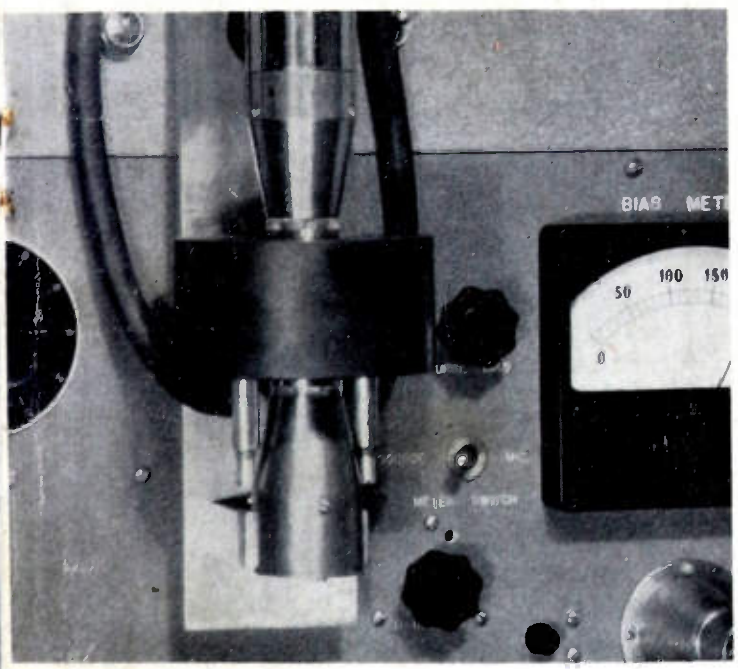
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

June, 1952

Volume 40 Number 6



Shure Brothers, Inc.

MICROPHONE CALIBRATION

Basic to all technological advance is the ability to perform primary measurements. Pictured above is the newest integrated setup for reciprocity calibration of laboratory standard microphones in a hydrogen-filled coupler.

PROCEEDINGS OF THE I.R.E.

TABLE OF CONTENTS INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 64A

The Institute of Radio Engineers

Humidity . . . temperature . . . altitude testing

Tracing filter curves electronically

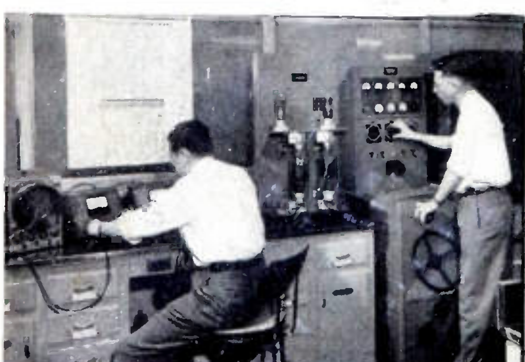
Development tests . . . ½ cycle filter

Filter development

Engineering Leadership

UTC has the reputation for exceptional quality and the ability to produce units previously considered impossible. This position of engineering leadership has been effected through a continuous program of research and development at the UTC laboratories. A few views of these laboratories are shown on this page.

Audio development



Testing modulation transformer under operating conditions

Q meter measurements on low frequency coils

Pulse transformer development

United Transformer Co.

150 VARICK STREET NEW YORK 13, N. Y.

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This complete guide to Sprague dry electrolytic capacitors designed to meet military requirements will gladly be sent to electronic engineers and purchasing agents on letterhead request. Sprague's new Catalog 11 is printed in large, clear type to facilitate ready reference to its 24 pages of military capacitor information. Write for your copy today to the Application Engineering Dept., Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

PIONEERS IN ELECTRIC **SPRAGUE** AND ELECTRONIC DEVELOPMENT

World's Largest Capacitor Manufacturer

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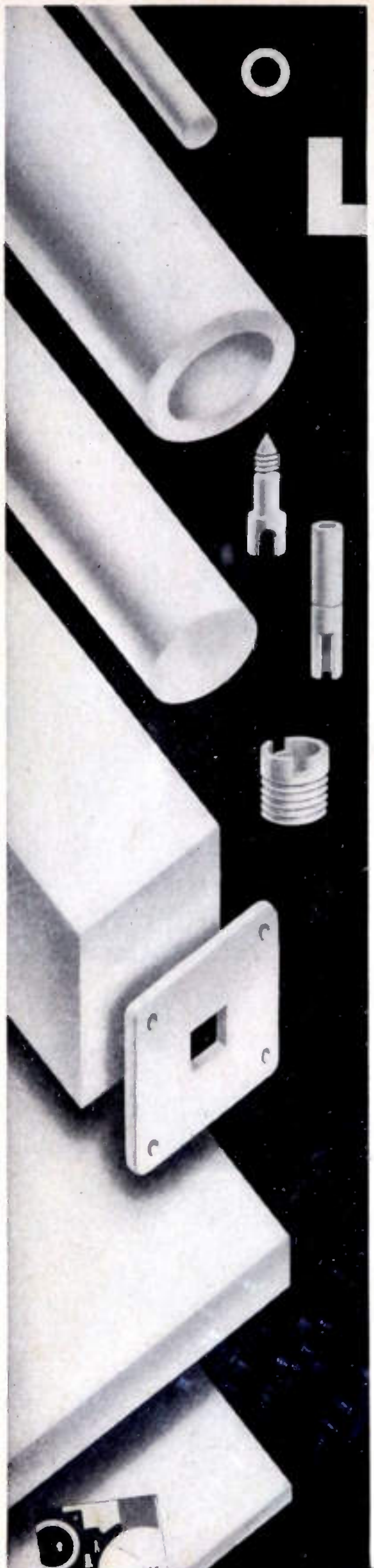
Table of Contents will be found following page 64A



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Meetings with Exhibits

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

△

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

●

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

△

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

△

Joint IRE-AIEE Computers Conference
December 10, 11, & 12, 1952
Park Sheraton Hotel
Exhibits: Perry Crawford
373 Fourth Avenue
New York City

△

Southwestern IRE Conference
February 5, 6, & 7, 1953
Plaza Hotel, San Antonio, Tex.
Accept Exhibits

△

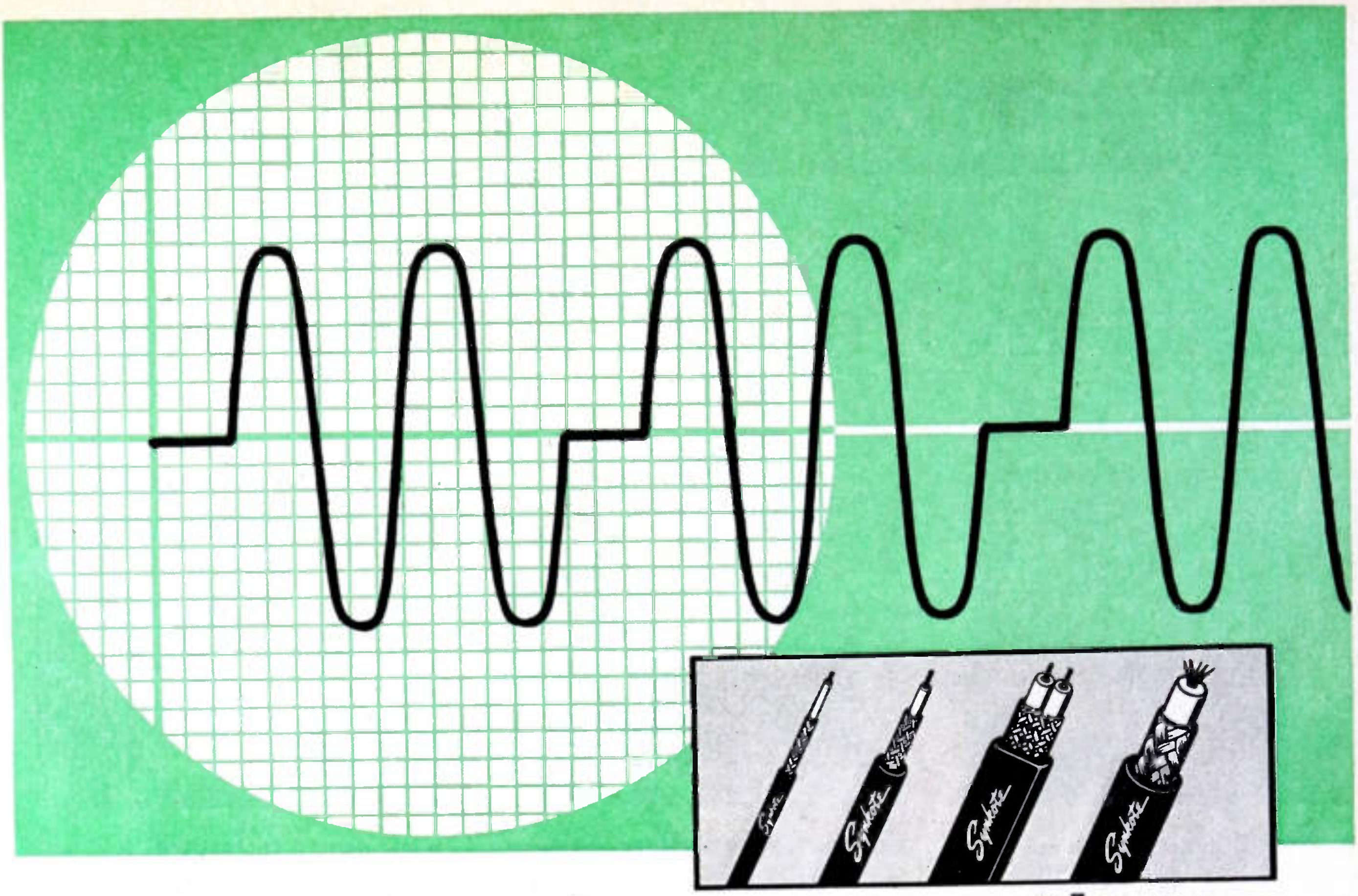
Radio Engineering Show
March 23, 24, 25, 26, 1953
Grand Central Palace
New York City
Exhibits Manager: Wm. C. Copp
303 W. 42nd St., New York 36, N.Y.

△

NEREM—New England Radio Engineering Meeting
Announced for April 25, 1953
University of Connecticut
Storrs, Conn.

△

National Conference on Airborne Electronics
May 11, 12 & 13, 1953
Hotel Biltmore, Dayton, Ohio
Exhibits: Paul D. Hauser
1430 Gascho Drive, Dayton 3



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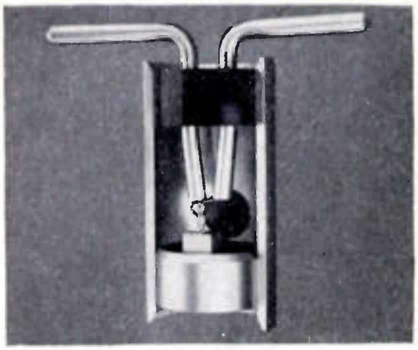
THE TRANSISTOR

A picture report of progress



FIRST TRANSISTORS were of this point contact type (picture three times life size). Current is amplified as it flows between wires through a wafer of germanium metal. These transistors are now being made at the Allentown plant of Western Electric, manufacturing unit of the Bell System. They will be used in a new selector which finds the best routes for calls in Long Distance dialing.

NEW JUNCTION TRANSISTORS, still experimental, also use germanium but have no point contacts. Current is amplified as it flows through germanium "sandwich" —an electron-poor layer of the metal between two electron-rich ends. This new transistor runs on as little as *one-millionth* of the power of small vacuum tubes.

MUCH HAD TO BE LEARNED, especially about the surface of germanium and the effect of one part in a million of alloying materials. Transistors promise many uses—as amplifiers, oscillators, modulators . . . for Local and Long Distance switching . . . to count electrical pulses.

ASSEMBLY PROBLEMS, such as fixing hair-thin wires to barely visible germanium wafers, have been solved through new tools and mechanized techniques. Finished transistors withstand great vibration and shock. Engineers see many opportunities for these rugged devices in national defense.

MOIST PAPER AND COIN generate enough current to drive audio oscillator using junction transistors. Half as big as a penny matchbox, an experimental two-stage transistor amplifier does the work of miniature-tube amplifiers ten times larger.

A tiny amplifying device first announced by Bell Telephone Laboratories in 1948 is about to appear as a versatile element in telephony.

Each step in the work on the transistor . . . from original theory to initial production technique . . . has been carried on within the Laboratories. Thus, Bell scientists demonstrate again how their skills in many fields, from theoretical physics to production engineering, help improve telephone service.

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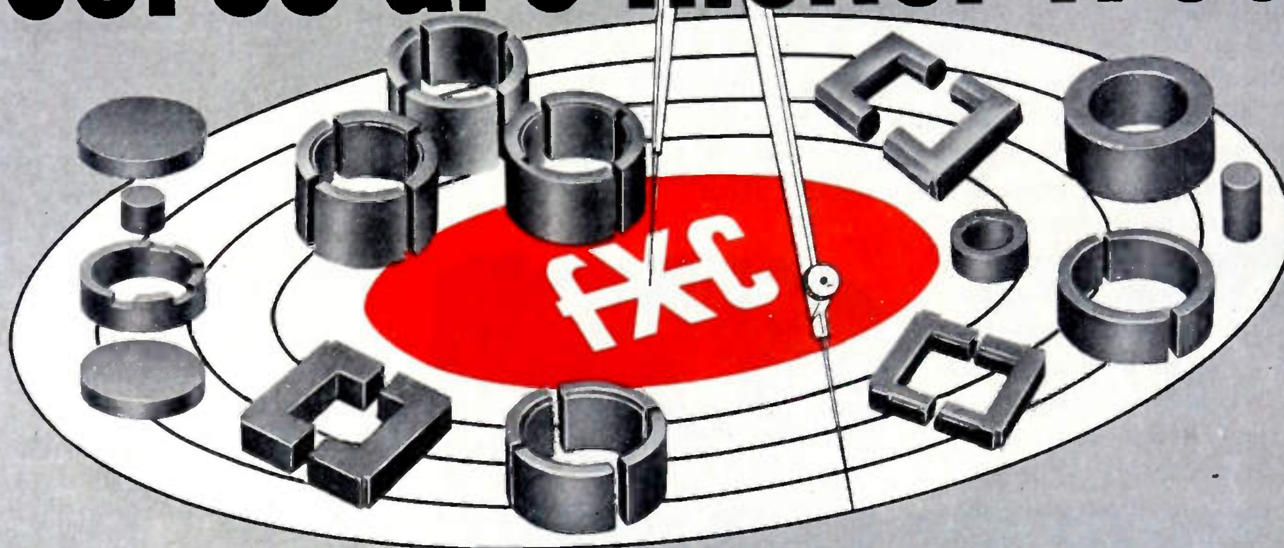
Improving telephone service for America provides careers for creative men in scientific and technical fields.

FERROXCUBE 3C cores are nickel-free



APPLICATIONS:

- I-F TRANSFORMERS
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- LOW-LOSS INDUCTORS
- SATURABLE CORE REACTORS
- HORIZONTAL OUTPUT TRANSFORMERS
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BUSINESS IN MOTION

To our Colleagues in American Business . . .

For several years this space has been used to tell how Revere has collaborated with its customers, to mutual benefit. Now we want to talk about the way our customers can help us, again to mutual benefit. The subject is scrap. This is so important that a goodly number of Revere men, salesmen and others, have been assigned to urge customers to ship back to our mills the scrap generated from our mill products, such as sheet and strip, rod and bar, tube, plate, and so on. Probably few people realize it, but the copper and brass industry obtains about 30% of its metal requirements from scrap. In these days when copper is in such short supply, the importance of adequate supplies of scrap is greater than ever. We need scrap, our industry needs scrap, our country needs it promptly.

Scrap comes from many different sources, and in varying amounts. A company making screw-machine products may find that the finished parts weigh only about 50% as much as the original bar or rod. The turnings are valuable, and should be sold back to the mill. Firms who stamp parts out of strip have been materially helped in many cases by the Revere Technical Advisory Service, which delights in working out specifications as to dimensions in order to minimize the weight of trimmings; nevertheless, such manufacturing operations inevitably produce scrap. Revere needs it. Only by obtaining scrap can Revere, along with the other companies in the copper and brass business, do the utmost possible in filling orders. You see, scrap helps us help you.

In seeking copper and brass scrap we cannot appeal to the general public, nor, for that matter, to the small businesses, important though they are, which have only a few hundred pounds or so to dispose of at a time. Scrap in small amounts is taken by dealers, who perform a valuable service in collecting and sorting it, and making it available in large quantities to the mills. Revere, which ships large tonnages of mill products to important manufacturers, seeks from them in return the scrap that is generated, which runs into big figures of segregated or classified scrap, ready to be melted down and processed so that more tons of finished mill products can be provided.

So Revere, in your own interest, urges you to give some extra thought to the matter of scrap. The more you can help us in this respect, the more we can help you. When a Revere salesman calls and inquires about scrap, may we ask you to give him your cooperation? In fact, we would like to say that it would be in your own interest to give special thought at this time to all kinds of scrap. No matter what materials you buy, the chances are that some portions of them, whether trimmings or rejects, do not find their way into your finished products. Let's all see that everything that can be re-used or re-processed is turned back quickly into the appropriate channels and thus returned to our national sources of supply, for the protection of us all.

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COMPOSITION-ELEMENT POTENTIOMETERS

Only 5/8" diameter. 1/8" diameter shaft. 1/4" x 32 threaded bushing for mounting.
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Housed in low-loss phenolic. Convenient terminal lugs.
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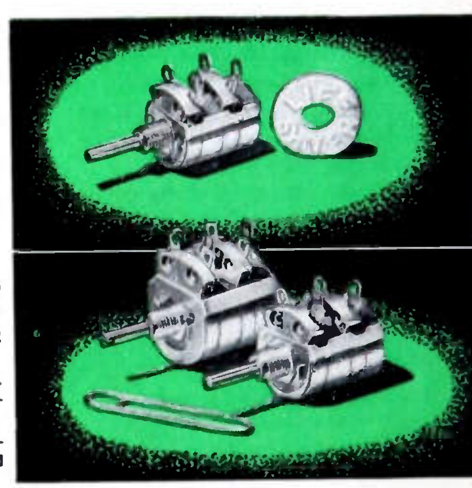

Again, you can stand pat with CLAROSTAT... Let us collaborate on your miniaturization or any other problems involving resistors, controls or resistance devices. Engineering data on request. Get our quotations!

