 of September.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:[621.392.029.64+621.396.611.4 3033

Study of the Transient Regimes in Waveguides and Cavity Resonators—Kahan and Colombo. (*See* 3019.)

621.314.2:[621.395.623.7 3034

Study of the Output Transformer for Feeding Loudspeakers—L. Chrétien. (*TSF Pour Tous*, vol. 24, pp. 185–188; July and August, 1948.) Discussion of the conditions which should be satisfied for optimum performance, with faithful reproduction of both high and low frequencies. Practical design details are considered briefly.

621.314.2:621.396.813 3035

Nonlinear Distortion in Low-Frequency Transformers with Strong Premagnetization—F. Böttcher. (*Frequenz*, vol. 2, pp. 140–143; May, 1948.) The distortion occurring in the load resistance is calculated under certain simplifying assumptions. The effective magnetization curve showing the relation between induction and field strength, taking account of the airgap, is thereby replaced approximately by a parabola. Values are thus derived for the harmonic and combination tones of the quadratic distortion in the secondary load, when two sine-wave tones are applied to the transformer primary.

621.316.86 3036

Thermistors and Their Applications—B. S. Sotskov. (*Avtomatika i Telemekhanika*, vol. 9, pp. 39–58; January and February, 1948. In Russian.) The properties of various types are tabulated, and discussed in detail under the following headings: (a) main characteristics, (b) effect of size and shape on parameters, (c) determination of steady-state and transient regimes in thermistor circuits, and (d) applications to (i) time delay relays, (ii) starting resistances for motors, (iii) shunting resistances for lamps and other current-carrying devices connected in series, (iv) voltage stabilizers, (v) temperature measurements, (vi) gas and liquid flow measurements, (vii) bolometers and (viii) rheostats. See also 765 or 3552 of 1947 (Becker, Green, and Pearson), and 3044 of 1947 (Rosenberg).

621.318.572:621.385.38 3037

A Fast, Noiseless Thyratron Switch—S. W. Kitchen. (*Rev. Sci. Instr.*, vol. 19, pp. 370–371; May, 1948.) The cathode resistor of a conventional thyratron switch is replaced by a capacitor and a resistor of only $100\ \Omega$ is used in the anode circuit. With this arrangement, if the desired current drain through the switch is small compared with the rated average thyratron current, inherent noise is eliminated while fast action is retained.

621.392:518.4 3038

Design Curves for Parallel-T Network—D. Espy. (*Electronics*, vol. 21, pp. 114–115; July, 1948.) Generalized curves from which transmissions at various deviations from resonant frequency can be read directly. Design equations are also given.

621.392:621.396.645.029.64 3039

Very-High-Frequency Triode Oscillator and Amplifier Circuits—G. Lehmann. (*Elec. Commun.*, vol. 25, pp. 50–61; March, 1948.) Translation of paper abstracted in 368 of 1947.

621.392.011.2 3040

A Contribution to the Approximation Problem [for impedance functions]—R. F. Baum. (Proc. I.R.E., vol. 36, pp. 863–869; July, 1948.) An approximation to the curve of a given impedance function can be obtained by combining a finite number of "semi-infinite slopes," represented by Butterworth functions. The labor normally involved in the calculation of impedance zeros and poles is thus greatly reduced. Tschebyscheff functions are more appropriate for obtaining approximations to filter curves. Approximations to resistance, reactance, and phase curves can be obtained similarly.

621.392.088.7 3041

The Theory and Design of Thermal Compensation of a Circuit for a Given Frequency Range—S. S. Arshinov. (*Radiotekhnika* (Moscow), vol. 3, pp. 21–39; March and April, 1948. In Russian.) The problem of reducing as much as possible the maximum value of the temperature coefficient for a given frequency range is a particular case of finding the best approximation to a continuous function. Conditions for optimum thermal compensation in simple and complex circuits are here derived from a formula due to Tchebyschev. Complete compensation is possible for the whole operating range. A numerical example is given.

621.392.43:621.314.2 3042

Impedance Matching Half-Wave Transformer—H. E. Dinger and H. G. Paine. (*Tele-Tech.*, vol. 7, part 1, pp. 41–43, 77; May, 1948.) A transformer for matching a balanced load to an unbalanced source of the same impedance, using a $\lambda/2$ coaxial line. Since such lines can be used at 20 per cent off resonance with an impedance change of only 5 per cent, continuous coverage for the frequency range 100 to 400 Mc is achieved with 4 interchangeable lines.

621.392.5:621.385.3:512.831 3043

The Application of Matrices to Vacuum-Tube Circuits—J. S. Brown and F. D. Bennett. (Proc. I.R.E., vol. 36, pp. 844–852; July, 1948.) The matrix equations for triode circuits are derived for linear operation with any one electrode grounded. A table relating matrix elements is calculated, allowing for the fact that the networks are not bilateral. Formulas are obtained for the gain of an amplifier having m identical stages. Two examples showing the advantages of the matrix method are included.

621.396.611:621.316.729 3044

Pseudosynchronization in Amplitude-Stabilized Oscillators—P. R. Aigrain and E. M. Williams. (Proc. I.R.E., vol. 36, pp. 800–801; June, 1948.) If a frequency f is injected into an amplitude-stabilized oscillator of frequency f_0, then over a band of frequencies for which $|f-f_0|$ is small, f_0 disappears and the system acts as a regenerative amplifier for the injected signal and "pulling" does not exist. These results are contrasted with those of Adler (2522 of 1946).

621.396.611.015.4 3045

Notes on a Property of the Voltage Resonance of an Oscillatory Circuit—E. Fromy. (*Onde Élec.*, vol. 28, pp. 218–221; June, 1948.) Starting from the elementary equations of an oscillatory circuit fed by a sinusoidal source, it is shown that the terminal voltage can be regarded as the sum of two voltages, whose amplitudes are constant and equal to half the terminal voltage of the circuit at resonance; one voltage is of constant phase, while the phase of the other varies on either side of the phase at resonance, according to the value of the tuning capacitance. This property is quite general and is established without making any simplifying assumptions; it can have numerous practical applications, particularly in the design of dephasing circuits.

621.396.611.1 3046

The Application of the Symbolic Method to the Analysis of the Free Régime of Linear Systems—A. A. Rizkin. (*Radiotekhnika* (Moscow), vol. 3, pp. 56–63; May and June, 1948. In Russian.) A general analytical method is proposed which is based on breaking the circuit under consideration at some point and determining the input impedance at this point. By equating this "characteristic" impedance to zero or infinity (depending on the properties of the circuit) a characteristic equation of the circuit is obtained. Five examples of the application of this method are given.

621.396.611.1 3047

The Starting of Oscillations in RC Generators—G. Gillard. (*TSF Pour Tous*, vol. 24, pp. 132–134; May, 1948.) A nonmathematical explanation.

621.396.611.21 3048

Series Mode Crystal Circuits—H. Goldberg & E. L. Crosby, Jr. (*Tele-Tech*, vol. 7, part 1, pp. 24–27, 86; May, 1948.) The equivalent circuit of a piezoelectric quartz crystal near resonance is considered; in a typical crystal, the series and parallel resonant frequencies are close together. A 118-Mc direct-operation circuit including a 7F8 twin triode is described; the eleventh mechanical harmonic of the crystal is used. This circuit can be modified for use as a converter.

621.396.611.21 3049

On the Equivalent Circuit and Performance of Plated Quartz Bars—J. K. Clapp. (*Gen. Radio Exp.*, vol. 22, pp. 1–7; March and April, 1948.) The impedance and equivalent electric circuit are derived for a long bar. Both mechanical and electrical stability for low-frequency modes of vibration are improved by the use of tension mounting and by operation at the second harmonic. Zero temperature coefficient can be obtained for a particular temperature.

621.396.611.4 3050

On the Design of Circuits Equivalent to Cavity Resonators—N. N. Malov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 421–430; April, 1948. In Russian.) The conditions of equivalence are defined and a general design method is discussed; this is used to determine a circuit equivalent to a cylindrical resonator in which a standing wave of the H_{011} type is present.

621.396.611.4 3051

A Method of Feeding Microwave Power into a Resonator Having a Fine Mode Structure—G. R. Newbery and W. E. Willshaw. (*Nature* (London), vol. 161, pp. 519–520; April 3, 1948.) Power was fed into a 24-cavity resonator 120 cm long, the frequency separation between the required mode of operation and the adjacent mode being only 0.2 per cent (6 Mc). The required mode had a phase difference of π between currents in adjacent cavities. A pulsed magnetron, Type E1944, was driven at 500 pulses per second, with 2 μs pulses, by an Admiralty Type 277 modulator with a peak output of 1.25 Mw. An estimated peak power of 390 kw was fed into the resonator.

621.396.615 3052

Use of Ordinary Valves as Microwave Oscillators—G. Fonda-Bonardi. (*Alta Frequenza*, vol. 17, pp. 69–73; April, 1948. In Italian with English, French, and German summaries.) The power output of an oscillator using an ordinary tube becomes zero when the oscillation frequency is so high that the electron transit time is a substantial fraction of the oscillation period. Oscillations can be obtained, however, when the transit time is a multiple of the oscillation period. A circuit using a 6V6 tube with a Lecher-line oscillatory circuit is described, with which frequencies from 1000 Mc to 3000 Mc have been obtained.

621.396.615 3053

The Self-Modulation of Auto-Oscillations in a Valve Oscillator with Automatic Biasing in the Cathode Circuit—N. A. Zheleztsov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 495–508; April, 1948. In Russian.) An analysis of the operation of the oscillator depicted in Fig. 1, using simplified tube characteristics and the van der Pol method of "abbreviated" equations. Methods are indicated for determining the amplitude of auto-oscillations and the conditions necessary for the appearance of self-modulation are found. Photographs are shown of phase trajectories on the van der Pol plane obtained with a cro. The gradual appearance of self-modulation is thus confirmed experimentally.

621.396.615 3054

An Investigation of a Relaxation Oscillator of the Transitron Type—V. V. Migulin and T. N. Yastrebtsova. (*Zh. Tekh. Fiz.*, vol. 18, pp. 603–614; May, 1948. In Russian.) An experimental investigation in which an external sinusoidal emf is applied to a *RC* oscillator operating in the regime of free auto-oscillations. Results are in good agreement with the theory of such oscillators, which is also discussed.

621.396.615.14:621.316.726.078.3 3055

Recent Developments in Frequency Stabilization of Microwave Oscillators—W. G. Tuller, W. C. Galloway, and F. P. Zaffarano. (Proc. I.R.E., vol. 36, pp. 794–800; June, 1948.) Pound's stabilizer (1690 of 1947 and 1311 of June) is discussed. The usable frequency range of the stabilizing circuit can be increased by changing the microwave circuit; this change is considered in detail and a graphical method of obtaining the frequency versus voltage characteristic is explained. The amount of harmonic distortion in the output of the stabilizing circuit is calculated for the case where the microwave oscillator has FM.

621.396.615.17 3056

Electrical Sawtooth Oscillations—H. Hertwig. (*Funk und Ton*, vol. 2, pp. 300–307; June, 1948.) Discussion of the conditions governing the rise and fall for both linear and exponential wave forms, with formulas for rise and fall times, frequency, and mean value of the current.

621.396.619.16 3057

On Some Characteristic Pulse Circuits—J. Moline. (*Radio Franç.*, pp. 11–15; May, 1948.) Details of circuits for: (a) low-power pulses, using phasing or dephasing, (b) frequency multipliers and dividers, and (c) high-power pulses suitable for the modulation of high-frequency transmitters.

621.396.619.23:621.395.44 3058

Linear Theory of Bridge and Ring Modulator Circuits—V. Belevitch. (*Elec. Commun.*, vol. 25, pp. 62–73; March, 1948.) The question of impedance mismatch between filters and modulators, and the losses in CuO-rectifier type modulators, are considered with reference to single-sideband carrier telephone systems. With stated assumptions and approximations, a simplified, practical theory is developed for bridge and ring modulators working between various ideal filter terminations. The transmission loss of a bridge modulator is measured experimentally under different terminating conditions approximating to those already assumed. Means of partially compensating for parasitic capacitance are described; these provide a basis for verifying the theory. An expression is given for the extra loss caused by the addition of capacitance to a capacitance-compensated modulator.

621.396.622.6:546.28 3059

The Silicon Crystal Detector—(*Bell Lab. Rec.*, vol. 26, pp. 152–155; April, 1948.) A short historical survey of development.

621.396.645 3060

The Use of Cathode Followers in the Penultimate Stage of Power Amplifiers—S. E. Glikman. (*Vestnik Svyazi*, no. 6, pp. 9–11; 1948. In Russian.)

621.396.645 3061

A Low-Noise Amplifier—H. Wallman, A. B. Macnee, and C. P. Gadsden. (Proc. I.R.E., vol. 36, pp. 700–798; June, 1948.) The "cascode" amplifier consists of a grounded-cathode triode followed by a grounded-grid triode. The combination is noncritical and has the low noise factor of a triode with the amplification and stability of a pentode. Noise factors averaging 0.25 db at a carrier frequency of 6 Mc and 1.35 db at 30 Mc have been achieved. Typical circuit details are given. Summary noted in 2475 of October.

621.396.645 3062

Time Response of an Amplifier of *N* Identical Stages—E. F. Grant. (Proc. I.R.E., vol. 36, pp. 870–871; July, 1948.) The response to a unit-step voltage is calculated, assuming that the pass band is so narrow compared to the center frequency that a low-pass equivalent circuit will describe the behavior of the amplifier with sufficient accuracy.

621.396.645 3063

Amplifier Load Impedance Reduction—B. M. Hadfield. (*Tele-Tech*, vol. 7, part 1, pp. 33–35, 82; May, 1948.) Design equations for amplifiers to deliver power into low-value load impedances over a wide range of frequencies, without the use of high-ratio output transformers. A specific example is considered in detail.

621.396.645:518.3 3064

Exact Design and Analysis of Double- and Triple-Tuned Band-Pass Amplifiers—M. Dishal. (*Elec. Commun.*, vol. 25, pp. 100–102; March, 1948.) Reprint of discussion on 3065 of 1947 already noted in 1315 of June. Reprint of original paper noted in 2493 of October.

621.396.645:537.533.9 3065

Use of Photo-Conductive Semiconductors as Amplifiers—E. S. Rittner. (*Phys. Rev.*, vol. 73, pp. 1212–1213; May 15, 1948.) The advantages of using a photoconductive semiconductor instead of an insulator as an amplifier are enumerated. Optimum results will probably be achieved with thin specimens of the same order of thickness as the penetration depth of the primary electron beam. Preliminary results for a polycrystalline layer of Se on glass are described. Further work on Se, Si, Ge, and PbS is in progress. See also 2757 of November.

621.396.645:539.16.08 3066

A Battery-Fed Amplifier for Counter Tubes with Continuous Generation of the Counter-Tube Voltage from an Accumulator Battery—A. Flammersfeld. (*Z. Naturf.*, vol. 1, pp. 168–170; March, 1946.) From the counter impulses the amplifier produces current pulses of constant length. The circuit is an improvement on that described by Neher (1697 of 1939). In the high-voltage generator, the tetrode anode current of a tetrode-triode sawtooth-wave generator is used to feed the primary of an ordinary mains transformer, whose secondary gives a series of high-voltage peaks.

621.396.645:621.396.822 3067

An Investigation of Amplifier Noise at Ultra High Frequencies—V. I. Siforov. (*Radiotekhnika* (Moscow), vol. 3, pp. 5–24; May and June, 1948. In Russian.) The theory of the one-stage amplifier is developed from consideration of an equivalent circuit consisting of a series of quadripoles in which an active noise-generating quadripole is included. Conditions for obtaining the maximum signal-to-noise ratio are established. The necessary and sufficient conditions for neutralizing tube noise are derived. The fundamental noise relationships in a multistage amplifier are determined; the noise factor of the amplifier can be deduced.

621.396.662 (083.74) 3068

Standardization of Tuning Units and Oscillators, 1948—Radionyme. (*Toute la Radio*, vol. 15, pp. 190–193. June, 1948.) A general discussion of the standards for 1939 and 1940 adopted by the Syndicat professionnel des Industries radioélectriques, with details of the standards proposed for 1948 by the SNIR (Syndicat national des Industries radioélectriques), which has taken the place of the SPIR. The new standards include variable capacitors of 490 pF and 130+360 pF max. respectively, and 3- and 4-waveband tuning units for which suitable capacitance values are tabulated.

621.396.662.21 3069

Design Calculations for Short-Wave Band-Spread Coils—J. Henry. (*Radio Franç.*, pp. 4–10; May, 1949.) Detailed calculations are made for tapped coils for the band 7 to 7.3 Mc; general formulas applicable to the whole of the short-wave bands are given. Tapped coils enable a single standard 460-pF capacitor to be used, and give a more linear spread than the series-capacitor method.

621.396.662.3 3070

An Experimental Investigation of the Wave-Guiding Properties of Multi-Terminal Filters—V. V, Potemkin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 447–454; April, 1948. In Russian.) Experiments were carried out with the ladder-type filters depicted in Figs. 1 and 2, to verify theoretical conclusions regarding various phenomena occurring in them, such as the appearance of different types of waves, dispersion of natural waves, band-pass filtering, and interruptions in the propagation of certain waves.

621.396.662.3 3071

Band Stoppers with Oscillating Crystals—W. Herzog. (*Arch. Elek.* (Übertragung), vol. 2, pp. 22–38; January, 1948.) Theory is given and design formulas are derived for both narrow- and broad-band stoppers. Suitable designs are given, with their functional characteristics, for (a) a broad-band stopper circuit using 2 crystals and 2 inductors, and (b) a narrow-band stopper using 3 crystals and 2 inductors.

621.396.662.3:621.396.611.21 3072

Quartz Filter Crystals with Low Inductance—J. J. Vormer. (Proc. I.R.E., vol. 36, pp. 802–804; June, 1948.) English version of 388 of March.

621.396.662.3:621.396.645 3073

A 3-Stage Coupling Filter for Wide-Band Amplifiers—P. G. Violet. (*Funk und Ton*, vol. 2, pp. 290–299; June, 1948.) Theory, design formulas, and practical details of a filter whose response curve shows three slight humps of equal height within the pass band, with relatively sharp cutoff at both limits.

621.396.665 3074

Adjustment Speed of Automatic-Volume-Control Systems—A. W. Nolle. (Proc. I.R.E., vol. 36, pp. 911–916; July, 1948.) The behavior of an avc amplifier, following a sudden change of input level, is analyzed on the basis of certain assumptions which are usually justified in practice. Equations are developed for the overload case and applied to a particular amplifier.

621.396.69:06.064 Paris 3075

Components, 1948—G. Giniaux. (*TSF Pour Tous*, vol. 24, pp. 57–60, 92–96, and 142–143; March to May, 1948.) Discussion of the special features of the Paris exhibition. See also 1898 of August.

621.396.621.001.4+621.397.62.001.4 3076

Most-Often-Needed F.M. and Television Servicing Information [Book Review]—M. N. Beitman. Supreme Publications, Chicago, 1948, 191 pp., $2.00. (Proc. I.R.E., vol. 36, p. 884; July, 1948.)

GENERAL PHYSICS

53.081+621.3.081 3077

On Systems of Electrical Units—É. Brylinski. (*Rev. Gén. Élec.*, vol. 57, pp. 200–204; May, 1948.) The use of the absolute systems of electrical units leads to certain difficulties, which are discussed. It is considered preferable to abandon the absolute systems and adopt electrostatic and electromagnetic systems defined by the condition that the dielectric constant of a vacuum is the unit of dielectric constant, or that the permeability of a vacuum is the unit of magnetic permeability. These systems give the same numerical measures of electrical quantities as the absolute systems. See also 2504 of October (Dzung and Meldahl) and back references.

530.145 3078

On the Quantum Theory of Wave Fields—F. J. Belinfante. (*Physica, 's Grav.*, vol. 7, pp. 765–778; October, 1940. In English, with German summary.)

535.215.6:621.383.5 3079

Theory of the Barrier-Layer Photoeffect—K. Lehovec. (*Z. Naturf.*, vol. 1, pp. 258–264; May, 1946.) Differential equations are established involving the relation between photocurrent and voltage, intensity of illumination, frequency of the light waves, the properties of the semiconductor and of the electrode, and the resistance of the external circuit. The properties of defect and excess semiconductors are discussed.

535.37 3080

Relation between Photoconduction and Luminescence in Zinc Sulphide—M. H. F. Wilkins and G. F. J. Garlick. (*Nature* (London), vol. 161, pp. 565–566; April 10, 1948.)

537.213:517.1:532.58 3081

On the Harmonic and Biharmonic Problems of a Region Bounded by a Circle and two Paral-

lel Lines—R. Westberg. (*Ingen. Vetensk. Akad. Handl.*, no. 197, 66 pp.; 1948. In English.) Analysis of the potential field of a circular cylinder between two parallel conducting planes, and also of the motion of a circular cylinder in a viscous fluid between two parallel planes.

537.228.1 3082
Equations of Piezoelectricity—R. K. Cook. (*Nature* (London), vol. 161, pp. 524–525; April 3, 1948.) In Cady's book on piezoelectricity (2084 of 1947), some confusion between generalized co-ordinates and forces in the Lagrange function results in erroneous piezoelectric equations. This gives rise to errors in the basically important linear equations which give the elastic stresses and electric field intensities as functions of the elastic strains and electric displacements.

537.228.1+539.3]:512.9 3083
Applications of Tensor Analysis to Elasticity and Piezoelectricity—J. H. Jurmain. (*Jour. Frank. Inst.*, vol. 245, pp. 475–500; June, 1948.) A systematic theory of elasticity, based on tensor analysis, is developed and shown to give results consistent with those derived by other methods. The theory is extended to piezoelectricity and, if boundary conditions are carefully selected, the general solutions obtained are applicable to practical problems. Examples are given.

537.312.62 3084
On the Theory of Superconductivity—F. J. Wiśniewski. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1964–1965; June 14, 1948.) A possible mechanism for superconductivity is described. The fundamental equation of London is deduced from the principles of mechanics.

537.52 3085
The Breakdown of Gases in High Frequency Electrical Fields—D. H. Hale. (*Phys. Rev.*, vol. 73, pp. 1046–1052; May 1, 1948.) "It is assumed that breakdown occurs when the electrical field and the frequency are such that an electron acquires the ionizing energy at the end of one mean free path. The field for breakdown is thus a function of the frequency of the applied potential and the ionization potential and pressure of the gas. The fields for breakdown of argon and xenon are calculated and expressed as functions of the frequency and the gas pressure." Good agreement between calculated potentials and experimental results is found for frequencies >10 Mc.

537.523.5 3086
On the Theory of the High-Current Arc Column—P. Schulz. (*Z. Naturf.*, vol. 2a, pp. 662–666; November and December, 1947.)

537.533.9:621.396.645 3087
Use of Photo-Conductive Semiconductors as Amplifiers—Rittner. (*See* 3065.)

538.1 3088
On the Current and the Density of the Electric Charge, the Energy, the Linear Momentum and the Angular Momentum of Arbitrary Fields—F. J. Belinfante. (*Physica, 's Grav.*, vol. 7, pp. 449–474; May, 1940. In English, with German summary.)

538.213:538.221 3089
Determination of Magnetic Permeability from Resistance Measurements on Iron Wires of Different Structure at Frequencies of the order of 100 Mc/s, in relation to the Weiss Elementary Domains—M. J. O. Strutt and K. S. Knol. (*Physica, 's Grav.*, vol. 7, pp. 635–654; July, 1940. In German, with English summary.) The high-frequency permeability calculated from the ratio of the ac and dc resistances is practically constant at room temperature up to about 300 Mc, but shows a marked decrease with increasing frequency at −183° C. An explanation of this is based on a simple model, which permits the determination of the order of magnitude of the Weiss domains for the different wires.

538.21 3090
Effect of Reversible and Irreversible Variations of Magnetization on the Thermoelectric Power of Ferromagnetic Materials—J. Bouchard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1895–1897; June 7, 1948.)

538.23 3091
On the Effective Length of a Small Barkhausen Discontinuity—J. L. Snoek. (*Physica, 's Grav.*, vol. 7, pp. 609–624; July, 1940. In English, with German summary.) The effective length is calculated for a wire of square cross section and found to be equal to the product of the reversible permeability and the wire thickness.

538.242 3092
On the Theory of Gyromagnetic Effects—C. J. Gorter and B. Kahn. (*Physica, 's Grav.*, vol. 7, pp. 753–764; October, 1940. In English, with French and German summaries.)

538.56+621.385.029.63/.64+621.384.6 3093
Waves and Electrons Traveling Together—A Comparison between Traveling Wave Tubes and Linear Accelerators—L. Brillouin. (*Phys. Rev.*, vol. 74, pp. 90–92; July 1, 1948.) Traveling-wave tubes operate with high space charge and weak waves of constant velocity, while "synchro-" devices use very low space charge and high-power waves whose velocity is progressively increased. In one case, there is an energy transfer from the electrons to the wave, which is strongly amplified; in the other case, energy is transferred from the wave to the particles, which are thereby accelerated. A general theory is needed that would include both extreme cases of traveling-wave amplifiers and linear accelerators without introducing any restriction on the magnitudes of either fields or space charge. A composite wave is considered which has discontinuities in the derivative of the longitudinal field distribution and represents a sort of electromagnetic shock wave. It yields a rigorous solution of the wave equation for the case of space charge, and represents a generalization of the two extreme cases.

538.566 3094
Remarks on Spherical Electromagnetic Waves—V. Sorokin. (*Zh. Eksp. Ter. Fiz.*, vol. 18, pp. 228–235; February, 1948. In Russian.) Using vectorial spherical functions, simple formulas are derived for spherical electromagnetic waves in vacuo and for electromagnetic fields of multipoles.

538.566 3095
A New Formula for Calculating the Phenomena of Diffraction—É. Durand. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1812–1814; May 31, 1948.)

538.566 3096
Electromagnetic Theory of Diffraction by Black Screens—É. Durand. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1972–1974; June 14, 1948.)

538.566:535.42 3097
Diffraction of Centimeter Electromagnetic Waves at an Aperture in a Metal Sheet—H. Severin. (*Z. Naturf.*, vol. 1, pp. 487–495; September, 1946.) The Huyghens-Fresnel principle, formulated for radio waves, is applied to the accurate quantitative determination of diffraction effects along the normal through the center of a circular aperture and also in the field near the aperture. Experiments with apertures of radius ranging from $\lambda/2$ to 8λ, and with wavelengths of 10 cm. and 6 cm, confirm the theory, which is in agreement with Kirchhoff's diffraction theory and Maxwell's equations. See also 2326 of September (Tranter).

538.569.4:535.61-15:525.7 3098
Transmission of the Atmosphere in the 1-5 Micron Region—A. Elliott and G. G. MacNeice. (*Nature* (London), vol. 161, p. 516; April 3, 1948.) Details of equipment for measurements over a 600-yard path. A typical transmission curve shows high values between 3μ and 4μ, where transmission is comparable with that at 2.2μ.

538.569.4:546.331.31-145.1 3099
Absorption of U.H.F. Waves in Salt Solutions—S. K. Chatterjee and B. V. Sreekantan. (*Indian Jour. Phys.*, vol. 22, pp. 229–242; May, 1948.) The absorption index is measured directly by a free-wave method at frequencies between 300 and 500 Mc for concentrations ranging from N/2 to N/16. Reflection coefficients at different frequencies are calculated; the values of ionic relaxation time, dielectric constant, and loss angle obtained are compared with theoretical values deduced from the Debye-Falkenhagen theory. The product of the wavelength for maximum absorption and the concentration of the solution is approximately constant.

538.569.4.029.64:546.171.1 3100
Collision Broadening of the Ammonia Inversion Spectrum at High Pressures—B. Bleaney and J. H. N. Loubser. (*Nature* (London), vol. 161, pp. 522–523; April 3, 1948.) Measurements have been extended over the frequency range 0.1 to 1.2 cm^{-1} at pressures up to 6 atmospheres. See also 1916 of August and back references.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.53:621.396.96 3101
Velocity of Meteors Measured by Diffraction of Radio Waves from Trails during Formation—C. D. Ellyett and J. G. Davies. (*Nature* (London), vol. 161, pp. 596–597; April 17, 1948.) The radiation scattered from the electron trail was observed as the meteor crossed the antenna beam; the variation in received power is compared with Fresnel's integrals for diffraction at a straight edge. The mean geocentric velocity deduced for the Geminid meteors is 34.4 km, in good agreement with Whipple's value of 34.7 km obtained from photographic measurements of visible meteors.

523.72+523.16]:621.396.822 3102
Radio Noise of Extra-Terrestrial Origin and Its Effect on Telecommunication Technique—G. Lehmann. (*Onde Élect.*, vol. 28, pp. 164–172 and 200–205; April and May, 1948.) The present state of knowledge of cosmic noise is reviewed, reference being made to investigations on wavelengths ranging from decameters to millimeters. The fundamental principles of thermodynamics necessary for the interpretation of the experimental results are recalled, receiver noise factor is defined and a method of measuring it is described. The performance of actual uhf receivers is compared with that which is theoretically possible.

523.72.029.63:523.746 3103
A Solar Noise Outburst at 600 Mc/s and 1200 Mc/s—F. J. Lehany and D. E. Yabsley. (*Nature* (London), vol. 161, pp. 645–646; April 24, 1948.) Bursts similar to those at 200 Mc described in 412 of March (McCready, Pawsey, and Payne-Scott) are not normally observed at 600 and 1200 Mc; noise level rises gradually as the sunspots appear, and shows correlation with sunspot area. On one occasion only, large bursts occurred at 600 and 1200 Mc, but not at 200 Mc.

523.746 3104

Magneto-Hydrodynamic Waves and Sunspots: Part 3—H. Alfvén. (*Ark. Mat. Astr. Fys.* vol. 34, part 4, section A, 20 pp.; April 13, 1948. In English.) The instability factor θ_s is zero at the center of the sun, rises to a maximum as distance from the center is increased radially, and thereafter falls again to zero at the surface of the unstable core.