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A page from the note-book of Sylvania Research

Special coil eliminates effects of magnetic fields in Electron Optical Studies

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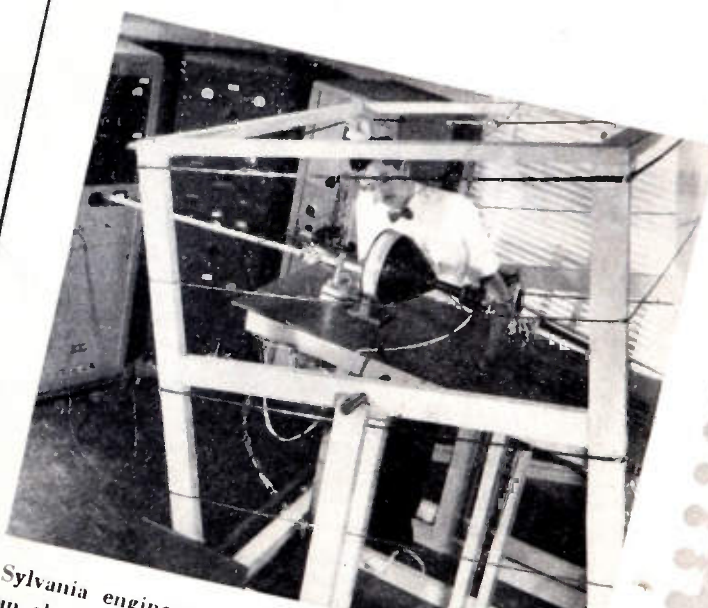
This is now accomplished at Sylvania Research Laboratories by a specially designed modified Helmholtz type coil which produces a uniform magnetic field which compensates for the effects produced by the earth's field and local fields. The coil frame is made entirely of wood and is oriented exactly along the earth's field. It is uniform to 1 part in 500 over a volume enclosed by a cylinder 20 inches in diameter and 30 inches long. The axis of this cylinder is the axis of the coil.

This field is then explored by a special Cathode Ray Tube to make sure the net field registers zero over the entire working region within the coil.

Such strict attention to fundamental research in every phase of electronics and radio development pays off in the outstanding performance of all Sylvania Tubes.



Sylvania engineer measuring spherical aberration of an electrostatic lens inside special Sylvania-built coil. Note, all controls are outside the coil.

SYLVANIA

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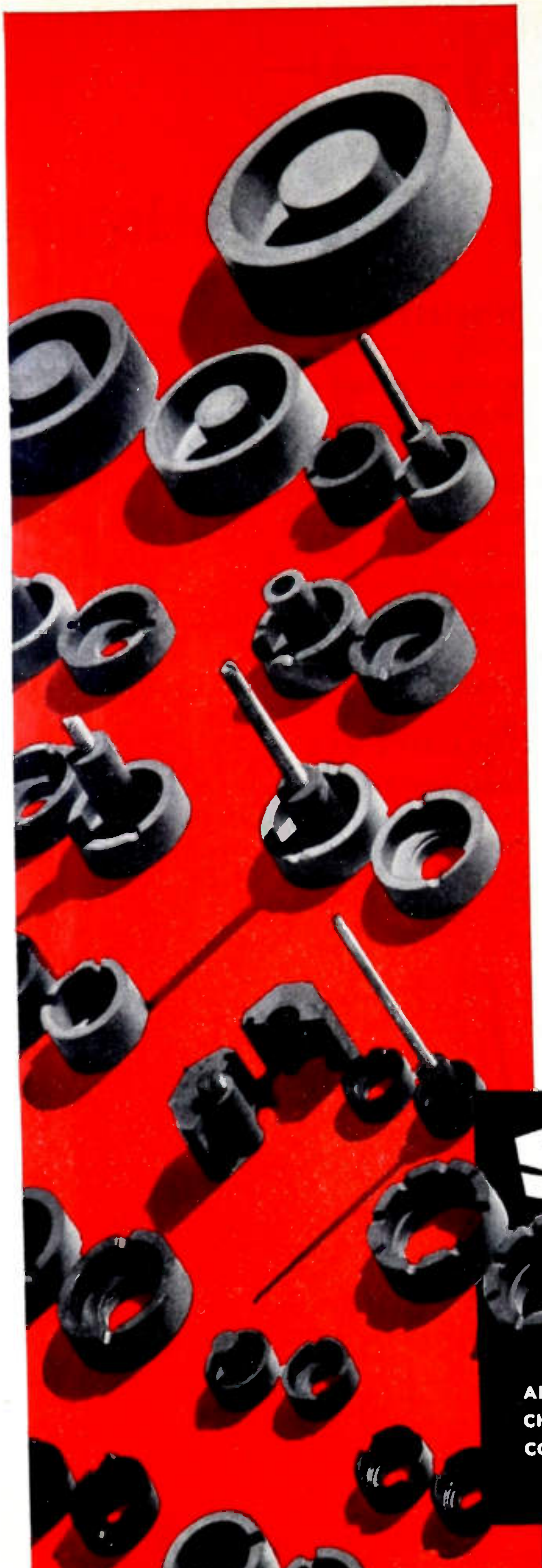

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St. Marys, Pa.

STACKPOLE

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WIRELESS ENGINEER—the magazine of radio research and progress—is produced for research engineers, designers and students in the fields of radio, television and electronics. Its editorial policy is to publish only original work, and its highly specialized content is accepted as the authoritative source of information for advanced workers everywhere. The magazine's Editorial Advisory Board contains representatives of the National Physical Laboratory, the British Broadcasting Corporation, and the British Post Office. Regular features include an Abstracts and References Section compiled by the Radio Research Organization of the Department of Scientific and Industrial Research.

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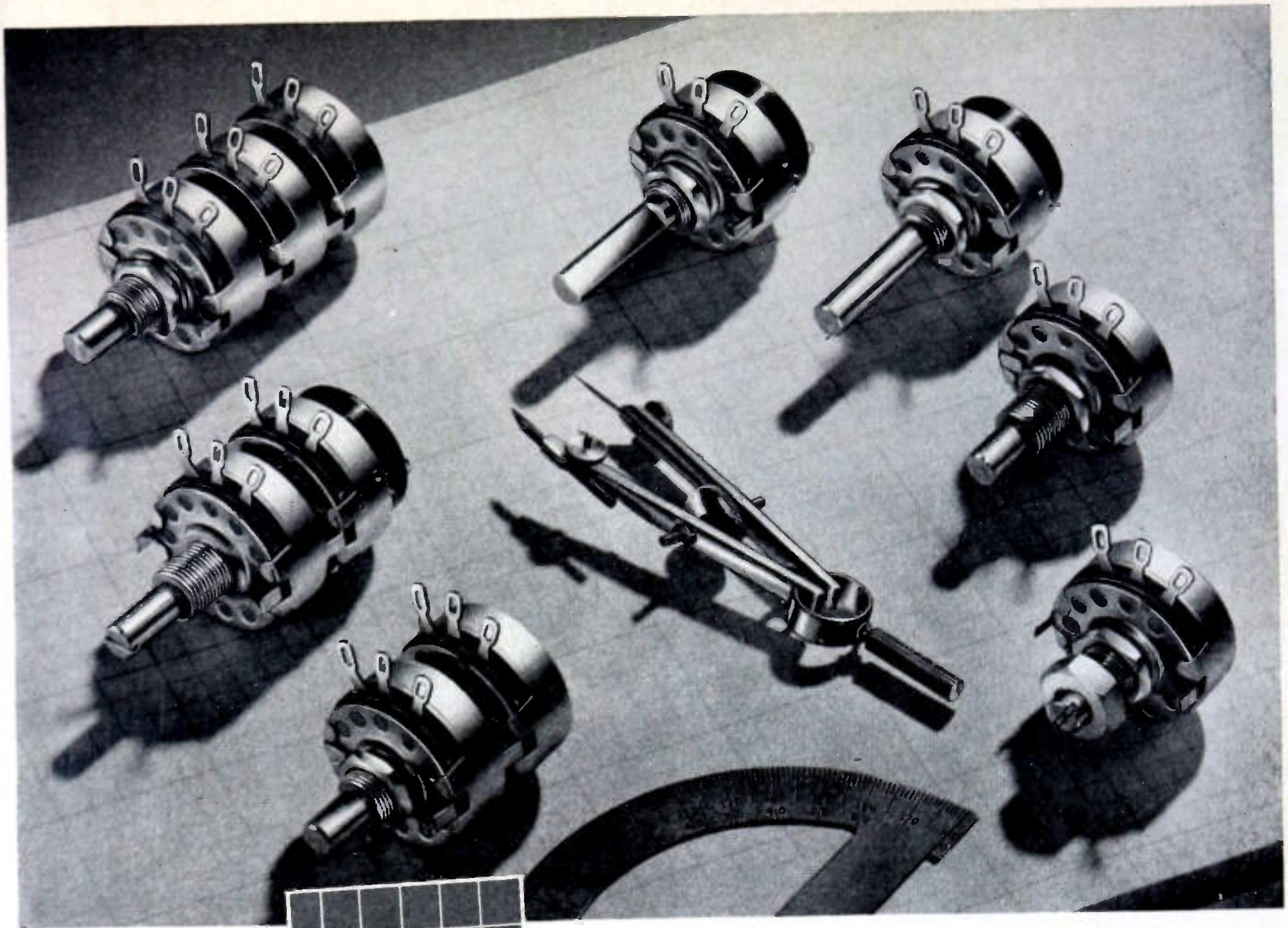

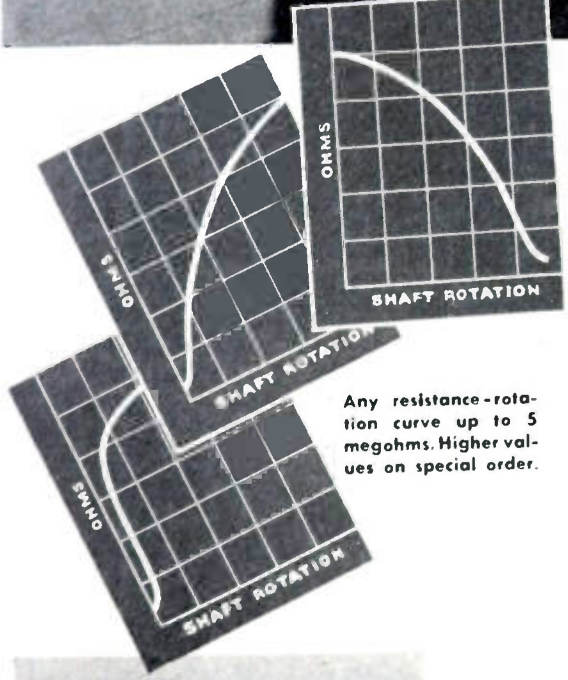

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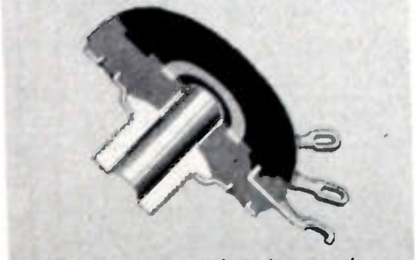

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from . . .
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with accuracy of . . .
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300	10 to 150,000 cycles	1 millivolt to 100 volts	1/2 meg. shunted by 30 mmfds.	2% up to 100 KC 3% above 100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

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

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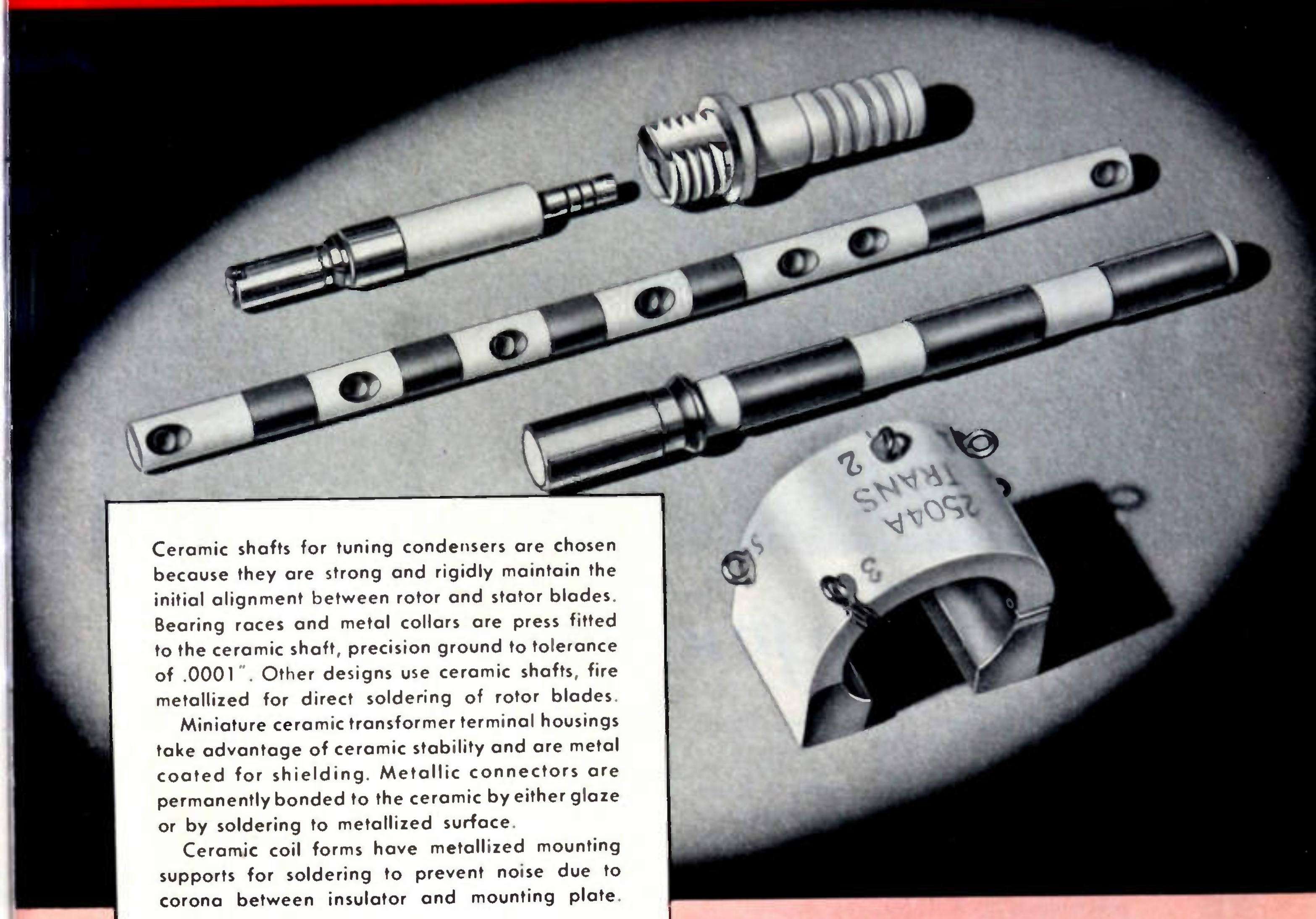

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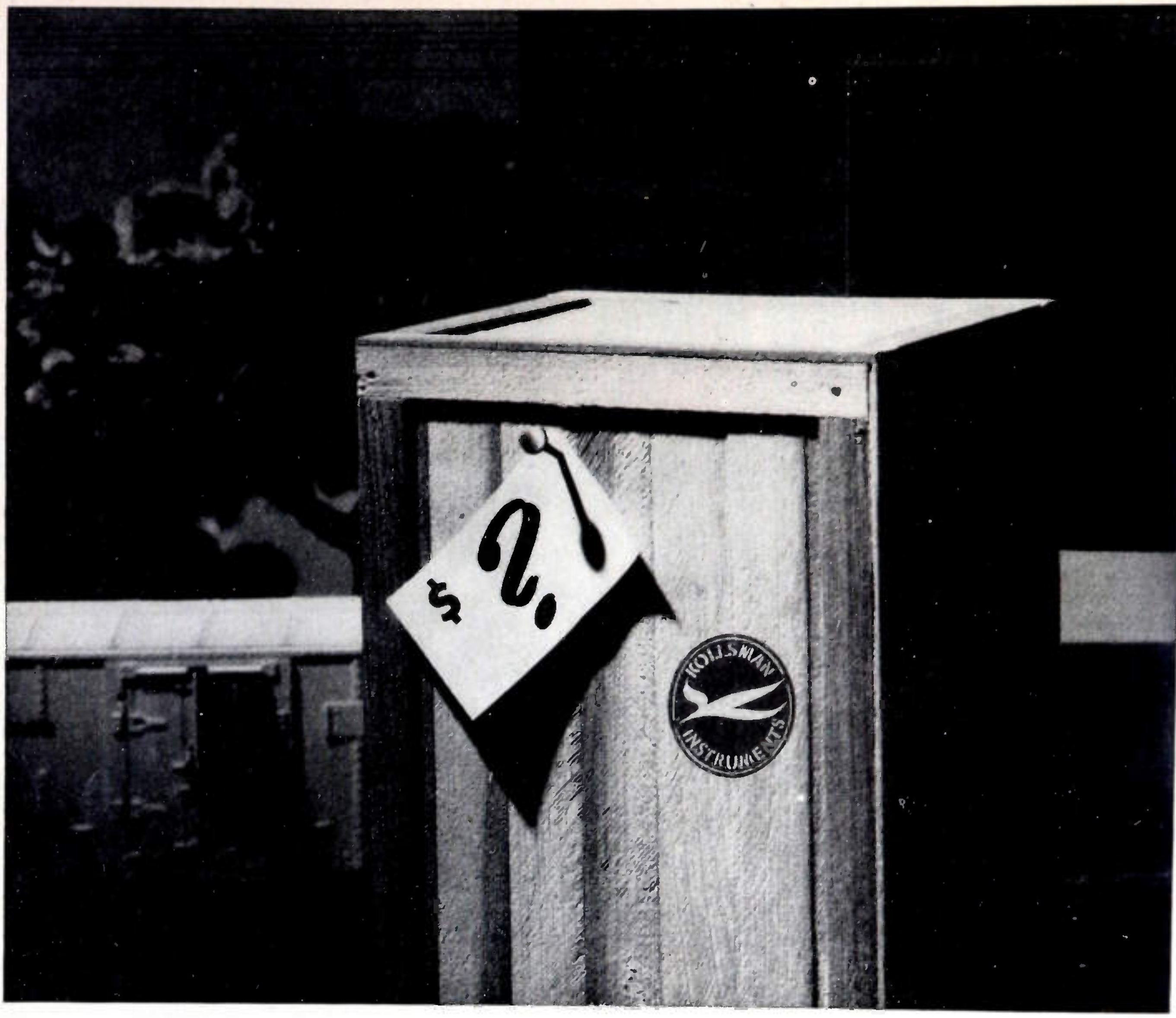

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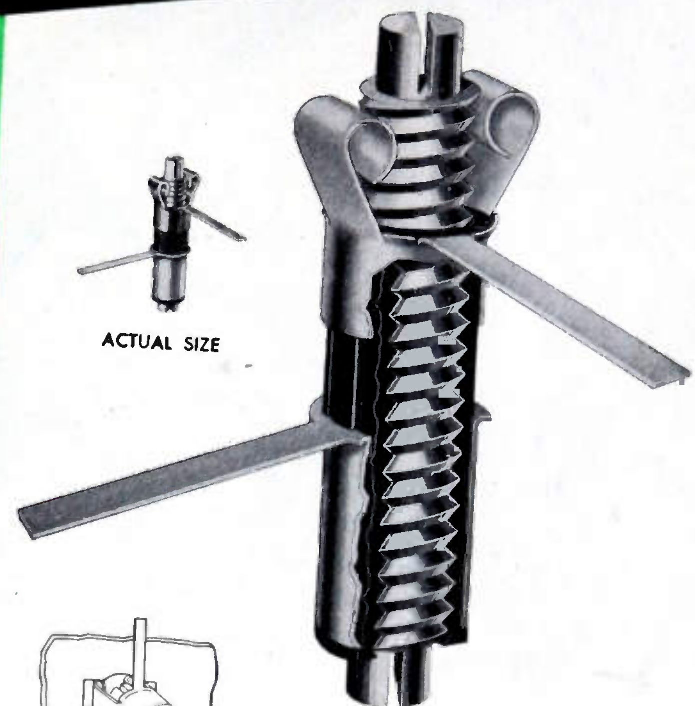
A "Natural" for Miniaturization and Low Inductance in UHF Circuits

THE ERIE Style 535 Tubular Trimmer combines economical price, compact size, and easy mounting, with features for UHF operation. Capacitance range is 0.7 to 3.0 MMF. When mounted it extends only 17/32" from the underside of the chassis. It is 7/32" in diameter, and high terminal is conveniently available to tube socket terminals at a level 1/4" from the underside of the chassis.