A disturbance in the zone of maximum instability in, say, the northern hemisphere gives rise to two hydrodynamic waves traveling in opposite directions parallel to the magnetic field. One waves escapes into the stable zone and on reaching the surface of the sun gives rise to a spot. The other passes through the unstable core and is subject to an acceleration, which is a function of θ_s, and is negligible until the southern zone of maximum instability is reached. Any acceleration of the wave results in a new disturbance and a new pair of waves having amplitudes proportional to the acceleration. Thus there are two strong waves emitted from this southern zone: one forms a spot in the southern hemisphere of the sun and the other retraces the path through the core. The process is repeated until it is attenuated to negligible proportions. We should thus expect a harmonic process of period T, the time taken for the wave to cross the core. T can be shown to be about 11 years.

The theory would also predict a correlation between sunspot activity at a given latitude in one hemisphere at a given epoch of one cycle and that at the same latitude in the other hemisphere in adjacent cycles. Statistical analysis, noted in 3105 below, confirms this hypothesis. For previous parts see *Mon. Not. R. Ast. Soc.*, vol. 105, pp. 3–16 and 382–394; 1945; abstracts of these will appear shortly. See also 2231 of September (Cowling) and back references.

523.746:519.251.8 3105

Statistical Tests of H. Alfvén's Theory of Sunspots—I. Galvenius and H. Wold. (*Ark. Mat. Astr. Fys.*, vol. 34, part 4, section A, 9 pp.; April 13, 1948. In English.) See 3104 above.

523.746:621.396.11 3106

Sunspots and Radio Weather—A. Arzinger, H. E. Hallborg, and J. H. Nelson. (*RCA Rev.*, vol. 9, pp. 229–244; June, 1948.) The sunspots producing disturbance in high-frequency radio communication are found to lie mainly within a semicircle of about 26° radius in the eastern half of the solar disk; its center coincides with that of the disk. The spots have maximum effect when they cross either the circle or the central meridian. While they are within the semicircle they cause erratic conditions. Studies of the polarities of the spots show that, for the northern hemisphere, "reds" (positive) have a preponderant effect and tend to lower frequencies, while "violets" (negative) tend to raise them. The reverse is true for the southern hemisphere. These observations, in combination with Washington F-layer data, are applied to short-period radio-weather forecasting.

523.752 3107

A Chromospheric Eruption of Extraordinary Size—A. Behr. (*Z. Naturf.*, vol. 1, pp. 537–539; September, 1946.) An account, with photographs, of a very large eruption observed at the Fraunhofer Institute on July 25, 1946, and of simultaneous recording of the field strength of a London 9.66-Mc transmission. The field strength fell to a low value at the commencement and remained low for the duration of the eruption. About 26½ hours after the start of the eruption, a geomagnetic storm was recorded at the Hamburg-Wingst observatory; this time lag corresponds to a velocity of the particle radiation from the sun of 1570 km. An aurora of medium brightness was observed for about an hour at Freiburg some 6 hours after the start of the magnetic storm. This was also seen in Paris and in Switzerland.

538.12:521.15 3108

Blackett's Hypothesis of the Magnetic Field of Rotating Bodies—N. Arley and J. Fuchs. (*Nature* (London), vol. 161, pp. 598–599; April 17, 1948.) Comment on 3112 of 1947 (Blackett). Difficulties involved in Blackett's interpretation are discussed by Arley and possible experimental tests suggested, such as those of Barnett on gyromagnetic magnetization, and of Einstein and de Haas on the inverse effect of rotation by magnetization.

Fuchs interprets Blackett's constant of proportionality by expressing it in a dimensionless form valid for the nonrational electromagnetic system of units as well as the usual nonrational electrostatic system.

550.384:551.510.5:525.624 3109

A Possible Influence of the Moon on Recurrent Geomagnetic Activity—O. R. Wulf and S. B. Nicholson. (*Phys. Rev.*, vol. 73, pp. 1204–1205; May 15, 1948.) Tidal air mounds due to the moon may be one of the factors which account for the observed 27-day period of geomagnetic activity. See also 3900 of 1947.

550.384:551.510.535:523.7 3110

Evidence of a Solar Effect on the Ionized Regions of the Upper Atmosphere—P. Lejay, A. Haubert, and J. Durand. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1768–1770; May 31, 1948.) Comparison of continuous records of F_2-layer critical frequency variations and magnetic K-indexes for September 1946, reveals a definite correlation between the two. The variations corresponding to magnetic storms have, in general, the same sense at stations as far apart as Slough, England, and Washington, D. C. The fact that the observed variations are generally greatest in the middle of the day, when the sun is high, suggests a quasi-instantaneous action of the sun.

550.384.3 3111

Basis and Preparation of the Magnetic Deviation Chart for the Epoch 1945—O.-F. Burmeister. (*Beitr. Geophys.*, vol. 60, nos. 3 and 4, pp. 177–195; 1944.) Discussion of all the data, from various sources, on which the chart was based.

550.384.3 3112

A Contribution to the Knowledge of Magnetic Secular Variation—A. Kiskyras. (*Beitr. Geophys.*, vol. 60, nos. 3 and 4, pp. 222–234; 1944.) Discussion based on the theory that the greater part of the secular variation is due to changes in the magnetization of the rocks in the upper 20 km of the earth's crust. Such changes are brought about by changes of temperature.

550.385 "1941.03.01":537.591 3113

Contribution to Material on the Effect of the Magnetic Storm of 1st March 1941 on Cosmic Radiation—A. Sittkus. (*Z. Naturf.*, vol. 1, pp. 204–208; April, 1946.) The results published by Kölhorster (12 of 1945) were obtained at Dahlem and Graz by coincidence methods. Ionization-chamber measurements at Freiburg i. Br. are here described and an estimate is made of the temperature effect at all three stations. The effect of the magnetic storm was qualitatively similar for the three places, though the amplitudes of the fluctuations differed.

551.510.53 (479) 3114

The Structure of the Upper Layers of the Atmosphere as determined by Twilight Observations—T. G. Megrelishvili and I. A. Khvustikov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 59, pp. 1283–1286; March 1, 1948. In Russian.) Values of upper-air density and pressure deduced from spectrophotometric observations have been obtained regularly since 1942 at Abbastuman observatory (South-West Caucasus) during morning and evening twilight. Results agree closely with those of other workers using different methods.

551.510.535 3115

An Approach to the Approximate Solution of the Ionosphere Absorption Problems—J. E. Hacke, Jr. (PROC. I.R.E., vol. 36, pp. 724–727; June, 1948.) A series of parabolic approximations has been obtained for portions of the Chapman distribution and its product with the collisional frequency. By their use, an improved approximate solution has been found for the "true" and the group height of reflection, and for the absorption in the region, under conditions of (a) vertical incidence, (b) wave frequency greater than the maximum collision frequency and less than the critical frequency for the region, and (c) with the earth's magnetic field neglected.

"These improved analytic approximations are compared with the usual parabolic approximation and with numerical approximations obtained by other workers."

551.510.535 3116

Electrical Conductivity of the Ionospheric D-Layer—T. G. Cowling and R. Borger. (*Nature* (London), vol. 161, p. 515; April 3, 1948.) Martyn's suggestion (1024 of May), that electric currents in the D layer must make a contribution to the solar and lunar geomagnetic variations roughly equal to that of the E and F layers, implies that the integral conductivity of the D layer must be as great as that of the E and F layers combined. Discussion shows this to be improbable.

551.510.535:621.396.11 3117

Upper-Atmosphere Circulation as Indicated by Drifting and Dissipation of Intense Sporadic-E Clouds—O. P. Ferrell. (PROC. I.R.E., vol. 36, pp. 879–880; July, 1948.) The motion of such clouds is illustrated by a series of maps for two recent occasions when the clouds were sufficiently intense to permit long-range communication in the 50 to 54-Mc amateur band.

551.510.535:621.396.11 3118

Triple Splitting of Ionospheric Rays—Eckersley. (*See* 3222.)

551.510.535:621.396.11 3119

Triple Splitting of Ionospheric Rays—Meek. (*See* 3221.)

551.513 3120

Intensity of the Zonal Circulation in the Atmosphere outside the Tropics—H. Flohn. (*Beitr. Geophys.*, vol. 60, nos. 3 and 4, pp. 196–209; 1944.) From new circumpolar charts of the atmospheric temperature and pressure distributions in the northern hemisphere, new mean values of the zonal component of the currents at heights of 5, 11, 16, and 22 km in temperate latitudes are deduced. Marked differences are found between the continental and maritime sectors.

551.578.1:621.396.96 3121

Measurement of Rainfall by Radar—J. S. Marshall, R. C. Langille and W. McK. Palmer. (*J. Met.*, vol. 4, pp. 186–192; December, 1947. Reprint.) Experiments using 10-cm radar equipment have confirmed theoretical conclusions that the radiation reflected from rain is proportional to Z, the sum of the sixth powers of diameters of raindrops in a representative volume. Correlation of Z with rainfall intensity R at ground level suggests that $Z \propto R^2$, approximately. It may be possible to determine rainfall intensity 100 km away by radar echo, since vertical scanning indicates that rain content varies only slightly with height. See also 2062 of August (Ryde) and back reference.

LOCATION AND AIDS TO NAVIGATION

621.396.93:551.594.6 3122

Distant Localization of Individual Atmospherics with a Cathode-Ray Direction-Finder of Unidirectional Type—W. Stoffregen. (*Ark. Mat. Astr. Fys.*, vol. 34, part 4, section A, 14 pp.; April 13, 1948. In English.) An improved form of cathode-ray direction finder is described in which a "sense antenna" eliminates an uncertainty of 180° by blacking out the negative half-cycle. Radar technique is used to determine the range.

621.396.933 3123

C.A.A. V.H.F. Omnidirectional Range at United Air Lines—F. J. Todd. (*Communications*, vol. 28, pp. 10–11, 34; April, 1948.) The new system referred to as VOR (vhf omnidirectional range) has the advantages of improved accuracy of indication, comparatively static-free operation, greater flexibility, and channel assignments adequate to accommodate the needs of expanding air traffic. The system provides the information necessary to enable an aircraft to fly along a definite path with respect to the VOR station. The component parts of the equipment are described.

621.396.96 3124

Frequency-Modulated Radar Techniques—I. Wolff and D. G. C. Luck. (*RCA Rev.*, vol. 9, pp. 352–362; June, 1948.) Continuation of 2798 of November. Two superheterodyne systems, namely a sideband system and a signal-following system, remove the transmitted modulation from the received if signal. An electromechanical device using a vibrating capacitor and an electronic device using a beam of electrons have proved useful for FM of radar oscillators. Several adaptations of the well-known cycle-rate counter for making use of beat-frequency data have made FM radar very useful where automatic indication of, or control by, the range and speed of a single target is needed. To be continued.

621.396.96 3125

Technique and Development of Radar—Demanche. (*Onde Élec.*, vol. 28, pp. 206–210; May, 1948.) Discussion of methods of eliminating permanent echoes. See also 1366 of June.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.5 3126

On the Theory of the Diffusion Pump—R. Jaeckel. (*Z. Naturf.*, vol. 2a, pp. 666–677; November and December, 1947.) The physical processes in the diffusion pump are treated theoretically and first- and second-approximation formulas for quantitative calculation are derived. From these formulas the power of a diffusion pump or the design characteristics for a pump of given power can be determined.

535.37 3127

The Effect of Electron Distribution at Localized Levels on the Course of Different Luminescence Processes in the CaS·SrS-Ce, Sm, La Phosphors and the Number of Repeated Electron Localizations—V. L. Levshin. (*Zh. Eksp. Teor. Fiz.*, vol. 18, pp. 149–163; February, 1948. In Russian.) An experimental investigation which shows that the luminescence processes depend not only on the number of ionized centers and localized electrons but also on the electron distribution at localized levels. It is shown that the number of repeated localizations of electrons is not great. Repeated flashes and phosphorescence are examined in detail.

535.37 3128

The Nature of Centres of Luminescence in Photochemically-Coloured Alkali-Halide Crystals—M. L. Katz. (*Zh. Eksp. Teor. Fiz.*, vol. 18, pp. 164–173; February, 1948. In Russian.)

535.37 3129

On a Group of Mixed Phosphors with Mixed Activators—P. Brauer. (*Z. Naturf.*, vol. 1, pp. 70–78; February, 1946.) An investigation of the properties of mixed sulphides of the alkaline earths with Sm as one activator and either Ce, Pr, Eu, Mn, or Pb as the other. The behavior of SrS·Eu-Sm differs considerably from that of other members of the group, whose general properties can be explained on the assumption that Sm acts as a "killer" substance. See also 762 of 1940.

535.37:535.61-31 3130

Miscellaneous Observations on the Rise and Decay of the Luminescence of Various Phosphors—W. de Groot. (*Physica*, *'s Grav.*, vol. 7, pp. 432–446; May, 1940. In English, with French and German summaries.) Results are given for periodic illumination of the following substances by intense ultraviolet light: ZnS-Ag, ZnS-Cu, ZnSCdS-Ag, ZnSMnS, CaWO$_4$-Sm, and Zn$_2$SiO$_4$-Mn.

535.37:621.385.832 3131

Fluorescent Compounds—(*Electronics*, Buyers' Guide Issue, vol. 21, p. M17; June, 1948.) Colors, luminous efficiencies, and other characteristics are tabulated for various materials used for cathode-ray screens.

535.61-15:621.383.4 3132

Spectral Sensitivity of Lead Telluride Layers—T. S. Moss. (*Nature* (London), vol. 161, pp. 766–767; May 15, 1948.) Curves of relative sensitivity and wavelength are shown for temperatures of 195° K, 90° K and 20° K; the results differ considerably from those obtained by earlier workers. The threshold of sensitivity at 90° K is similar to that for PbSe but the PbTe layers are easier to produce and form efficient infra-red detectors for wavelengths up to 5 or 6 μ at 20° K.

537.228.1+621.314.63 3133

Crystal—(*Electronics*, Buyers' Guide Issue, vol. 21, pp. M14–M16; June, 1948.) Electrical characteristics of quartz, EDT, and DKT crystals, and of Si, Se, and CuO rectifiers are tabulated or shown graphically.

537.228.1 3134

Compressional Piezoelectric Coefficients of Monoclinic Crystals—H. Jaffe. (*Phys. Rev.*, vol. 73, p. 1467; June, 15, 1948.) Experimental values are given for the coefficients relating to fields parallel to the polar axis. The hydrostatic piezoelectric effect for EDT appears to be much smaller than that derived from Mason's data (740 of April).

537.312.62 3135

Measurements of Radio Frequency Resistance of a Piece of Columbium Nitride through the Transition [from the normal to the superconducting state]—J. V. Lebacqz. (*Phys. Rev.*, vol. 73, p. 1476; June 15, 1948.) The measured values for frequencies between 600 and 1000 kc were the same as the dc values at the same temperatures.

538.221:538.213 3136

On the Dispersion of Initial Permeability—J. L. Snoek. (*Physica*, *'s Grav.*, vol. 7, pp. 515–518; June, 1940. In German, with English summary.) Purified and annealed Fe has an initial permeability, at frequencies of the order of 1 cps, which is many times higher than the value determined from measurements of the skin effect at frequencies of 1 to 100 Mc. Present theories offer no explanation of this.

538.27 3137

The Effect of Premagnetization on the Complex Permeability of Coil Cores—R. Feldtkeller and E. Stegmaier. (*Frequenz*, vol. 2, pp. 121–130; May, 1948.) The laws connecting premagnetization with the complex permeability of Si-Fe and Ni-Fe sheet are reviewed, with particular reference to curves relating the critical value of the complex permeability to the frequency and amplitude of the ac induction. With constant ac induction, premagnetization has an effect on the eddy currents in the material. This effect is explained by the action of the premagnetization on the reversible permeability of the surface layers and of the interior of the sheet.

546.42/.431].82 3138

Barium Titanate and Barium Strontium Titanate Resonators—H. L. Donley. (*RCA Rev.*, vol. 9, pp. 218–228; June, 1948.) Values of the equivalent circuit elements are given for such resonators vibrating normal to the direction of the polarizing and rf fields. An increase in the effective piezoelectric constant with increasing polarizing field is observed. At 25 v/mil an average piezoelectric constant of about 200×10^{-8} electrostatic units, a Q, of 80 and an electromechanical coupling factor of 0.2 are found for BaTiO$_3$ resonators. Lower activity, but higher Q, results with (Ba/Sr)TiO$_3$ resonators. A frequency constant of 211 kc-cm is found for BaTiO$_3$ resonators; for mixtures this rises to 275 kc-cm as the Sr content increases to 30 per cent. The piezoelectric activity ceases as the Curie temperature of a particular composition is approached. A large temperature coefficient of frequency of 1 part in 400 per 1° C limits the use of BaTiO$_3$ for resonators but it compares favorably with Rochelle salt, for use in pickups.

546.431.82:537.228.1/.2 3139

Piezoelectric or Electrostrictive Effect in Barium Titanate Ceramics—W. P. Mason. (*Phys. Rev.*, vol. 73, pp. 1398–1399; June 1, 1948.) Resonance effects in multicrystalline BaTiO$_3$ ceramics are shown to be electrostrictive; they are analogous to magnetostrictive effects in ferromagnetic materials and not to the "quadratic" piezoelectric effect discussed by Mueller (2186 of 1940).

549.623.5(94):621.315.613 3140

Australian Mica—(*Engineering* (London), vol. 165, p. 379; April 16, 1948.) The output of high-grade mica could be increased to 250 tons per year, equivalent to 15 per cent of world production, by provision of improved equipment and amenities for workers.

621.3.015.5:546.217 3141

Electrical Breakdown Strength of Air at Ultra High Frequencies—J. A. Pim. (*Nature* (London), vol. 161, pp. 683–684; May 1, 1948.) A series of experiments in the frequency range 100 to 300 Mc, using cw throughout and an improved method of measuring the rf voltage at the point of breakdown. The relations obtained between the electrical breakdown stress and gap width, frequency and gas pressure suggest that with an alternating field of a particular frequency, some charged particles, probably electrons, are unable to cross the gap before the field is reversed.

621.315.612+666 3142

Ceramics and Glass—(*Electronics*, Buyers' Guide Issue vol. 21, pp. M4–M5; June, 1948.) Electrical, thermal, and mechanical properties are tabulated for various materials, with brief remarks about each.

621.315.615 3143

A Brief Outline of Insulating Oil Problems—E. S. Lane. (*Distrib. Elec.*, vol. 20, pp. 337–338; April, 1948.) Summary of paper read at an Australian Electricity Supply Engineers' Association conference. The presence of oxygen causes acidity and sludging. Oxidation can be reduced by using inert gases, silica-gel, or conservators.

621.315.616:533.5 3144

Vacuum Properties of Synthetic Dielectrics—B. G. Hogg and H. E. Duckworth.

(*Rev. Sci. Instr.*, vol. 19, pp. 331–332; May, 1948.) Wide differences of vapor pressure are found in vacuum tests on 28 commercial materials. Many of these, such as polystyrene, teflon, and mycalex, are particularly suitable for use in vacuum systems.

621.315.616:678 3145

Rubber Dielectrics—Some Chemical Aspects—B. B. Evans. (*Distrib. Elec.*, vol. 21, pp. 2–6; July, 1948.) Qualitative accelerated aging tests for the study of rubber dielectric deterioration are described. Vulcanizing processes are hastened by the use of organic accelerators, the desirable properties of which are discussed. Life can be increased by incorporating small amounts of "anti-oxygen" or "disactivator" materials.

621.318.22+621.318.32 3146

Magnetic Materials—(*Electronics*, Buyers' Guide Issue, vol. 21, pp. M20–M23; June, 1948. Characteristics are tabulated for materials used for laminated, solid, and powdered metal cores and for permanent magnets.

621.383 3147

On the Absorption, Light Sensitivity and Electrical Conductivity of CdS Layers—K. Weiss. (*Z. Naturf.*, vol. 2a, pp. 650-652; November and December, 1947). A short review of the properties of layers produced by a particular process. The sensitivity is 10^{-2}A/lumen and a light power of 10^{-10}w can be detected. Completely insulating layers of CdS, PbS and Sb_2S_3 have been produced.

621.383:546.817.221 3148

On the Effect of Gases, particularly of Traces of Oxygen, on the Electrical Properties of Evaporated PbS Layers—H. Hinterberger. (*Z. Naturf.*, vol. 1, pp. 13–17; January, 1946.) An investigation of the effect of N, Ar, O, H, and air. N and Ar have no effect, but air and especially small traces of O have a marked effect, which is qualitatively the same as that of a sulphur treatment. Tempering in H increases the conductivity of PbS layers with excess Pb.

621.395.625.3 3149

Test Characteristics of Recording Wire—G. S. Carter and R. Koontz. (*Tele-Tech*, vol. 7, part 1, pp. 38–40, 75; May, 1948.)

669 3150

Metals and Alloys—(*Electronics*, Buyers' Guide Issue, vol. 21, pp. M24–M29; June, 1948.) Brief details of various compositions used for solders, resistors, switches and contacts, thermocouples, and tube parts.

678.72 3151

New Synthetic Rubber Has High Oil and Heat Resistance—(*Materials and Methods*, vol. 27, pp. 72–74; June, 1948.) Discussion of a polyacrylic ester, commercially named Hycar P.A., which withstands dry heat to 400° F, has excellent resistance to deterioration by oils or sunlight, forms a good gas barrier and has a good flex life. Methods of preparing cured varieties of varying hardness are described. The relative merits of Hycar P.A., Lactoprene EV and styrene rubber GR-S are discussed.

679.5 3152

Plastics—(*Electronics*, Buyers' Guide Issue, vol. 21, pp. M30–M35; June, 1948.) Electrical and mechanical properties of moulding and casting compounds, laminated materials, and synthetic and natural rubbers.

679.5:620.193 3153

The Effect of Fungi and Humidity on Plastics—J. Leutritz, Jr. (*ASTM Bull.*, pp. 88–90; May, 1948.) Discussion of experimental results for various materials exposed to tropical conditions. Fungus without moisture is a fairly good insulator. Elimination or mitigation of moisture difficulties will automatically control fungus and reduce insulation failures.

679.5:621:397.5:535.88 3154

Plastic Lenses in Television Projection—D. Starkie. (*Jour. Telev. Soc.*, vol. 5, pp. 86–92; September, 1947. Discussion, p. 93.) The properties of suitable plastic optical materials are discussed, and a method of manufacturing lenses by a "surface-finishing" process is described. A preform plastic optical component is first moulded, using light polymerization. The preform is removed from the mould and annealed. Any departures from the form of the mould found after the annealing are corrected by casting a thin film of polymer on the surface of the preform.

679.5:621.793 3155

Metal-Coated Plastics Combine Advantages of Both Materials—H. R. Clauser. (*Materials and Methods*, vol. 27, pp. 79–82; June, 1948.) Metal coatings can be applied by electro-plating, vacuum evaporation or metal spraying. The relative merits of these methods are discussed, and various practical applications are mentioned. See also 2544 of October (Narcus).

679.5:681.42 3156

Plastics as Optical Materials—H. R. Moulton. (*ASTM Bull.*, pp. 75–77; March, 1948. Discussion, p. 77.) Discussion of: (a) properties of optical materials, (b) the relative merits of glass and plastics, (c) methods of processing plastics, (d) future possibilities of using complex mixtures.

533.5 3157

A Manual of Vacuum Practice [Book Review]—L. H. Martin and R. D. Hill. Melbourne University Press, Melbourne; Oxford University Press, London; 1947. 120 pp., 10s.6d., (*Nature* (London, vol. 161, p. 665; May 1, 1948.) "An excellent addition to the all too few modern books in the English language on vacuum technique . . . the commercial apparatus referred to in the text includes both English and American types of the most modern pattern."

MATHEMATICS

517.392:621.396.67 3158

Concerning a New Transcendent, its Tabulation and Application in Antenna Theory—C. J. Bouwkamp. (*Quart. Appl. Math.*, vol. 5, pp. 394–402; January, 1948.) Some of the features of the function

$$E_1(z)=\int_0^1(1-e^{-izst})ds\ dt/st$$

are discussed with particular reference to Hallen's antenna theory. A short table of numerical values is given.

517.512.2:621.396.67 3159

Fourier Transforms in Aerial Theory: Part 6—J. F. Ramsay. (*Marconi Rev.*, vol. 11, pp. 45–50; April to June, 1948.) Conclusion of 2270 of September. The earlier parts are summarized and further developments indicated. A classified bibliography of 82 references is included.

518.5 3160

The Univac—(*Electronic Ind.*, vol. 2, pp. 9, 19; May, 1948.) Summary of IRE 1948 Convention paper noted in 2548 of October. A short description of a high-speed electronic digital computer. The method of storing 1000 12-digit numbers as supersonic pulses in Hg columns is described. Magnetic tape recording is used to pass controling instructions and data to and from the computer. Possible applications varying from the solution of integral and partial differential equations to the classification of statistical information are discussed.

518.5 3161

Compact Analog Computer—S. Frost. (*Electronics*, vol. 21, pp. 116–122; July, 1948.) A technical description of the Reeves Electronic Analog Computer (REAC) together with circuit diagrams and photographs. The REAC comprises a computer unit, servomechanism unit, recorder unit, and associated power supplies. Solutions of both linear and nonlinear simultaneous differential equations, such as occur in design engineering, can be readily obtained.

518.61:621.396.611.1 3162

The Escalator Process for the Solution of Damped Lagrangian Frequency Equations—J. Morris. (*Phil. Mag.*, vol. 38, pp. 275–287; April, 1947.) Extension of 1636 of 1945 (Morris and Head) to equations of the Lagrangian type in which each element is a polynomial instead of merely a linear function of the unknown latent root. The "modes" of such equations have orthogonal properties analogous to those for ordinary Lagrangian frequency equations. See also 2167 of 1947.

518.61:621.396.611.1:512.831 3163

Note on the Morris Escalator Process for the Solution of Linear Simultaneous Equations—R. A. Frazer. (*Phil. Mag.*, vol. 38, pp. 287–289; April, 1947.) The process described in 463 of 1947 (Morris) is here expressed in matrix form. See also 3162 above.

519.27:530.16 3164

Distribution of the Sum of Randomly Phased Components—W. R. Bennett. (*Quart. Appl. Math.*, vol. 5, pp. 385–393; January, 1948.) A method of finding the distribution of the sum of n vectors with given moduli and randomly distributed arguments for values of n of the order of 10. A convergent Fourier-Bessel series is derived. The special case where all the vectors have equal moduli is used to illustrate the method, which is later extended to the general case. The envelope of a group of sine waves is discussed briefly.

MEASUREMENTS AND TEST GEAR

531.761 3165

A Portable Electronic Chronometer—G. T. Baker. (*Jour. Sci. Instr.*, vol. 25, pp. 194–198; June, 1948.) A crystal-controlled chronometer, which measures intervals up to 1 sec with an accuracy of 0.1 ms and can be used by a semi-skilled operator for measuring circuit transients, operation times of relays, etc.

621.3.018.4(083.74) 3166

Two Portable Substandards of Frequency—R. Terlecki and J. W. Whitehead. (*Jour. Sci. Instr.*, vol. 25, pp. 237–239; July, 1948.) Details of two instruments, using crystal-controlled multivibrators, which give spot frequencies (a) 10 kc, (b) 1 kc apart over the range 10 kc to 35 Mc. Check points are available at intervals of 1 Mc and 5 Mc and in the second instrument the output may be modulated.

621.3.018.4(083.74) 3167

Portable Crystal-Controlled Frequency Standard—R. I. B. Cooper. (*Elec. Commun.*, vol. 25, pp. 30–34; March, 1948.) Illustrations and performance details of a substandard using a 100-kc GT-cut crystal with plated electrodes, mounted in an evacuated bulb.

621.317.3:621.392.029.64 3168

On the Representation and Measurement of Waveguide Discontinuities—Marcuvitz. (*See* 3020.)

621.317.3.011.5 3169

The Plate Method for Determining Dielectric Constants and Loss Angles—I. A. El'tsin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 657–664; May, 1948. In Russian.) A rigorous theory is given of a method in which a thin plate of the material under investigation is placed at the middle of a tuned Lecher system. Formulas are derived for calculating, from measurements at λ 3.36 m, the dielectric constants and loss angles for various solid dielectrics.

621.317.336 3170

Accuracy of Impedance Measurements—B. Secker. (*Elec. Commun.*, vol. 25, pp. 74–83; March, 1948.) Impedance-unbalance and precise impedance measurement systems are considered. Sources of errors in bridges of various types are discussed and comparisons made between different bridges for particular measurements. Curves show the limits of measurement of series- and parallel-resonance bridges; the fundamental balance conditions for different types of bridge are derived.

621.317.35 3171

A New Method in the Analysis of Complex Electric Waves—W. E. Rogers (*Rev. Sci. Instr.*, vol. 19, pp. 332–335; May, 1948.) A potentiometer method for direct measurement of the phase and amplitude of the harmonic components of complex waves. A known complex wave form having a periodic relationship with the wave to be analyzed is used as a reference and both waves are fed to a tunable, frequency-selective detector which acts as a null indicator. Comparison is made by adjusting the amplitude and phase of the reference wave.

621.317.44:538.632:546:289 3172

A Magnetic Field Strength Meter Employing the Hall Effect in Germanium—G. L. Pearson. (*Rev. Sci. Instr.*, vol. 19, pp. 263–265; April, 1948.) The essential components include a small Ge probe and a microammeter which is calibrated directly in gauss. Accuracy is within 2 per cent for fields between 100 and 8000 gauss. At higher field strengths the readings are too low, the error being about 9 per cent at 20,000 gauss. Advantages are (a) small size and portability, (b) steady reading, (c) small nonmagnetic probe which can be used in very narrow gaps.

621.317.7 3173

New Long-Scale Instruments—R. M. Rowell and N. P. Millar. (*Gen. Elec. Rev.*, vol. 51, pp. 14–19; April, 1948.) Construction details of dc and ac ammeters and voltmeters, wattmeters, power-factor meters, and frequency meters with pointer movement of 250°.

621.317.715 3174

The Construction of Micro-Galvanometer Systems—A. C. Downing. (*Jour. Sci. Instr.*, vol. 25, pp. 230–231; July, 1948.) Systems of period 0.01 second, resistance 20Ω, 0.3 to 0.5 mm wide and weighing 3 to 5 mg are described, with methods of winding and inserting connecting tags. See also 3175 below.

621.317.715 3175

Moving-Coil Galvanometers of Short Period and their Amplification—A. V. Hill. (*Jour. Sci. Instr.*, vol. 25, pp. 225–229; July, 1948.) The design and performance of a short-period galvanometer and photoelectric amplifier are discussed. Potential changes of a few microvolts in a low-resistance circuit can be measured to within 0.5 per cent. See also 3174 above.

621.317.72:621.392.43 3176

A Modified Micromatch—D. N. Corfield and C. W. Cragg. (*RSGB Bull.*, vol. 23, pp. 211–213; May, 1948.) An instrument similar to that described in 2853 and 3188 of 1947 (Jones and Sontheimer) but suitable for use at frequencies up to 60 Mc. Details of its construction and operation, and the selection of suitable components, are discussed.

621.317.723 3177

An Electrometer Tube Amplifier Circuit—E. Lindholm and E. Hullegård. (*Ark. Mat. Astr. Fys.*, vol. 34, part 4, section B, 5 pp.; April 13, 1948. In English.) A 2-tube circuit which gives high stability and uses ordinary T114 tubes. The same battery is used for filament current and anode voltage.

621.317.725 3178

Valve Voltmeter and Galvanometer with D.C. Amplifier—R. L. Schupp and R. Mecke. (*Funk und Ton*, vol. 2, pp. 285–289; June, 1948.) A 3-stage amplifier with push-pull output stage is described which uses two Type EDD11 double triodes. Stabilized supply voltages give good zero stability. With a 4000-Ω output meter the voltage sensitivity is 0.1 mv.

621.317.729:537.58:621.385.1 3179

Representation in the Electrolyte Tank of the Effect of Space Charge in Valves—R. Musson-Genon. (*Onde Élec.*, vol. 28, pp. 236–242; June, 1948.) An account of a method for determining the distribution of potential, taking account of the space charge $\Delta U_{(x,y)} = -4\pi\rho$. The bottom of the tank is suitably contoured. The method involves successive approximations which converge very rapidly, it can be extended to the study of the functions $\Delta\phi_{(x,y)} = f(x, y, \phi, \ldots)$. Examples of its use are given.

621.317.733 3180

Remarks on A. C. Wheatstone Bridges—F. Perrier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1806–1808; May 31, 1948.) Simple theory shows that maximum sensitivity is obtained when the bridge arm between the source and the impedance *Z* to be measured consists of an impedance whose modulus is equal to that of *Z* and whose argument differs from that of *Z* as much as possible.

621.317.733 3181

Capacitance Bridge with Mechanical Rectifier and Moving-Coil Galvanometer as Indicator—F. Koppelmann. (*Frequenz*, vol. 2, pp. 100–105; April, 1948.) Theory and description of a practical bridge of the Schering type. A mirror galvanometer gives greater sensitivity than is obtained with a vibration galvanometer as indicator. A disadvantage is the sensitivity to harmonics. The disturbing effect of any single harmonic can be eliminated by adjustment of the rectifier contacts, but other harmonics may still mask the balance to some extent.

621.317.738 3182

Capacitance Meter with Neon-Lamp Indicator—L. Grillet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 226, pp. 1968–1969; June 14, 1948.) Details of a method particularly suitable for the comparison of capacitances less than 1000 pF. The resistor usually fitted in the neon lamp socket is removed and the lamp then serves as a very sensitive current indicator, becoming luminous for currents much less than 1 μamp.

621.317.75.015.3:621.395.625.6 3183

Sweeping Device for the Display of Transient Phenomena and Nonlinear Distortion—W. Meyer-Eppler. (*Arch. Elek.* (Übertragung), vol. 2, pp. 1–14; January, 1948.) A photomechanical projection apparatus for either visual observation or photographic recording, and particularly suitable for use with sound films. See also 2854 of 1942.

621.317.755 3184

Multigun C.R. Oscillography—H. S. Bamford. (*Electronic Ind.*, vol. 2, pp. 10–13; May, 1948.) A description of cathode-ray tubes having up to ten complete gun and deflecting systems mounted inside the same envelope. The methods of shielding each assembly and connecting the deflecting plates by short leads to terminals spaced round the neck of the tube are illustrated. Individual assemblies can have either standard orthogonal deflecting systems or a radial deflecting system with two conical electrodes. Possible applications in the fields of bio-electricity and color television are discussed.

621.317.755:531.767 3185

Measurement of [c.r.] Spot Trace Speed—C. Besle. (*Rev. Gén. Élec.*, vol. 57, pp. 189–192; May, 1948.) A method is described in which the spot is made to describe a spiral with constant angular velocity. The maximum recording speed is found by noting, on the photographic record of the spiral, the point where the spot ceases to be visible.

621.317.761 3186

Direct-Reading Superheterodyne Frequency Meter—L. M. Berman. (*Toute la Radio*, vol. 15, pp. 176–178; June, 1948.) Details of an instrument, constructed by the Laboratoires Radioélectriques, which has definite advantages as regards both sensitivity and selectivity over wavemeters using direct amplification. Ranges are 550 kc to 5 Mc and 5 Mc to 30 Mc. The absolute accuracy of the quartz crystal reference frequency is within 1 part in 10^6 and the possible error in reading the difference frequency directly can be reduced to 1 cps. A signal can be measured in the presence of an interfering signal of equal strength only 100 cps away. If the interfering signal is about 50 times as strong as the signal to be measured, a separation of 400 cps is necessary.

621.317.761:621.385.38 3187

Study of the Discharge of a Capacitor through a Thyratron. Application to the Study of the Operation of a Direct-Reading Valve Frequency Meter—R. Legros. (*Rev. Gén. Élec.*, vol. 57, pp. 193–200; May, 1948.) The operational characteristics of thyratrons are reviewed; graphs show the theoretical variations of thyratron current and voltage during the discharge. An expression is derived for the mean current registered by a thyratron/capacitor type of frequency meter. Experimental data are in agreement with theory.

621.317.78 3188

Method of Measuring H.F. Power—L. Liot. (*Radio Franç.*, pp. 21–25; May, 1948.) Two methods are described, one using lamps with short straight filaments, the other using thermojunctions. One lamp is connected to a voltage source derived from the circuit to be measured; the second is fed from a dc source, which is adjusted till the two filaments are equally bright. An accuracy within about 10 per cent can be obtained at frequencies up to 1000 Mc for mean powers of about 20w. The thermojunction method gives comparable results up to frequencies of the order of 300 Mc.

621.317.79:621.396.822 3189

A Noise Meter for Broadcasting Stations—M. A. Slutsker and M. A. Studitski. (*Vestnik Svyasi*, no. 6, pp. 12–13; 1948. In Russian.)

621.317.79:621.396.97 3190

CBS Transmission Measuring Set—D. F. Maxwell. (*Audio Eng.*, vol. 32, pp. 16–19, 46; April, 1948.) A new instrument designed for precision af testing in broadcast service. The set is a combination of calibrated attenuators, matching devices, and power-level indicators. Response/frequency measurements are accurate to within 0.1 db. Input and output power levels are given to within 0.2 db.

621.396.615.17 3191

A Precision Double-Pulse Generator—D. J. Medley and H. D. Rathgeber. (*Jour. Sci. Inst.*, vol. 25, pp. 234–236; July, 1948.) Description of an instrument for producing either single or repeated pairs of electrical pulses, similar to Geiger-counter discharges, separated by a time interval which is continuously adjustable from 0 to 74 μ seconds.

621.396.822:621.385.2 3192

[Diode] Noise Generator for Receiver Measurements—P. G. Sulzer. (*Electronics*, vol. 21, pp. 96–98; July, 1948.) The receiver bandwidth need not be known and measurement is independent of the response curve of the receiver.

621.396.822:621.396.621 3193

The Estimation and Measurement of the Sensitivity of Radio Receivers in Terms of *kT* Units—Slepyan. (*See* 3232.)

621.317.79:621.396.615 3194

Test Gear for Frequency Modulation and Television [Book Notice]—B.I.O.S. Final Report No. 1269 and addendum. H.M. Stationery Office, London, February 6, 1947, 22 and 3 pp., 3s.6d. and 1s. Signal generators were the only forms of test equipment designed in Germany specifically for FM; they were conventional in design, as were the methods of modulation and demodulation. The addendum describes a test oscillator which may be swept cyclically over three frequency bands (0 to 8, 6 to 14, and 12 to 20 Mc) in order to measure the if characteristics of television receivers, the frequency response being displayed on a cathode-ray tube.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9.001.8:614.8 3195

Notes on Using High-Power Ultrasonics—S. Y. White. (*Audio Eng.*, vol. 32, pp. 26–27, 37; April, 1948.) Discussion of possible dangers to health in high-intensity ultrasonic fields, and of the conditions necessary for ultrasonic coagulation of fine particles in air.

535.33:535.61-15 3196

Electronic Eyepiece for Spectroscopy of Near Infra-Red—Z. V. Harvalik. (*Rev. Sci. Instr.*, vol. 19, pp. 254–257; April, 1948.)

539.16.08 3197

Self-Quenching Halogen-Filled Counters—S. H. Liebson and H. Friedman. (*Rev. Sci. Instr.*, vol. 19, pp. 303–306; May, 1948.) Counters filled with inert gases containing small amounts of one of the halogens are selfquenching and have an apparently unlimited life. Their characteristics are similar to those of counters with argon-alcohol filling.

539.16.08 3198

The Effect of the Composition of the Gas Mixture in Self-Quenching Geiger-Müller Tubes on Their Plateau Characteristics—S. J. du Toit. (*Phys. Rev.*, vol. 73, p. 1473; June 15, 1948.)

59.16.08 3199

Geometric Factors Underlying Coincidence Counting with Geiger Counters—H. E. Newell, Jr. (*Rev. Sci. Instr.*, vol. 19, pp. 384–389; June, 1948.) The coincidence rate of a perfectly efficient 2-counter telescope embedded in an isotropic field of radiation is obtained in terms of the radiation intensity and the effective dimensions and separation of the two counters. The formulas determine the counting rate within limits which in most practical cases differ by only a few per cent.

539.16.08 3200

Characteristics of the Parallel-Plate Counter—L. Madansky and R. W. Pidd. (*Phys. Rev.*, vol. 73, pp. 1215–1216; May 15, 1948.)

615.84:616.853 3201

The Electroshock—B. Roger. (*Radio Franç.*, pp. 37–40; May, 1948.) Some details of apparatus for producing shocks of various intensities for therapeutic purposes, particularly the treatment of epilepsy.

621.365:621.385.1 3202

Tube Trends in Field of Industrial Heating—Doolittle and Steinberg. (*See* 3278.)

621.38.001.8:786.6 3203

Design of Electronic Organs: Part 3—W. Wells. (*Audio Eng.*, vol. 32, pp. 24–25, 40; April, 1948.) A detailed discussion of the Hammond organ. The tone generator assembly contains 91 phonic wheels, gear-driven from a synchronous motor operated from ac mains, so that the instrument is always in tune. Manual and pedal contact troubles are avoided by using Pt-Ir and Pd for the switch contacts. The switch assembly is sealed in a metal housing, and requires neither cleaning nor adjustment. Novel tone controls are provided for addition of harmonics. See also 2583 of October (Long).

621.384.6+538.56+621.385.029.63/.64 3204

Waves and Electrons Traveling Together—A Comparison between Traveling Wave Tubes and Linear Accelerators—Brillouin. (*See* 3093.)

621.384.6 3205

Effect of the Electron Beam on the Voltage Distribution of a High-Voltage Multi-Stage Electron Accelerator—F. W. Waterton. (*Nature* (London), vol. 161, pp. 563–564; April 10, 1948.)

621.384.6 3206

A New Method for Displacing the Electron Beam in a Betatron (Synchrotron)—R. Wideröe. (*Rev. Sci. Instr.*, vol. 19, pp. 401–402; June, 1948.) Description of a method proposed in 1945 similar to that described in 832 of 1947 (Clark, Getting, and Thomas), and of an improvement designed to reduce orbital instability.

621.384.6 3207

Investigations on a 15-MeV Betatron—R. Kollath and G. Schumann. (*Z. Naturf.*, vol. 2a, pp. 634–642; November and December, 1947.) An account of its construction and operation, and of tests regarding its X-ray output.

621.385.833 3208

Summarized Proceedings of Conference on Electron Microscopy—Leeds, September, 1947—V. E. Cosslett. (*Jour. Sci. Instr.*, vol. 25, pp. 167–170; May, 1948.) Discussion of (a) biological applications, (b) new instruments and technical methods. See also 1713 of July (Reed).

621.385.833 3209

The French Electrostatic Microscope—P. Grivet. (*Ann. Radioélec.*, vol. 3, pp. 144–145; April, 1948.) A review of its development and of relevant theory.

621.385.833:535.317.25 3210

A New Microscopic Principle—D. Gabor (*Nature* (London), vol. 161, pp. 777–778; May 15, 1948.) A note giving a broad explanation of the principle which may enable the resolving power of electron microscopes, at present limited by spherical aberration, to be increased by dispensing with the objective. Micrographs are obtained by electronic analysis followed by optical synthesis. The principle is used in the "X-ray microscope," but can be applied much more generally. It has been tested by means of an optical model.

621.391.63:526.9 3211

Surveying with Pulsed-Light Radar—W. W. Hansen. (*Electronics*, vol. 21, pp. 76–79; July, 1948.) An adaptation of radar techniques for surveying over inaccessible terrain. Light pulses from a flash lamp at one site are reflected back from the remote site by a system of mirrors. The reflected pulses operate a photo cell and produce pips on the trace of a cro. Distances and angles can be measured with an accuracy comparable to that of a third-class survey.

621.396.9 3212

Apparatus for Finding Pieces of Iron in Timber—J. Hacks. (*Z. Angew. Phys.*, vol. 1, pp. 11–19; January, 1948.) A permanent magnet carrying a multiturn search coil and compensating coil is used. Any iron near the search coil causes a variation in the pitch of the note given by an audio beat-frequency generator. A 2-cm shot can be detected at a distance of 25 cm and a 2-mm iron nut at 5.5 cm.

PROPAGATION OF WAVES

538.566 3213

Calculation of the Potential from the Asymptotic Phase—C. E. Fröberg. (*Ark. Mat. Astr. Fys.*, vol. 34, part 4, section A, 16 pp.; April 13, 1948. In English.) Full paper. Summary abstracted in 223 of February.

621.396.11 3214

Investigations on Short-Wave Echo Signals: Part 1—H. A. Hess. (*Z. Naturf.*, vol. 1, pp. 499–505; September, 1946.) Measurements for frequencies in the range 10 to 20 Mc give a value of 0.137788 second for the time of travel round the earth. This value appears to be independent of frequency, time of day and time of year, and for distances of over 1000 km gives distances correct to within about 25 km. The results in general favor the theory of guided propagation and not the theory of zigzag reflection between the ionosphere and earth. For part 2 see 3215 below. See also 2889 of November.

621.396.11 3215

Investigations on Short-Wave Echo Signals: Part 2—H. A. Hess. (*Z. Naturf.*, vol. 2a, pp. 528–534; September, 1947.) An account of investigations for distances between transmitter and receiver less than 1000 km, where multipath effects are pronounced. These may give rise to path time differences of several milliseconds. The strength of the interfering echoes was found to depend on the directional characteristics of the receiving antenna and on the polarization of the incoming waves. The results are discussed. Part 1: 3214 above.

621.396.11 3216

Long-Distance Propagation of Short Waves—L. Hamberger and K. Rawer: H. A. Hess. (*Z. Naturf.*, vol. 2a, pp. 521–527; September, 1947.) Hamberger and Rawer consider that the approximate constancy of the time taken for short-wave propagation right round the earth can be explained, contrary to the views of Hess (3214 above), by zigzag reflection between the ionosphere and the earth. The values given by Hess would correspond to 12 to 17 reflections.

Hess points out that the observed angle of arrival of waves which have traveled round the earth does not favor the zigzag reflection theory.

621.396.11:523.746 3217

Sunspots and Radio Weather—Arzinger, Hallborg, and Nelson. (*See* 3106.)

621.396.11:551.510.535 3218

Upper-Atmosphere Circulation as Indicated by Drifting and Dissipation of Intense Sporadic-E Clouds—Ferrell. (*See* 3117.)

621.396.11:551.510.535 3219

A Study of the Interaction of Radio Waves—J. A. Ratcliffe and I. J. Shaw. (*Proc. Roy. Soc. A*, vol. 193, pp. 311–343; July 2, 1948.) Experiments have been carried out to provide information about the mechanism by which modulation can be transferred from one wave to another during transmission through the ionosphere. Earlier theories, indicating that this phenomenon is due to nonlinear absorption, are restated in current nomenclature and are shown to be confirmed by the experimental data. The results provide a measure of the heights at which absorption takes place at different frequencies and of the electron collision frequency in the absorbing regions.

621.396.11:551.510.535 3220
Restricted-Range Sky-Wave Transmission—J. E. Hacke, Jr., and A. H. Waynick. (Proc. I.R.E., vol. 36, pp. 787–793; June, 1948.) 2.4-Mc signals from an omnidirectional antenna can normally be received by reflection from the *E* layer at ranges up to 2000 km. By designing an antenna array to give increased vertical directivity, the maximum range can be reduced to 500 km. The limitations and advantages of the system are discussed.

621.396.11:551.510.535 3221
Triple Splitting of Ionospheric Rays—J. H. Meek. (*Nature* (London), vol. 161, p. 597; April 17, 1948.) The theory of ionospheric reflection of medium-frequency and high-frequency radio waves at vertical incidence indicates the existence of a triple splitting effect in polar regions, due to the nearly vertical magnetic field of the earth. In Canada a complete trace is often seen extending from the *E* region through the F_1 region to the F_2 region. A sample record is given showing observed characteristics of the third split. See also 3222 below.

621.396.11:551.510.535 3222
Triple Splitting of Ionospheric Rays—T. L. Eckersley. (*Nature* (London), vol. 161, pp. 597–598; April 17, 1948.) The triple splitting of rays reflected from the *F* region is thought, on the basis of the observed polarization, to be due to coupling between the ordinary and extraordinary rays. See also 2595 of October (Newstead) and 3221 above.

621.396.812 3223
On the Wave-Guiding Properties of Non-Uniform Media—P. E. Krasnushkin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 431–446; April, 1948. In Russian.) By using the method of normal modes, a complete picture of the wave propagation beyond the horizon in the case of a nonuniform troposphere and ionosphere can be given. The method consists in representing the wave field in the form of a spectrum of normal modes, the discontinuous portion of which is analogous to the spectrum of normal waves in a hollow pipe. An exact solution of the problem is given for the case of a plane multiplayer medium and an approximate solution for the case of a spherical multilayer medium. See also 516 of 1947 (Booker), 2892 of 1947 (Booker and Walkinshaw) and 2211 of 1947 and 224 of February (Pekeris).

621.397.812.029.62/.63 3224
Comparative Propagation Measurements: Television Transmitters at 67.25, 288, 510, and 910 Megacycles—G. H. Brown, J. Epstein, and D. W. Peterson. (*RCA Rev.*, vol. 9, pp. 177–201; June, 1948.) The transmitting antennas were near to or at the top of the Empire State Building. Two 50-mile paths were chosen, one over hilly terrain with maximum elevation 1200 feet and the other over fairly flat terrain with no hills above 230 feet. The receiver antennas were mobile, and 30 feet above ground. Measurements were made at 2-mile intervals along each path.

Theoretical and experimental field-strength values agreed more closely for the flat path than the hilly one, and more closely for 67.25 Mc than for 288 Mc. The field strengths for 510 and 910 Mc were usually far below the theoretical values.

Shadowing from obstacles increased steadily with frequency and was severe at 910 Mc. Curves are given from which television transmitter power requirements can be estimated for different frequencies.

Multipath effects were slight at 67.25 and 288 Mc, but severe in obstructed areas at 510 and 910 Mc. A clean picture could usually be obtained by suitable orientation of the receiver antenna.

621.396.812.029.64 3225
Results of Horizontal Microwave Angle-of-Arrival Measurements by the Phase-Difference Method—A. W. Straiton and J. R. Gerhardt. (Proc. I.R.E., vol. 36, pp. 916–922; July, 1948.) The measurements were made on λ 3.2 cm over a 7-mile path lying along a shore line in the Gulf of Mexico. Small deviations of the angle of arrival (of the order of 0.02°) in a a landward direction from the geometric path were very frequently noted. Meteorological soundings showed overwater ducts to be present nearly all the time and there was a general correlation between the angular deviations and the horizontal gradient of radio refractive index. See also 1182 of 1947 (Sharpless) and 1183 of 1947 (Crawford and Sharpless).

621.396.11:551.510.535 3226
Radio Research Special Report No. 17. Fundamental Principles of Ionospheric Transmission [Book Notice]—H. M. Stationery Office, 1s.6d. (*Govt. Publ.* (London), p. 18; June, 1948.) Joint publication of the Department of Scientific and Industrial Research and the Admiralty.

RECEPTION

621.396.621 3227
Converting the 1147B for 50–250 Mc/s Operation—B. W. St. L. Montague. (*RSGB Bull.*, vol. 23, pp. 214–216, 218; May, 1948.) Details of the necessary modifications of the RAF receiver R.1147 B, which include rewinding the if transformers and oscillator coils. Provision is made for the addition of an FM discriminator, which will be described later.

621.396.621 3228
The Marconi High Discrimination Communication Receiver, Type R.G. 44—L. R. Mullin. (*Marconi Rev.*, vol. 11, pp. 33–38; April to June, 1948.) A receiver designed for war-time mass production. The printed tuning scale forms a spiral on a drum, and is calibrated at 10-kc intervals between 2 and 20 Mc. Mechanical and electrical features are discussed, including methods of temperature compensation for the oscillator and an avc system having improved noise characteristics.

621.396.621 3229
A High-Fidelity Receiver—L. Chrétien. (*TSF Pour Tous*, vol. 24, pp. 84–86, 127–129, and 152–154; April to June, 1948.) Continuation of 2336 of September. Discusses high-frequency amplification, coupling circuits, and frequency changing.

621.396.621 3230
Single-Signal Single-Sideband Adaptor—E. W. Rosentreter. (*Electronics*, vol. 21, pp. 124, 143; July, 1948.) Full circuit details are given for the General Electric single-sideband selector. Principles of operation are discussed, with reference to the work of Villard (2597 of October) and Dome (1021 of 1947). The selector unit is connected to the last if stage of an existing AM receiver by means of a small probe and a short length of low-capacitance shielded cable.

621.396.621:06.064 Paris 3231
French Receiver Construction at the Paris Fair, 1948—G. Giniaux and J. Rousseau. (*TSF Pour Tous*, vol. 24, pp. 159–164; June, 1948.) A general account, with a table giving particulars of about 130 receivers, and also a discussion of receivers with special characteristics.

621.396.621:621.396.822 3232
The Estimation and Measurement of the Sensitivity of Radio Receivers in Terms of *kT* Units—L. B. Slepyan. (*Radiotekhnika* (Moscow), vol. 3, pp. 3–10; March and April, 1948. In Russian.) Starting from the well-known Nyquist formula (1) determining the signal level in an equivalent antenna circuit necessary for producing the same effect at the output of the receiver as that due to noise originating in it, formulas are derived for calculating the sensitivity of a receiver in *kT* energy units, *k* being Boltmann's constant and *T* the absolute temperature. The difference between the coefficient of insensitivity *D* (formula 10) and noise factor *F* (formula 12) is established and methods are indicated for measuring them. See also 1656 and 2337 of 1942 (North).

621.396.621.53:621.385.2 3233
The Fundamental Relationships in the Diode Frequency Changer—L. S. Gutkin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 615–638; May, 1948. In Russian.) A theoretical analysis of a diode mixer stage is given, neglecting the effects of parasitic inductance, capacitance, etc. The various relationships determining the operation of the stage are derived and conditions are established for obtaining the maximum amplification when the diode is loaded with one- or two-section filters. The effect of noise is investigated in detail. Formulas derived by Strutt (1573 of 1947) are criticized.

621.396.621.54:621.316.72 3234
On the Automatic Stabilization of the Amplification of a Superregenerative Receiver for the Reception of Pulse Signals—M. K. Belkin. (*Radiotekhnika* (Moscow), vol. 3, pp. 25–35; May and June, 1948. In Russian.) The theory of the receiver is discussed and the necessity of automatic stabilization is pointed out. The total amplification *N* of the receiver is equal to $N_p\eta$ where N_p is the gain due to regeneration and η the gain given by superregeneration as compared to regeneration. An experimental investigation in which stabilization of the receiver was effected by controlling N_p or η is discussed. The second method is preferable but neither method is quite satisfactory. It is suggested that stabilization within the circuit of the receiver should be combined with that of the power supplies.

621.396.622 3235
Theory of Frequency Counting and Its Application to the Detection of Frequency-Modulated Waves—É. Labin. (Proc. I.R.E., vol. 36, pp. 828–839; July, 1948.) A theoreitcal study of the operation of electronic circuits which produce an output proportional to the frequency of the input signal. If the input signal is modulated, the output should reproduce the modulation. As a detector, the frequency counter has low over-all sensitivity, but this appears to be unimportant; it is much more rugged than ordinary "differential" discriminators, and does not require close tolerances, or adjustment when installed in a receiver.

621.396.662 3236
The Browning RV-10 F.M. Tuner—F. A. Spindell. (*FM and Telev.*, vol. 8, pp. 37–40, 59; May, 1948.) Circuit details, performance characteristics and alignment procedure for a straight tuner using the Armstrong circuit.

621.396.82 3237
Interference between Very-High-Frequency Radio Communication Circuits—W. R. Young, Jr. (Proc. I.R.E., vol. 36, pp. 923–930; July, 1948.) Various common causes of interference are discussed and sample measurements are quoted to illustrate their relative magnitudes. Formulas are given for computing the frequency of the disturbances. A method is described for making charts, suitable for a given type of equipment, from which spurious frequencies can be read directly as a function of the operating frequency.

621.396.822 3238

Some General Results in the Theory of Noise through Non-Linear Devices—D. Middleton. (*Quart. Appl. Math.*, vol. 5, pp. 445–498; January, 1948.) The Fourier series method of Rice (2169 of 1945 and back references) is applied to the following unsolved problems; (a) Passage of a modulaated signal in the presence of noise through a general nonlinear apparatus, with a sinusoidally modulated carrier, or with narrow-band noise, symmetrically distributed in frequency about the carrier. (b) The biased μ th-law rectifier, for modulated and unmodulated carriers; limiting cases of large noise or signal voltages are also discussed. (c) The case of a modulated signal and narrow-band noise, with a determination of the various probability densities associated with the envelope of the wave. (d) The correlation function and mean power associated with the envelope of signal and noise. The low-frequency output of the half-wave μ th-law device is considered. (e) The μ th-law, half-wave rectification of noise alone. This is relevant to the measurement of noise and to the detection of pulse signals in the presence of noise. (f) A general "small-signal" theory in which the peak values of the incoming wave, whether noise or signal plus noise, are so small that overloading and cutoff do not occur.

621.396.822 3239

Thermal Noise in Resistors—S. Rodda: D. A. Bell. Correction to 2072 of July. The paper by Bell there mentioned should have been 2780 of 1938. See also 1914 of 1939 (Moullin).

621.396.822:523.72+523.16 3240

Radio Noise of Extra-Terrestrial Origin and its Effect on Telecommunication Technique—Lehmann. (*See* 3102.)

STATIONS AND COMMUNICATION SYSTEMS

534.861.1:621.316.345:621.396.664 3241

Modern Design Features of CBS Studio Audio Facilities—R. B. Monroe and C. A. Palmquist. (Proc. I.R.E., vol. 36, pp. 778–786; June, 1948). Description of a small, space-saving, single-unit console which is easily accessible for maintenance. See also 2316 of 1946 (Chinn).

534.861.1:621.396.712.3 3242

Speech-Input Equipment for New Oslo Broadcasting House—Julsrud and Weider. (*See* 3008).

621.39 3243

Telecommunication Services for the Fifth Olympic Winter Sports at St. Moritz, from 30th January to 8th February 1948—A. Wettstein. (*Tech. Mitt. Schweiz. Telegr.-Teleph-Verw.*, vol. 26, pp. 99–115; June 1, 1948. In German and French.) Details of the general arrangements, with particular reference to telephone facilities.

621.391.63:534.321.9 3244

Distortion in Light Modulation by an Ultrasonic Cell—Sette. (*See* 3000.)

621.394.441 3245

A Multi-Channel Carrier Telegraph System—A. L. Matte. (*Bell Sys. Tech. Publ. Monogr.* B-1529 4 pp; *Railway Signaling*, vol. 40, pp. 778–781; December, 1947.) Description of the 40AC1 voice-frequency system, which is specifically designed to meet railway requirements and will provide 12 duplex or simplex telegraphy channels on a 4-wire circuit, or 6 on a 2-wire circuit.

621.396.41 3246

Theoretical Analysis of Various Systems of Multiplex Transmission—V. D. Landon. (*RCA Rev.*, vol. 9, pp. 287–351; June, 1948.) A systematic method of classifying and specifying such systems is discussed. Definitions of 77 associated terms are included. The basic types of frequency division, time division, and triple modulation systems are considered in detail. Formulas for the signal-to-noise ratio for each system relative to that for single-channel AM are obtained and tabulated; due allowance is made for pre-emphasis. More detailed discussion of the relative merits of the systems will be given later.