Design simplicity results in very low inductance, and uniform, straight-line, and noiseless adjustment. It can be mounted close to associated circuit elements, and the ribbon type leads help to minimize inductance in UHF circuits.

The Style 535 Trimmer as shown at the right, is unique in requiring work from only one side of the chassis when mounting. Ground terminal is provided for soldering to chassis when desired.

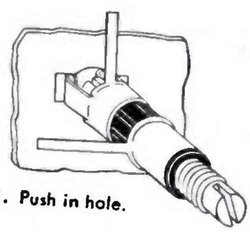
Write for descriptive literature and samples.



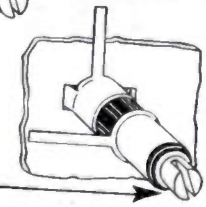
ACTUAL SIZE

STYLE 535

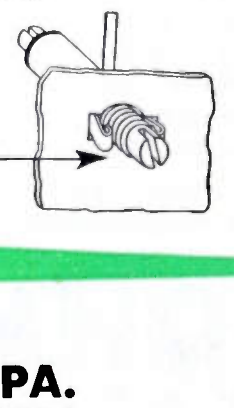
PATENTED



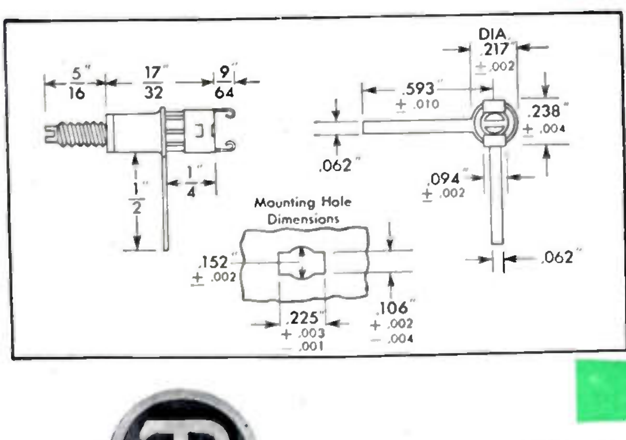
1. Push in hole.



2. Lock in hole by turning adjusting screw through top terminal.



3. Adjust capacitance from top at final test station.



Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

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Ohmite wire-wound resistors have earned a world-wide reputation for *dependability* . . . the ability to give unfailing performance under adverse operating conditions.

These fine units are available in the most complete line of types and sizes on the market. Included are fixed, tapped, adjustable, non-inductive, and precision units—in more than 60 wattage sizes and 18 types of terminals, and in a wide range of resistance values. For *extra* dependability, specify Ohmite resistors, overwhelmingly the first choice of industry, today.

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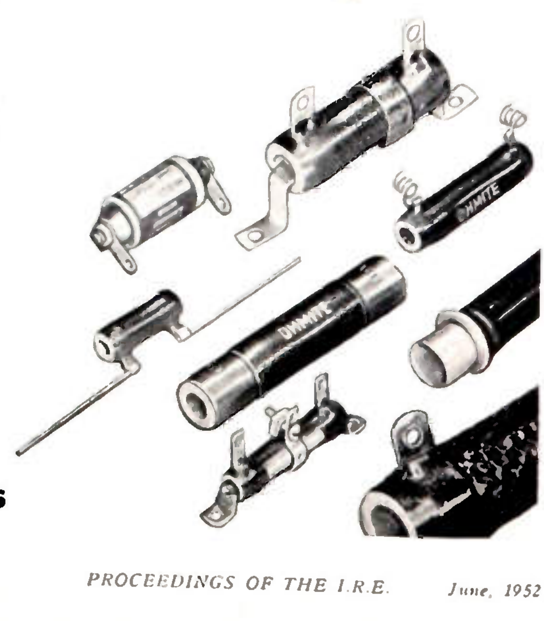
Ohmite Manufacturing Co., 4862 Flournoy St., Chicago 44, Illinois

OHMITE®

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



VOLTAGE REGULATOR AND REFERENCE TUBES

TYPE	MAX. DIMENSIONS INCHES		MIN. STARTING VOLTAGE SUPPLY	OPERATING VOLTAGE (Approx.)	MIN. OPERATING CURRENT MA.	MAX. OPERATING CURRENT MA.	MAX. REGULATION VOLTS
	HEIGHT	DIAM.					
OA2	2.63	.75	185	150	5	30	6
OB2	2.63	.75	133	108	5	30	4
CK1017	2.69	.75	750	700	.005	.055	15
CK1022	2.69	.75	1100	1000	.005	.055	20
CK1037	1.75	.40	720	700	.005	.100	15
CK1038	1.75	.40	925	900	.005	.055	15
CK1039	1.75	.40	1230	1200	.005	.100	25
CK5651*	2.13	.75	115	87	1.5	3.5	3
CK5783*	1.63	.40	115	87	1.5	3.5	3
CK5787	2.06	.40	141	100	5	30	6
CK5962	2.69	.75	730	700	.005	.055	15
CK6213	1.38	.40	200	130	1.0	2.5	2

*Voltage Reference Tube

RAYTHEON MANUFACTURING COMPANY

Receiving Tube Division

Newton, Mass., Chicago, Ill., Atlanta, Ga., Los Angeles, Calif.

RAYTHEON ®

Excellence in Electronics

RELIABLE SUBMINIATURE AND MINIATURE TUBES • GERMANIUM DIODES AND TRANSISTORS • RADIAC TUBES • RECEIVING AND PICTURE TUBES • MICROWAVE TUBES

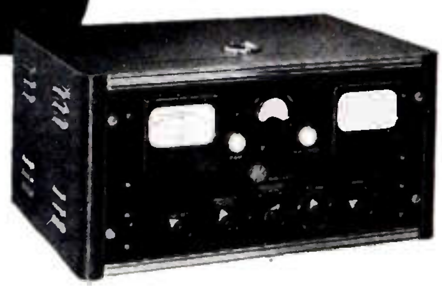
Keeping communications "ON THE BEAM"

JAMES KNIGHTS

Formerly Manufactured by DOOLITTLE RADIO, INC.

FREQUENCY & MODULATION MONITOR

The JK FD-12 monitors any four frequencies anywhere between 25 mc and 175 mc, checking both frequency deviation and amount of modulation. A truly precise instrument for communication systems!



When used for different bands, plug-in type antenna coils provided. Crystal accuracy guaranteed to be ± .0015% over range of 15° to 50° C. Meets or exceeds FCC requirements.

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Well known to every communications man is the famous JK Stabilized H-17, with a frequency range of 200 kc to 100 mc. But this is just one crystal in the JK line. Write for complete crystal catalog!

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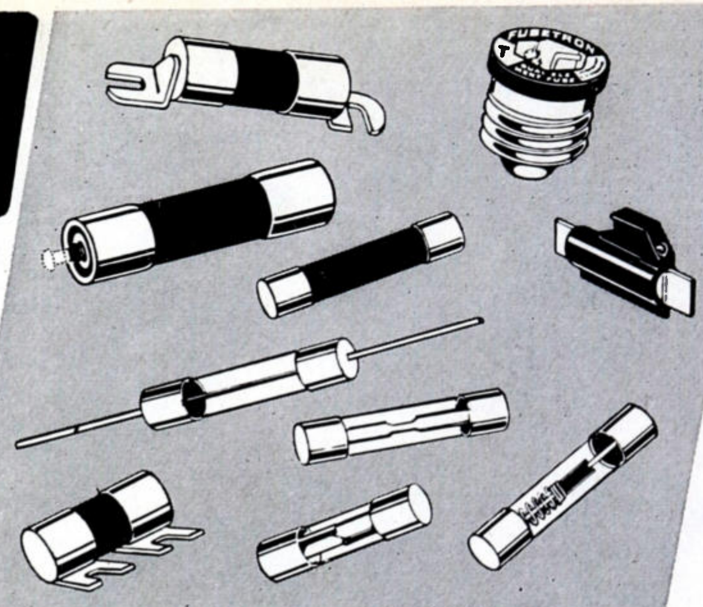
Behind each fuse or fuse mounting are 37 years of know-how in building products of unquestioned high quality, the world's largest fuse research laboratory and the world's largest fuse production capacity.

Each BUSS Fuse Electronically Tested.

To assure proper operation in the field, each and every BUSS fuse is tested in a highly sensitive electronic device that rejects any fuse that is not correctly calibrated — properly constructed and right in physical dimensions.

BUSS Fuses are made to Protect — not to Blow.



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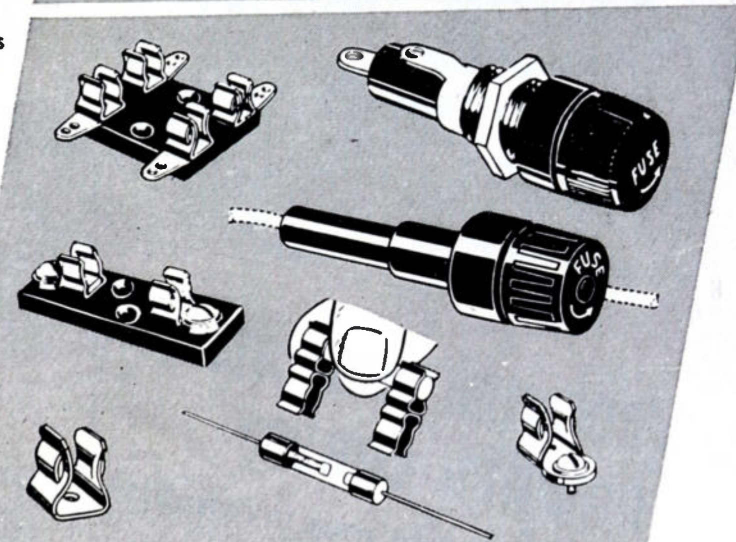
RENEWABLE FUSES, ONE TIME FUSES, SPECIAL FUSES



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We welcome requests to help you in selecting the proper fuse or in designing a special fuse or fuse mounting best suited to your conditions. Submit sketch or description showing type of fuse contemplated, number of circuits, type of terminals and the like.

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Division of McGraw Electric Company
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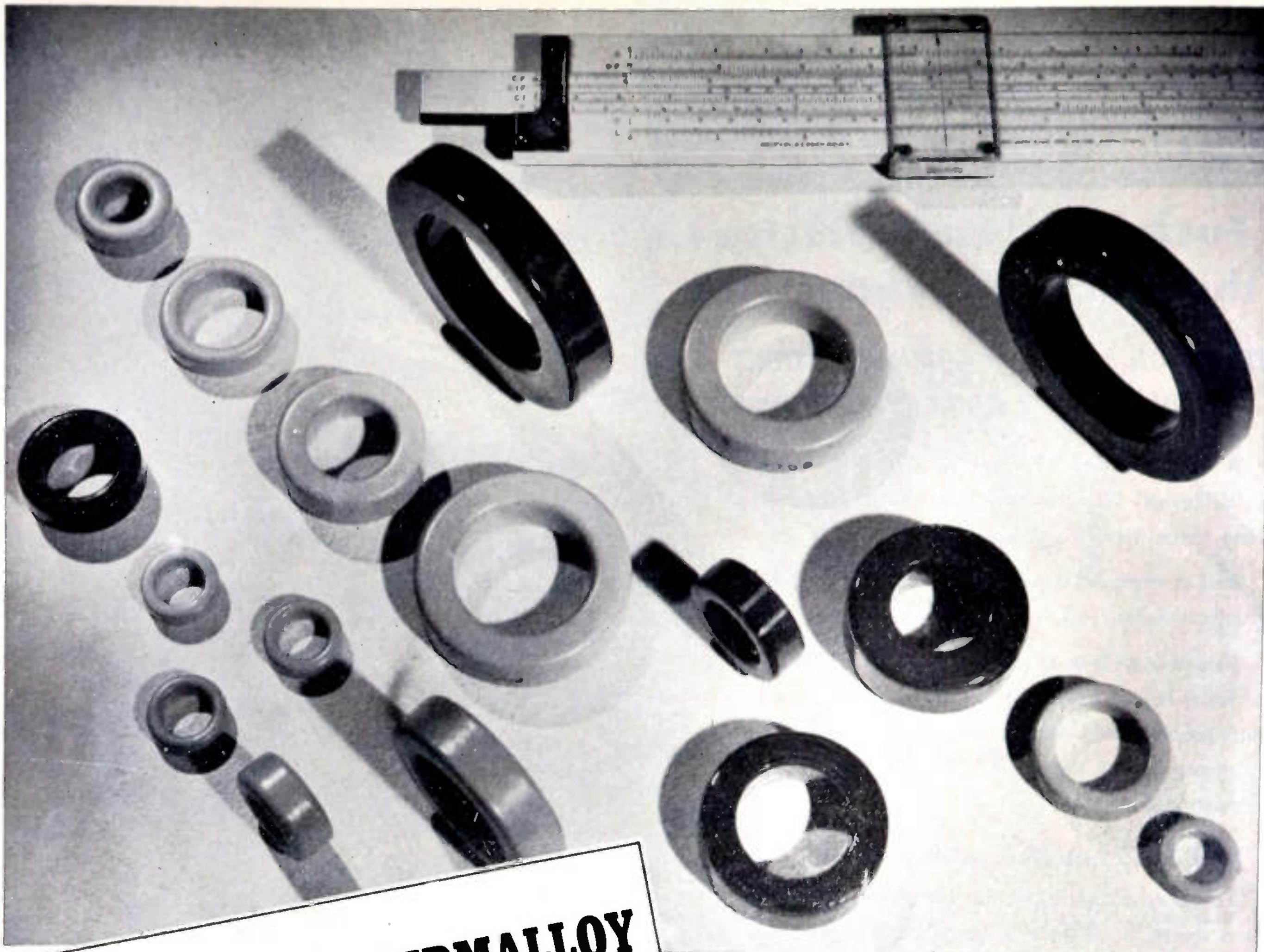
FOR HOME, FARM, COMMERCIAL AND INDUSTRIAL USE.



MOLYBDENUM PERMALLOY POWDER CORES*

(New technical data now available)

HIGH Q TOROIDS for use in Loading Coils, Filters, Broadband Carrier Systems and Networks— for frequencies up to 200 K C

COMPLETE LINE OF CORES TO MEET YOUR NEEDS

★ Furnished in four standard permeabilities — 125, 60, 26 and 14.

★ Available in a wide range of sizes to obtain nominal inductances as high as 281 mh/1000 turns.

★ These toroidal cores are given various types of enamel and varnish finishes, some of which permit winding with heavy Formex insulated wire without supplementary insulation over the core.

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*Manufactured under license arrangements with Western Electric Company

W&D 4127

THE ARNOLD ENGINEERING COMPANY

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

General Office & Plant: Marengo, Illinois


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at Hermetic
As Shown by These
Recent Notable
Hermetic Seal Firsts
The design and production of hermetic seals is a job for experts, steeped in this highly specialized activity. Hermetic Seal Products Co. has consistently pioneered and developed the outstanding firsts in ceramic-metal headers. Their experience, know-how and engineering talent are unrivaled in this field. Such specialization assures you of quality hermetic headers in unlimited shapes that will withstand mass spectrometer leak tests, —55°F. conditions, swamp test, temperature cycling, high vacuum, high pressure, salt water immersion and spray, etc. They are the only headers you can hot tin dip at 525°F. for easy assembly soldering for a strain and fissure-free sealed part with resistance of over 10,000 megohms.
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—available in two grades:

MYCALEX 410 priced comparable to mica-filled phenolics. Loss factor is only .015 at 1 mc., insulation resistance 10,000 megohms. Approved fully as Grade L-4B under N.M.E.S. JAN-I-10 "Insulating Materials Ceramic, Radio, Class L".

MYCALEX 410X — low in cost but insulating properties greatly exceed those of general purpose phenolics. Loss factor is only one-fourth that of phenolics (.083 at 1 mc.) but cost is comparable. Insulation resistance 10,000 megohms.

PREMIUM INSULATION — Bodies are MYCALEX glass-bonded mica, the dielectric that combines every characteristic required in a modern insulation including low dielectric loss, high dielectric strength, high arc resistance, non-hygroscopic and great dimensional stability.

COMPETITIVELY PRICED — Although manufacture is to the most exacting quality standards and fully meets RTMA recommendations, an exclusive MYCALEX manufacturing process permits pricing at a level competitive with low cost phenolic types.

PRECISION MOLDED — An exclusive MYCALEX injection molding technique affords great dimensional accuracy, exact uniformity, superior low loss characteristics and perfect homogeneity.

MYCALEX TUBE SOCKET CORPORATION
Under Exclusive License of Mycalex Corporation of America
30 ROCKEFELLER PLAZA • NEW YORK 20, N. Y.

INFORMATIVE DATA SHEETS

Include them in your files — Complete information including dimensional data, specifications and other pertinent facts on MYCALEX low-loss, low-cost, tube sockets. Write for your set complete with loose-leaf binder that permits the inclusion of subsequent releases and data sheets.

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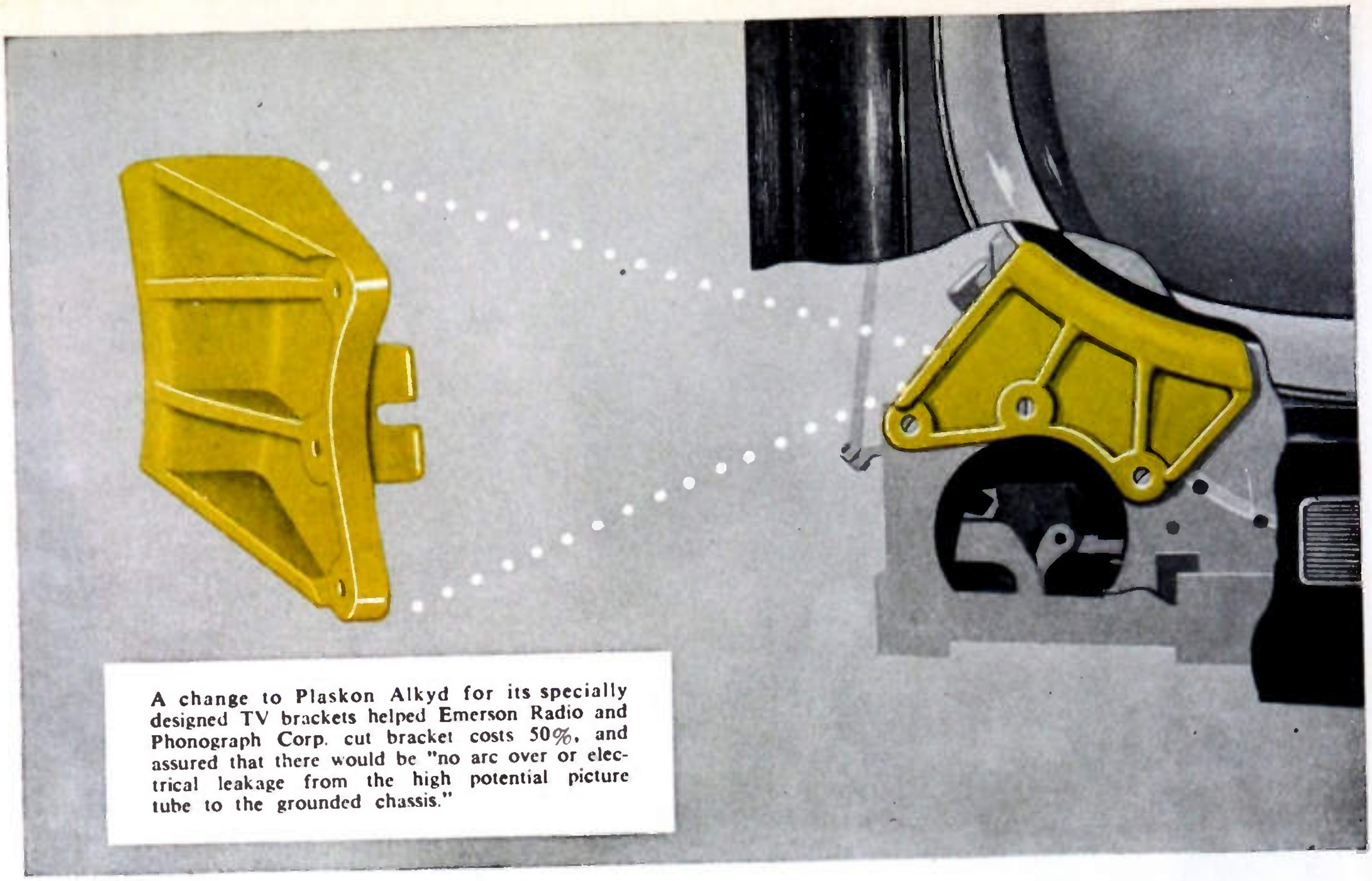
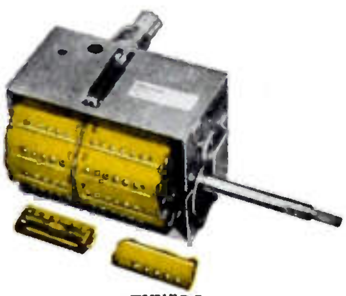
Executive Offices: 30 ROCKEFELLER PLAZA, NEW YORK 20 — Plant & General Offices: CLIFTON, N.J.



A change to Plaskon Alkyd for its specially designed TV brackets helped Emerson Radio and Phonograph Corp. cut bracket costs 50%, and assured that there would be "no arc over or electrical leakage from the high potential picture tube to the grounded chassis."

TV parts that resist high heat and arcing... hold precise dimensions!



TUNERS

CONDENSERS

When your TV parts are molded of Plaskon Alkyd, you can meet the extremely close tolerances demanded in television assemblies. That's because Plaskon Alkyd has exceptional dimensional stability with no after-shrinkage.

And the high heat resistance prevents parts molded of Plaskon Alkyd from breaking down, even under short-time contact with molten solder when connections are made.

What's more, Plaskon Alkyd combines a number of outstanding properties so essential for superior electrical insulating parts: high dielectric strength, superior arc resistance, excellent resistivity. In addition, it can be molded faster and at lower temperatures, giving increased production and greater savings.

Before you redesign, look into the advantages Plaskon Alkyd can offer. Write today for full information on television and electronic uses.

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INSIST ON
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FOR SUPERIOR ELECTRICAL PARTS

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

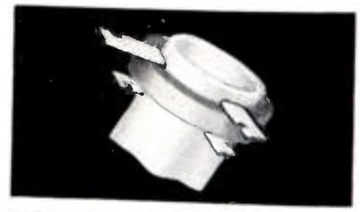
Take advantage of one of C.T.C.'s most popular and useful services . . . the winding of slug tuned coils to exact specifications. Single layer or pie types furnished. You can be sure your specs—military or personal—will be faithfully followed to the last detail of materials and methods, and with expert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

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

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

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

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



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

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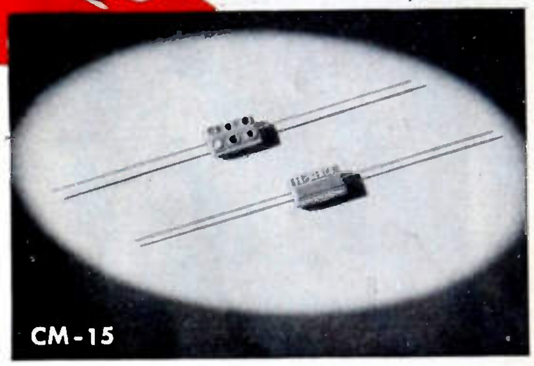
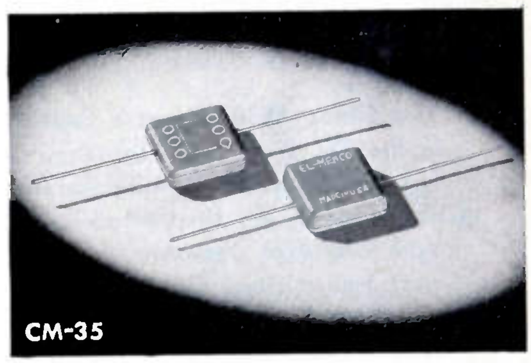
Instruments of war must be unerringly dependable, and every part used in their construction must contribute to this standard. That is why El-Menco Capacitors have won such wide recognition in their particular field . . . Because of their margin of extra wide safety factor they are absolutely reliable.

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



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

Write on your business letterhead for catalog and samples.



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MOLDED MICA **El-Menco** **MICA TRIMMER**

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

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

Advancement in Emergency Communication

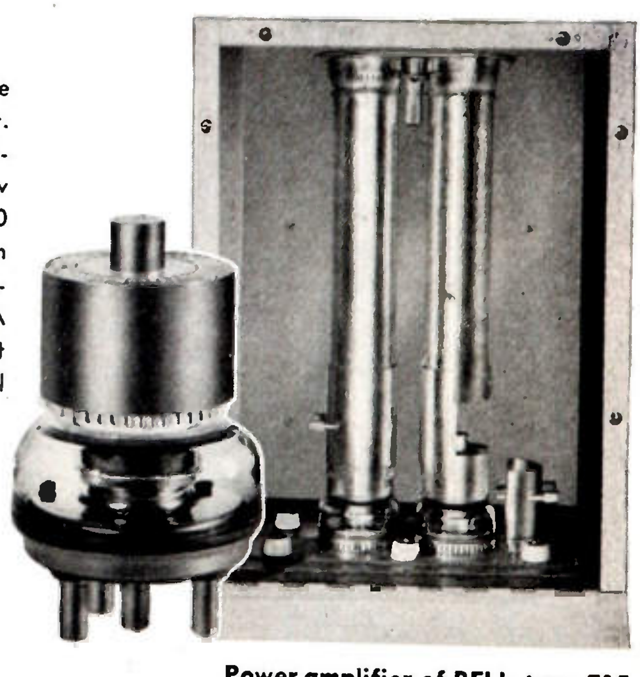
Eimac tubes fill key sockets In continuous service transmitters

REL type 757C transmitter

New and unique in civil emergency communication systems is the New York City Fire Department's five boraugh radio network planned to meet the threat of any emergency, including atomic attack. Transmitters designed by Radio Engineering Laboratories to give continuous operation are significant contributions to this electronic accomplishment.

Eimac's 2C39A triode is utilized in REL's type 757C point-to-point radio relay transmitter operating in continuous around the clock service at 900 mc. The 2C39A is used in two stages —as a tripler from 150 mc. to 450 mc. and as a doubler from 450 mc. to 900 mc. The 2C39A is a natural to serve in REL's 757C where it can perform as a frequency multiplier at ultra high frequencies with excellent operating efficiency. This compact, rugged, high-mu tube is designed for a variety of uses as a power amplifier, oscillator or frequency multiplier wherever dependability and durability are demanded.

Two Eimac 4X500A's give dependable performance in the REL type 715 emergency service transmitter. These external-anode tetrodes are in the power output stage of the final amplifier in each of the New York City Fire Department's eight main station 350 watt transmitters. Operating in the 150 mc. region the 4X500A's meet the challenge of 24-hour performance. Designed for application the 500 watt 4X500A has small size and low inductance leads which permit efficient operation at relatively large outputs well into VHF.



Power amplifier of REL's type 715

● ***Write our application engineering department for the latest information and technical data about these and other Eimac tubes.***

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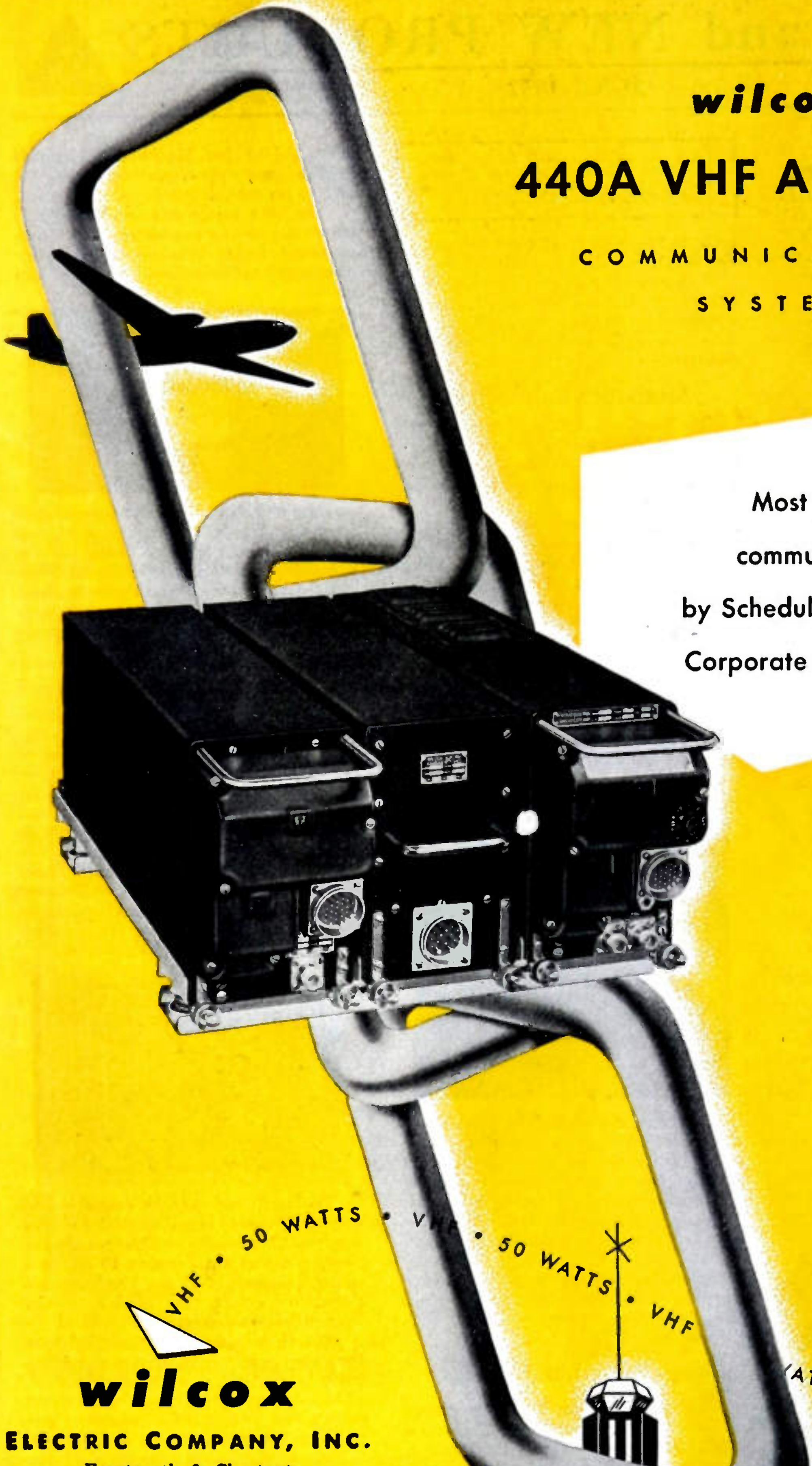
440A VHF AIRBORNE

COMMUNICATIONS SYSTEM



Most *dependable* communications used by Scheduled, Non-scheduled, Corporate Aircraft Operators

Advanced techniques and design are used to insure a maximum of performance and reliability with minimum maintenance time and expense.

Covers all 180 channels (118-136 mc/s) assigned to world-wide Civil Aviation — your protection against obsolescence.

The most powerful transmitter available, plus a highly sensitive and selective receiver, provides dependable communications for low altitude operations and under adverse conditions.

Write for specifications today.

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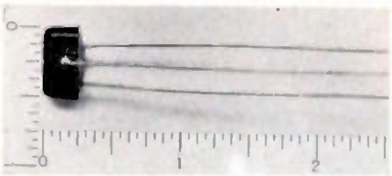
ELECTRIC COMPANY, INC.
Fourteenth & Chestnut
Kansas City 1, Missouri

NEWS and NEW PRODUCTS

JUNE 1952

Junction Transistors

Germanium Products Corp., 28 Cornelison Ave., Jersey City, N. J., (a subsidiary of Radio Development & Research Corp.) licensed by Western Elec. Corp., and S. I. Weiss, has announced the availability of a junction type n-p-n germanium crystal amplifier, Type RD 2517.



The space saving possibilities of this new transistor are expected to be utilized to such an extent that the chassis of a TV receiver will be reduced to the size of a cigar box.

Power requirements are said to be 1 millionth of a watt (0.000001).

The specifications for the Type RD 2517 are:

Ec −45 v supply voltage-collector circuit;
ic 400 μa collector current;
ie −175 μa emitter current;
Ee 1 to 3 v may be operated as self-biasing device by placing suitable resistor between emitter and base;
A 20 db minimum gain at 1,000 cps;
Zin 500 ohms input impedance;
R1 60 k load resistance;
frequency response: ½ db from 30 cps to 20 kc.
ambient temperature: 50°C.

The transistors are being merchandized through **Federated Semi-Conductor** Co., 66 Dey St., New York, N. Y.

Toroids

The **Raytheon Manufacturing Co.**, Waltham, Mass. announce that they have complete facilities for large-volume production, as well as for engineering design and production of models, of custom-made transformers.

More than 10 years' experience in designing and building toroid-L units enables Raytheon to design toroid-L-coils from the problem stage up, or to wind to specified *C, L,* and *Q.* values, precision

wound on temperature stabilized, powdered permalloy cores, high-permeability solid materials, or stamped "O" cores. They are able to wind No. 20 to No. 42

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

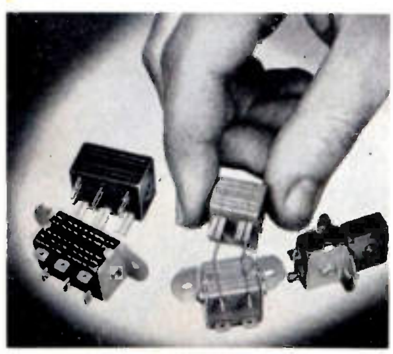
wires on "wedding-ring" cores to small ultimate ID.

Raytheon is equipped for litzendraht coil windings. Facilities for all types of winding are available, including square coils from strip materials for improved geometry.

Miniature Connectors

The **Elco Corp.**, 190 W. Glenwood Ave., Philadelphia 40, Pa. is introducing Varicons, miniature connectors that have various production and application features.

The connectors are rated at 30 amperes, and at 110 volts, withstanding voltage between closest terminals of 4,000 volts. Voltage is rated at 1,330 volts. Using four basic components, it is possible to assemble male and female connectors with any required number of contacts. The connectors can be furnished assembled, or can be put together by the user to suit his own requirements. Stocking only the four components, the user can produce finished connectors on a mass-production basis and yet make changes in the number of contacts or polarity of any connector as needed.



Capacitance is negligible. Contact resistance is 0.0001 ohm and contact spacing is suitable for 300 ohm lines. Contacts are made of brass and phosphor bronze on beryllium copper, and are silver-plated; the body sections are of molded phenolic in general purpose or mica filled and alkyd resins.