621.396.41.029.64 3247

A Duplex System of Communications for Microwaves—R. V. Pound. (Proc. I.R.E., vol. 36, pp. 840–844; July, 1948.) A single oscillator is used both as transmitter and as beating oscillator of a superheterodyne receiver. The oscillator frequency is stabilized at the frequency of a high-Q cavity; FM takes place about this stabilization frequency. Duplex communication can easily be arranged by this method and an experimental set is described; the operator initiating communication can ascertain that his signals are being received. The application of the system to ground/aircraft communication and to a booster system for a relay link is discussed. See also 1690 of 1947 and 1311 of June.

621.396.61/.62].029.63 3248

Practical Experiments on 2350 Megacycles—N. T. J. Bevan and L. Grimshaw. (*RSGB Bull.*, vol. 24, pp. 2–4; July, 1948.) Construction and brief performance details of the transceivers used. An oscillator consisting of a type CV90 tube with the modified cavity of a Sutton tube forms the basis of the transmitter. Provision is made for radio-telephone or mcw operation. The associated receiver is superregenerative.

621.396.619.16(083.72) 3249

Standardization of Nomenclature for Pulse Modulation—H. H. Heeroma. (Proc. I.R.E. vol. 36, p. 880; July, 1948.) Statement of terms proposed by the Netherlands Electrotechnical Committee, for use internationally. Terms favored are: pulse-rate, pulse-width, pulse-position, pulse-height, pulse-slope, and pulse-code modulation. Terms deprecated are pulse-frequency, pulse-length, pulse-phase, pulse-displacement, pulse-time, pulse-delay, and pulse-amplitude modulation.

621.396.65.029.63 3250

S-T [studio-transmitter] Link on 920 to 980 Mc/s—R. H. DeWitt. (*FM and Telev.*, vol. 8, pp. 22–25; May, 1948.) A radio link for use where wire lines are either lacking or uneconomic. Block diagrams, numerous illustrations, and a short description of the equipment are given, with performance details. Corner-reflector antennas are used for line-of-sight distances up to 12 miles. Paraboloids are necessary for greater distances up to the maximum of 35 miles.

621.396.933 3251

V.H.F. for Civil Aircraft—(*Wireless World*, vol. 54, pp. 242–243; July, 1948.) The aircraft set weighs only 12 lb, takes 3.75 amp from a 12-v battery and delivers about 300 mw to the antenna. It comprises two crystal multiplier chains, two rf and frequency-changer chains, a common if amplifier with detector, agc and af stages, and a modulator chain. For airport control, a 5-w rack-mounted installation is provided, comprising two separate transmitters and receivers and their power supplies. All four are crystal-controlled and have provision for remote or local operation. The receiver used for direction-finding is very similar to the communication receiver. Two-way communication has been maintained at ranges up to 33 miles with the aircraft at 2000 feet and up to 70 miles with the aircraft at 10,000 feet.

SUBSIDIARY APPARATUS

621-526:621.396 3252

Servomechanisms in Connection with Radio Problems: Part 1—G. Lehmann. (*Onde Élec.*, vol. 28, pp. 213–217; June, 1948.) A general discussion of a wide variety of applications in radio technique.

621.314.634 3253

Small Selenium Rectifiers—J. J. A. Ploos van Amstel. (*Philips Tech. Rev.*, vol. 9, no. 9, pp. 267–276; 1947 and 1948.) Discussion of: (a) the general properties of Se rectifiers, (b) the construction and measurement of the dynamic characteristics of three types, (c) some applications.

621.316.722 3254

Voltage Regulators of the Shunt Type—W. G. Hoyle. (*Rev. Sci. Instr.*, vol. 19, pp. 244–246; vol. 19, April, 1948.) A general equation is derived, giving the relation between the nominal required input voltage and the maximum shunt regulating current for any variation of the input voltage and for any load. The necessary relation between the shunt regulating current and the output current for maximum electrical efficiency is obtained. For a fixed load, the maximum average efficiency of any shunt regulator is $1-k^{1/2})^2$, where k is the maximum variation, per unit, of the supply voltage.

621.316.722 3255

The Effect of Frequency Variations on the Operation of Ferro-Resonant Voltage Stabilizers—B. V. Belyaev. (*Avtomatika i Telemekhanika*, vol. 9, pp. 59–73; January and February, 1948. In Russian.) The effects of various parameters of stabilizers, such as the inductance value, the capacitance of capacitors and the compensating coil, the value and type of the load resistance, etc., on the output voltage variation with frequency are investigated by a graphical/analytical method. The main conclusions are confirmed by experimental results.

621.316.722.1 3256

Electromagnetic Voltage Stabilizers for Valve Apparatus—W. Geyger. (*Funk und Ton*, vol. 2, pp. 308–314; June, 1948.) Discussion of basic principles, with description, characteristics, and performance of some modern types.

621.316.722.1 3257

An Inductively Coupled Series Tube D.C. High Voltage Regulator—R. Pepinsky and P. Jarmotz. (*Rev. Sci. Instr.*, vol. 19, pp. 247–254; April, 1948.) A stabilizer operating in the range 5 to 50kv for currents up to 50 ma. The series regulator tube, which is at high voltage, is coupled inductively to the feedback amplifier at earth potential. The signal from the amplifier provides AM for a rf oscillator, whose output voltage is fed to a transformer, rectified, filtered, and then applied as a dc correcting signal to the grid of the series tube. Insulation between high-voltage and low-voltage circuits is provided in the rf transformer.

621.319.339 3258

Van de Graaff Electrostatic Generator—J. M. Ferguson, E. W. Webster, and T. E. Calverley. (*Elec. Times*, vol. 113, pp. 575–579; May 13, 1948.) A short, illustrated description.

TELEVISION AND PHOTOTELEGRAPHY

621.397.331:513.3 3259

The Application of Projective Geometry to the Theory of Color Mixture—F. J. Bingley. (Proc. I.R.E., vol. 36, pp. 709–723; June, 1948.) IRE 1948 National Convention paper; summary noted in 2647 of October.

621.397.335 3260
Phasing of Remote TV Signals—R. C. Palmer. (*Communications*, vol. 28, pp. 14–16; April, 1948.) An instrument designed to provide phase synchronization between the vertical synchronization intervals of a remote composite picture signal and the studio synchronizing generator.

621.397.5 3261
Electro-Optical Characteristics of Television Systems: Part 2—Electro-Optical Specifications for Television Systems—O. H. Schade. (*RCA Rev.*, vol. 9, pp. 245–286; June 1948.) Part 1, 2940 of November.

621.397.5:535.88:679.5 3262
Plastic Lenses in Television Projection—Starkie. (*See* 3154.)

621.397.5:778.53 3263
Motion Picture Photography of Television Images—R. M. Fraser. (*RCA Rev.*, vol. 9. pp. 202–217; June, 1948.) A description of the apparatus and methods developed for photographing television cathode-ray tube images. 16-mm film is used because 35-mm film costs more and is subject to rigorous fire regulations.

A ZnS blue-fluorescing screen is desirable for photographic recording. In the experimental equipment considered, a 5-inch tube with a flat screen and aluminized blue phosphor was used. A rf high-voltage supply delivered 29 kv to the second anode of the cathode-ray tube, which was mounted at one end of a lathe bed, with the camera at the other. A 2-inch Eastman Anastigmat F 1.6 lens, with apertures from F 2.0 to F 2.8, was used.

The recordings can be retransmitted satisfactorily.

621.397.6:621.398:629.135 3264
Television Equipment for Aircraft—(*Télév. Franç.*, pp. 8–9; May, 1948.) A short account of the complete equipment for the Roc guided projectile, using the mimo pickup tube, and of the "ring" equipment fitted in United States reconnaissance aircraft. See also 2959 to 2962 of 1947.

621.397.61 (083.74) 3265
RMA Standards—(Proc. I.R.E., vol. 36, pp. 932–938; July, 1948.) Electrical performance standards, definitions, and recommended methods of measurement for television broadcast transmitters at frequencies between 44 and 216 Mc.

621.397.62 3266
Television Receiver with Screen Projection (Philips' Receivers).—(*Radio Franç.*, pp. 42–48; May, 1948.) A general description, with details of the folded Schmidt optical system and of certain special features. The picutre size is 31 cm×41 cm.

621.397.62 3267
"Surplus" Television Receiver—L. J. Dalby. (*Wireless World*, vol. 54, pp. 251–252; July, 1948.) Brief description and circuit details of a simple and cheap television receiver which gives good results. The cathode-ray tube is readily obtainable from war surplus equipment.

621.397.8 3268
Cause and Cure of Spurious TV Receiver Oscillations—R. T. Cavanaugh. (*Tele-Tech*, vol. 7, part 1, pp. 36–37, 79; May, 1948.) Causes of spurious oscillations in pentode output stages of line scanning generators are investigated. The use of a magnet to modify the electron paths within the tube is advocated as a cure.

621.397.812.029.62/.63 3269
Comparative Propagation Measurements; Television Transmitters at 67.25, 288, 510, and 910 megacycles—Brown, Epstein, and Peterson. (*See* 3224.)

TRANSMISSION

621.396.61 3270
Transmitters in Parallel—V. O. Stokes. (*Marconi Rev.*, vol. 11, pp. 39–44; April to June, 1948.) The use of transmitters in parallel and the attendant problems are discussed. Methods of phase adjustment and monitoring are described and details given of a completely automatic system of synchronization. Although the case of two transmitters is discussed, there appears to be no technical objection to the use of more than two in parallel.

621.396.61 3271
A New Approach to Single Sideband—D. E. Norgaard. (*QST*, vol. 32, pp. 36–42; June, 1948.) Discussion of practical methods of generating a single-sideband suppressed-carrier signal without the need for sharp filtering and multiple heterodyning. One of the sidebands is removed by a process in which two audio channels with a constant phase difference of 90° are balanced. See also 1805 of July (Nichols).

621.396.61:621.316.729 3272
Synchronization of Low-Power Transmitters—Chamagne and G. Guyot. (*Télév. Franç.*, Supplement *Électronique*, pp. 12–16; May, 1948.) Details of a system suitable for a network of low-power transmitters serving a limited area, such as a large town and its suburbs. A single pilot frequency of 106.6 kc is used; multipliers convert this to the transmission frequency of 960 kc. The quality of the low-frequency modulation could be improved by using FM instead of AM, but the quality with AM is quite satisfactory for the transmission of news, dance music, etc.

VACUUM TUBES AND THERMIONICS

621.385.029.63/.64+538.56+621.384.6 3273
Waves and Electrons Traveling Together—A Comparison between Traveling Wave Tubes and Linear Accelerators—Brillouin. (*See* 3093.)

621.385.029.63/.64 3274
Small-Signal Analysis of Traveling-Wave Tube—C. Shulman and M. S. Heagy. (*RCA Rev.*, vol. 9, p. 366; June, 1948.) Corrections to 2103 of August.

621.385.1:621.365 3275
Tube Trends in Field of Industrial Heating—H. D. Doolittle and E. B. Steinberg. (*Electronic Ind.*, vol. 2, pp. 4–7; May, 1948.) Industrial operating requirements impose severe conditions on heater tubes. Improvements in tube construction technique are described which give increased life and reliability under such exacting conditions.

621.385.38:621.317.761 3276
Study of the Discharge of a Capacitor through a Thyratron. Application to the Study of the Operation of a Direct-Reading Valve Frequency Meter—Legros. (*See* 3187.)

621.396.615.142.2 3277
Millimetre Wavelengths—(*Wireless World*, vol. 54, p. 258; July, 1948.) The Clarendon Laboratory has designed a reflex klystron tunable over the wavelength range 8 to 9 mm. The volume of the klystron cavity is altered by means of a cam mechanism. With a resonator potential of 2.4 kv and a reflector negative potential of 200 v, 10 to 20 mw of continuous-wave power is obtained.

621.396.622.6:546.28 3278
The Silicon Crystal Detector—(*See* 3059.)

621.385.1 3279
Vacuum Tubes [Book Review]—K. R. Spangenberg. McGraw-Hill, London, 860 pp., 45s. (*Wireless Eng.*, vol. 25, p. 237, July, 1948.) "This important book is designed to give a comprehensive account of vacuum tubes and the physical laws on which their behavior depends. It contains a splendid collection of potential-distribution diagrams, nomograms, and design charts pertaining to electron optics, thermionic receiving tubes and the more recent developments, such as klystrons and magnetron oscillators."

MISCELLANEOUS

016.621.39 3280
Technical Bibliographies—(*Electronics*, vol. 21, p. 128; July, 1948.) A list of unpublished bibliographies compiled by the Special Libraries Association can be obtained from R. H. Hopp, Bettelle Memorial Institute, Columbus, Ohio. Other sources of bibliographical information are the Office of Technical Services United States Department of Commerce, and the Engineering Societies Library, New York, N. Y.

06.051 3281
International Radio Conferences—(*Nature* (London), vol. 161, pp. 695–696; May 1, 1948.) An account of the activities of the Union Radio Scientifique Internationale (URSI) and of the Comité Consultatif International des Radio-Communications (CCIR).

061.3:621.396 3282
Scientific Radio—R. L. Smith-Rose. (*Nature* (London), vol. 161, pp. 793–796; May 22, 1948.) A brief account of the proceedings of the Convention on Scientific Radio organized by the IEE in co-operation with the British National Committee for Scientific Radio. The proceedings were divided into four sessions, corresponding in scope to the four URSI commissions. A preliminary survey was made of the British contribution to be presented at the URSI meeting at Stockholm in July, 1948. Four or five short papers were read at each session, surveying progresses in the past few years in various parts of the field of fundamental radio science.

621.39 3283
High-Frequency, Communications, and Remote Control Engineering—(*Brown Boveri Rev.*, vol. 34, pp. 50–57; January to March, 1947.) Illustrations and a few details of (a) multichannel beam telegraphy and telephony equipment, (b) FM usw radio-telephone equipment for communication with mobile stations, (c) medium- and low-power transmitters for commercial operation, (d) high- and medium-power broadcast transmitters for medium, short, and ultrashort waves, (e) transmitting and special tubes, (f) carrier-current telephony and industrial control equipment.

027 3284
British Sources of Reference and Information [Book Review]—T. Besterman (Ed.). Association of Special Libraries and Information Bureaux, London, 56 pp., 6s. (*Jour. Sci. Instr.*, vol. 25, p. 255; July, 1948.) A description of the organization through which books in Great Britain can be borrowed, whether from inside or outside the country. Brief accounts and select lists of the leading library and book organizations, and a list of indispensable works of reference, are given.

621.396 3285
Radio Data Charts [Book Review]—R. T. Beatty, revised by J. McG. Sowerby. Iliffe and Sons, London, 4th edition, 1947, 93 pp., 7s. 6d. (Proc. I.R.E., vol. 36, p. 637; May, 1948.) A revision of the original collection of abacs published in 1930. "The charts are most useful to a radio receiver designer, but are also commendable for student use since many of the nomograms present a physical picture of what would otherwise be a complex formula difficult to comprehend."

PROCEEDINGS OF THE I.R.E.

(Including WAVES AND ELECTRONS Section)

Published Monthly by
The Institute of Radio Engineers, Inc.

Index to Volume 36—1948

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NOTE

The Journal of The Institute of Radio Engineers is officially known as the PROCEEDINGS OF THE I.R.E. The WAVES AND ELECTRONS section is to be regarded solely as a portion thereof, and not as a separate publication.

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The Institute

The Institute of Radio Engineers serves those interested in radio and allied electronics and electrical-communication fields through the presentation and publication of technical material.

Membership has grown from a few dozen in 1912 to approximately 23,000. There are several grades of membership, depending on the qualifications of the applicant, with dues ranging from $3.00 per year for Students to $15.00 per year for Members, Senior Members, Fellows, and Associates of more than five years' standing.

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In 1939, the name of the PROCEEDINGS of The Institute of Radio Engineers was changed to the PROCEEDINGS OF THE I.R.E. and the size of the magazine was enlarged from six by nine inches to eight and one-half by eleven inches.

Subscriptions

Annual subscription rates for the United States of America, its possessions, and Canada, $18.00; to college and public libraries when ordering direct, $13.50. Other countries, $1.00 additional for postage.

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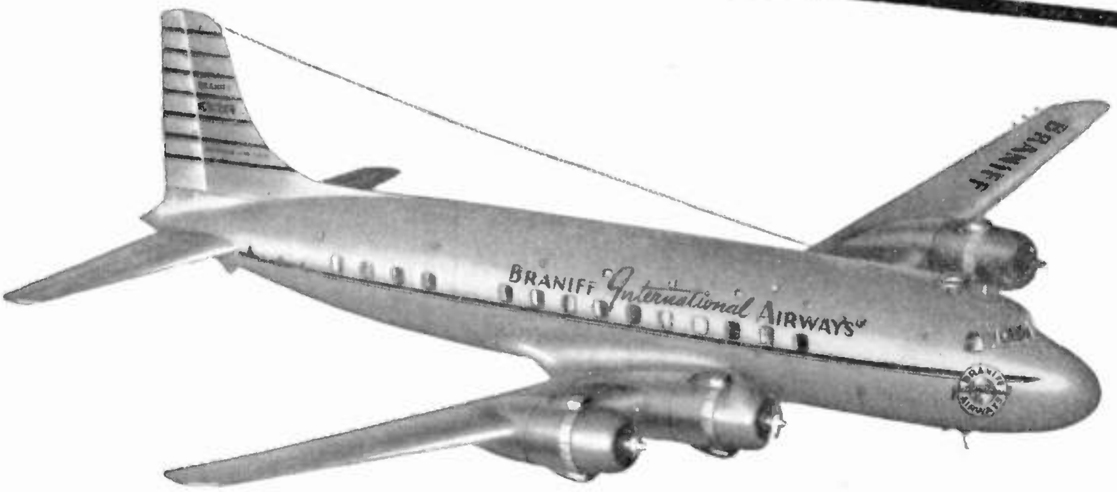

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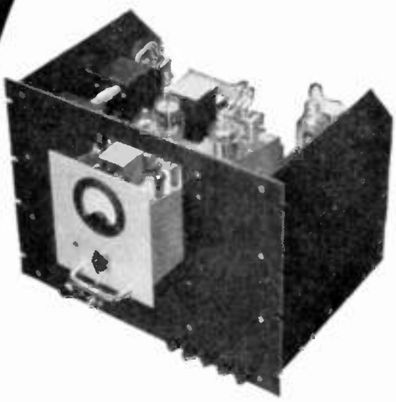

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ATLANTA

"Railroad Radio Communication Equipment," by H. B. Hildreth, Central of Georgia Railroad; June 25, 1948.

"Installation and Equipment of WSB New TV Station," by C. F. Daugherty, Station WSB; September 17, 1948.

BALTIMORE

"Columbia Long-Playing Microgroove Recording System," by P. C. Goldmark, Columbia Broadcasting Company; September 28, 1948.

BUFFALO-NIAGARA

"Horizontal Deflection and High Voltage in Television Receivers," by H. R. Shaw, Colonial Radio Corporation; September 15, 1948.

CLEVELAND

"Television Studio Lighting," by R. Blount, General Electric Company; September 23, 1948.

DALLAS-FORT WORTH

"Some Problems of Television," by B. E. Shackelford, President, The Institute of Radio Engineers; October 11, 1948.

HOUSTON

"Printed Circuits, Subminiature Radio Transmitters, and Receivers," by C. Brunetti, National Bureau of Standards; September 20, 1948.

"Recent Developments in Semi-Conductors and the Transitor," by J. A. Becker, Bell Telephone Laboratory; September 20, 1948.

"Basic and General Problems of Television as a Service," by B. E. Shackelford, President, The Institute of Radio Engineers; October 8, 1948.

KANSAS CITY

"Magnetic Recording," by C. G. Barker, Magnecord, Inc.; September 20, 1948.

NEW MEXICO

"Some Problems of Teleivision," by B. E. Shackelford, President, The Institute of Radio Engineers; October 5, 1948.

NORTH CAROLINA-VIRGINIA

"Detection by Heat Radiation," by N. C. Jamison, Phillips Laboratories; May 14, 1948.

"Electronic Computers," by C. N. Hoyler, Radio Corporation of America; September 10, 1948.

PHILADELPHIA

"A New Long-Playing Disk Recording System," by P. C. Goldmark, Columbia Broadcasting Company; October 7, 1948.

ST. LOUIS

"Sound Level Measurements," by J. L. Glaser, Washington University; September 23, 1948.

SAN ANTONIO

"Radar Reflections from the Lower Atmosphere," by W. E. Gordon, University of Texas; May 26, 1948.

"The Electron Microscope," by L. L. Antes, University of Texas; June 24, 1948.

SAN FRANCISCO

"The Measurement of Faint Light by Photoelectric Methods," by G. E. Kron, Lick Observatory; September 8, 1948.

SEATTLE

"Cathode-Ray Sweep Circuits," by M. A. Starr, University of Portland; June 11, 1948.

"Problems in Development of Television," by B. E. Shackelford, President, The Institute of Radio Engineers; September 24, 1948.

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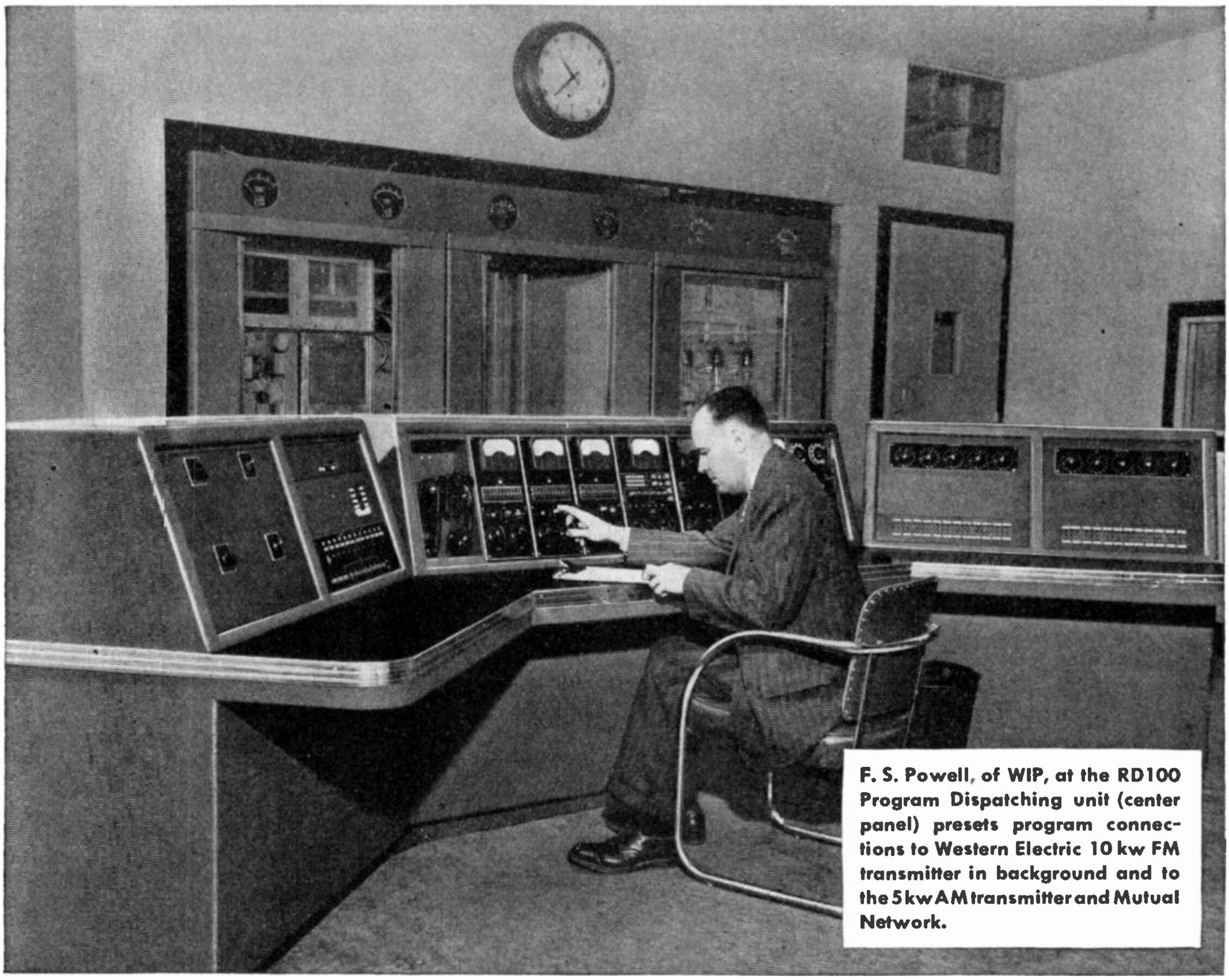

F. S. Powell, of WIP, at the RD100 Program Dispatching unit (center panel) presets program connections to Western Electric 10 kw FM transmitter in background and to the 5 kw AM transmitter and Mutual Network.

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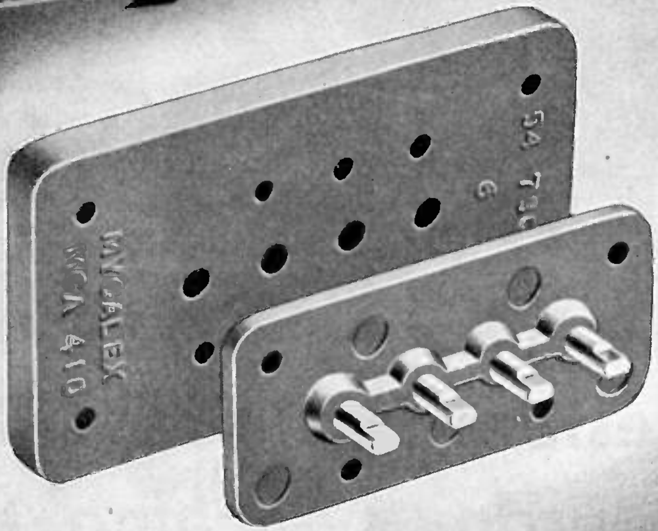

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(*Continued from page 34A*)

SYRACUSE

"Nuclear Power Production," by H. Stevens, General Electric Company; October 7, 1948.

TOLEDO

"Microwave Communication," by J. W. McRae, Bell Telephone Laboratories, Inc.; September 16, 1948.

TORONTO

"The Design and Installation of a 5-Kw High-Channel Television Transmitter," by R. E. Fisk, General Electric Company; October 4, 1948.

TWIN CITIES

"Television—Its Mechanism and Promise," by W. Lawrence, Radio Corporation of America; September 21, 1948.

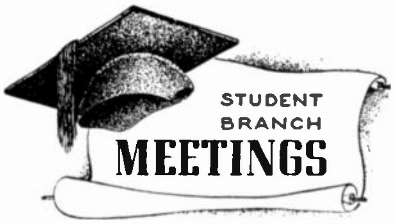
WASHINGTON

"Magnetic Tape Recording," by R. H. Ranger, Rangertone, Inc.; September 20, 1948.

SUB-SECTIONS

NORTHERN NEW JERSEY

"The WATV Television Statión," by T. M. Gluyas, Radio Corporation of America, and F. Bremer, Bremer Broadcasting Company; September 15, 1948.



UNIVERSITY OF ARIZONA—IRE-AIEE BRANCH

Election of Officers; September 15, 1948.

"Grand Coulee Power Project," by J. H. Pfeiffer, Student, University of Arizona; October 6, 1948.

UNIVERSITY OF CALIFORNIA—IRE-AIEE BRANCH

"The Engineering Industry Looks at the College Graduate," by M. P. O'Brien, University of California; October 5, 1948.

"Advantages of Membership in a Professional Society," by J. R. Whinnery, University of California; October 5, 1948.

"The Plans for the New Electrical Engineering Building," by T. C. McFarland, University of California; October 5, 1948.

CARNEGIE INSTITUTE OF TECHNOLOGY—
IRE-AIEE BRANCH

"Carnegie Tech Cyclotron," by H. E. DeBolt and L. Johnson, Graduate Students, Carnegie Institute of Technology; October 11, 1948.

CITY COLLEGE OF NEW YORK—IRE BRANCH

"Student Activity in the Institute of Radio Engineers," by E. K. Gannett, The Institute of Radio Engineers; September 28, 1948.

"Co-ordination of Student Branches in the Metropolitan Area," by S. G. Lutz, New York University; September 28, 1948.

UNIVERSITY OF FLORIDA—IRE-AIEE BRANCH

"Our Institute," by E. S. Lee, President, AIEE; September 29, 1948.

(*Continued on page 38A*)

STUDENT BRANCH MEETINGS

(Continued from page 37A)

IOWA STATE COLLEGE—IRE-AIEE BRANCH

"Registration of Engineers," by J. S. Dodds, Iowa State College; September 29, 1948.

STATE UNIVERSITY OF IOWA—IRE-AIEE BRANCH

"Radio Station WMT—Electrolysis Transmission Lines," by J. A. O'Brien and D. A. McMillon. October 13, 1948.

NEW YORK UNIVERSITY—IRE BRANCH

"Electrical Analogues of Acoustical Speakers," by C. Rehberg, New York University; October 14, 1948.

NORTH CAROLINA STATE COLLEGE—IRE BRANCH

"High-Fidelity Reproduction of Sound," by D. K. Briggs, Western Electric Company; September 29, 1948.

UNIVERSITY OF NORTH DAKOTA—
IRE-AIEE BRANCH

Election of Officers; September 29, 1948.

PRATT INSTITUTE—IRE BRANCH

"Cathode Follower," by W. Heacock, Student, Pratt Institute; October 7, 1948.

STANFORD UNIVERSITY—IRE-AIEE BRANCH

"Industrial Research Today," by J. E. Hobson, Stanford Research Institute; October 6, 1948.

UNIVERSITY OF TENNESSEE—IRE BRANCH

"Dimensional Analysis," by J. D. Tillman, Faculty of University of Tennessee; October 5, 1948.

WAYNE UNIVERSITY—IRE-AIEE BRANCH

"Student Papers," by E. Spring, Wayne University; October 7, 1948.