Specific information regarding the application of the new Varicons in product design may be obtained from the Elco Corporation.

Resistance Limit Bridge

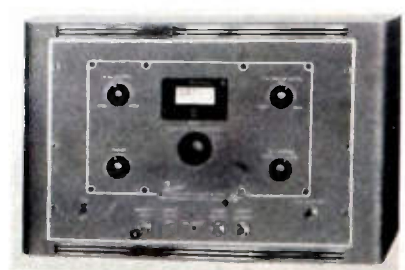
The General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass., has a new Type 1652-A resistance limit bridge, which indicates on a panel meter the percentage deviation of unknown resistors. Deviation can be measured from an external standard or from an internal standard adjustable from 1 to 1,111,111 ohms in steps of 0.1 ohm. Maximum deviation is ±20 per cent and accuracy is between 0.2 and 0.5 per cent, depending on application.

The limit bridge can also be used to match one resistor to another, or as a conventional decade Wheatstone bridge for resistance measurements by the null method.

Production applications were of primary importance in the design of this bridge, although it is as useful in the laboratory and model shop as in manufacturing departments. To facilitate the speed and convenience of production measurements, test jigs can be connected directly to panel terminals.

Oscillator

The Type 907, a fundamental oscillator continuously tunable over the frequency range of 35 to 900 mc has been developed by the **Polytechnic Research and Development Co.**, 55 Johnson St., Brooklyn, N. Y. The unit features a tank circuit design that permits a 30-to-1 tuning range with an output voltage of not less than 1 volt across 75 ohms at all frequencies. Other features include a video-type blanking circuit which yields a true horizontal zero base line and provisions for the introduction of an external frequency marker.



The oscillator may be swept in frequency by means of a sinusoidally vibrating "tank" capacity which provides a sweep width of not less than 10 mc for a center frequency of 35 mc. The frequency deviation is greater than 20 mc for carrier frequencies above 60 mc.

The rf output power is coupled from the sweep generator by means of a waveguide beyond the cutoff-type attenuator. The attenuator, which is terminated on the output end with a BNC connector, permits continuous adjustment of the output voltage from 10 μv to 1 volt at all frequencies. A resistance pickup loop provides a low vswr source impedance for the sweep generator over the operating frequency range.

(Continued on page 39A)

Q: Are all brands of Resistors similar in Quality, Specifications and Performance?

A: Naturally, no! That's why DAVEN has earned the right to add a superlative in naming its line

SUPER DAVOHM

PRECISION WIRE WOUND

RESISTORS

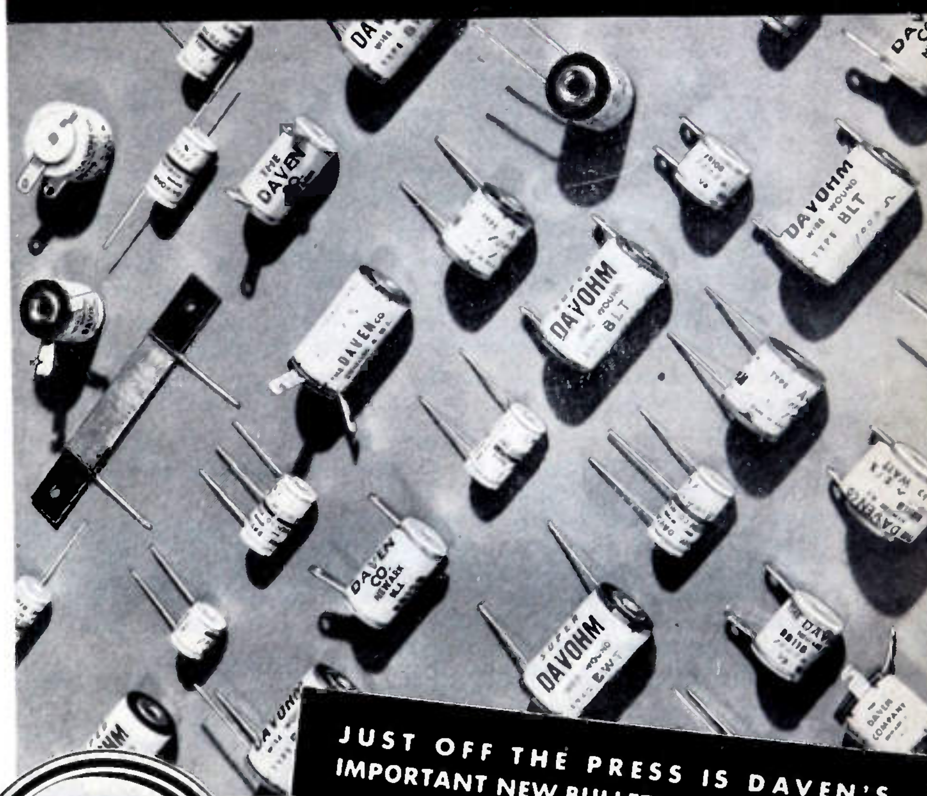
● Brands of Resistors vary as widely in the completeness of a line and in performance, as do brands of any other product.

DAVEN originated the first pie-type, wire wound Resistor more than a generation ago. Since that time, DAVEN has designed and manufactured Precision Wire Wound Resistors of every conceivable type to meet the increasing demands of the electronics industry.

SUPER DAVOHM RESISTORS are noted for their high stability and accuracy under extreme temperature and humidity conditions. DAVEN Resistors are made in accordance with JAN-R-93 specifications and are in use in all types of Army, Navy and Air Force electronic equipment.

DAVEN has developed special small precision Resistors for use in miniaturized assemblies. All types of mountings, sizes, tolerances and temperature coefficients are available from a large variety of standard types. That's why DAVEN can fill your precision Resistor needs.

Take advantage of DAVEN's advanced engineering and manufacturing techniques to help with any Resistor problem confronting you.



JUST OFF THE PRESS IS DAVEN'S IMPORTANT NEW BULLETIN ON RESISTORS.

Write for your copy today and prove to yourself, from the information supplied, that when you order Resistors you should specify only . . .

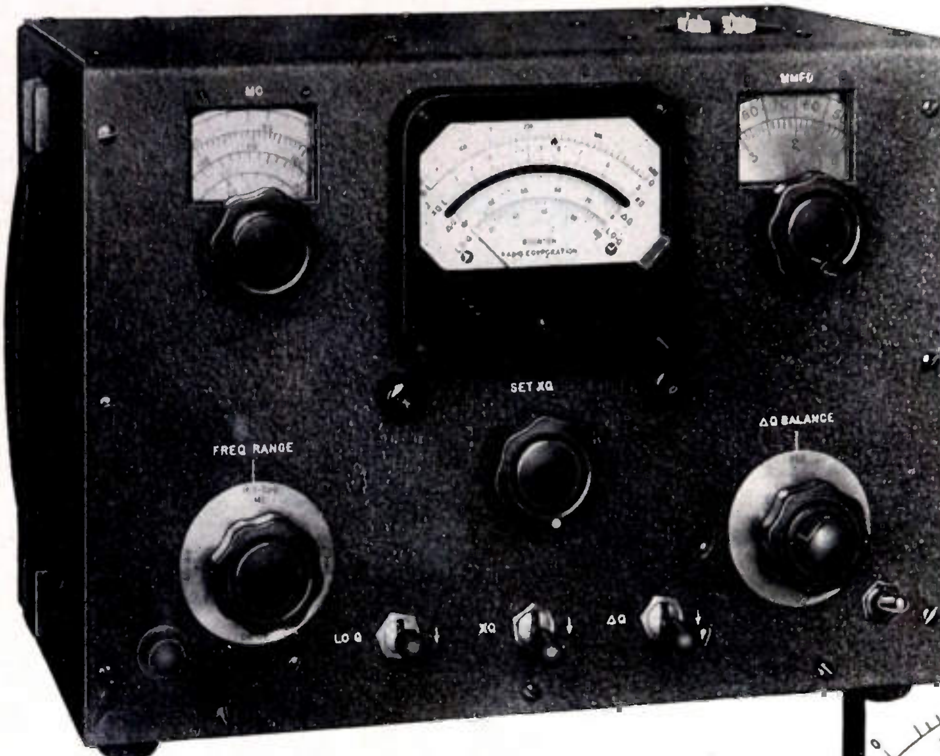
SUPER DAVOHM Precision Wire Wound RESISTORS

THE **DAVEN** CO.

195 CENTRAL AVENUE
NEWARK 4, NEW JERSEY

Measure Differential Q

with The Q METER Type 190-A



In Designing Tuned Circuits the effect on Q of adding capacitors, iron cores, or resistors must frequently be determined. The Q of the separate components is also often needed. These measurements made on Q Meters formerly available required the use of a small difference between two large Q values in various formulæ. This led to large errors. The Q Meter Type 190-A reads the difference between the Q of a reference circuit and the Q of this circuit when new components are added. The scale that indicates this *Differential Q* has a sensitivity 4 times as great as the scale which reads *Q*. The accuracy and ease with which *Differential Q* can be read is greatly improved by use of the *190-A Q Meter*.

The Q Meter Type 190-A has a "Lo Q" scale which reads Q down to a value of 5. The internal resonating capacitor is directly read and has a vernier arrangement for accurate reading of capacitance. The dial rotates approximately 10 times in covering the capacitance range. All readings are made on a single meter corrected for parallax.

SPECIFICATIONS

FREQUENCY COVERAGE: 20 mc to 260 mc. Continuously Variable in Four Ranges.
FREQUENCY ACCURACY: Calibrated to ± 1%.
RANGE OF Q MEASUREMENTS: 5 to 1200.
RANGE OF DIFFERENTIAL Q MEASUREMENTS: 0 to 100.
ACCURACY OF Q MEASUREMENTS: Circuit Q of 400 read directly on meter can be determined to accuracy of ± 5% to 100 mc and to ± 12% to 260 mc.
INTERNAL RESONATING CAPACITANCE RANGE: 7.5 mmf to 100 mmf (direct reading) calibrated in 0.1 mmf increments.
ACCURACY OF RESONATING CAPACITOR: ± 0.2 mmf to 20 mmf
± 0.3 mmf to 50 mmf
± 0.5 mmf to 100 mmf
POWER SUPPLY: 90-130 volts—60 cps (internally regulated). Power Consumption—55 watts.

(Specifications subject to change without notice)

PRICE: $625.00 F.O.B. Factory

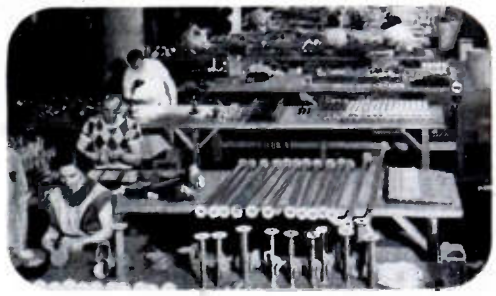
SINGLE, EASY-TO-READ METER WITH PARALLAX CORRECTION FOR ALL FUNCTIONS

- Q indicating voltmeter: 50 to 400.
- Multiply Q scale: 0.5 to 3.0.
- A differential Q scale for accurately indicating the difference in Q between two test circuits.
- Additional accurate expanded scale for measuring low values of Q.
- A counter type resonating capacitor dial for improved setting and reading accuracy.
- Regulated power supply for increased stability and accuracy.
- Careful design to minimize instrument loading of circuit under test.

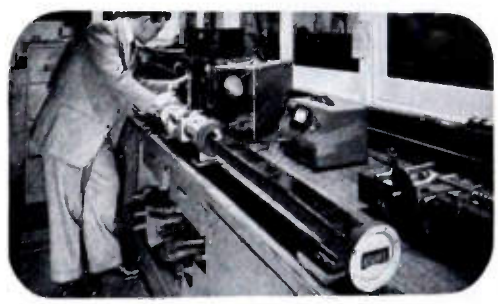
BOONTON RADIO Corporation
BOONTON·N·J·U·S·A·

A Molehill of Difference Can Make a Mountain of Trouble in Waveguides

Remember this traffic-stopper at the 1952 IRE Show in Grand Central Palace? It's a torture test. Flexed well over 1,000,000 times, Waveflex flexible Waveguides gave no evidence of failure or loss of physical or electrical properties.



Fabricated to precision methods, Titeflex flexible and rigid Waveguides are produced to the closest tolerances and to exacting specifications. Titeflex maintains strict quality control and inspection from raw materials to finished products.



No Waveguide gets by this department without a thorough electrical check-up. Every single Titeflex Waveguide is tagged before shipment with its test score on JAN-W-287 specifications for flexible Waveguides or JAN-W-85A for rigid Waveguides.

. . . A little difference in waveguides—imperceptible to the eye—can jeopardize a costly investment.

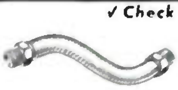
If you want to be sure of your electronic equipment, if you want to reduce operational failures, insist upon Titeflex microwave components.

Send for free catalog of uses, properties, and specifications.

Let Our Family of Products Help Yours

Titeflex

✓ *Check products you are interested in.*



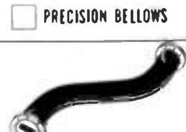
☐ SEAMED AND SEAMLESS METAL HOSE

☐ PRECISION BELLOWS

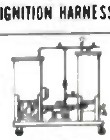
☐ IGNITION HARNESS

☐ IGNITION SHIELDING

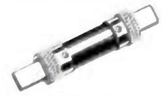
☐ ELECTRICAL CONNECTORS



☐ RIGID AND FLEXIBLE WAVE GUIDES



☐ FILTERS



☐ FUSES

TITEFLEX, INC.
511 Frelinghuysen Ave.
Newark 5, N.J.

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THE COURSE OF YOUR CAREER may depend upon what you do about your future—now. A sure way to miss success is to miss opportunity.

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Whatever your plans for the future—you will find the booklet "The Role of the Engineer in RCA" interesting reading. Write for your *free* copy.

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If you qualify for any of the positions listed above, send us a complete resumé of your education and experience, also state your specialized field preference. Send resumé to:

MR. ROBERT E. McQUISTON,
Specialized Employment Division, Dept. 94F
Radio Corporation of America,
30 Rockefeller Plaza,
New York 20, N.Y.

RCA RADIO CORPORATION of AMERICA

International RECTIFIER CORPORATION

EL SEGUNDO CALIFORNIA

Selenium Diodes

D-1224

1/8" diameter
1/4" length
Potted in thermo-setting compound.

2 Times actual Size

D-1290

RMS applied voltage, max.	26 volts per cell
Peak inverse voltage	60 volts per cell
RMS input current, max.	3.75 milliamperes
DC output voltage	20 volts per cell
Voltage drop at full load	1 volt per cell
DC output current, avg.	1.5 milliamperes
DC output current, peak	20 milliamperes
Max. surge current	80 milliamperes
Reverse leakage at 10V RMS	2.4 microamperes
Reverse leakage at 26V RMS	12 microamperes
Frequency max. CPS	100 KC

Also available in 2, 3 and 4-cell Diodes.

D-1224

RMS applied voltage, max.	26 volts per cell
Peak inverse voltage	60 volts per cell
RMS input current, max.	500 microamperes
DC output voltage	20 volts per cell
Voltage drop at full load	1 volt per cell
DC output current, avg.	200 microamperes
DC output current, peak	2.6 milliamperes
Max. surge current	10 milliamperes
Reverse Leakage at 10V RMS	0.6 microampere
Reverse Leakage at 26V RMS	3 microamperes
Frequency max. CPS	200 KC

Also available in 2-cell Diodes.

2 Times actual Size

D-1290

5/32" diameter
9/32" length
Potted in thermo-setting compound.

International RECTIFIER CORPORATION

GENERAL OFFICES:
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Phone El Segundo 1890

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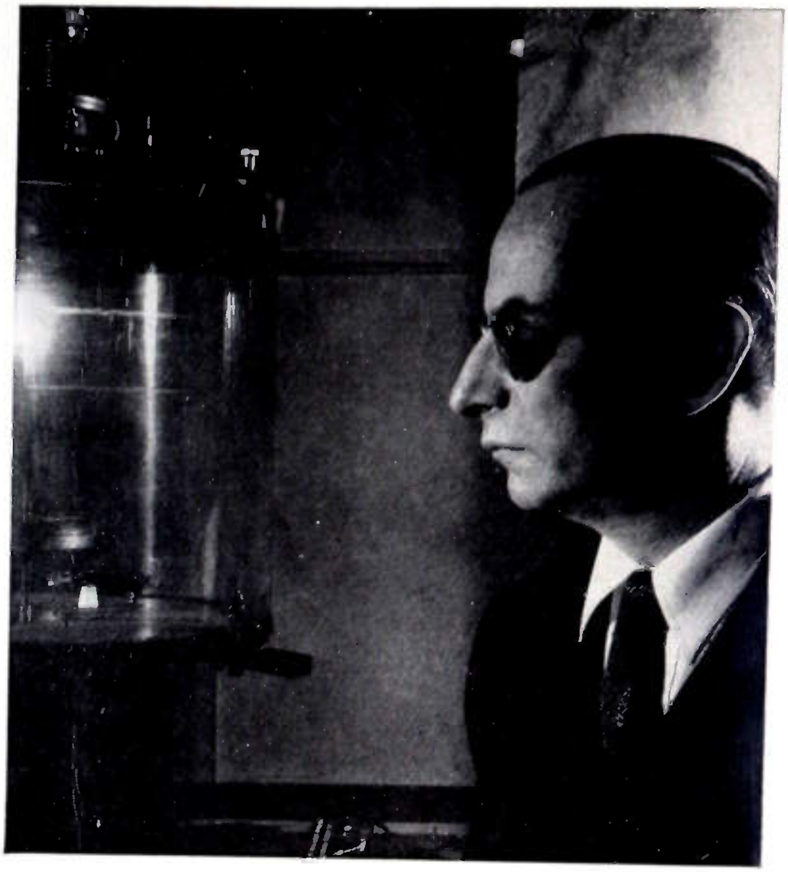
Adventurers in Research..

Dr. Joseph Slepian

INVENTOR-SCIENTIST



One of the world's foremost authorities on the behavior and control of the electric arc. He left his job as mathematics instructor at Cornell University in 1916 to work as a coil winder in the Westinghouse East Pittsburgh, Pa., plant. But his brilliant handling of engineering problems won immediate attention. In 1922 he was named head of the general research section, four years later Research Consulting Engineer, and in 1938 was appointed Associate Director of the Research Laboratories.

His colleagues at the Westinghouse Research Laboratories say of Dr. Joseph Slepian that "he can look at an electric arc and see not fire and heat, but all of the atoms, ions, and molecules arranged in a neat mathematical formula". They also say that if you want to know anything about arcs, Slepian is your man.