UNIVERSITY OF WISCONSIN—IRE BRANCH

"Job Opportunities and How to be Properly Interviewed," by H. Goehring, University of Wisconsin; October 5, 1948.

WORCESTER POLYTECHNIC INSTITUTE—
IRE-AIEE BRANCH

"Frequency Modulation," by G. E. Stannard, Worcester Polytechnic Institute; October 7, 1948.

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Beggs, G. E., Jr., School Lane, Warrington, Pa.
Brewer, A. F., 11605 Crenshaw Blvd., Inglewood, Calif.
Cerrillo, M. V., Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.
Cornes, R. W., 231 Northern Pkwy., East Hempstead, L. I., N. Y.
Sievert, A. H., Canadian Westinghouse Co., Ltd., Hamilton, Ont., Canada
Sobel, A. D., 2939 Ocean Ave., Brooklyn, N. Y. (November 1, 1948.)

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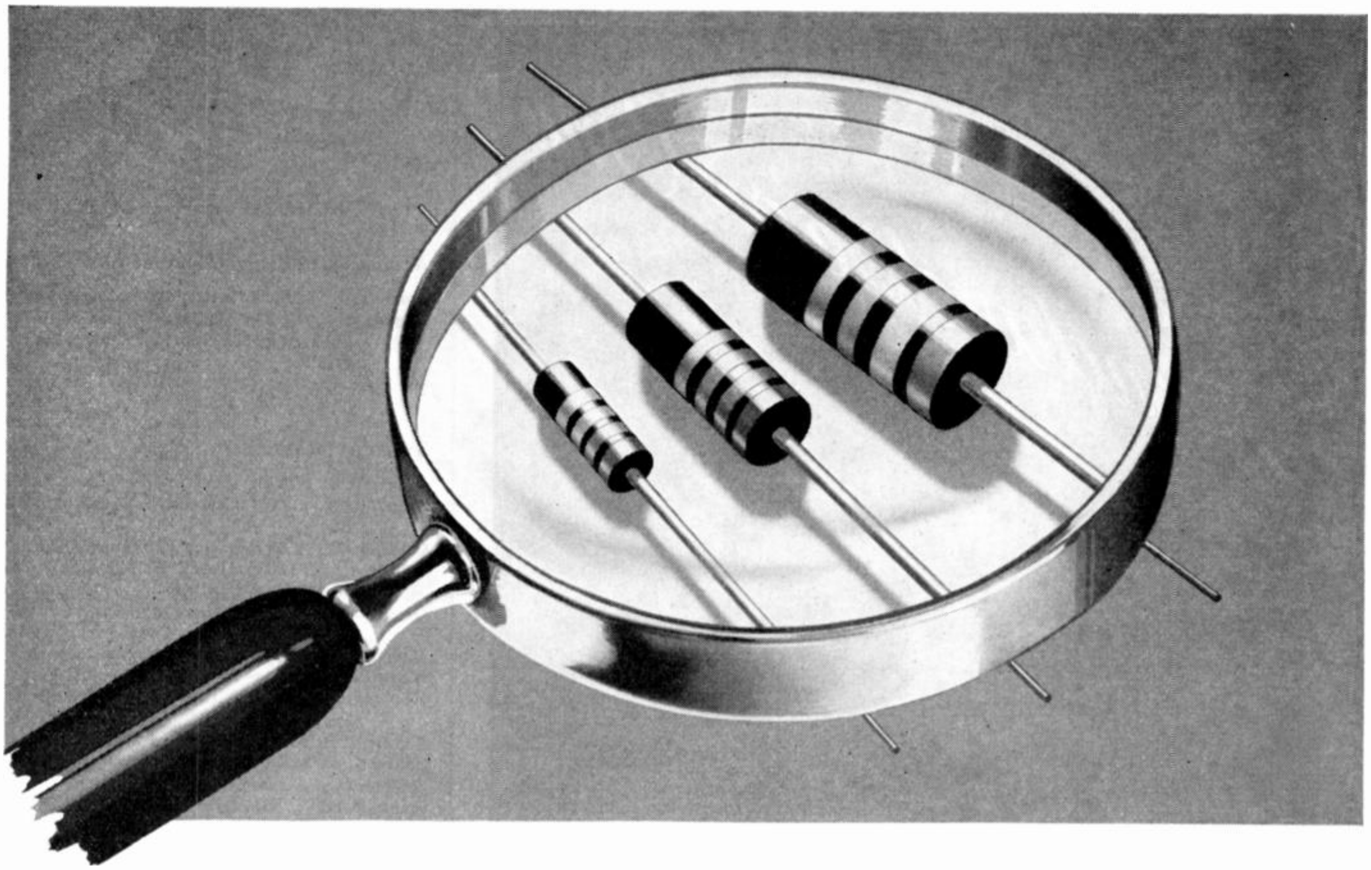

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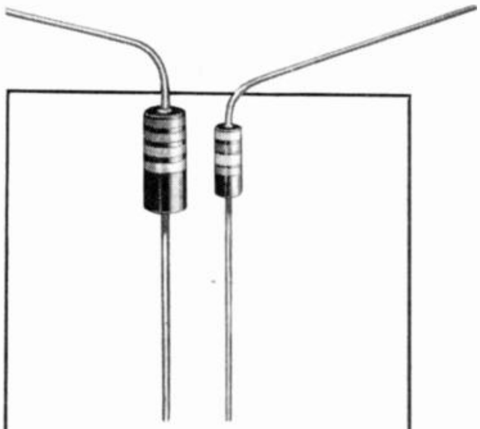

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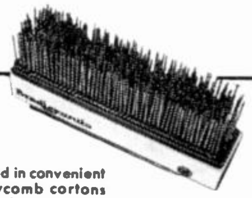

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(Continued from page 38A)

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Jones, T. F., Jr., 41 Atherton St., Roxbury, Mass.
Mayberry, L. A., The Hallicrafters Co., 4401 W. Fifth Ave., Chicago, Ill. (November 1, 1948.)
Sinninger, D. V., 2662 W. Jarlath Ave., Chicago, Ill.

Transfer to Member

Abbott, S. L., 1929 Davidson Ave., New York, N. Y.
Arand, E. M., 3125 N. Emerson St., Franklin Park, Ill.
Cameron, E. G., 10 Randolph Pl., Verona, N. J.
College, C. H., 11 Duvall Dr., W. Moreland Hills, Washington, D. C.
Goodman, P. D., Scott Laboratory, Wesleyan University, Middletown, Conn.
Houghton, R. W., Foster St., Littleton, Mass.
Moulton, A. B., 2554—24 St., N., Arlington, Va.
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Raycer, P. M., 333 E. 69 St., New York, N. Y.
Seaton, J. W., 3430—39 St., N.W., Washington, D.C.
Shapiro, O., 7 Sunset Dr., E., Nutley, N. J.
Smith, E. K., Jr., 4313 Ninth Ave., Brooklyn, N. Y.
Taylor, R. S., 2012 N. Upland St., Arlington, Va.
Toye, J. M., 51 St. Clair Ave., Toronto, Ont., Canada
Woster, G. W., 3018 W. 51 Ter., Kansas City, Kan.

Admission to Member Grade

Benson, K. B., Harbor View, South Norwalk, Conn.
Bobb, L. J., The International Electronics Co., 808 N. Broad St., Philadelphia, Pa.
Bose, K. K., 28/1/3 Gariahat Rd., Calcutta, P.O. Ballygunge, India
Corner, K. T., 108 North Ave. 19, Los Angeles, Calif.
Coullard, J. B., 111 Hazelhurst Ave., R.F.D. 1, North Syracuse, N. Y.
Field, L. M., 695 Columbia St., Palo Alto, Calif.
Gray, H. F., Jr., 21 Studio Lane, Bronxville, N. Y.
Green, J. R., 535 W. 110 St., New York, N. Y.
Johnston, O. B., 1012 E. 35 St., Minneapolis, Minn.
Jones, L. M., 1092 Beverly Way, Altadena, Calif.
Kerawalla, R. D., 6 Elsham Rd., London W. 14, England
Ralston, G., Apt. 201, 607 Hay St., Wilkinsburg, Pa.
Ray, H. B., 223 Newcomb St., S.E., Washington, D. C.
Riggin, J. D., Box 11, Del Monte, Calif.
Rooney, J. D., Box 3, ComServPac Staff, c/o FPO, San Francisco, Calif.
Schultz, C. W., R.F.D. 2, Storrs, Conn.
Shannon, C. E., Bell Telephone Laboratories, Murray Hill, N. J.
Tosh, W. M., c/o Electronics Institute Inc., 21 Henry St., Detroit, Mich.
Widerquist, V. R., SEES, Georgia Institute of Technology, 225 North Ave., N.W., Atlanta, Ga.

The following admissions to Associate grade have been approved and were effective as of November 1, 1948:

Ainsworth, M. G., 4803 Cornelia, Chicago 41, Ill.
Allen, J. R., 746 W. 104 Pl., Los Angeles 44, Calif.
Alpert, N., 55 E. Mosholu Pkwy., N., New York 67, N. Y.
Backstrand, W. C., c/o Physics Department, Whitman College, Walla Walla, Wash.
Barnes, A. W., 1917 Virginia, Berkeley 9, Calif.
Beaumont, G. F., 91 Charlton Ave., W., Hamilton, Ont., Canada
Berry, R. M., 2307 Cascade Trail, Bremerton, Wash.

(Continued on page 42A)

FREED

"PRODUCTS of EXTENSIVE RESEARCH"

No. 1030 by Freed

No. 1110 by Freed

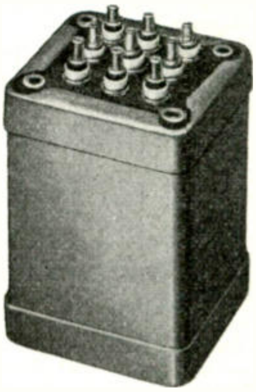
"Q" INDICATOR

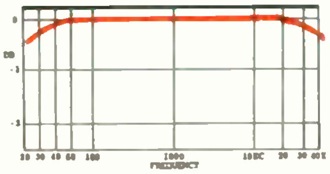
Frequency range from 20 cycles to 50 kilocycles. "Q" range from .5 to 500.
"Q" of inductors can be measured with up to 50 volts across the coil. Indispensable instrument for measurement of "Q" and inductance of coils, "Q" and capacitance of capacitors, dielectric losses, and power factor of insulating materials.

INCREMENTAL INDUCTANCE BRIDGE

IMPEDANCE RANGE: One millihenry to 1000 henries in five ranges. Inductance values are read directly from a four dial decade and multiplier switch. This range can be extended to 10,000 henries by the use of an external resistance.
INDUCTANCE ACCURACY: Within plus or minus 1% through the frequency range from 60 to 1000 cycles.

A NEW LINE OF HIGH FIDELITY OUTPUT TRANSFORMERS





Type No.	Primary matches following typical tubes	Primary Impedance	Secondary Impedance	± ½db from	Maximum level
F1950	Push pull 2A3's, 6A5G8s, 300A's, 275A's, 6A3's, 6L6's.	5000 ohms	500, 333, 250, 200, 125, 50	20-30000 cycles	15 watts
F1951	Push pull 2A3's, 6A5G8s, 300A's, 275A's, 6A3's, 6L6's.	5000 ohms	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20-30000 cycles	15 watts
F1954	Push pull 245, 250, 6V6, 42 or 2A5 A prime	8000 ohms	500, 333, 250, 200, 125, 50	20-30000 cycles	15 watts
F1955	Push pull 245, 250, 6V6, 42 or 2A5 A prime	8000 ohms	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20-30000 cycles	15 watts
F1958	Push pull 6B5, 6A6, 53, 6F6, 59, 79, 89, 6V6, Class B 46, 59	10,000 ohms	500, 333, 250, 200, 125, 50	20-30000 cycles	15 watts
F1959	Push pull 6B5, 6A6, 53, 6F6, 59, 79, 89, 6V6, Class B 46, 59	10,000 ohms	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20-30000 cycles	15 watts
F1962	Push pull parallel 2A3's, 6A5G's, 300A's, 6A3's, 6L6	2500 ohms	500, 333, 250, 200, 125, 50	20-30000 cycles	36 watts
F1963	Push pull parallel 2A3's, 6A5G's, 300A's, 6A3's, 6L6	2500 ohms	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20-30000 cycles	36 watts
F1966	Push pull 6L6 or Push pull parallel 6L6	3800 ohms	500, 333, 250, 200, 125, 50	20-30000 cycles	50 watts
F1967	Push pull 6L6 or Push pull parallel 6L6	3800 ohms	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20-30000 cycles	50 watts

FREED TRANSFORMER CO., INC.

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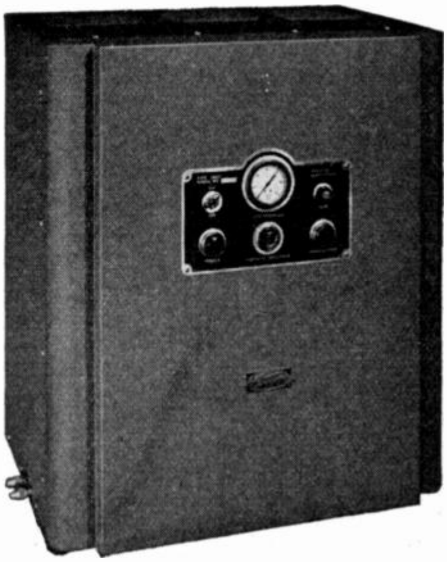
1718-36 WEIRFIELD ST.,
BROOKLYN 27, NEW YORK



MEMBERSHIP

(Continued from page 40A)

Beymer, E. H., 605 B. Anacada Dr., Oxnard, Calif.
Bluhm, R. W., 32 Woodcrest Ave., Short Hills, N. J.
Bombe, A. A., 29-01—159 St., Flushing, L. I., N. Y.
Browning, C. L., 3617 Montrose Blvd., Houston 6, Tex.
Brown, T. J., 1542 Marine Dr., W. Vancouver, B. C., Canada
Burgess, H. R. C., 2621 Laguna St., San Francisco 23, Calif.
Byers, A. E., 50 North Broadway, White Plains, N. Y.
Carney, J. J., Y.M.C.A., 117 W. Monument St., Dayton 2, Ohio
Chamberlain, E., 102 Woodlands Rd., Hull, E. Yorks, England
Chapin, E. L., Jr., 4711 W. Main St., Belleville, Ill.
Chase, W., Barnard Hall, North Brother Island, New York 54, N. Y.
Clark, L. G., 6227—21, N.E., Seattle 5, Wash.
Corbett, H. L., 124½ Percy St., Ottawa, Ont., Canada
Daleo, S. L., 198 First Ave., New York 3, N. Y.
DeVoe, W. D., 3312 Technical Training Squadron, Officers Comm., Scott Field, Belleville, Ill.
Douglas, M. D., 76 Concord Rd., R.F.D. 3, Chagrin Falls, Ohio
Dunlap, R. H., 408 Baltimore, San Antonio, Tex.
Edwards, M. J., R.F.D. 3, Box 483, Glendale, Ariz.
Eisner, W. K., 486 Ashford St., Brooklyn, N. Y.
Elikann, L. S., 1439 East 19 St., Brooklyn, N. Y.
Favreau, R. R., 6765 W. 86 Pl., Los Angeles 45, Calif.
Fielding, B. L., 5019½ Lankershim Blvd., North Hollywood, Calif.
Fisher, S., 11 Westcott Rd., Princeton, N. J.
Froke, L. C., Radio Station KELO, Sioux Falls, S. Dak.
Gold, M., 254 Fountain Ave., Dayton, Ohio
Goldstein, I., 80 Brantwood Rd., Worcester, Mass.
Gones, R. D., 4568 Mary Ave., St. Louis 7, Mo.
Gudaitis, W. V., 1617 Morrell, Detroit 9, Mich.
Harpole, J. L., Live Oak Plant, Ravenel S. C.
Holmquist W. L. 116 E. Park Ave., Libertyville, Ill.
Horn, R. H., 1370 Washington St., San Francisco 9, Calif.
Jones, H. G., 3776 Milan St., San Diego 7, Calif.
Jorgensen, S. W., 88-36—187 St., Hollis 7, L. I., N. Y.
Kasparek, J. J., 309 N. Wayne St., Angola, Ind.
Kliopera, M. F., 3221 Charleston Circle, Houston 4, Tex.
Kreis, R. J., Box 97, College Park, Md.
Lanwell, L. W., 313½ Main Ave., McCook, Neb.
Lapidos, R. W., 134 Penn Ave., Collingswood, N. J.
Larsen, N. J., Radio Station KMMJ, Grand Island, Neb.
Lauterbach, R. E., 845 S. Williams St., Denver 9, Col.
Lidz, S., 451 Washington Ave., Brooklyn 5, N. Y.
Munn, E. H., Jr., 306 N. West St., Hillsdale, Mich.
Olander, R. O., 64 Homestead Ave., Bridgeport 3, Conn.
Osteyee, W. W., QTRS, 349 Doniphan, Fort Bliss, Tex.
Ott, L. O., 220 S. Kenwood St., Glendale 5, Calif.
Parode, L. C., 711 Stepney St., Inglewood, Calif.
Pellock, J., 10745 Channock Rd., Los Angeles, Calif.
Petit, F. W., North Woodbury, Conn.
Prabhu, K. S., Department of Communication Engineering, Indian Institute of Science, Bangalore 3, India
Pritchett, W. C., 400 E. 32, Austin, Tex.

(Continued on page 46A)

For Perfected Large-Size Home Projection-PROTELGRAM

The 2½" magnetic projection triode 3NP4 has a face as small as a compact and is only 10½" long.

HERE'S THE OPPORTUNITY THAT MANUFACTURERS OF TELEVISION RECEIVERS HAVE BEEN AWAITING!

10 SIGNIFICANT FEATURES

1. Flat 16" x 12" non-reflecting picture provides fatigueless viewing from less than 5 feet and upward!
2. Wide-angle visibility — *square corners.*
3. True photographic black and white picture quality—no discoloration.
4. Compact unit—suitable for table model cabinets.
5. Long-life, low-cost picture tube.
6. Manufacturers can most economically extend their product range into projection television by adapting their 10" EM chassis for use with PROTELGRAM.
7. Easy to service.
8. High contrast ratio and broad gray tone range.
9. Simple optical adjustment system.
10. *Quality built* after more than 10 years of development.

NORELCO PROTELGRAM consists of a projection tube, an optical box with focus and deflection coils, and a 25 kv regulated high-voltage supply unit, making possible large-size home projection. More than ten years of exhaustive research resulted in this ideal system for reproducing a projected picture. The optical components are designed to produce perfected projection for a 16" x 12" image, the optimum picture size for steady, distant observation and also for *proper viewing at less than 5 feet.*

Other NORELCO products include standard 10" direct-viewing tubes and special-purpose cathode-ray tubes for many applications.

NORTH AMERICAN PHILIPS COMPANY, INC.
Norelco

PROTELGRAM IS PICTURE PERFECTION IN PROJECTION

NORTH AMERICAN PHILIPS COMPANY, INC.
DEPT. TP12, 100 EAST 42nd STREET, NEW YORK 17, N. Y.

IN CANADA: PHILIPS INDUSTRIES LTD., 1203 PHILIPS SQUARE, MONTREAL ★ EXPORT REPRESENTATIVE: PHILIPS EXPORT CORPORATION, 100 EAST 42ND STREET, NEW YORK 17, N. Y.

This NEW IRC type

NOW an ADVANCED insulated fixed composition resistor that offers new opportunities to radio, television, electronic and electrical engineers.

At desks and drawing boards across the country resistor requirements are being reviewed in the light of this *advanced* resistor. Quiet huddles in engineering departments and research labs are rapidly disclosing present performance standards for fixed composition resistors to be obsolete.

The new IRC Type BT resistor *meets* JAN-R-11 specs. At ⅓, ½, 1 and 2 watts, the *new* IRC Type BT is an *advanced* resistor in every respect.

You may specify this *advanced* IRC resistor immediately. It is in production now, with hundreds of thousands coming off production lines daily.

BT *means* ***B****etter* ***T****echnically* **BTS** *means* ***B****eats* ***T****oughest* ***S****pecs* **BTR** *means*

BT resistor will change your standards of performance for fixed composition resistors

Get the full performance facts on this ADVANCED resistor

Standards for resistor performance set by this *new* IRC Type BT are so advanced, you need complete information to fully evaluate its significance. Test result charts will show you how this *advanced* resistor easily performs the rigorous requirements of television. Performance curves prove its excellence in every characteristic . . . particularly in high ambient temperatures.

Technical Data Bulletin B-1 gives the full story. We shall be glad to rush it to your desk or drawing board . . . or to have our representative review your requirements in the light of this *advanced* resistor. Use the handy coupon below.

IRC MEETS JAN-R-11


IRC CATALOG DATA BULLETIN

TYPE BT
ADVANCED FIXED COMPOSITION RESISTORS
½ - ¾ - 1 and 2 watts

IRC MEETS JAN-R-11

IRC

power resistors • precisions • insulated composition resistors • low wattage wire wounds • rheostats • controls • voltmeter multipliers • voltage dividers • HF and high voltage resistors.

INTERNATIONAL RESISTANCE CO., 401 N. Broad Street, Philadelphia 8, Pa.
IN CANADA: *International Resistance Co., LTD., Toronto, Licensee*

Better Test Results BT means Better Television

International Resistance Co.
401 N. Broad St., Phila. 8, Pa.

I want to know more about IRC's *advanced* BT Resistor:

☐ Send me Technical Data Bulletin B-1.
☐ Have your representative call—no obligation.

Name ....
Title ....
Company ....
Address ....
....

MEMBERSHIP

(Continued from page 42A)

Rapp, R. M., 8 Kinmore St., Warren, Pa.
Rasmussen, C. F., 201 West 90 St., Los Angeles, Calif.
Rayasa, N. G., Nagappa Block 1321 Sreerampur, Bangalore 3, India
Rayasa, R. G., c/o Radio Sonde Section, Observatory, Lodi Rd., New Delhi, India
Reeves, E. H., 200 Franklin St., Bloomfield, N. J.
Robey, R. E., 142-36—32 Ave., Flushing, L. I., N. Y.
Robinson, R. J., 326 Chicopee St., Chicopee, Mass.
Rogers, J. B., 449 Highbrook Ave., Pelham Manor, N. Y.
Seshadri, K. V., c/o Prof. K. Sreenivasan, Department of Electric Communication Engineering, Indian Institute of Science, Bangalore 3, India
Seybold, J. F., 3278 West 44 St., Cleveland 9, Ohio
Shaw, H. H., 1811 Roberta Ave., Willow Grove, Pa.
Smith, J. D., 3220 Duval St., Houston, Tex.
Soorin, A. J., R.F.D. 1, Box 20, Long Branch, N. J.
Srinivasan, L. S., c/o Rao Bahdur T.V.V. Aiyer, Conservator of Forests, Thycaud, Trivandrum, India
Steen, P. W., 2 Matthew Lane, Buffalo 21, N. Y.
Stewart, G. L., 194 Colbeck St., Toronto, Ont., Canada
Tariot, J. N., 1 Highland St., Cambridge 38, Mass.
Theurer, D. L., 900 W. Maumee St., Angola, Ind.
Thomson, R. D., 510 E. Wynnewood Rd., Wynnewood, Pa.
Tims, E. F., Louisiana State University, Electrical Engineering Department, Baton Rouge 3, La.
Triplett, T. R., 18 Chauncey St., Apt. 6, Cambridge 38, Mass.
Tuksal, M. A., 3053 W. Grand Blvd., Detroit 2, Mich.
Warwick, J. D., 25 Marathon Ave., Dayton 5, Ohio
Weinberg, M. D., 81 Ocean Pkwy., Brooklyn 18, N. Y.
Whelchel, G. O., Jr., 531 Stinard Ave., Syracuse, N. Y.
Whitman, K. W., 5752 Seventh Ave., Los Angeles 43, Calif.

The following transfers to Associate grade were approved to be effective as of November 1, 1948:

Cassutt, R. J., 1811 Chandler St., S.W., Cedar Rapids, Iowa
Dimasi, L. A., 641 Highland Ave., Greensburg, Pa.
Giloth, P. K., 2515 Ashland Ave., Evanston, Ill.
Goodill, J. J., 544 East 7 St., Erie, Pa.
Hooper, W. D., 1103 S. University, Ann Arbor, Mich.
Irons H. R., 7929 Georgia Ave., Silver Spring, Md.
Kazanowski, H. F., 228 Winchester St., Brookline, Mass.
Lewis, F. C., 336 S. Market St., Galion, Ohio
Licata, J. P., 1063 Forest Rd., Schenectady, N. Y.
Lipin, B., 4142 Paseo, Kansas City 4, Mo.
Lovell, J. A., 5 Croyden Rd., Mineola, L. I., N. Y.
Meyer, J. A., Box 2018, University of Arkansas, Fayetteville, Ark.
Panetta, A. R., 1865 East 81 St., Cleveland 3, Ohio
Ross, D. O., 88 Percival Ave., Montreal West, Que., Canada
Sabol, R. W., 17 S. Oxford St., Brooklyn 17, N. Y.
Slagter, H. C., 47 Lincoln St., Pittsfield, Mass.
Spielberger, S. C., 90-28—148 St., Jamaica 2, N. Y.
Wall, V. W., 253 Elmwood Ave., Maplewood, N. J.
Wolff, J. L., Jr., R.F.D. 4, Greensburg, Pa.
Wright, J. R., 9600 St. Lawrence Blvd., Montreal, Que., Canada
Yard, R. R., 1 West 68 St., New York 23, N. Y.

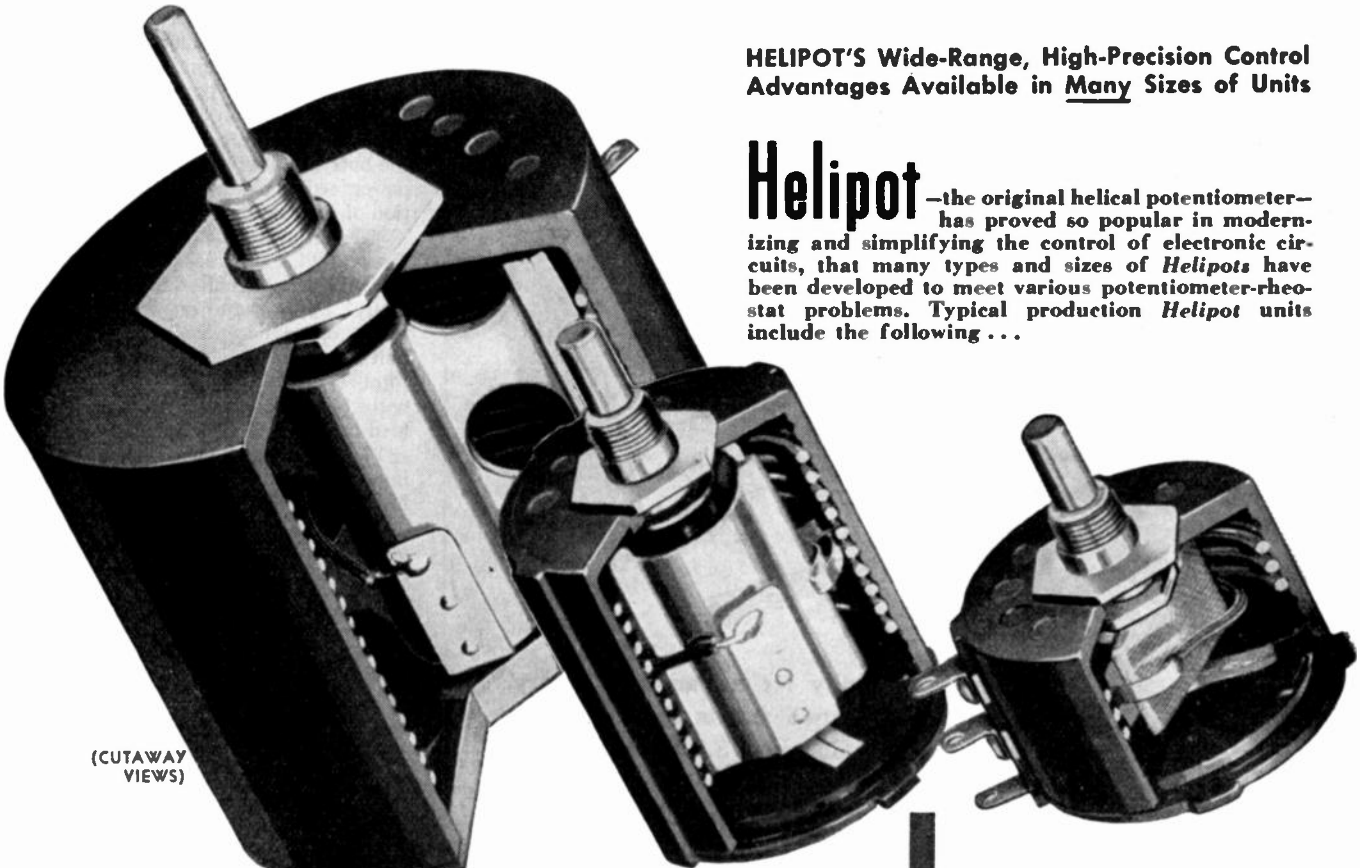
There's a Beckman Helipot

(Trade Mark of the **HEL**ical **POT**entiometer)

to simplify YOUR Potentiometer—Rheostat Problems!

HELIPOT'S Wide-Range, High-Precision Control Advantages Available in Many Sizes of Units

Helipot—the original helical potentiometer—has proved so popular in modernizing and simplifying the control of electronic circuits, that many types and sizes of *Helipots* have been developed to meet various potentiometer-rheostat problems. Typical production *Helipot* units include the following . . .



(CUTAWAY VIEWS)

MODEL B—Case diameter—3.3"; Number of turns—15; Slide wire length—140½"; Rotation—5400°; Power rating—10 watts; Resistance ratings—50 to 300,000 ohms.

MODEL A—Case diameter—1.8"; Number of turns—10; Slide wire length—46½"; Rotation—3600°; Power rating—5 watts; Resistance ratings—10 to 100,000 ohms.

MODEL C—Case diameter—1.8"; Number of turns—3; Slide wire length—13.5"; Rotation—1080°; Power rating—3 watts; Resistance ratings—5 to 30,000 ohms.

WIDE CHOICE OF DESIGN FEATURES

Not only are *Helipots* available in a wide range of sizes and ratings, but also can be supplied with various design features to meet individual requirements . . .

- ▶ Available with special length shafts, flatted shafts, screw-driver slots, etc.
- ▶ Can be supplied with shaft extensions at each end to permit coupling to indicating instruments or other devices.
- ▶ May be provided in ganged assemblies of two or three units, all operating from a common shaft.
- ▶ Available with linearity tolerances of 0.1%—and even less.
- ▶ Models A & B can be modified to include additional taps at virtually any point on windings.

. . . and many other special features.

Investigate the many important advantages to be gained by using the *Helipot* in your electronic control applications. Write outlining your problem!

SPECIAL MODELS

In addition to the above standard *Helipot* units, special models in production include . . .

MODEL D—Similar to Model B, above, but longer and with greater length of slide wire. Case diameter—3.3"; Number of turns—25; Slide wire length—234"; Rotation—9000°; Power rating—15 watts; Resistance ratings—100 to 500,000 ohms.

MODEL E—Similar to Model B, but longer and with greater length of slide wire than Model D. Case diameter—3.3"; Number of turns—40; Slide wire length—373"; Rotation—14,400°; Power rating—20 watts Resistance ratings—150 to 800,000 ohms.

Send for HELIPOT Literature!

THE Helipot CORPORATION
SOUTH PASADENA 6, CALIFORNIA

Have you latest data on DUODIAL—Tthe turns-indicating knob dial? If not, write!

News—New Products

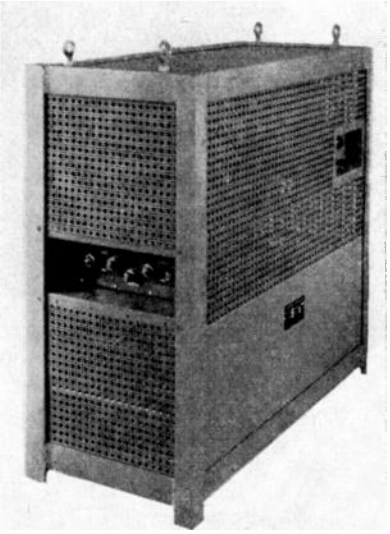
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 30A)

Voltage Regulation

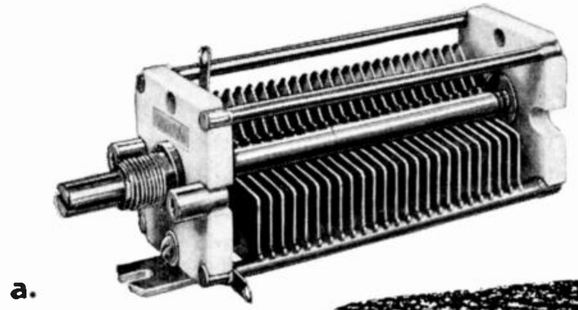
Increasing development of precise electronic components will require a more widespread use of power supplies for electron tube and experimental operations possessed of good regulating qualities. Each month sees announcement of one or more of these units, for applications where the user does not care to undertake their construction. Too, fuller recognizance of this principle is finding application in the supply of primary power to complete installations of radio and electronic equipment. Broadcast transmitters located in the outlying districts or at the end of long power feeder lines are a point in demonstration of this thesis.

The Howard Co., 934 Argyle Road, Drexel Hill, Pa. comes up with a new unit of sufficiently small dimensions to find ready application on the laboratory bench, or for incorporation into completed units. It is capable of delivering 60 watts total regulated power (300 volts, 200 ma), with 1-volt regulation from zero to full load and with a line-voltage variation of ±10 volts from the design center. Ripple is so low as to serve well when operating high-gain amplifiers, and the constancy of regulation is exceptional over long periods of operation.



Sorensen & Co., Inc., 375 Fairfield Ave., Stamford, Conn., have announced models for primary power regulation, of 5 and 10 kva capacity. These will serve the low-power stages of a high-powered transmitter, thus affording constant voltage to the frequency regulating stages, or will provide sufficient capacity to power a whole laboratory of test equipment for precise measurement techniques. The phase distortion is low in all models, but a special design is offered where this factor may prove an important consideration. Frequency shift from 60 to 50 cycles will not affect the accuracy of these units. Full descriptive literature is available from the makers.

(Continued on page 59A)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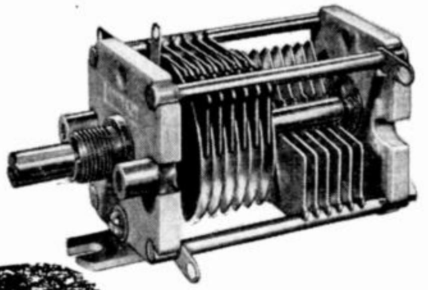
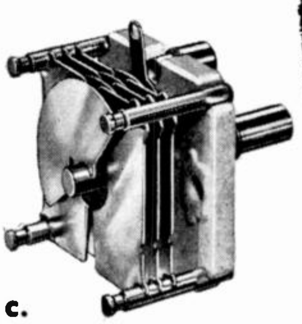
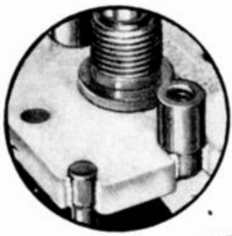
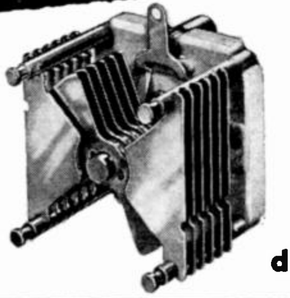

b.

CERAMIC SOLDERED . . .


c.



FOR STABILITY
AND STRENGTH


d.

Announcing the new JOHNSON 167 VARIABLES

BETTER FOR UHF - VHF - LF

With the introduction of this new line of air variables, JOHNSON brings you many important design advantages never before available.

Outstanding of these is the use of perfected ceramic soldering which assures absolute — and permanent — rigidity and strength, absolute — and permanent — maintenance of capacities!

There are no eyelets, nuts or screws to work loose, causing stator wobble and fluctuations in capacity. JOHNSON ceramic soldering leaves a bond which is stronger than the rugged steatite end plates themselves. There's nothing to come loose, because the stator terminals, mounting posts and rotor bearings are ceramic soldered!

Silent operation on the highest frequencies is assured with a split sleeve tension bearing that also prevents fluctuations in capacity.

These new variables are ideal for peak efficiency even under the severest conditions, such as portable — mobile operation. They are available in .030" and .080" spacings.

Two sets of stator contacts are provided for connecting components to either side of condenser without appreciably increasing inductance of the circuit. New bright alloy plating is used. It has high corrosion resistance, is easily soldered and possesses lower electrical resistance than other common platings.

These variables are available for all types of communications equipment having tuned circuits operating as high as 500 mc.

Features

1. Ceramic soldered for stability and strength
2. Soldered plate construction, heavy .020" plates, new bright alloy plating
3. Beryllium copper contact spring, silver plated
4. Split sleeve rotor bearings — no wobble to shaft
5. Steatite end plates
6. Long creepage paths
7. Low minimum capacity — maximum tuning range
8. Small size — end plate only 1⅞" square

Other capacities and spacings available on special order.

a. SINGLE SECTION VARIABLES

.030" Spacing

Cat. No.	Cap. Per Section Max.	Cap. Per Section Min.	Length Behind Panel
167-101	11	2.8	15/16
167-102	27	3.5	1-9/64
167-103	51	4.6	1-7/16
167-104	75	5.7	1-¾
167-151	99	6.8	2-7/32
167-152	202	11.6	3-33/64

Also Available In .080" Spacing

b. DUAL SECTION VARIABLES

.030" Spacing

Cat. No.	Cap. Per Section Max.	Cap. Per Section Min.	Length Behind Panel
167-501	27	3.5	1-13/16
167-502	51	4.6	2-27/64
167-503	99	6.8	3-⅝

Also Available In .080" Spacing

c. DIFFERENTIAL VARIABLES

.030" Spacing

Cat. No.	Cap. Per Section Max.	Cap. Per Section Min.	Length Behind Panel
167-301	11	2.8	15/16
167-302	27	3.5	1-9/64
167-303	51	4.6	1-7/16

Also Available In .080" Spacing

d. BUTTERFLY VARIABLES

.030" Spacing

Cat. No.	Cap. Per Section Max.	Cap. Per Section Min.	Length Behind Panel
167-201	10.5	2.8	1-3/64
167-202	26	4.3	1-7/16
167-203	51	6.5	1-15/16

Also Available In .080" Spacing

Write For NEW JOHNSON 167 VARIABLE CATALOG

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

E. F. JOHNSON CO., WASECA, MINNESOTA

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

ELECTRONIC ENGINEERS—ELECTRONIC TECHNICIANS

Men with production, inspection or re-design experience preferred. Direct inquiries to SLX-1, P.O. Box 5800, Albuquerque, New Mexico.

TELEVISION ENGINEERS

Well established electronics manufacturer, located in suburban New York City area needs high grade television development engineers. Company is expanding its television activities and seeks capable men with background sufficiently complete to merit responsible positions. Send details to Chief Engineer, Box 536.

PROJECT ENGINEER

Stamford firm engaged in development and research work with government agencies requires project engineer with technical and practical background in UHF radar work. Salary open depending upon background. Company is young and growing with promising future for able engineers in this kind of work. Box 538.

ELECTRONIC ENGINEERS

College graduates with 3-5 years of development engineering experience in circuit design. Well versed in magnetic circuits, non-linear circuit operation and electronic theory. Send resume and all particulars to Personnel Department, General Precision Laboratory, Inc., Pleasantville, New York.

RESEARCH AND DEVELOPMENT ENGINEERS

Engineers with considerable experience in RF and UHF circuits wanted by large, well established radio company in New York area. Send resume of education and experience to Box 539.

GRADUATE PHYSICIST

Graduate physicist or electronic engineer with good background in gaseous conduction wanted by established New England radio tube manufacturer, for development work on gas filled tubes. Box 540.

MATHEMATICIANS, ENGINEERS, PHYSICISTS

Men to train in oil exploration for operation of seismograph instruments, computing seismic data, and seismic surveying. Beginning salary $250.00 to $300.00 per month depending upon background. Excellent opportunity for advancement determined by ingenuity and ability. The work requires changes of address each year; work indoors and out; general location in oil producing locations. Send resume and include snapshot to National Geophysical Co., Inc., 8800 Lemmon Ave., Dallas 9, Texas.

(Continued on page 52A)

Your enjoyment climbs to new altitudes through radio and television achievements of RCA

Radio and Radar Development and Design Engineers

RCA Victor Division has openings for men with experience in:

- **Mobile Transmitters**
- **Radar**
- **Audio Communications and Supersonics**

Please furnish complete resume of experience to:

Employment Manager
Camden Personnel Division
RCA Victor Division
Camden, New Jersey

POSITIONS OPEN

(Continued from page 50A)

ELECTRONICS ENGINE[illegible]

Well known, 40 year old manu[illegible]rer of electrical and electronic instruments wants research engineer experienced in design of radio electronic apparatus at high frequencies, for the development of military and civilian test equipment. Box 542.

PHYSICISTS—ENGINEERS

Opportunities for physicists and electronic engineers at the Naval Ordnance Test Station, P.O. China Lake, California. Applicants should have college degree or equivalent, plus professional experience. Especially desired are applications from persons with experience in design and development of microwave radar components. Send application form 57 to Placement Officer U.S.N.O.T.S., Inyokern, P.O. China Lake, California.

ENGINEERS

A young rapidly expanding mid-western research laboratory has openings for the following types of personnel:

(1) ELECTRONIC DESIGN ENGINEERS—Experience circuit design enginees wanted for indicator timing circuit development.

(2) SERVO DESIGN ENGINEERS OR PHYSICISTS—Experience required for the development airborne computor equipment.

Top pay offered to those capable of project responsibility. Local university offers graduate courses in servo and computer design, and applied mathematics. Send complete resume. Our Engineering Dept. knows of this announcement. Box 544.

ELECTRONIC ENGINEERS—PHYSICISTS

A leading electronics company in Los Angeles, California offers permanent employment to persons experienced in advanced research and development. State qualifications fully. Box 545.

ELECTRONIC ENGINEER

An opportunity for a man with considerable experience to head a small development and engineering group in a growing company located in Chicago. Pulse experience a necessity and a background of work with nuclear radiation instruments and nuclear detectors very desirable. Please give full details. Box 546.

RADIO AND TELEVISION ENGINEERS

The Industry Service Laboratory (formerly License Laboratory), New York, has several positions open for Senior and Junior engineers having qualifications for development and consultation work in television and radio. Good technical education and some experience required. Interesting work, broadening experience, and wide contacts. Write fully to Director, Industry Service Lab., RCA Laboratory Division, 711 Fifth Ave., New York 22, N.Y.

CRYSTAL ENGINEER

Manufacturer of Piezo electric crystals desires experienced engineer familiar with quartz oscillating crystals and their applications to radio frequency control. Write full details. Box 548.

SENIOR ENG EER

with outstanding academic and practical experience and executive ability in the field of cathode ray tube display and indicators for radar systems, wanted by long established corporation. Knowledge of all types of projection systems and present types of service equipment desirable. Write, giving resume of education, experience, age and salary.

FREED RADIO CORPORATION
200 Hudson Street, New York, N.Y.

WANTED ELECTRONIC ENGINEERS AND PHYSICISTS

Excellent opportunities for graduates with research, design, and/or development experience in Communications and aerial navigation systems including direction finders, radar, FM, television, micro-wave.

Write complete details regarding education, experience and salary desired.

To Personnel manager
Federal Telecommunication Laboratories
500 Washington Ave.
Nutley, New Jersey

POSITIONS OPEN

RESEARCH AND DEVELOPMENT ENGINEERS

Wanted for advanced research and development. Should have extensive experience on analysis of electronics systems in the fields of microwaves, missiles, radar, servomechanisms communications, navigational devices. Outstanding ability in E.E. or physics required. Please furnish complete resume, salary requirements and availability to: Personnel Manager, W. L. Maxson Corporation, 460 West 34th Street, New York, N.Y.

ELECTRICAL ENGINEER

Nationally known Chicago company is in need of high grade, experienced electrical engineer. Must be a college graduate with either a B.S. or E.E. degree, with high scholastic record. Should have from 2 to 5 years experience in electronics, preferably with a minimum of two years in the design of audio amplifiers. In reply give all particulars and state expected salary. Address Box 549.

DESIGN AND DEVELOPMENT ENGINEER

Design and development engineer to take charge of engineering and development of receiving antennae and associated equipment. U.H.F. experience desirable. Upstate New York manufacturer. Reply giving age and qualifications to Box 450.

(Continued on page 54A)

ZENITH RADIO CORPORATION needs Research and Development Engineers and Physicists

Project, senior and junior engineers and physicists are required for prosecution of several very interesting developments. Openings exist for men experienced in various aspects of radio and television receiver and transmitter development, in radar, and in applications of electronics to ordnance problems. *Salaries commensurate with experience and ability.*

WRITE: Director of Research
Zenith Radio Corp.
6001 W. Dickens Ave.
Chicago 39, Illinois

Electronic Engineers

Bendix Radio Division
Baltimore, Maryland
manufacturer of
RADIO AND RADAR EQUIPMENT
requires:

PROJECT ENGINEERS

Five or more years experience in the design and development, for production, of major components in radio and radar equipment.

ASSISTANT PROJECT ENGINEERS

Two or more years experience in the development, for production, of components in radio and radar equipment. Capable of designing components under supervision of project engineer.

Well equipped laboratories in modern radio plant . . . Excellent opportunity . . . advancement on individual merit.

Baltimore Has Adequate Housing

Arrangements will be made to contact personally all applicants who submit satisfactory resumes. Send resume to Mr. John Siena:

BENDIX RADIO DIVISION
BENDIX AVIATION CORPORATION
Baltimore 4, Maryland

Wanted
Design Engineers
Physicists

Men with a college degree and two to four years design experience should investigate the opportunities offered by the Collins Radio Company. This well-recognized manufacturer of radio equipment has a limited number of positions available for qualified engineers and physicists. These men will work on the design and development of broadcast, communications, radar, and electronic circuits in the design and research departments. Give present position, nature of work, experience, and education in first letter.

ADDRESS DEPARTMENT EP
COLLINS RADIO COMPANY
CEDAR RAPIDS, IOWA

POSITIONS OPEN

(Continued from page 53A)

SALES ENGINEER

Several territories open east of Rocky Mountains for alert, experienced sales engineer representative capable of selling and installing FM two-way radiotelephone systems for mobile operations. Nationally advertised product. Exceptional opportunity for right man. Send detailed qualifications, education, past experience and territory desired. Radiotelephone Operators License, 1st or 2nd class preferred. Must have had previous experience in FM radiotelephone. Reply Box 451.

SCIENTISTS AND ENGINEERS

Wanted for interesting and professionally challenging research and advanced development in the fields of microwaves, radar, gyroscopes, servomechanisms, instrumentation, computers, and general electronics. Scientific or engineering degrees required. Salary commensurate with experience and ability. Direct inquiry to Manager, Engineering Personnel, Bell Aircraft Corporation, P.O. Box 1, Buffalo 5, New York.

RADIO PROJECT ENGINEER

Graduate engineer; 5 years recent experience design and development oscillator and amplifier circuits in VHF and UHF ranges. Familiar theoretical concepts and calculations circuit components, as well as practical design and layout work. Must have initiative and supervisory ability. Federal Manufacturing & Engineering Corporation, Brooklyn 5, N.Y.

AERO DYNAMICIST ENGINEERS

Aero Dynamicist engineers wanted to work on the design of Analog computers, to simulate the flight characteristics of specific aeroplanes. 3 years experience in stability and control essential. Knowledge of servomechanisms dynamics of free flight and applied mathematics desirable. Apply in person, or submit resume to Personnel Dept. Curtiss Wright Corp., Propeller Division, Route 6, Caldwell Township, New Jersey.

ELECTRONICS ENGINEERS

Top flight engineers. Must have 10 years design and development experience on servomechanisms and amplifiers, circuits and equipment layout. Apply in person or submit complete resume to Personnel Dept., Curtiss Wright Corp., Propeller Division, Route 6, Caldwell Township, New Jersey.

ELECTRICAL ENGINEER

Opening for a man who has the ability to teach advanced electronic circuits and theory, frequency modulation and television. Should eventually teach course in ultra-high frequency techniques and electric wave phenomena and pulse systems. Reply to: Director of Academic Administration, 1020 North Broadway, Milwaukee 2, Wisconsin.

SOUND ENGINEER

For large radio and television manufacturer in Chicago area. Experience in loud speaker design, audio circuits, and acoustics necessary. Fine opportunity for advancement. Please write giving full particulars. Box 552.

(Continued on page 55A)

Positions Open

PROFESSOR

Professor of communications engineering needed for fall 1949 by southeastern university. Will be in charge of graduate work and research activities. $6000.00 for nine months with extra income for summer teaching. Must have Ph.D. or D.Sc. degree. Write Box 553.

* * * *

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

ENGINEER

Engineer. 28. B.S.E.E. Columbia, also business degree. 3 years responsible business experience. Best references. New York area preferred. Call Lu. 8-9164 mornings or write. Box 182 W.

SALES ENGINEER OR EXECUTIVE ASSISTANT

Young, aggressive, hard-hitting design and development engineer invites inquiries from firms having need for addition to sales engineering staff or assistant to top executive. Thorough background of research, design and supervision in measurement apparatus; receivers, transmitters and audio equipment. Capable of handling engineering, purchasing, production, inspection and personnel. Prefer New York City location. Box 187 W.

PHYSICIST

Twenty-seven year old graduate physicist and mathematician with experience in control circuits desires either foreign or domestic employment. Box 197 W.

BROADCAST ENGINEER

Nine years experience. Now employed as assistant chief engineer 5 KW AM 50 KW FM, major network station, directional antennas. Desires chief engineer position or transmitter supervisor in east. Best references. All offers considered. Box 199 W.

ENGINEERING—ADVERTISING

Hard-hitting advertising executive who can talk an engineer's language.

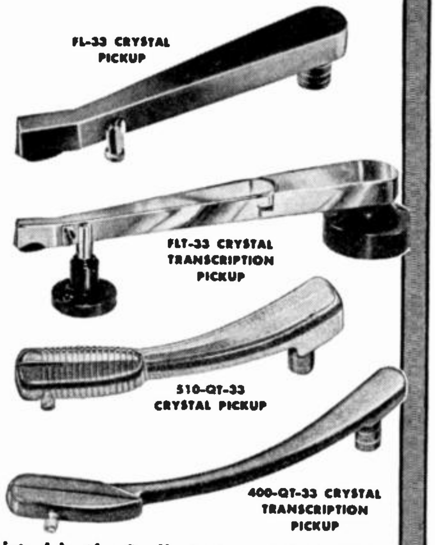
A.B., M.A., plus 3 years electrical engineering. Last 4 years with top electrical corporation creating sales campaigns, ad-

(Continued on page 56A)

ANNOUNCING...

A COMPLETE LINE OF ASTATIC LONG-PLAYING PICKUPS and CARTRIDGES that includes JUST WHAT YOU'RE LOOKING FOR!

- Want a two-in-one pickup that plays both LP and 78 RPM Records with the simple switching of cartridges . . . so simple that a child can make the change in a few seconds? Or, perhaps you are looking for comparable reproduction quality, with more emphasis on economy? Regardless of whether your requirements point to cartridges employing ceramic elements or magnetic-type units . . . whether you prefer permanent or replaceable needles, metal, sapphire or diamond . . . whether cost is first or secondary . . . whatever the conditions to be met —there now is a unit of Astatic precision engineering and construction that will exactly fill the bill. Space permits mention here of only a few. Why not write for new brochure, giving complete information, illustrations, on the Astatic Long-Playing Equipment Line?



Listed in the Radio Industry Red Book

FL-33 CRYSTAL PICKUP—Incomparable reproduction, utility and convenience. Employs LP-33 Crystal Cartridge for LP Records and LP-78 Crystal Cartridge for 78 RPM Records. Change cartridges in a second, like slipping modern fountain pen from its cap, nothing else to do. Special anti-resonance base mounting.

FLT-33 CRYSTAL TRANSCRIPTION PICKUP—Like the FL-33 Arm, plays either LP or 78 RPM Transcriptions with the LP-33 and LP-78 Cartridges. Anti-resonance base and arm-rest are adjustable to desired height. Five-gram needle pressure and perfect tracking assured by revolutionary hinged division of arm. Two-toned black and satin chrome finish.

FLT-TR CRYSTAL TRANSCRIPTION ARM—The same fine instrument as the FLT-33, except for 2.4 mill tip-radius needle necessary for lateral broadcast transcriptions. Employs the LP-TR Cartridge, instantly replaceable with LP-33 or LP-78 Cartridges.

510-QT-33 CRYSTAL PICKUP—Short mounting centers, gracefully curved lines and moderately offset head make this the ideal pickup for a host of applications. Famous "QT" Series Cartridge with replaceable, one mill tip-radius, precious metal or sapphire needle.

510-MI-2M-33 MAGNETO-INDUCTION PICKUP—Same as 510-QT-33, except for revolutionary Magneto-Induction Cartridge. Consistent service and adverse climatic conditions are no threat to the stability and troublefree operation of this magnetic type unit.

400-QT-33 CRYSTAL TRANSCRIPTION PICKUP—Graceful, slender-lined beauty of professional pickups. Employs QT Cartridge with replaceable precious metal or sapphire needle. Flawless reproduction at lower cost.

400-MI-2M-33 MAGNETO-INDUCTION TRANSCRIPTION PICKUP—Identical to 400-QT-33, except for Magneto-Induction Cartridge.

THE Astatic CORPORATION
CONNEAUT, OHIO
IN CANADA: CANADIAN ASTATIC LTD. TORONTO ONTARIO

Astatic Crystal Devices Manufactured Under Brush Development Co. Patents

Positions Wanted

(*Continued from page 55A*)

vertisements, displays, sales aids, technical literature. Desires connection as advertising manager or assistant sales manager for medium sized New York City area electrical concern. Box 200 W.

JUNIOR ENGINEER

R.C.A. Institute graduate desires position in production, design or allied fields in New York City vicinity. Age 27. Married. 1 year laboratory and 2 years flight (radio) experience. First class radiotelephone license. Call Da. 6-7203 or write Box 201 W.

TELEVISION ENGINEER

B.S.T.E. Age. 25. Married. First class radiotelephone license. Desires position in television station or development work. Trained in operation and maintenance of R.C.A. Image Orthicon, DuMont equipment and very high frequency techniques. Box 205 W.

JUNIOR ENGINEER

Graduate 2 year course television engineering. Married. Age 25. First class FCC license. Trained in all phases of television studio work. Desires position in television broadcasting field. Box 206 W.

JUNIOR ENGINEER

Syracuse University. B.E.E. June 1948. Age 26. Married with no children. Desires work with power company or motor manufacturing company located in the east. Prefer training program if possible. Interested in transmission and mathematical design. Box 207 W.

ENGINEER

University of Minnesota, communications major. B.E.E. with distinction, August 1948. 2½ years electronics experience in U.S. Army. Desires position in production or electronic development. Will work anywhere in U.S. Box 208 W.

ELECTRONICS ENGINEER

Will graduate March 1949 Iowa State College B.S.E.E. in communications. Married. Age 25. First class Radio Telephone license. Some servicing experience. Desires position in radio or electronics anywhere in U.S. Box 209 W.

ENGINEER

B.S.E.E. Northeastern University, Boston 1947. Two years experience as Navy radio technician with Navy radar and communication equipments. Valuable experience in UHF antenna and radiation research and development at Naval Air Test Center, Patuxent River, Maryland. Desires position in research and design of antennae. Box 210 W.

JUNIOR ENGINEER

B.S. Television Engineering, American Television Institute of Technology, January 1949. Age 25. Married, no children. Three years experience on Navy radar. Desires position in microwave research. Anywhere in U.S. Box 211 W.

JUNIOR ENGINEER

R.C.A. Institute's graduate seeks position in research and development, or production field in New York area. Age 29. Married. Studying for engineering degree at night. Air Force officer, 2½ years experience as instructor in bomb-sight and auto-pilot theory and operation. Box 212 W.

30 MC I.F. STRIPS

Overall gain: 25 db or more.
Bandwidth: 4 plus or minus .4 mc @ 3 db down.
Center freq: 30 plus or minus .5 mc.
Current drain: 30 plus or minus 5 ma.
New, less tubes $17.50

MICROWAVE PLUMBING

10 CENTIMETER

MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated $45.00
10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or d.c. arranged switching motor. Mfg. Raytheon. CRP 24AAS New and complete $150.00
10 CM. END-FIRE ARRAY POLYRODS ...$1.75 ea.
"S" BAND Mixer Assembly, with crystal mount, pickup loop, tunable output $3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers $12.50
10 CM. McNALLY CAVITY Type SG $3.50
WAVEGUIDE SECTION. MC 445A, rt. angle bend, 5½ ft. OA. 8" slotted section $21.00
10 CM. OSC. PICKUP LOOP, with male Homedell output $2.00
10 CM. DIPOLE WITH REFLECTOR in lucite ball, with type "N" or Sperry fitting $4.50
10 CM. FEEDBACK DIPOLE antenna, in lucite ball, for use with parabola $8.00

⅞" RIGID COAX—⅜" I.C.

RIGHT ANGLE BEND, with flexible coax output pick-up loop $8.00
SHORT RIGHT ANGLE bend, with pressurizing nipple $3.00
RIGID COAX to flex coax connector $3.50
STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length $5.00
RT. ANGLES FOR ABOVE $2.50
⅞" COAX. ROTARY JOINT $8.00
RT. ANGLE BEND 15" L. OA $3.50
FLEXIBLE SECTION, 15" L. Male to female .$4.25
MAGNETRON COUPLING to ⅞" rigid coax with TR pickup loop, gold plated $7.50

⅞" RIGID COAX—¼" I.C.