Dr. Slepian's work with the electric arc hasn't remained in the realm of pure mathematics, however, for he combines with it a practical knack for invention that has produced some 225 patentable ideas thus far in his career. This prolific record has prompted one of his associates to remark that "if Dr. Slepian takes a pencil out during lunch, it's almost a sure bet another patent is in the making".

He developed the "De-ion®" circuit breaker and the "De-ion" protector tube, which have helped pave the way for transmission of power at higher voltages and for the greatly improved defense of power lines against lightning. To cite just one instance, before "De-ion" flashover protectors were installed on a 47-mile stretch of line in a western state, there were 46 interruptions a year because of lightning. Afterwards, interruptions averaged less than two a year.

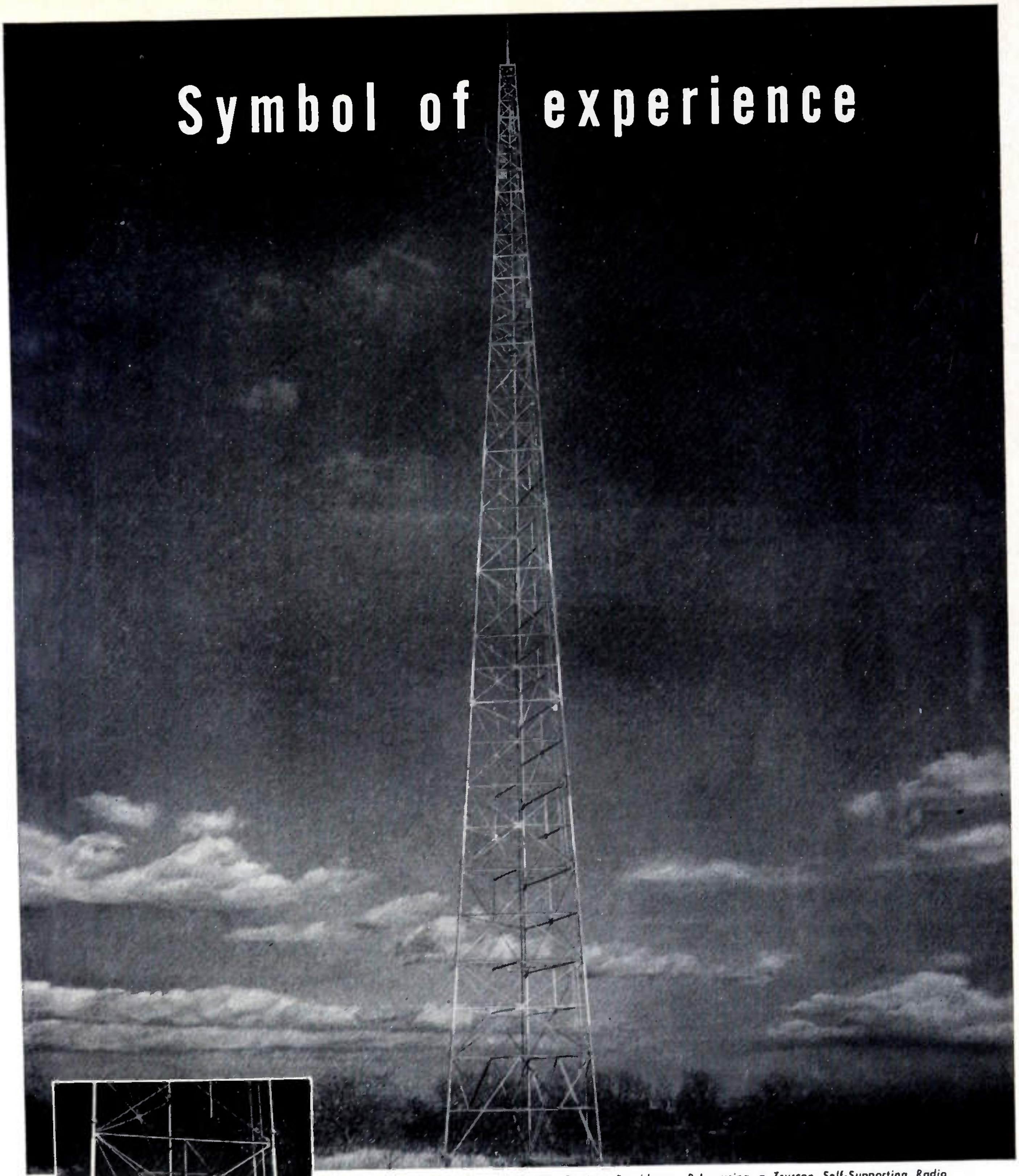
Similarly, Dr. Slepian's study of arc behavior led to the development of the Ignitron mercury-arc rectifier. Perfected in the 1930's, the Ignitron came into its own in 1940 when the requirements of aluminum production reached an all-time high. Now Ignitron installations provide the direct-current power for magnesium and aluminum plants the nation over. The Ignitron has also been adapted as the control element in electrical circuits that generate power for two of the nation's largest cyclotrons. And its most recent application is in the field of electrified locomotives, where it promises greater simplicity and economy of operation.

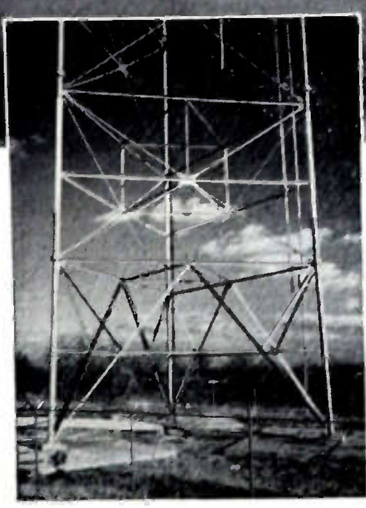
A keen and agile thinker, Dr. Slepian likes nothing better than to joust with younger researchers on scientific topics. One of his favorite hobbies is to devise plausible but impossible inventions and then challenge his colleagues to find the flaw.

Of the many honors bestowed on Dr. Slepian, nearly all have stressed the happy combination of pure science and practical inventiveness. It is the kind of combination that at Westinghouse has made for the continuous flow of new and improved equipment, while providing a fruitful source for the products of tomorrow. Westinghouse Electric Corporation, Pittsburgh, Pennsylvania.

G-10226

YOU CAN BE SURE... IF IT'S Westinghouse





WPJB-FM-TV Broadcasting Station, Providence, R.I., using a Truscon Self-Supporting Radio Tower 431 feet high, to support an RCA-TV Double Antenna 2-Section Pylon 14.7 feet high.

Years of research and development are represented in this Truscon Self-Supporting Radio Tower, designed and erected by Truscon for WPJB-FM-TV Broadcasting Station, Providence, Rhode Island.

In every corner of America, and in many foreign countries, there are outstanding examples of Truscon Tower design for AM, FM, TV and Microwave broadcasting. Truscon has designers, engineers, and fabricators with an unexcelled fund of practical knowledge to meet every tower requirement.

Your phone call or letter to any convenient Truscon district office, or to our home office in Youngstown, will bring prompt, capable engineering assistance on your tower problems. Call or write today.

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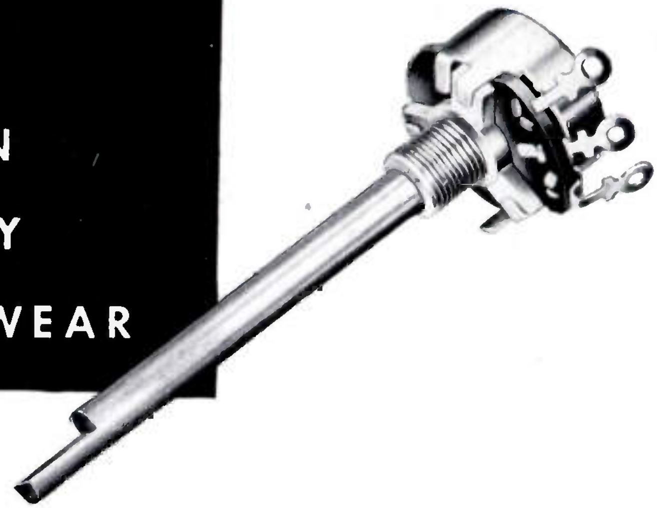

MALLORY

SERIES Q CARBON CONTROLS

withstand them all!

Now . . . from Mallory . . . you can get a carbon control that takes the toughest service conditions in stride. It's the Q series Midgetrol® . . . a new version of this outstanding control, with added features that make it applicable to the most severe requirements:

NEGLIGIBLE HUMIDITY DRIFT: carbon is deposited under precise control on a base material which affords greatly improved stability under humid conditions.

IMPROVED INSULATION: selected for unusually high insulation resistance and extremely low moisture absorption . . . thoroughly fungus-proofed.

SALT SPRAY RESISTANCE: all metal parts pass 100-hour salt spray test.

LONGER ROTATIONAL LIFE: hard nickel silver contacts limit wear, assure long service.

For special service, these additional features can be supplied:

WATERPROOFING: gasket-sealed shaft bushing packed with silicone grease, and gasket-sealed panel mounting.

VIBRATION-PROOFING: lock-type split bushing prevents shaft rotation even under heavy vibration.

Q series Midgetrols are supplied in values from 5000 ohms to 10 megohms in all standard JAN tapers. Single or dual units are available, with or without attached switch.

Wire Wound Controls
for Military applications

Look to Mallory for Q series wire wound controls made especially for Military service . . . now available with all the construction features listed for Q series Midgetrol carbon controls.

Be ready for those tough applications: find out about Mallory Q series Midgetrols now. Call or write Mallory today.

Television Tuners, Special Switches, Controls and Resistors

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SERVING INDUSTRY WITH THESE PRODUCTS:
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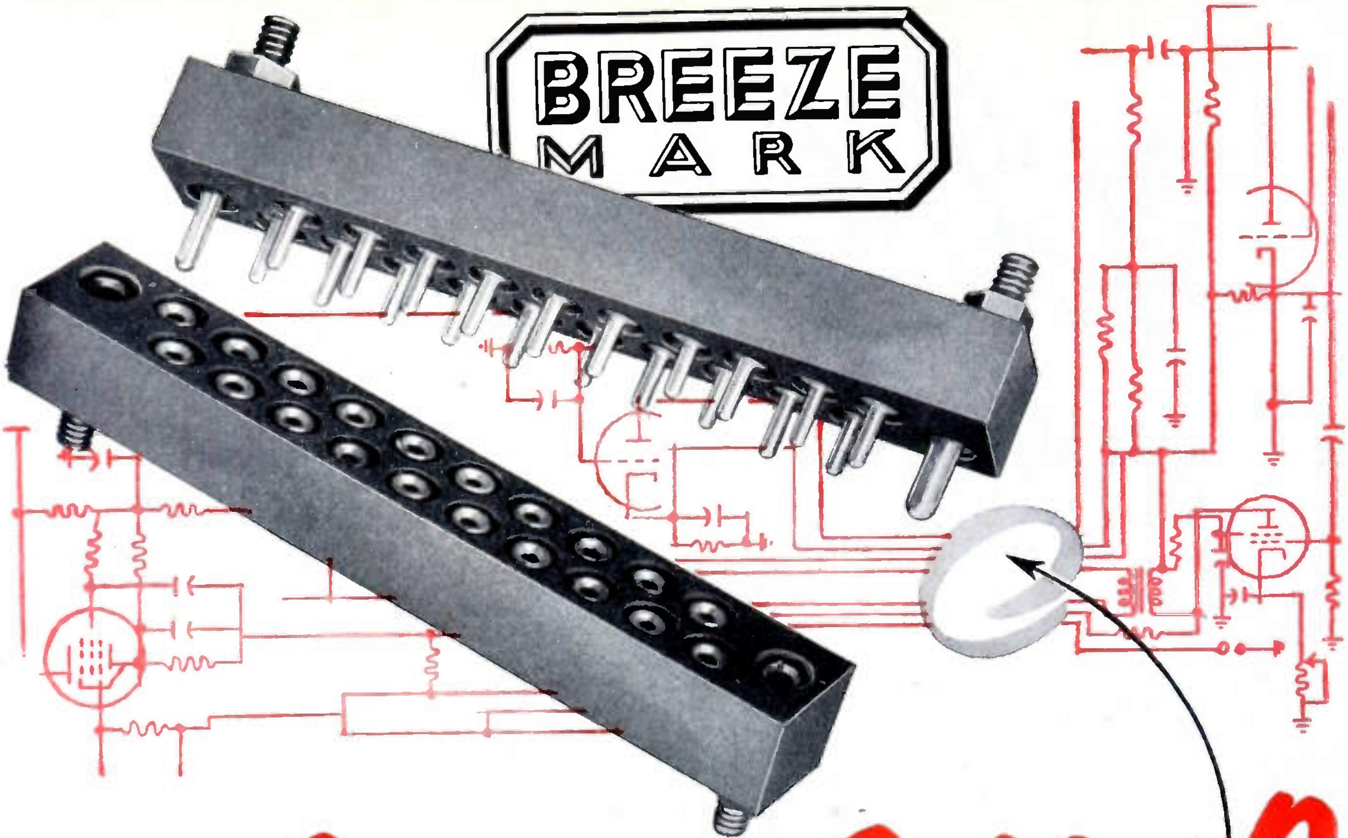


Connector Problem?

...We'll take it from HERE

Good ideas for electronic circuitry sometimes run afoul of connector problems. Maybe existing connector units won't hold air pressure gradients, won't stand the heat, aren't rugged enough for the job. Or maybe it's a question of altitude, or under-water application. But if you can sketch the circuit, we'll take it from there. We've engineered so many special connectors, solved so many "impossible" problems, that whatever the requirements are, we can usually provide the answer.


Lightweight actuators for any requirement.


Job engineered, welded-diaphragm bellows.

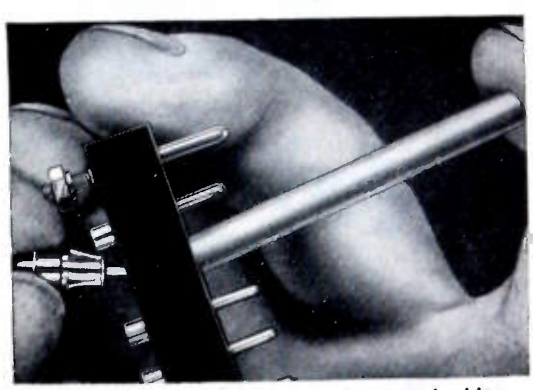

Flexible conduit and ignition assemblies.


Aero-Seal vibration-proof hose clamps.

WRITE TODAY for specific information, or send us your sketches. We'll forward recommendations promptly.

BREEZE
Special CONNECTORS

BREEZE CORPORATIONS, INC.


Removable pins in Breeze connectors speed soldering, save time, trouble. Pins snap back into block.

For *Critical* Applications

Triplett 630-A Has No Counterpart

Accuracy
to 1½%

Readability
with a Mirror-Scale

Adaptability
with ½% resistors

Try it at your distributor's

TRIPLETT ELECTRICAL INSTRUMENT CO., BLUFFTON—OHIO

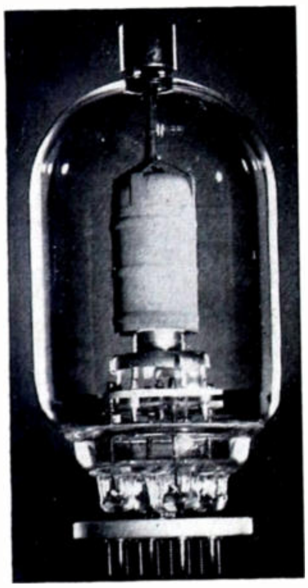
News—New Products

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

(Continued from page 28A)

Clipper Diode

Designed for high-voltage pulse applications, a new clipper diode, of the Los Gatos Type 719A, has been developed by **Lewis and Kaufman, Inc.**, Los Gatos, Calif. It is recommended for use in pulse generators where the pulse-repetition rate is 2,000 pps or less and the peak forward current is 10 amperes. Peak inverse rating is 25,000 volts.



Conservatively rated at 75 watts, the tube incorporates a new black-body heat-dissipating anode surface, termed Sinter-cote.

Of large size and designed for extreme sturdiness, the cathode is entirely supported from the base of the tube. The heater draws 7 amperes at 7 volts. The tube has a maximum height of $5\frac{7}{8}$ inches and a maximum diameter of $2\frac{9}{16}$ inches. It fits a standard No. 234 socket.

FM Signal Generator

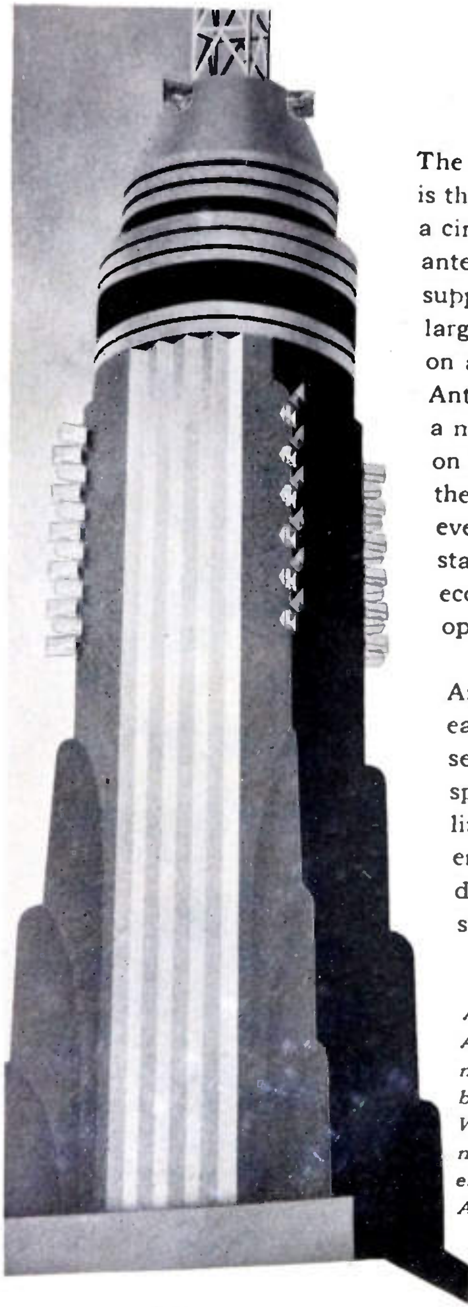
The **New London Instrument Co.**, P.O. Box 189, New London, Conn. announces the Type 100B FM signal generator, covering a 20-to-110-mc frequency range.

Features of this generator include: (1) Low distortion. At 150-kc deviation, typical distortion is 2 per cent at 1,000-cps modulation to a maximum of 4.5 per cent

(Continued on page 40A)

ANDREW "Skew" ANTENNA*

for VHF and UHF television



The ANDREW "Skew" Antenna is the *only* antenna which provides a circular radiation pattern from antenna elements placed around a supporting structure which is larger than a half wave-length on a side! With the "Skew" Antenna, it is possible to mount a multiplicity of TV antennas on the sides of tall buildings, on the sides of existing towers—even towers which also support a standard antenna on top. The economy offered by a joint operation of this type is obvious.

At present, the "Skew" Antenna is custom built for each installation and consequently general performance specifications cannot be delineated. However, ANDREW engineers will be glad to discuss its application to specific situations.

ANDREW four element "Skew" Antenna on the conical end of the mooring mast of the Empire State building, used as auxiliary by WJZ-TV. Lower on the mooring mast, artist's sketch shows the 48 element ANDREW "Skew" Antenna to be installed for WATV.

*Patents applied for

Andrew CORPORATION • ANTENNA SPECIALISTS • 363 EAST 75TH STREET, CHICAGO 19

TRANSMISSION LINES FOR AM-FM-TV-MICROWAVE • ANTENNAS • DIRECTIONAL ANTENNA EQUIPMENT • ANTENNA TUNING UNITS • TOWER LIGHTING EQUIPMENT

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 39A)

at 15,000-cps modulation. Distortions are proportionately lower for deviations smaller than 150 kc. (2) Precise Tuning. In addition to a fine tuning knob, the single-band frequency range can be more precisely adjusted with an incremental frequency dial which changes the carrier up to ±100 kc. (3) Precision piston attenuator, with a 100,000-to-0.02-μv output. (4) Low drift. No reactance tube is used. It includes a single-tube rf circuit and a temperature compensated oscillator. Low leakage. The rf compartment is enclosed in a silver-plated, cast-bronze cavity. There are minimal spurious outputs. No heterodyning, mixing, or multiplying is used. (5) External and internal (100- to 15,000-cps) modulation. The price is $950. F. O. B., New London, Conn.

Simplified Frequency Standard

Standard frequency harmonics of 1 mc, 100 kc, and 10 kc, with output frequencies as high as 1,000 mc, are available from the new GR Type 1213-A unit crystal oscillator, a product of the **General Radio Co.**, 275 Massachusetts Ave., Cambridge 39, Mass. Short-period stability (several hours) is about one part per million (0.0001 per cent).

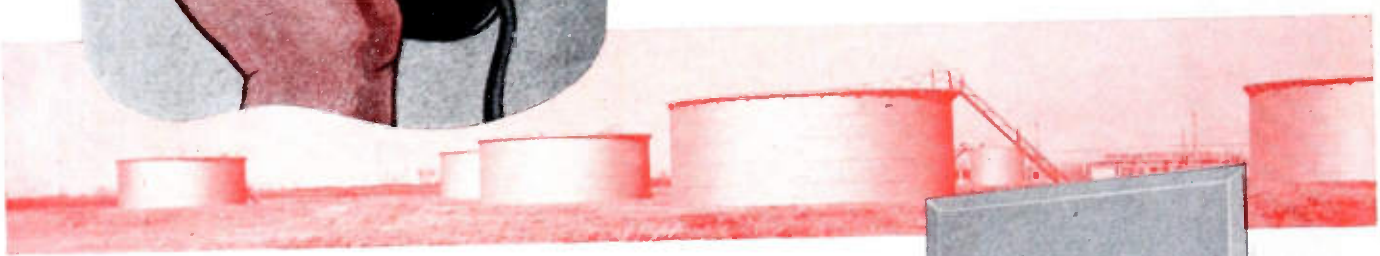
Usable 1-mc harmonics extend to 1,000 mc and the 100- and 10-kc harmonics to at least 250 and 25 mc, respectively. With good receiving equipment the 10-kc harmonics can be used to 30 mc and higher.

The 1-mc crystal is a plated, wire-mounted, hermetically sealed unit with a low temperature coefficient of frequency. The crystal and its series capacitor form a series-resonant circuit connected between two low-impedance cathode circuits. This circuit gives a very stable crystal oscillator with a minimum number of components. Following the oscillator are two 10:1 multivibrators which provide the 100-and 10-kc output frequencies.

The unit crystal oscillator is designed to be operated from a Type 1203-A unit power supply, which plugs onto the side of the case. However, any power supply capable of furnishing the proper voltages and currents can be used. The Type 1213-A unit crystal oscillator is priced at $130.00 and the Type 1203-A unit power supply at $47.50.

(Continued on page 54A)

for MARINE COMMUNICATIONS or TANK GAUGING



Relays BY GUARDIAN

Leading manufacturers of marine radio equipment specify Guardian Controls for high efficiency handling of military ship to ship and ship to shore communications. Perfected through twenty years of development and use in commercial travel ships and warcraft, Guardian controls have served so faithfully in marine radios that they've justified their selection for a wide variety of intricate applications. For example, Guardian Controls have been used *exclusively* over more than fourteen years to transmit and receive remote readings on Telematic Liquid Level gauge systems. At the touch of a button, instantaneous readings of remote storage tank levels appear on the indicator panel in the control room. Leading pipeline companies and oil refineries report not a single service call since original installations made more than fourteen years ago. Let Guardian Controls serve you . . . *just as well!*

Telematic Indicator Panel —
Telematic Corporation — Chicago 13, Ill.

AN-3320-1 D.C.

AN-3324-1 D.C.

Series 595 D.C.

Series 610 A.C.—615 D.C.

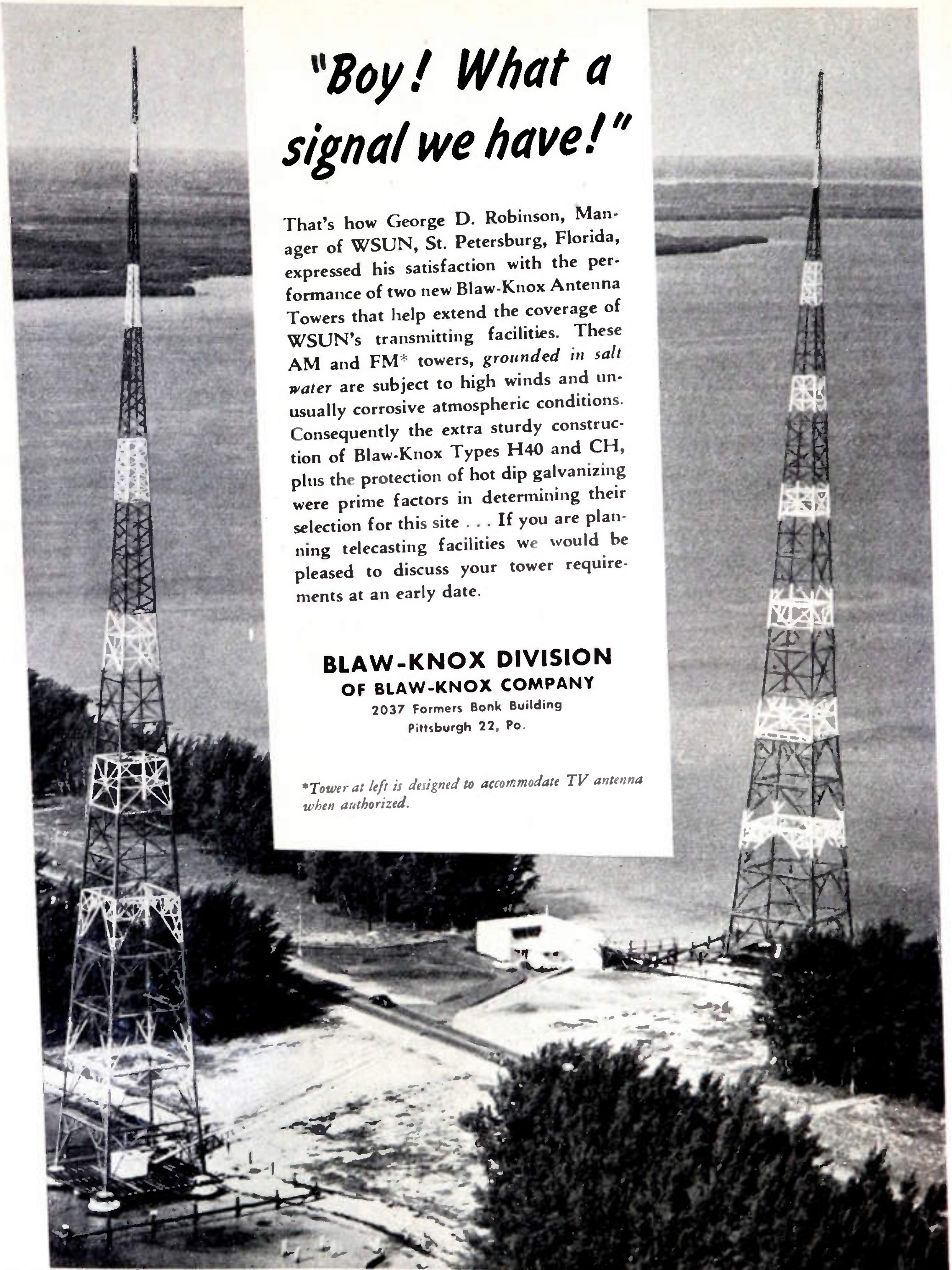
Series 695 D.C.

Get Guardian's New HERMETICALLY SEALED RELAY CATALOG Now!

GUARDIAN G ELECTRIC

1628-G W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY



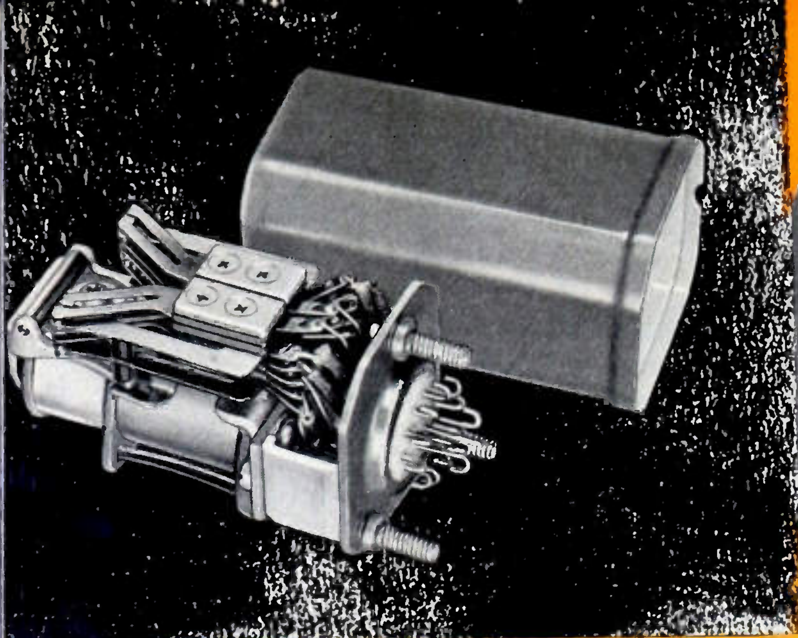
"Boy! What a signal we have!"

That's how George D. Robinson, Manager of WSUN, St. Petersburg, Florida, expressed his satisfaction with the performance of two new Blaw-Knox Antenna Towers that help extend the coverage of WSUN's transmitting facilities. These AM and FM* towers, *grounded in salt water* are subject to high winds and unusually corrosive atmospheric conditions. Consequently the extra sturdy construction of Blaw-Knox Types H40 and CH, plus the protection of hot dip galvanizing were prime factors in determining their selection for this site . . . If you are planning telecasting facilities we would be pleased to discuss your tower requirements at an early date.

BLAW-KNOX DIVISION
OF BLAW-KNOX COMPANY
2037 Formers Bonk Building
Pittsburgh 22, Po.

**Tower at left is designed to accommodate TV antenna when authorized.*

BLAW-KNOX *ANTENNA* TOWERS



CLARE *Hermetically Sealed* RELAYS

Clare Type "J" twin contact relay before sealing in enclosure.

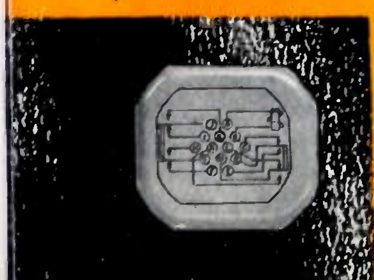
Cutaway view of Clare Type "K" hermetically sealed relay.

offer designers wide range of relays and enclosures

Another type of enclosure (cutaway) shows compactness of assembly.

"AN" approved Clare hermetically sealed relay with "AN" plug connector.

Connection diagrams are clearly and permanently imprinted on enclosures by silk screen process.



Long established as leaders in the design and manufacture of superior relays for all types of industrial use, C. P. Clare & Co. are pioneers in the development of a method of hermetic sealing which insures their long-life protection against unfavorable atmospheric and environmental conditions.

Hermetic sealing, as practiced by Clare, injects an ideal atmosphere of dry inert gas and seals it in to provide permanent immunity from the natural enemies of relays—moisture, pressure and density changes, salt, corrosive fumes, dust and fungi.

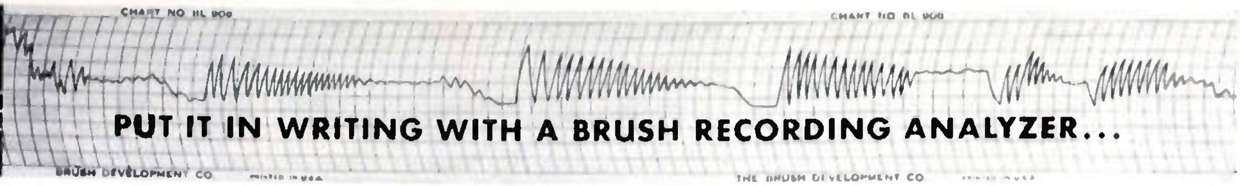
Fifty and more different series of Clare hermetically sealed relays are now available to relay users. Within each series innumerable variations of coil and contact specifications are possible.

For a full treatment of the subject of hermetic sealing—the Clare way—as well as a description of many types of Clare hermetically sealed relays, write for Clare Sales Engineering Bulletin No. 114.

Address C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials Ltd., Toronto 13.

CLARE RELAYS

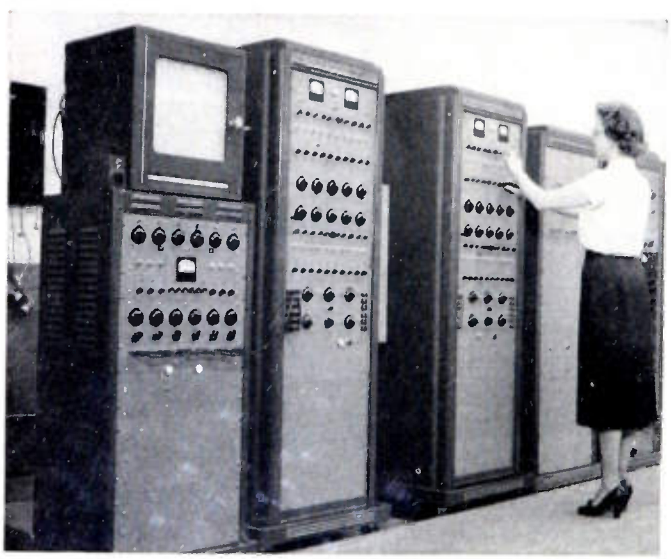
First in the Industrial Field



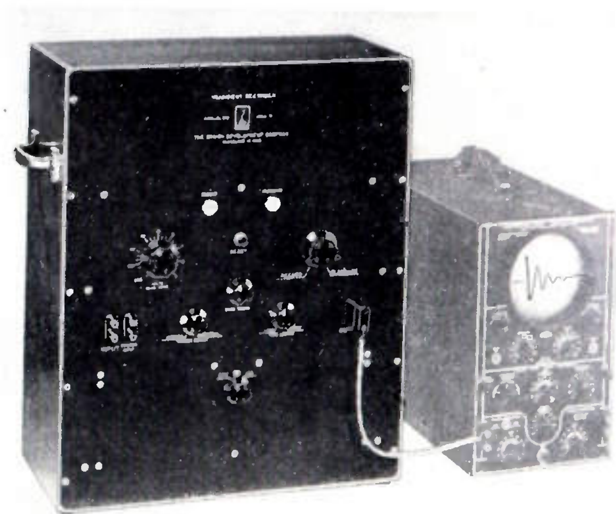
Records Voltage and Current On One Chart!

● This Brush Direct-writing Dual-channel Oscillograph plots starting voltage and current of a fluorescent lamp simultaneously . . . thus aids a leading manufacturer in design and test work.

Use the Brush Magnetic Oscillograph, in combination with the proper Brush Amplifier, to make an immediately available direct chart recording of physical and electrical phenomena. Direct-inking or electric stylus models available. Gear shift provides chart speeds of 5, 25, and 125 mm per second. An auxiliary chart drive is available for speeds of 50, 250, and 1250 mm per hour. Accessory equipment provides event markers where an accurate time base is required, or where it is desirable to correlate events.



RECORDS SIX VARIABLES SIMULTANEOUSLY. The Brush six-channel Magnetic Oscillograph is designed for simultaneous recording of six electrical phenomena, with a chart record immediately available. In this application the results of six different computations of an electronic differential analyzer are recorded. Instrument facilitates multiple strain measurement, vibration analysis, wind tunnel work, circuit analysis, etc. Either d-c or a-c phenomena up to 100 cycles can be recorded.



"PLAYS BACK" TRANSIENTS. The Brush Transient Recorder is designed to record and reproduce transient phenomena of ⅕ second or less. This instrument records transients on tape, then reproduces them for visual analysis on an oscilloscope. Signals can be shown complete, or expanded on the screen to show detail. Electrical transients or other transients which can be converted into electrical impulses can be studied.

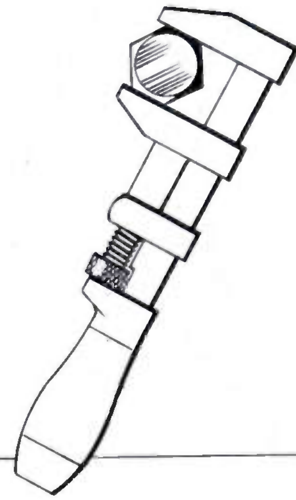
For Bulletin 618 giving details on these instruments, write The Brush Development Co., Dept. F-31, 3405 Perkins Ave., Cleveland 14, Ohio. Representatives located throughout the U. S. In Canada: A. C. Wickman Ltd., Toronto.