⅞" RIGID COAX, BEAD SUPPORTED per ft. $1.20
SHORT RIGHT ANGLE BEND $2.50
ROTATING JOINT, with deck mounting $6.00
RIGID COAX slotted section CU-60/AP..........$5.00

3 CENTIMETER PLUMBING

(STD. 1" x ½" GUIDE, UNLESS OTHERWISE SPECIFIED)

"X" Band pressurizing gauge section, with 15-lbs. gauge and pressurizing nipple $18.50
45 DEG. TWIST, 6" Long $10.00
12" SECTION, 45 deg. twist, 90 deg. bend $6.00
11" STRAIGHT WAVEGUIDE section choke to cover. Special heavy construction, silver plated $4.50
15 DEG. BEND, 10" choke to cover $4.50
5 FT. SECTIONS, choke to cover $14.50
18" FLEXIBLE SECTION $17.50
"E" and "H" PLANE BEND $12.50
BULKHEAD FEED THRU $15.00
"X" BAND WAVEGUIDE, 1¼" x ⅝" OD. 1/16" wall, aluminum ....per ft. $ .75
WAVEGUIDE, 1" x ½" I.D. per ft. $1.50
TR CAVITY for 724-A TR tube $3.50
8" FLEX SECTION, square flange to circular flange adapter $7.50
724 TR tube (41-TR-1) $2.50
WAVEGUIDE SECTION, CG 251/APS-15A, 26" long choke to cover, with 180 deg. bend of 2½" rad. at one end $6.00
SWR MEAS. SECTION, 4" L, with 2 type "N" output probes MTD full wave apart. Bell size guide. Silver plated $10.00
ROTARY JOINT with slotted section and type "N" output pickup $8.50
WAVEGUIDE SECTION. 12" long choke to cover, 45 deg. twist & 2½" radius, 90 deg. bend $4.50
SLUG, TUNER/ATTENUATOR, W.E. guide, gold plated $6.50
TR/ATR DUPLEXER section with iris flange ..$8.00
TWIST 90 deg., 5" choke to cover, w/press nipple $6.50
WAVEGUIDE SECTIONS 2½ ft. long, silver plated, with choke flange $5.75
WAVEGUIDE. 90 deg. bend E plane, 18" long ..$4.00
ROTARY JOINT, choke to choke $6.00
ROTARY JOINT, choke to choke, with deck mounting $6.00
S-CURVE WAVEGUIDE, 8" long cover to choke $3.50
DUPLEXER SECTION for 1B24 ....$10.00
CIRCULAR CHOKE FLANGES, solid brass .... .55
SQ. FLANGES, FLAT BRASS ....ea. .55
APS-10 TR/ATR DUPLEXER section with additional iris flange $10.00
CU 105/APS 31 Directional coupler, 25 db $15.00
CU 106/APS 33 Directional coupler, 25 db $15.00
CG 176/AP Directional coupler, 20 db $18.00
FLEX. WAVEGUIDE ....$4.00/Ft.
"X" BAND calibrated attenuator $85.00
SHIELDED KLYSTRON tube mounts with rough attenuator outputs $90.00
2½" FLEXIBLE SECTION, cover to cover $5.00
SHORT ARM "T" section, with additional choke output on vertical section $4.00

COAX CABLE

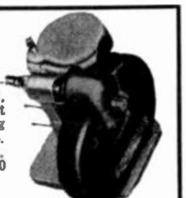
RG 18/U. 52 ohm im. armored ....$ .51/ft.
RG 23/U. twin coax. 125 ohm imp. armored ..$ .50/ft.
RG 28/U. 50 ohm imp. pulse cable. Corona min. starting voltage 17 KV ....$ .50/ft.
RG 35/U. 70 ohm imp. armored ....$ .50/ft.

1.25 CENTIMETER

MITRED ELBOW cover to cover $4.00
TR/ATR SECTION choke to cover $4.00
FLEXIBLE SECTION 1" choke to choke $5.00
KBAND Rotary joint $45.00
ADAPTER, rd. cover to sq. cover $5.00
MITRED ELBOW and 8 sections choke to cover .$4.50

AN/CPN-6, 3 CM RADAR BEACON INSTALLATIONS

MAGNETRONS



2J41 Magnetron-Magnet-Stabilizer Pkg. 9290-9330MC. 1.25 KW Pk. Pulse Output Power. 100MC Tuning Range. Refer Rad. Lab. Series Vol. 6, Pg. 766 ....(as shown) $75.00

TUBE	FRQ. RANGE PK.	PWR. OUT	PRICE
2J31	2820-2860 mc.	265 KW.	$25.00
2J21-A	9345-9405 mc.	50 KW.	$25.00
2J22	3267-3333 mc.	265 KW.	$25.00
2J26	2992-3019 mc.	275 KW.	$25.00
2J27	2965-2992 mc.	275 KW.	$25.00
2J32	2780-2820 mc.	285 KW.	$25.00
2J38 Pkg.	3249-3263 mc.	5 KW.	$25.00
2J39 Pkg.	3267-3333 mc.	87 KW.	$25.00
2J55 Pkg.	9345-9405 mc.	50 KW.	$25.00
2J61	3000-3100 mc.	35 KW.	$65.00
2J62	2914-3010 mc.	35 KW.	$65.00
3J31	24,000 mc.	50 KW.	$55.00
5J30			$39.50
714AY			$25.00
718DY			$25.00
720BY	2800 mc.	1000 KW.	$50.00
720CY			$50.00
725-A	9345-9405 mc.	50 KW.	$25.00
730-A	9345-9405 mc.	50 KW.	$25.00

Klystrons: 723A/B $12.50 707B W/Cavity $20.00

MAGNETS

For 2J21, 725-A, 2J22, 2J26, 2J27, 2J31, 2J32, and 3J31 ....Each $8.00
4850 Gauss, ⅝" bet. pole faces, ¾" pole diam. $8.00
1500 Gauss, 1⅛" bet. pole faces, 1⅝" pole diam. $8.00

TUNABLE PKGD. "CW" MAGNETRONS

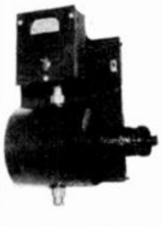
QK 61	2975-3200 mc.	QK 62	3150-3375 mc.
QK 60	2800-3025 mc.	QK 59	2675-2900 mc.

New, Guaranteed ....Each $65.00

GREAT TUBE VALUES

01-A	$ .45	5FP7	3.50	532	3.95	1629	.35
1B24	4.85	5JP2	8.00	559	4.00	1961	5.00
1H5	.55	5J30	$39.50	562	90.00	8012	3.95
1N5	.69	6AC7	1.00	615	.89	9002	.65
1T4	.69	6G	2.00	703-A	7.00	9004	.47
2C21	.69	6K7	.55	704-A	.75	9006	.47
2C22	.69	6L6GA	1.00	705-A	2.85	CEQ 72	1.95
2J21-A	25.00	6SC7	.70	†707-B	20.00	EF 50	.79
2J22	25.00	6SL7	1.00	714AY	25.00	F-127	20.00
2J26	25.00	6V6	.79	715-B	12.00	FC 258A	165.00
2J27	25.00	7C4	1.00	720BY	50.00	FC 271	40.00
2J31	25.00	7E5	1.00	720CY	50.00	GL 562	75.00
2J32	25.00	10Y	.60	721-A	3.60	GL 623	75.00
2J38	25.00	12A6	.35	723-A/B	12.50	GL 697	75.00
2J39	25.00	12GP7	14.95	724B	1.75	ML 100	60.00
2J55	25.00	12K8Y	.85	725-A	25.00	QK 59	65.00
3J31	55.00	12SF7	.49	726-A	15.00	QK 60	65.00
2X2/879	.69	12SR7	.72	800	2.25	QK 61	65.00
3A4	.65	15R	1.40	801-A	1.10	QK 62	65.00
3BP1	2.25	28D7	.75	804	9.95	*RCA932	.65
3C24	.60	30 (Spec.)	.70	815	2.50	VR 91	1.00
3C30	.70	45 (Spec.)	.59	836	1.15	VR 130	1.25
3D6	.79	39/44	.49	837	1.95	VR 135	1.25
3CP1/S1	3.50	35/51	.72	843	.59	VR 137	1.25
3D21-A	1.50	227A	3.85	860	15.00	VU 120	1.00
3DP1	2.25	225	8.80	861	40.00	VU 134	1.00
3EP1	2.95	268-A	20.00	874	1.95	WL 532	4.75
3FP7	1.20	355-A	19.50	876	4.95	WN 150	3.00
3GP1	3.50	417A	22.50	1005	.35	WT 260	5.00
3Q5	.79	539	$90.00	1613	.95	†with cavity	5.00
5BP1	1.20	531	45.00	1619	.21	*Photocell	
5BP4	4.95			1624	$ .85		
5CP1	3.75						

MICROWAVE TEST EQUIPMENT



TS-238 GP. 10 cm. Echo box with resonance indicator and micrometer adjust cavity. 2700 to 2900 Mcs calibrated ...$85.00
TS 108-AP dummy load $65.00
W. E. I 138. Signal generator. 2700 to 2900 Mc range. Lighthouse tube oscillator with attenuator & output meter. 115 VAC input, reg. Pwr. supply. With circuit diagram $50.00
3 cm. wavemeter: 9200 to 11,000 mc transmission type with square flanges $15.00
3 cm. stabilizer cavity, transmission type ..$20.00
3 cm. **Wavemeter. Micrometer** head mounted on X-Band guide. Freq. range approx. 7900 to 10,000 Mc. $75.00

COAX CONNECTORS

831SP	$.35	UG 254/U	$.75
831AP	$.35	UG 255/U	$1.25
831HP	$.15	UG 146/U	$1.00
UG 21/U	$.85	UG 85/U	$1.25
UG 86 U	$.95		

Homedell male to type "N" male adapter $1.25
D-166366 Baby "N" ....$ .85
Adapter Cable Ass'y, Type "N" Male to Type "N" Female $2.25
Adapter Cable Ass'y, Sperry Male to Type "N" Male $2.25
Connector, #49579 for RG 18/U $1.80

VARISTORS W.E.

D-171121	$.95	D-168549	$.95
D-171631	$.95	D-162482	$3.00
D-167176	$.95	D-99136	$1.65
D-168687	$.95	D-166271	$2.50
D-171812	$.95	D-162356	$1.50
D-171528	$.95	D-161871A	$2.85
D-163298	$.95	D-99946	$2.00

THERMISTORS—W.E.

D-167332 (tube)	$.95	D-164699 FOR MTG. in "X" Band Guide	$2.50
D-170396 (bead)	$.95	D-167018 (tube)	$.95
D-167613 (button)	$.95		
D-166228 (button)	$.95		

MICROWAVE GENERATORS

AN/APS-15A "X" Band compl RF head and modulator, incl. 725-A magnetron and magnet, two 723A/B klystrons (local osc & beacon), 1B24 TR, revr-ampl, duplexer, HV supply, blower, pulse xfmr. Peak Pwr Out: 45 KW apx. Input: 115, 400 cy. Modulator pulse duration .5 to 2 micro-sec apx. 13 KV Pk Pulse. Compl with all tubes incl. 715-B, 829B, RKR 73, two 72's. Compl pkg. new $210.00

APS-15B. Complete pkg. as above, less modulator $150.00

PULSE EQUIPMENT

MODULATOR UNIT BC 1203-B

Provides 200-4,000 PPS. Sweep time: 100 to 2,500 microsec. in 4 steps, fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse, sweep voltages, calibration voltages, fil. voltages. Operates 115 vac, 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio #804B or #804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New $125.00
MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 114 KW (12 KV at 12 amp). Duty Ratio: .001 max. Pulse duration: .5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-'72's, 1-'73. New $110.00
APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro Sec. Rep. rate 624 to 1348 Pps. Pk. pwr. out 35 KW. Energy 0.018 Joules $49.00
TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50
APS-10 MODULATOR DECK, Complete, less tubes $75.00
APS-10 Low voltage power supply, less tubes ..$18.50

PULSE NETWORKS

G.E. #25E5-1-350-50P2T, 25 KV, 5 sections, "E" circuit, 1 microsecond pulse length, 350 PPS, 50 ohms impedance $45.00
G.E. #6E3-5-2000-50P2T, 6KV, "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance $6.50
G.E. #3E (3-.84-810; 8-2.24-405) 50P4T: 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec, 810 PPS, 50 ohms imp.: Unit 2, 8 Sections, 2.24 Microsec, 40 PPS, 50 ohms imp. $6.50

PULSE TRANSFORMERS

W.E. #D166173 HI-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc to 2 mc. 2 sections parallel connected, potted in oil $12.00
W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1. Frequency range: 380-520 c.p.s. Permalloy core $2.00
G.E. #K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms, Sec. Imp: 450 Ohms. Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK. Sec. Output: 28 KV PK. Peak Output: 800 KW. Bifilar 2.75 Amp. $19.50
W.E. #D169271 Hi Volt input pulse Transformer..$9.95
G.E. K2450A Will receive 13KV. 4 micro-second pulse on pri., secondary delivers 14KV. Peak power out 100KW GE $15.00
G.E. #K2748A Pulse Input, line to magnetron ..$12.00
#9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1½" 19/32 ....$1.50

PE218 INVERTERS

Input: 25-28 VDC @ 92 amps. Output: 115 volts @ 1500 volt-amps. 380-500 cycles. New, Hermetically Sealed $49.95

MICROWAVE ANTENNAS

AN MPG-1 Antenna. Rotary feed type high speed scanner antenna assembly, including horn parabolic reflector. Less internal mechanisms, 10 deg. sector scan. Approx. 12'L x 4'W x 3'H. Unused. (Gov't Cost—$4500.00) $280.00
APS-4 3 cm. antenna. Complete. 14½" dish. Cutler feed dipole directional coupler, all standard 1" x ½" waveguide. Drive motor and gear mechanisms for horizontal and vertical scan. New, complete ..$65.00
AN/TPS-3. Parabolic dish type reflector approx. 10' diam. Extremely lightweight construction. New, in 3 carrying cases $89.50
RELAY SYSTEM PARABOLIC REFLECTORS: approx. range: 2000 to 6000 mc. Dimensions: 4½' x 3', rectangle, new $85.00
TDY "JAM" RADAR ROTATING ANTENNA, 10 cm. 30 deg. beam. 115 v.a.c. drive. New $100.00
SO-13 ANTENNA, 24" dish with feedback dipole 360 deg. rotation, complete with drive motor and selsyn. New .... $120.00 Used $45.00

All merchandise guaranteed. Mail orders promptly filled. All prices, F.O.B. New York City. Send Money Order or Check. Only shipping charges sent C.O.D. Rated Concerns send P.O. Prices subject to change without notice.

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Crystal users appreciate the complete service James Knights Co. offers.

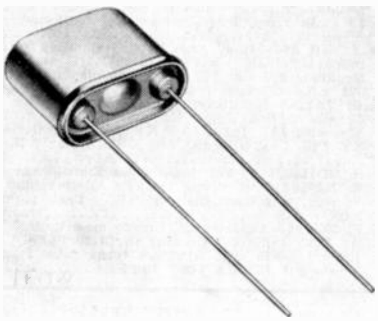
If you have a special crystal problem, James Knights Co. is equipped to build crystals to your exact specifications—no matter what they may be. Because of a special production line for short runs, the price is right—whether you need one, ten, or several thousand crystals!

In addition, James Knights Co. fabricates a complete line of "Stabilized" crystals to meet every ordinary need—precision built by the most modern methods and equipment.

Fast service is yours, too! Two company planes save hours when speed is important.

Your inquiries—and crystal problems—are invited.

Send For New James Knights Co. Catalog



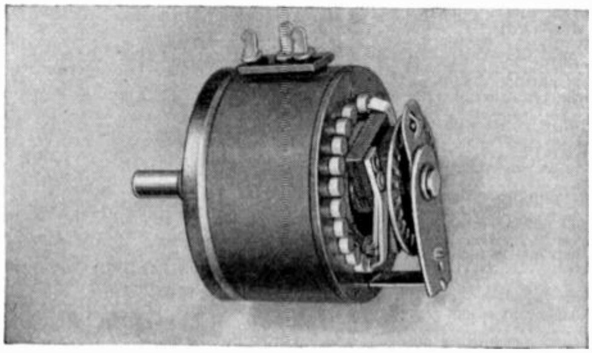
A large AIRCRAFT RADIO MANUFACTURER needed extremely small and light weight 3105 kc crystals. We designed and built one that weighed less than two ounces, now our Type H-17W.

The JAMES KNIGHTS Co.

SANDWICH, ILLINOIS

JK CRYSTALS

Shallcross ATTENUATORS



BRIDGED 'T' ATTENUATOR

Type 410-4B1

10 steps, 4 db/step. Linear attenuation with detent. 2⅛" diameter, 2 1/16" depth.

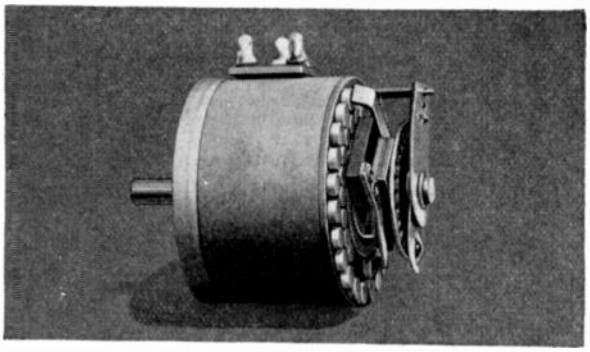
$11.50
LIST PRICE

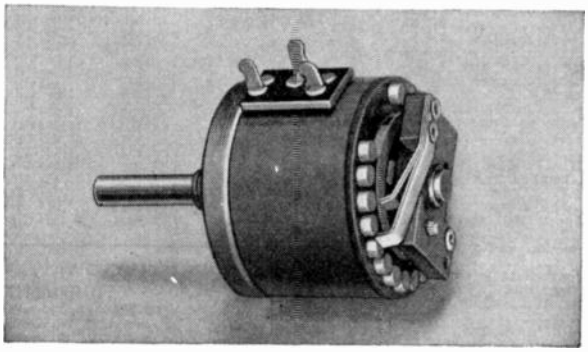
BRIDGED 'T' ATTENUATOR

Type 420-2B2

20 steps, 2 db/step. Linear attenuation with off position and detent. 2⅛" diameter, 2 1/16" depth.

$16.00
LIST PRICE





POTENTIOMETER

Type C720-2A3

20 steps, 2 db/step, tapered on last three steps to off, composition resistors. 1¾" diameter, 1¾" depth.

$8.00
LIST PRICE

These Shallcross Features Mean Better Performance —Better Value!

Off position attenuation well in excess of 100 db.

25% to 50% fewer soldered joints.

Noise level ratings that are factual. (130 db. or more below zero level.)

Non-inductive Shallcross precision resistors used throughout assure flat attenuation to and beyond 30 kc.

Types and sizes engineered for all needs. Attentuation accuracies of 1%. Resistor accuracies of 0.1%, on special order.

SHALLCROSS ATTENUATORS

Shallcross variable attenuators have proved their remarkable quietness and serviceability in dozens of applications for leading users in all parts of the world. Such important details as the use of spring-temper silver alloy wiper arms, silver alloy collector rings and contacts, non-inductive precision resistors, and sturdy, substantial mounting plates have made possible the high standard of performance attributed to Shallcross.

Standard types include ladder and bridged T mixer controls, bridged T and straight T master gain controls and V.U. meter multipliers, wirewound and composition potentiometers for grid control. Cueing attenuators, and fixed pads, both composition and wirewound, in all circuit configurations are also available.

Write for Catalog and Attenuator Specification Sheet

SHALLCROSS MANUFACTURING COMPANY

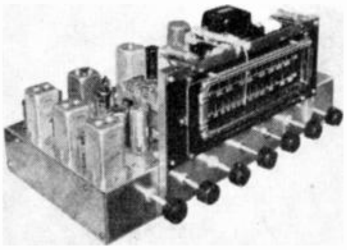
Department PR-128 Collingdale, Pa.

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)

FM and AM Tuner



Browning Laboratories, Winchester, Mass., has just announced their latest model RJ-20 FM and AM tuner, with inbuilt preamplifier and tone-compensating networks.

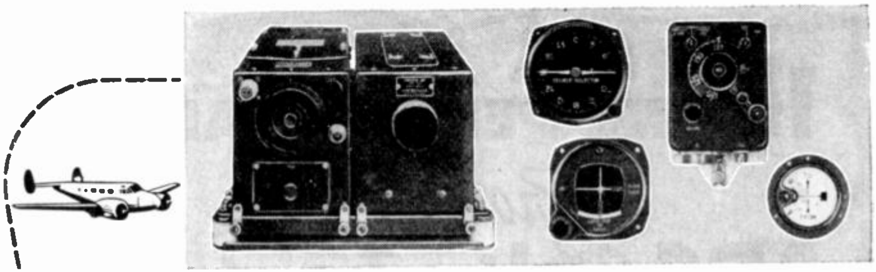
This is a precision type of instrument appealing to the laboratory type of application, the custom builder of equipment, and for broadcast monitoring of program content and station performance. Two separate tuners are mounted on one chassis, the AM section incorporating a set of band-expanding if transformers, with an 8-kc and 18-kc bandwidth position, controlled from a panel switch. The FM section employs separate oscillator and mixer triode tubes, dual limiters and discriminator in a fully licensed Armstrong circuit, for finest FM reception. Sensitivity is such that 10 microvolts input signal will produce 32 db of quieting at full audio fidelity.

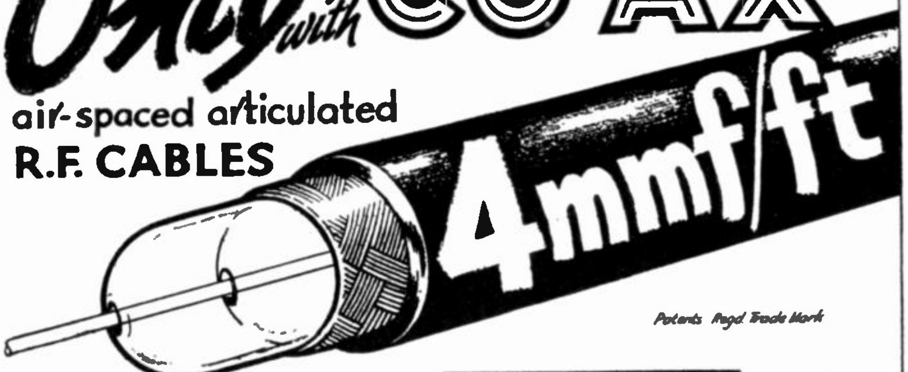
The audio amplifier is a two-stage cascade unit, common to both tuning sections, and also providing a phonograph input channel. Between the two tubes are a series of $R-C$ tone compensating networks, to provide for attenuation and equalization of both ends of the audio spectrum. As the phonograph input operates through this system, it can be equalized too. A power supply for the tuner is integral with the chassis.

Plant Expansions

Tech Laboratories, Inc., has announced the purchase and occupancy of their own building at Bergen & Edsall Boulevards, Palisades Park, N. J. Long occupying rented space in Jersey City, their increased sales volume necessitated this move to more than 17,000 square feet of office and production space. The location is 1½ miles south of George Washington bridge, on route S-1. On the hilltop location, unobstructed light is afforded to all areas of the 100% air-conditioned building. These factors will contribute to the high standards of technical excellence and engineering integrity long recognized as a feature of the attenuators, resistances, and switches associated with this firm. Increased laboratory facilities for special contract work have been provided in the new quarters.

(Continued on page 61A)

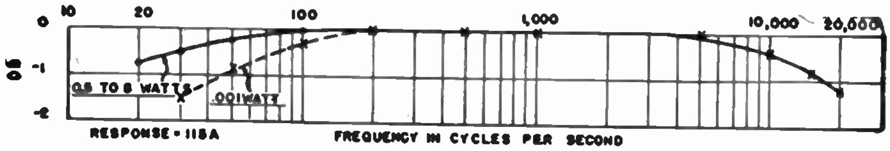


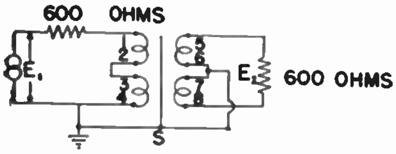


1st LINE PERFORMANCE *Proved in* ADC 2nd Line Transformer

An **ADC** 115A (Industrial Series) impedance matching transformer, picked at random from stock, was submitted to tests to compare its performance with that of other makes of *1st line* transformers. Here are the results. Compare performance of the **ADC** transformer with that of other makes.

FREQUENCY RESPONSE

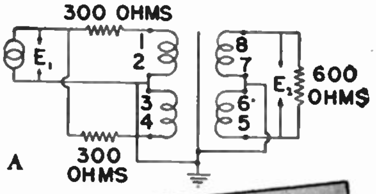




It may be noted that altho the permeability of magnetic materials drops at low flux densities, the **ADC** transformer has sufficient reserve inductance to allow for this even at low power levels. At 40 db below maximum power level it exceeds the response guarantee. Insertion loss at 1,000 cps was 0.75 db.

LONGITUDINAL BALANCE

The most common interference voltages encountered in telephone line transmission are longitudinal; that is, the induced voltages in both wires are in phase with respect to ground. These can be removed from the signal voltage only by means of a well balanced line transformer. Illustration "A" shows the test circuit used to measure the degree of removal of these interference voltages. Level reduction on the **ADC** 115A transformer was 67 db at 100 cps and 56 db at 10,000 cps.



A

CONSULT **ADC** for your engineered transformer where exacting specifications require positive results. **ADC**'s policy assures you the finest available materials and workmanship to give you the very best electronic components.

ADC QUALITY PLUS TRANSFORMERS

Finest transformer made. For AM and FM broadcast stations and recording studios. ± ½ db 30-15,000 cps.

MANUFACTURERS, JOBBERS:
Write today for catalog of ADC electronic components or for information on units engineered to your requirements.

AUDIO DEVELOPMENT

Audio Development Co.

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"Audio Develops the Finest"

RUGGED!

POWER RHEOSTATS

★ Yes, RUGGED — mechanically and electrically. That's why Clarostat Power Rheostats are found in so many radio-electronic and electrical assemblies that must stand up.

Insulated metal core for winding. Element imbedded in exclusive cold-setting cement. Maximum heat dissipation for cooler, longer-lasting operation. Smoothest rotation. Positive conduction, always.

25- and 50-watt ratings. 1-5000 and 0.5-10,000 ohms, respectively.

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Write for Bulletin 115. Let us quote on your control and resistor needs.

Controls and Resistors

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In Canada: CANADIAN MARCONI CO., Ltd.
Montreal, P.Q., and branches

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 59A)

Plant Expansions

Sylvania Electric Products Inc., commenced production about October 1 in their new television tube manufacturing plant located at Ottawa, Ohio. Increased interest over the nation in television, translated into increased receiver production and sales, has called for the construction of increased capacity for making of cathode-ray tubes used in direct viewing receiving sets. Opening of this new plant will supplement the output of the two factories currently in production at Emporium, Pa., and afford facile distribution of the finished product to the makers in the Chicago area, largest center of video receiver production, in this country.

At Waterford, N. Y., the **General Electric Co.** has opened a new plant for the production of its many silicone products to commence production at full capacity, this fall. Partial operation has been going on for about a year. Many forms of this new series of chemical products will be formulated at this new manufacturing facility, including the famous "bounding putty" used as a center material for golf balls, with signal success.

(Continued on page 64A)

DIALCO

PILOT LIGHT ASSEMBLIES

PLN SERIES—Designed for NE-51 Neon Lamp

Features

- THE MULTI-VUE CAP
- BUILT-IN RESISTOR
- 110 or 220 VOLTS
- EXTREME RUGGEDNESS
- VERY LOW CURRENT

Write for descriptive booklet

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BUILD BETTER ELECTRONIC EQUIPMENT *with* STANDARDIZED

PAR-METAL ELECTRONIC HOUSINGS

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Par-Metal Equipment is preferred by Service Men, Amateurs, and Manufacturers because they're *adaptable, easy-to-assemble, economical.* Beautifully designed, ruggedly constructed by specialists. Famous for quality and economy.

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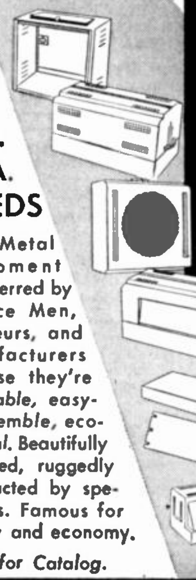


TWIN Power Supply

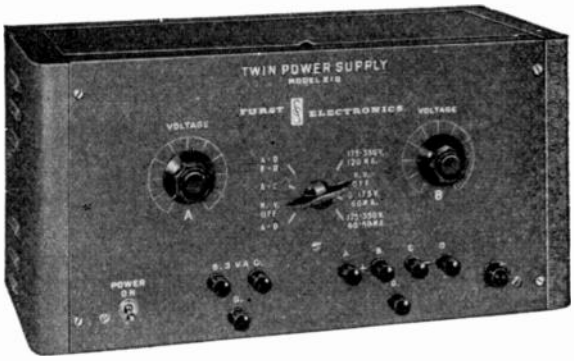
Electronically Regulated for Precise Measurements

Two independent sources of continuously variable D.C. are combined in this one convenient unit. Its double utility makes it a most useful instrument for laboratory and test station work. Three power ranges are instantly selected with a rotary switch:
175-350 V. at 0-60 Ma., terminated and controlled independently, may be used to supply 2 separate requirements.
0-175-V. at 0-60 Ma. for single supply.
175-350 V. 0-120 Ma. for single supply.
In addition, a convenient 6.3 V.A.C. filament source is provided. The normally floating system is properly terminated for external grounding when desired. Adequately protected against overloads.

- *Output voltage variation less than 1% with change from 0 to full load.*
- *Output voltage variation less than 1 V. with change from 105 to 125 A.C. Line Voltage.*
- *Output ripple and noise less than .025 V.*

Twin Power Supply Model 210
Complete $130.00
Dimensions: 16" X 8" X 8" Shipping Wt. 35 lbs.
(Other types for your special requirements)

FURST ELECTRONICS
North Avenue at Halsted St., Chicago 22, Illinois



ANNOUNCING! . . .

NEW SLIDE WIRE RESISTANCE BOXES!

Technology Instrument Corporation's newly developed Type 110 Slide Wire Resistance Boxes represent a big step forward in the design of specialized instruments for student and general laboratory lse. A combination of high accuracy, wide resistance range and convenient size, the Type 110 is suitable for use at audio and super-sonic frequencies.

Its compactness combined with its low cost make it possible to provide more of these important instruments for college and industrial laboratories. Yet it is suitable for use in most cases where a more elaborate decade box is used.

The Type 110 Slide Wire Resistance Boxes consist of a precision non-inductive decade resistor and a continuously adjustable slide wire resistor which provide a useful, direct-reading resistance range ratio of 1000 to 1. Two models are now available: Type 110-A, with a range of 0-11,000 ohms; and Type 110-B, with a range of 0-110,000 ohms. Send a trial order today—or if you prefer, ask for detailed information.

SPECIFICATIONS

Accuracy—Decade resistance cards adjusted to within 1% of nominal values. Slide wire resistors direct-reading to within 1% of maximum resistance.

Temperature Co-efficient—Slide wire and decade resistors have temperature co-efficient of less than 0.00002 parts per degree C. at room temperature.

Mounting—Cast aluminum cabinet, aluminum panel. All resistance elements and switches completely enclosed. Dimensions: 4" wide 8⅝" long, 5⅛" high. Net weight, 4 lbs.

PRICES

Type 110-A: 0-11,000 ohms, 2 dials $42.50
Type 110-B: 0-11,000 ohms, 2 dials $45.00

TECHNOLOGY INSTRUMENT CORP.
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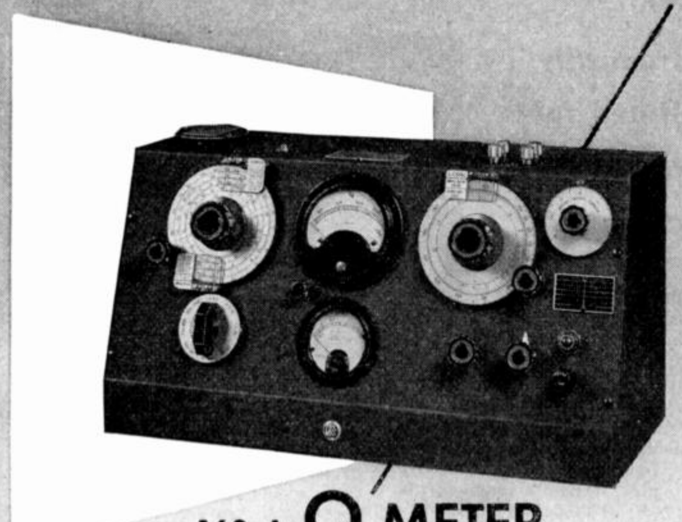
Midwest Office: Alfred Crossley & Associates, Chicago, Ill., Phone STate 2-7444
Eastern Offices: Holliday-Hathaway Co., Cambridge, Mass. Phone: ELiot 4-1751, Canaan, Conn., Phone 649

MEASUREMENT • TEST • CONTROL



160-A Q METER

The 160-A Q-Meter is unexcelled for laboratory and development applications, having received world wide recognition as the outstanding instrument for measuring Q, inductance, and capacitance at radio frequencies.

Frequency Range: 50 kc. to 75 mc. (8 ranges)
Q Measurement Range: 20 to 250 (20 to 625 with multiplier)
Range of Main Q Capacitor: 30-450 mmf.
Range of Vernier Q Capacitor: +3 mmf., zero, −3 mmf.

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

For further specifications and descriptive details, write for Catalog F

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Phone: Little Falls 4-1000
Established 1926

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RADIOTELETYPE · FREQUENCY SHIFT
INK SLIP RECORDING · TELETYPE NETWORKS
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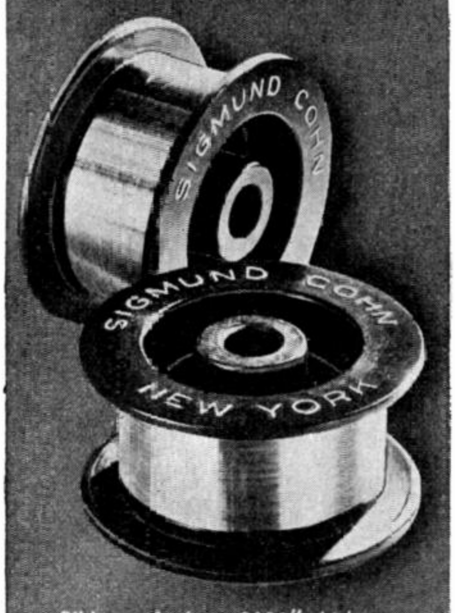
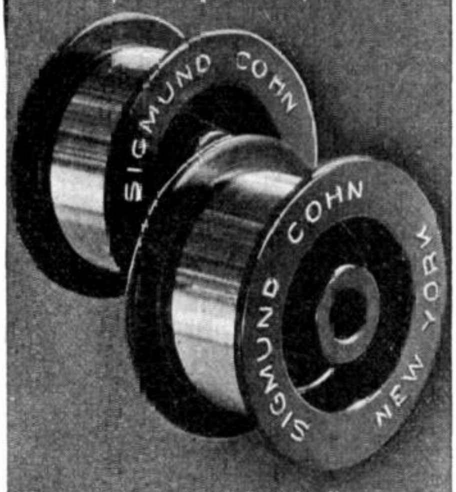
Wires drawn to .0004" diameter.



Ribbon rolled to .0001" thickness.

WIRE & RIBBON
MICRODIMENSIONAL
FOR VACUUM TUBES

Alloys for Special requirements



WRITE for list of stock alloys.

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The Tektronix Type 512 Oscilloscope is a truly NEW quantitative measuring instrument.

The combination of DC amplifiers and single, recurrent or triggered sweeps ranging from 3 seconds to 30 microseconds is of particular interest to geo-physical, mechanical and biological research groups.

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Outstanding Type 512 Features

- Sweep and vertical amplifiers are P.P. and **direct coupled** throughout.
- Vertical amplifier band width 3 mc., 75 volts to 0.25 volt sensitivity; 1 mc. from 0.25 volt to .0075 volt sensitivity.
- DC Amplifier stability is achieved by operating heaters of first two stages from an electronically regulated DC supply.
- Revolutionary carrier type blanking circuit, overcoming deficiencies of capacity coupling for long blanking pulses.
- Delayed trigger pulse, variable over entire length of sweep, available at front panel.
- Sweep time calibration accuracy 5%. Conveniently read directly from dial, obviating need for timing markers.
- Any 20% of sweep may be expanded 5 times for detailed signal study.
- All DC voltages, including accelerating potential, electronically regulated against line voltage changes.

Price $950 f.o.b. Portland

Your inquiry will bring more detailed information and name of the nearest Field Engineering Representative.

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Cables, TEKTRONIX
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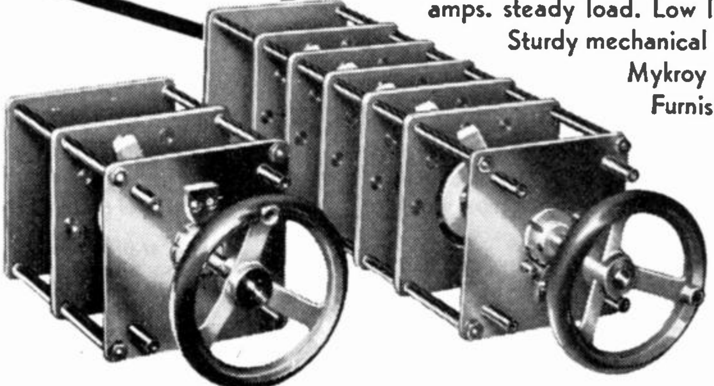
AN IMPORTANT TECH LAB DEVELOPMENT

New Type 1250
R. F. SWITCH

High r. f. current carrying capacity 50 amps. max. intermittent load; 30 amps. steady load. Low loss factor. Sturdy mechanical design . . . Mykroy insulation Furnished in any number of decks.



Write for Bulletin No. 472

TECH LABORATORIES, INC.

Manufacturers of Precision Electrical Resistance Instruments

BERGEN BLVD., PALISADES PARK, N.J.
Tel: LEonia 4-3106

Laboratory and Research Instruments

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SWEEP CALIBRATOR Model GL-22

This versatile source of timing markers provides these requisites for accurate time and frequency measurements with an oscilloscope:

Positive and negative markers at 0.1, 0.5, 1.0, 10, and 100 microseconds ● Marker amplitude variable to 50 volts ● Gate having variable width and amplitude for blanking or timing ● Trigger generator with positive and negative outputs.

Further details in Bulletin RE-812.

OSCILLOSYNCHROSCOPE Model OL-15B

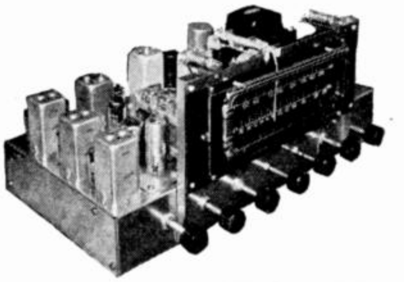
Provides a variety of time bases, triggers, phasing and delay circuits, and extended-range amplifiers in combination with all standard oscilloscope functions.

Extended-range amplifiers: vertical, flat within 3 db 5 cycles to 6 megacycles; horizontal, flat within 1 db 5 cycles to 1 megacycle ● High sensitivity: vertical, 0.05 RMS volts per inch; horizontal 0.1 RMS volts per inch ● Single-sweep-triggered time base permits observation of transients or irregularly recurring phenomena ● Variable delay circuit usable with external or internal trigger or separate from 'scope ● Sawtooth sweep range covers 5 cycles to 500 kilocycles per second ● 4,000-volt acceleration gives superior intensity and definition. Request Bulletin RO-812.

Enjoy High-fidelity

RADIO RECEPTION WITH THE

Browning FM or FM-AM *Tuner*



This is the new Model RJ-20 FM-AM Tuner . . . designed for high-fidelity reception on both FM and AM, and built to meet your highest performance requirements. Its features include:

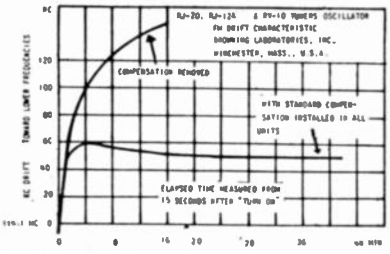
- Armstrong FM circuit for maximum noise reduction and full frequency response to 15,000 cycles.
- Separate RF and IF systems for FM and AM . . . no coil switching.
- Variable bandwidth IF gives AM bandwidths from 9 kc. to 4 kc.
- Two-stage audio system allows 20 db. boost in bass or treble range.
- New 6AL7 tuning eye for precise tuning on strong or weak FM stations.
- Self-contained power supply.

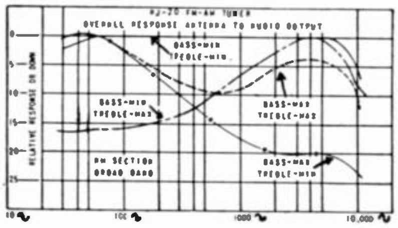
See, hear, and handle this new Browning Tuner . . . and enjoy new satisfaction in your radio and music reproduction.

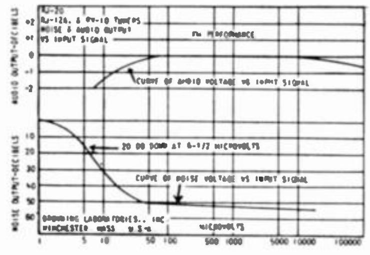
Write today for
Data Sheet RT-812

HERE'S PERFORMANCE to satisfy the man who knows radio . . . provable by both instrument and listening tests.

CHECK THESE CURVES and you'll see why Browning Tuners are chosen by those who insist upon the best.







To feed a separate high-fidelity audio system, use the Browning RJ-12A for FM/AM or the Browning RV-10 for straight FM. They're all "tops" in the high-fidelity field.

BROWNING LABORATORIES, INC.
WINCHESTER, MASS.

ENGINEERED FOR ENGINEERS

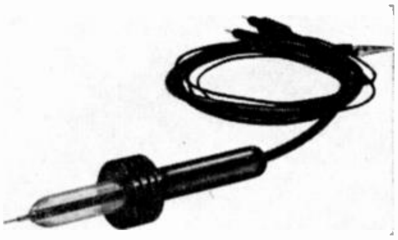
News—New Products

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

(Continued from page 61A)

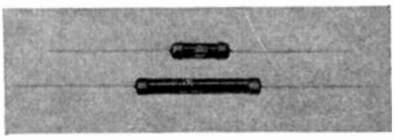
Safety Test Probes

Precision Apparatus Co., Inc., 92-27 Horace Harding Blvd., Elmhurst, L. I., N. Y., has announced high-voltage testing probes, constructed with internal shielding and flashover protection to safeguard the life of the user.



When the proper interchangeable resistor is installed, they will extend the range of many makes of test apparatus to 30 kilovolts, and beyond, still affording complete safety in the hands of the operating personnel. Sensitive test instruments and vacuum-tube voltmeters can be extended in the range over which they will perform through incorporation of these new accessories.

Recent Catalogs



International Resistance Co., 401 N. Broad St., Philadelphia Pa., announce in their bulletin B-4 a new series of deposited carbon resistors, in ranges of 200 ohms to 20 megohms. Multiple layers of lacquer protect the units from mechanical damage, high stability of resistance, and low voltage co-efficient, are advantages claimed for these 1- and 2-watt units. They are supplied with soft copper leads, securely anchored to the silvered end caps.

Ebert Engineering & Manufacturing Co., 185-09 Jamaica Ave., Hollis, L. I., N. Y., announce data on their line of high current-carrying-capacity mercury relays, approved by the Underwriters Laboratories. These are available in 2- and 3-pole units, the current-carrying parts being totaly glass-enclosed, and capable of handling up to 35 amperes. Coil operation is designed for d.c. or half-wave rectified, unfiltered a.c. Small physical size increases designer interest in this new line, where space requirements impose rigid limitations.

(Continued on page 65A)

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We specialize in Equipment and Methods for the Manufacture of

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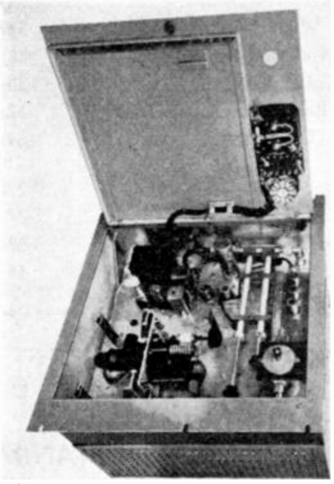
News—New Products

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

(Continued from page 64A)

STL FM Broadcast Transmitter

Federal Telephone & Radio Corp., 100 Kingsland Road, Clifton, N. J., has entered the studio-to-transmitter radio link market with the announcement of their 3-watt direct-frequency-modulated klystron-oscillator transmitter cut shown, and receiver units, incorporating parabolic antenna units. Because of this feature, low power will afford reliable line-of-sight transmission over approximately 30 miles. The single-superheterodyne receiver, employing a similar klystron as a local oscillator, and with automatic-frequency-control circuits, affords a stability of 0.01%.



Provision on both units, which are relay rack mounted, is made for full metering of tube circuits and aural monitoring. Characteristics meet the RMA and FCC requirements for this service, when used as a broadcasting station studio-to-transmitter location link. Small physical size and accessibility of all components are features that will appeal to the operating and servicing personnel of the station using this equipment.

Recent Catalogs

American Phenolic Corp., 1830 S. 54 Ave., Chicago 50, Ill., have issued Bulletin A-1, a complete listing of their line of AN connectors for power, signal and control circuits in aircraft and electronic applications. Thousands of types are listed in detail, under group headings, together with an index to locate specific fittings by their designation numbers. Assembly data on many types are included, as well as a listing showing some typical cable-harness arrangements which have been built at this firm's factory to specific order, using many of the connectors listed in the catalogue. Design engineers will appreciate the engineering factors shown as advantages of the Amphenol line.

keep abreast of the latest developments in

MICROWAVES and RADAR ELECTRONICS

By ERNEST C. POLLARD and JULIAN M. STURTEVANT, both of the Yale University faculty

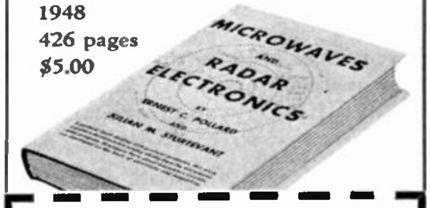
•

This book by two experts in the field is the most up-to-date treatment of the subject of microwave electronics. It not only covers wartime work in radar, but also includes developments that have taken place as recently as the last few months. In this volume, the authors treat pulse circuits as a unified field, and discuss radar as only one of many possible applications of microwave electronics. The clear, concise language of the book makes it understandable to those with only a basic physical background. The rapid progress in the field of microwave electronics, during and since the war, makes this book essential reading for the communication engineer.

Contents

Electromagnetic Fields and Microwaves; Coaxial Lines, Wave Guides, and Cavities; The Production of Microwaves; Microwave Technique; Pulse Circuits; Cathode Ray Tube Indicators; Tubed Amplifiers; Amplification of Very Weak Signals; Servomechanisms and Computers; Miscellaneous Circuits; Radar and Its Accessories; Microwave Communications; Microwaves and Physical Research; Appendix 1—The Fourier Integral. Appendix 2—Curl and Stokes Theorem. Appendix 3—Units.

1948
426 pages
$5.00



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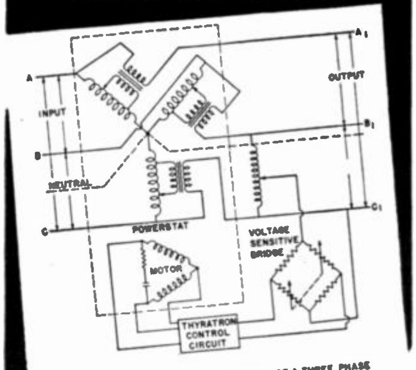

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In the plant—as in the laboratory—constant voltage is a necessity. It can be maintained to large industrial loads by using STABILINE (Electromechanical) Automatic Voltage Regulators.

These units feature zero waveform distortion; complete insensitivity to magnitude and power factor of load; no effect on system power factor; no critical adjustments; high efficiency; adjustable output voltage. Available for a wide range of applications in 115, 208, 230, 440 volt, single and three phase ratings; capacities to 100 KVA.

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INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS

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STANDARD SIGNAL GENERATOR

2 to 400 MEGACYCLES

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FM Signal Generators
Square Wave Generators
Vacuum Tube Voltmeters
UHF Radio Noise & Field Strength Meters
Capacity Bridges
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MODULATION: Amplitude modulation is continuously variable from 0 to 30%, indicated by a meter on the panel. An internal 400 or 1000 cycle audio oscillator is provided. Modulation may also be applied from an external source. Pulse modulation may be applied to the oscillator from an external source through a special connector. Pulses of 1 microsecond can be obtained at higher carrier frequencies.

FREQUENCY ACCURACY ±.5%

OUTPUT VOLTAGE
0.1 to 100,000 microvolts

OUTPUT IMPEDANCE
50 ohms

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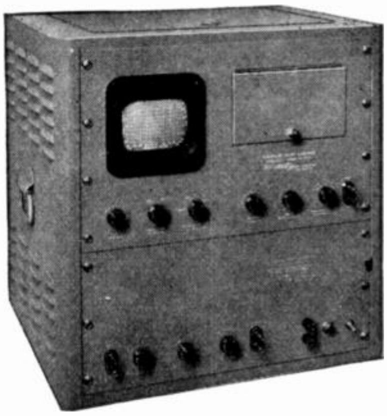

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Model AP-1 assures faster, simpler audio analysis by *automatically* separating the components of complex audio waves and *simultaneously* measuring their frequency and amplitude.

Whether your problem is investigation of harmonics, intermodulation, transmission characteristics, high frequency vibration, noise or acoustics, it will pay to look into the unusual advantages offered by the Panoramic Sonic Analyzer.

- Indications are spectrographic-frequency versus voltage
- Quick overall views of the 40-20,000 cps spectrum are provided once per second
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- Observations of random changes in energy distribution are possible
- Chances of missing components are removed
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Write NOW for complete technical data, price and delivery.

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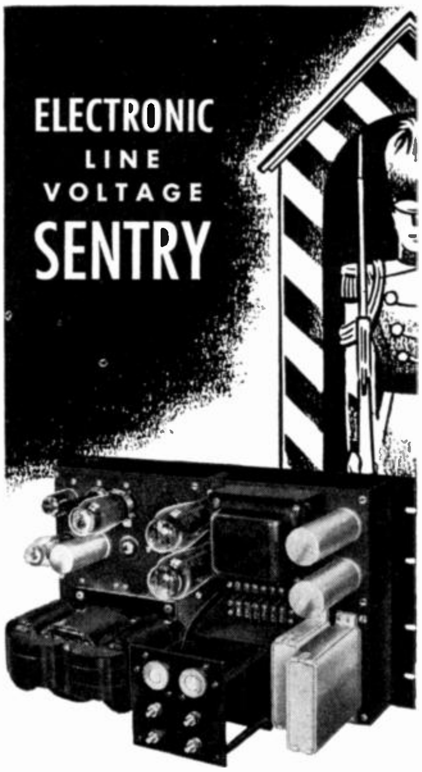

Rear View STABILINE Type IE

The Superior Electric STABILINE Automatic Voltage Regulator acts as a sentry — always "on duty" to maintain a constant voltage to electrical apparatus. Type IE, an *all-electronic* regulator, acts instantaneously to keep delivered voltage to within ±0.1 volts of the preset value, regardless of line variations. STABILINE Type IE will hold the output to within ±0.15 volts for any load current change or load power factor change from lagging .5 to leading .9. Waveform distortion never exceeds 3 percent — a negligible error not recognized in instrument readings.

Bulletin 547 gives you information on this and other Superior Electric voltage control equipment. Write for your copy today.

THE SUPERIOR ELECTRIC COMPANY
812 MEADOW STREET
BRISTOL, CONN.

Making television history, first coverage of air-sea maneuvers demonstrates value of research by RCA Laboratories to our armed forces.

Now television "stands watch" at sea

Picture the advantage—in military operations—when commanding officers can watch planes, troops, ships maneuver at long range . . .

This new use of television was seen by millions when the aircraft carrier *Leyte*—as Task Force TV—maneuvered at sea before a "battery" of 4 RCA Image Orthicon television cameras.

Seventy planes—Bearcats, Avengers, Corsairs—roared from *Leyte's* flight deck and catapult . . . dived low in mock attack . . . fired rockets. And an escorting destroyer stood by for possible rescues.

Action was beamed by radio to shore, then relayed over NBC's Eastern television network. Reception was sharp and clear on home television receivers . . .

Said high officials: "The strategic importance of television in naval, military, or air operations was dramatically revealed" . . . "There is no doubt that television will serve in the fields of intelligence and combat."

Use of television as a means of military communications is only one way in which radio and electronic research by RCA Laboratories serves the nation. All facilities of RCA and NBC are available for development and application of science to national security . . . in peace as well as war.

• • •

When in Radio City, New York, you are cordially invited to see the radio, television and electronic wonders at RCA Exhibition Hall, 36 West 49th Street. Free admission. Radio Corporation of America, RCA Building, Radio City, New York 20.

RCA RADIO CORPORATION of AMERICA

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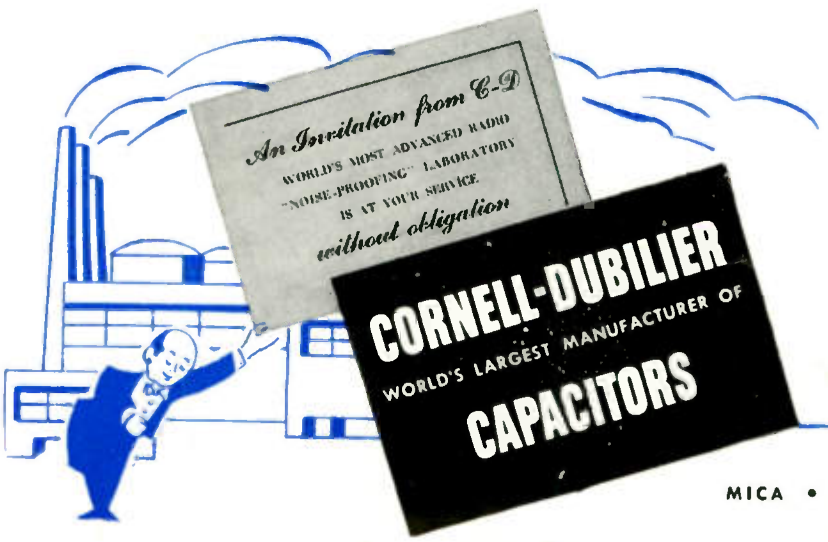
You may build the best appliance of its kind on the market — but if it sets up local radio interference—you'll have tough sledding against today's keen competition. Your customers are *demanding* radio noise-free performance in the electrical equipment they buy.

The answer, of course, is to equip *your* products with C-D Quietones. Why Quietones? First, because they're the best-engineered noise filters — second, because they guard your product's reputation by giving long trouble-free service — third, because they're designed and built to meet manufacturers' specific needs — efficiently and economically.

Speed up sales — build prestige — boost profits with C-D Quietones. Your inquiries are invited. Cornell-Dubilier Electric Corporation, Dept. M-12, South Plainfield, New Jersey. Other large plants in New Bedford, Brookline and Worcester, Mass., and Providence, Rhode Island.

Make Your Product More Saleable with C-D Quietone Radio Noise Filters and Spark Suppressors



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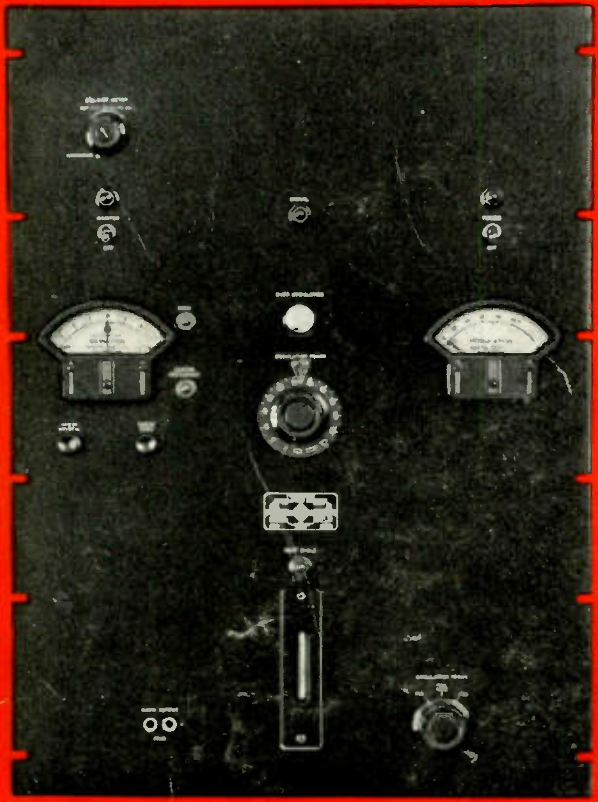
COMPLETE MONITORING EQUIPMENT By GR

for TV and FM TRANSMITTERS

WITH our recent announcement of the Type 1182-T Video Frequency Monitor, complete transmitter monitoring equipment by General Radio Company for TV as well as FM stations is now available. The FM Monitor has FCC approval; the TV equipment meets all of the tentative proposed specifications of the FCC.

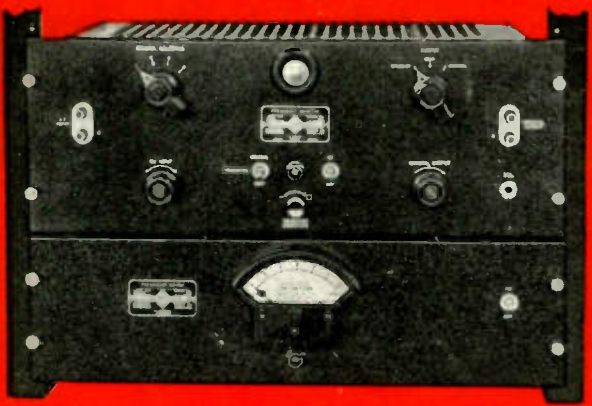
Simple to install, practically self-explanatory in operation, direct-reading in frequency deviation, these monitors have the same high accuracy and stability and years of reliable operation ahead of them that all G-R broadcast monitoring instruments have provided since the beginning of broadcasting.

This equipment can be bought with confidence . . . confidence that it will give trouble-free operation with the minimum of routine maintenance. The G-R trade mark insures that.



TYPE 1170-A FM MONITOR

For both FM and audio channel monitoring of TV stations, features: **TRANSMITTER RANGES** of either 30 to 162 Mc or 160 to 220 Mc. **CONTINUOUS MONITORING:** center-frequency indication continuous; requires restandardization only once a day. **REMOTE MONITORING:** equipped with circuits and terminals for remote indicators of Center-Frequency Indication, Percentage-Modulation Meters, Over-Modulation Lamp, 600-Ohm Unbalanced Aural Monitor, Recorder and with the Type 1932-A Distortion and Noise Meter will measure distortion. **HIGH STABILITY:** 200 cycles (2 parts per million) or better with daily check of electrical zero of meter. **LOW INPUT POWER:** 1 volt at high impedance, amplified to several hundred volts for high-level operation. **LOW RESIDUAL DISTORTION:** less than 0.2% at 100 k.c. swing; accurate for measurements to as low as ½ per cent. ***75-25 KC DEVIATION*** provided with a single internal adjustment for either TV or FM monitoring. **TYPE 1170-A FM MONITOR . . . $1625**



TYPE 1182-T VIDEO MONITOR

For video channels of TV transmitters, features include: **TRANSMITTER RANGE:** 54 to 220 Mc. **DEVIATION RANGE:** 3-0-3 and 6-0-6 kc for all present TV channels. **DEVIATION** from assigned channel is shown continuously on large-scale meter, unaffected by TV video modulation. **HIGH STABILITY:** ± 0.001%. **HIGH ACCURACY:** crystal frequency when monitor is delivered is within ± 0.001% (10 parts per million). Adjustment provided to set monitor in agreement with frequency measuring service. **HIGH IMPEDANCE INPUT CIRCUIT** for channels 2 to 6; coaxial line for channels 7 to 13. **REMOTE FREQUENCY DEVIATION METER** terminals provided. **THREE SPARE CRYSTAL POSITIONS** selected by panel switch. **TYPE 1182-T VIDEO FREQUENCY MONITOR . . . $675**

WRITE FOR COMPLETE ENGINEERING DATA

GENERAL RADIO COMPANY Cambridge 39, Massachusetts

90 West St., New York 6 920 S. Michigan Ave., Chicago 5 950 N. Highland Ave., Los Angeles