THE Brush DEVELOPMENT COMPANY

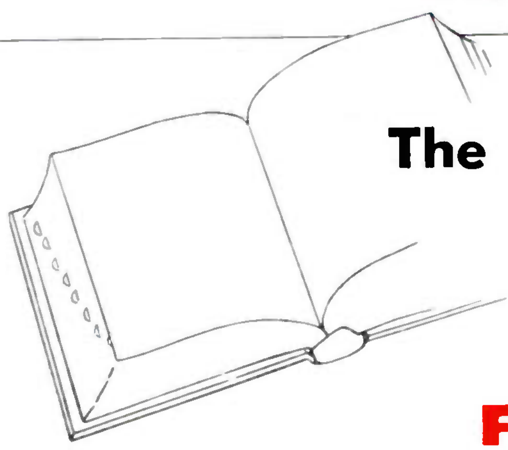
BRUSH

Piezoelectric Crystals and Ceramics
Magnetic Recording
Acoustic Devices
Ultrasonics
Industrial & Research Instruments

You can dye it
and call it MINK
but it's NOT



You can tighten it up
and call it "HERMETIC"
but it's NOT



The dictionary says "hermetic"
means made airtight by
fusion or soldering.

FUSITE GLASS-TO-METAL TERMINALS PERMIT A TRUE HERMETIC SEAL, QUICKLY, EASILY AND INEXPENSIVELY.

INSULATE ELECTRICALLY
GLASS TO METAL
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TERMINALS
PROTECT PRODUCT PERFORMANCE
SEAL HERMETICALLY

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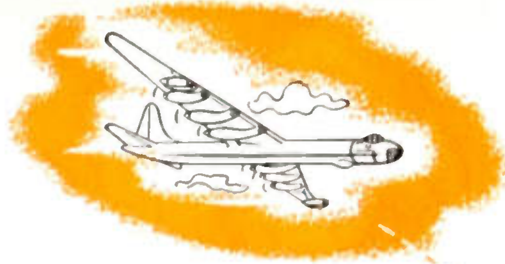
6028 FERNVIEW AVENUE - CINCINNATI 13, OHIO



MILITARY dehydrators

FOR AIR, LAND AND

TYPICAL APPLICATIONS IN WHICH CP DEHYDRATORS PROVIDE YEAR 'ROUND TROUBLE-FREE AUTOMATIC SERVICE:

- Purging and pressurizing transmission lines, waveguides and associated apparatus.
- Pressurizing large cavities and other radio and radar equipment enclosures.
- Fog prevention in precision optical systems.
- Corrosion prevention in precise servo amplifier assemblies.
- For raising and maintaining the power handling capacity of high voltage systems and apparatus and innumerable other similar applications.

CP DEHYDRATORS OFFER THE FOLLOWING UNIQUE FEATURES:

Low dewpoint • operating pressure up to 100 lbs. per square inch fully automatic operation • continuous duty performance • low noise level • minimum vibration • long service life with minimum maintenance

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SEABORNE SERVICE

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CP dehydrators are readily adaptable to the critical requirements of the Armed Forces. Standardized parts permit rapid assembly of equipments suitable for practically any specialized need at minimum cost and without prolonged delay. Over a decade of CP experience in dehydrator design and manufacture insures products of long life and dependable service with an absolute minimum of maintenance. Inquiries are invited.

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MARLBORO, NEW JERSEY
Telephone: FReehold 8-1880

DIPOLE ANTENNAS, SWITCHES, Q-MAX LACQUER AND CEMENT

Oscillograms tell the story of the NEW DUMONT Type 303-A

EXCELLENT FREQUENCY RESPONSE . . . Figure 1 shows faithful reproduction, lack of overshoot of 0.8 μsec, 7 volt peak pulse through attenuator and amplifier (middle waveform) compared with same pulse directly to deflection plates (upper) . . . internally generated 1 MC timing signal is imposed below . . . note that high sensitivity of Type 5YP- Cathode-Ray Tube is responsible for large deflection of directly connected pulse . . . gradual drop-off of frequency response permits viewing of sine-wave signals greater than 20 MC.

PULSE RISE TIME MEASUREMENTS . . . Rise time of the 0.8 μsec pulse seen in Figure 1 is easily measured . . . Figure 2 shows the rise time at a sweep speed of 10"/μsec (25.4 cm/μsec) determined by the 10 MC internally generated timing signal . . . Between 10% and 90% amplitude points, pulse rise time measures 0.4" or 0.04 μsec . . . Y-amplifier rise time of the new Type 303-A is 0.033 μsec . . . pulse is found to be of 0.02 μsec rise time from the relation:

$$T_{pulse} = \sqrt{T^2_{measured} - T^2_{amplifier}}$$

WIDE-RANGE POSITIONING CONTROL . . . Fall time of the 0.8 μsec pulse seen in Figure 3 is easily positioned on screen . . . writing rate remains at 10"/μsec, fall time occurring 8" after rise time on this time base . . . sweep is expanded to 6 times full screen diameter without appreciable distortion and any portion of sweep may be positioned on screen.

HIGH SWEEP SPEEDS . . . Sweep speeds considerably in excess of the rated 10"/μsec are available as shown by Figure 4 where a single cycle of 10 MC timing signal covers 2" on screen . . . above 10"/μsec, some sacrifice in positioning range and sweep linearity is experienced but measurements are still made accurately by time-calibration substitution.

ACCURATE TIME AND AMPLITUDE MEASUREMENTS . . . In Figure 5 sweep speed is 2"/μsec (5.08 cm/μsec) as shown by the 10 MC timing signal . . . vertical sensitivity is set at 5 volts/inch (2 volts/cm) by the 10 volt internally-generated amplitude marker . . . The pulse is seen to be 0.8 μsec duration measured between 50% amplitude points and 7.2 volts peak amplitude . . . note the 1.5" of undistorted deflection from the unidirectional signal.

The *illuminated calibrated scale* seen in all the oscillograms is supplied with the instrument as well as *suitable filter* for visual contrast. A new Du Mont Type 2592-52 *Shielded Coaxial Adapter* with 52 ohm termination is also supplied for use in connecting to the Type 303-A signals that are carried on coaxial lines.

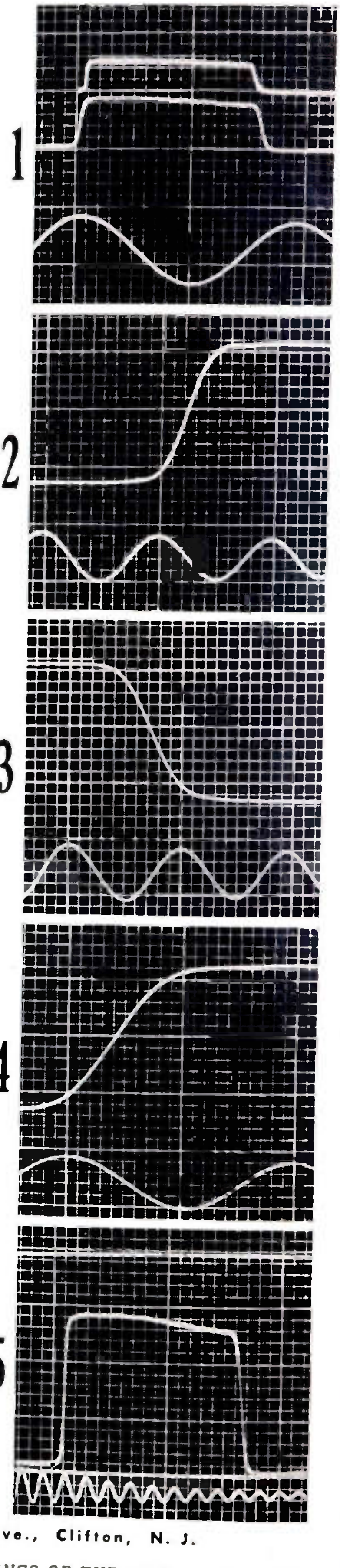
Let us make this demonstration for you . . .

Write to

Instrument Division
Allen B. Du Mont Laboratories, Inc.
1500 Main Avenue, Clifton, New Jersey

SPECIFICATIONS

- Y-Sensitivity: 0.1 p-p v/in (0.04 p-p v/cm.)
- Y-Frequency Response: Down less than 30% at 10 cps and 10 MC.
- Pulse Response: 0.033 μsec.
- X-Frequency Response: d-c to 700 KC (30% down.)
- Sweep Speeds: 0.1 sec to 2 μsec; expansion on all ranges to 6 times full screen; max. linear sweep speed better than 10"/μsec (25.4 cm/μsec.)
- Amplitude Calibration: 0.1, 1.0, 10, 100 volts, better than ± 5% accuracy.
- Time Calibration: 0.1, 1.0, 10, 100 μsec, better than ± 3% accuracy.
- Illuminated scale with dimmer control.
- Du Mont Type 2592-52 shielded coaxial adapter with 52 ohm termination included.



1

2

3

4

5

DUMONT *for Oscillography*

PRICE $825

Allen B. DuMont Laboratories, Inc. Instrument Division, 1500 Main Ave., Clifton, N. J.

You can use yoga to make it fit...

BUT it's simpler to design the radio around the battery!

"Eveready" No. 950 "A" batteries and the No. 467 "B" battery make an ideal combination for small portable receivers.

"EVEREADY" "Nine-Lives" radio batteries offer you a complete range of *standard* types and sizes. You can *start* with the batteries and design *around* them . . . regardless of the type or size of new-model receiver.

Compact and long-lasting, "EVEREADY" radio batteries give better radio performance with fewer replacements. And, when replacements *are* necessary, they're a cinch for the user to obtain because "EVEREADY" brand batteries are available everywhere.

Write to our Battery Engineering Department for full details and specifications of "EVEREADY" radio batteries.

EVEREADY
TRADE-MARK
RADIO BATTERIES

The terms "Eveready", "Nine Lives" and the Cat Symbol are trade-marks of Union Carbide and Carbon Corporation

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A Division of Union Carbide and Carbon Corporation
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District Sales Offices: Atlanta, Chicago, Dallas, Kansas City, New York, Pittsburgh, San Francisco
In Canada: National Carbon Limited, Montreal, Toronto, Winnipeg

Ultra-Sensitive

DC MICROAMMETER

RCA WV-84 A

Battery-operated . . . completely portable

FEATURES:

- Reads from 0.0002 microampere to 1000 microamperes in six ranges. Will indicate current flow below one-billionth ampere.
- Can be used with external battery to measure extremely high resistance values in the order of billions of ohms.
- Meter movement electronically protected against burnout.
- Can be used as a voltmeter (external multipliers included) to measure voltages from 0.1 volt to 10 volts at input resistances from 100 to 1000 megohms.
- Voltage drop for full-scale deflection on all ranges is only 0.5 volt. Has 50-megohm input resistance on lowest range.
- Battery-operated for excellent stability and complete freedom from effects of power-line voltage fluctuations. Readily portable.

$100.00, Suggested User Price (Batteries not included).

SPECIFICATIONS:

SIX DC CURRENT RANGES:

0 to 0.01, 0.1, and 1 microampere: 0 to 10, 100, and 1000 microamperes.

ACCURACY:

On 0.01-Microampere Range . ± 5% of full scale
On All Other Ranges . . . . ± 4% of full scale
Voltage Drop for Full-Scale Deflection . . . . 0.5 volt on all ranges

INTERNAL SHUNT RESISTANCE:

0.01-μamp Range . . . . 50 megohms
0.1-μamp Range . . . . 5 megohms
1-μamp Range . . . . 0.5 megohms
10-μamp Range . . . . 50,000 ohms
100-μamp Range . . . . 5,000 ohms
1000-μamp Range . . . . 500 ohms

POWER SUPPLY:

"A" Batteries . . . . 2, 1½ volts (RCA-VS106)
"B" Batteries . . . . 2, 22½ volts (RCA-VS102)

DIMENSIONS:

9½" High, 6¼" Wide, 5¾" Deep

WEIGHT:

9½ lbs. (incl. batteries).

Useful in laboratories, industrial plants, broadcast stations . . . for the measurements of minute currents, critical voltages, high resistance

The RCA-84A Ultra-Sensitive DC Microammeter is a battery-operated vacuum-tube microammeter designed for the measurement of minute direct currents. The instrument has six scales for reading currents from 0.0002 microampere to 1000 microamperes; a ratio of 5,000,000 to 1.

The amplifier circuit is designed so that the maximum meter current is limited to a safe value. This feature protects the instrument against meter burnout. The meter has a large face with wide scale divisions that are easy to read accurately. The meter movement is suitably damped to bring the pointer quickly to its reading position with negligible overswing and without oscillation. The selector switch opens the battery circuits when in the "off" position, and, in addition, functions as a polarity-reversing switch to eliminate the need for reversing leads when the current polarity changes.

The vacuum tubes employed have low-drain filaments. In addition, the circuit has been designed to keep the plate current low. Consequently, batteries have an exceptionally long life.

• • •

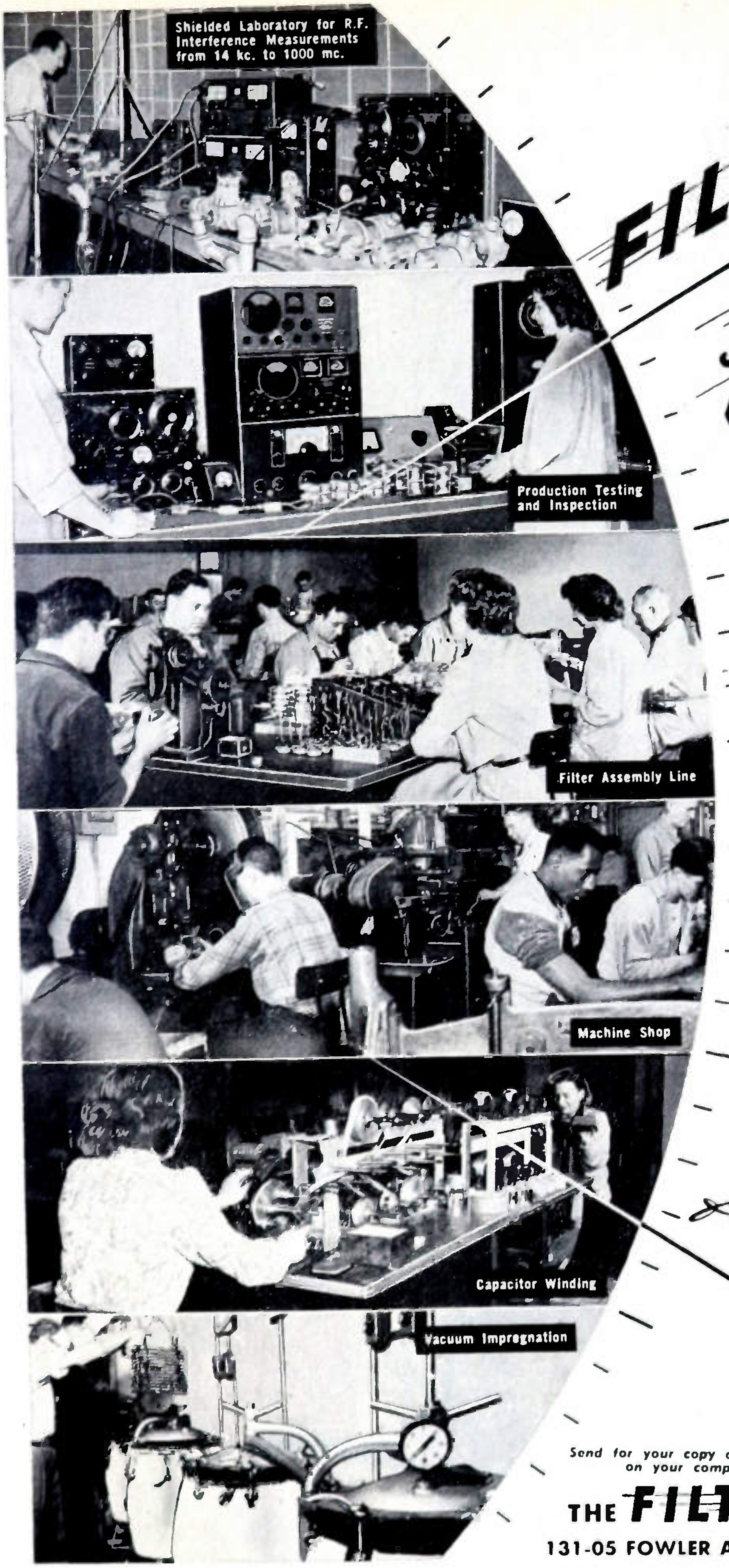
Ask your *RCA Test Equipment Distributor* for descriptive bulletin, or write RCA, Commercial Engineering, Section FX 47, Harrison, N. J.

TMKS. ®

RADIO CORPORATION of AMERICA

TEST EQUIPMENT

HARRISON, N. J.



FILTRON'S *Ahead*

in

- QUALITY
- PRODUCTION
- ENGINEERING

FILTRON'S COMPLETE FACILITIES: capacitor manufacturing, coil winding, stamping department, tool and die shop, and assembly department, together with its Engineering and Research laboratories insures quality production and ON SCHEDULE DELIVERY.

FILTRON'S *advanced* engineering, due to constant research and development by engineers with years of R.F. INTERFERENCE experience has resulted in *smaller, lighter,* and more *efficient* R.F. Interference Filters.

FILTRON'S completely equipped shielded laboratory is available for the R.F. Interference testing and filter design for your equipment *...to specification requirements.*

With over *400 standard filters* to choose from, and the engineering know-how to custom design filters to meet specific size and mounting requirements, Filtron's engineers will specify the *right filter* for your application.

FILTRONS are suppressing R.F. Interference in modern Military Aircraft, Naval Equipment and Ground Signal Installations.

FILTERED BY FILTRON means ...
RADIO INTERFERENCE FREE PERFORMANCE.

RF INTERFERENCE FILTERS FOR:

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Electric Motors	Business Machines
Electric Generators	Electric Appliances
Electronic Equipment	Electronic Signs
Fluorescent Lights	Electronic Heating Equipment
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LARGEST EXCLUSIVE MANUFACTURERS OF RF INTERFERENCE FILTERS

Training FOR Tactical Defense

GUIDED MISSILES have become one of the major military elements in the over-all tactical defense picture.

Troops of the U. S. Army are being trained in the handling and application of these ground-to-air missiles, designed and produced specifically for tactical training purposes by Fairchild.

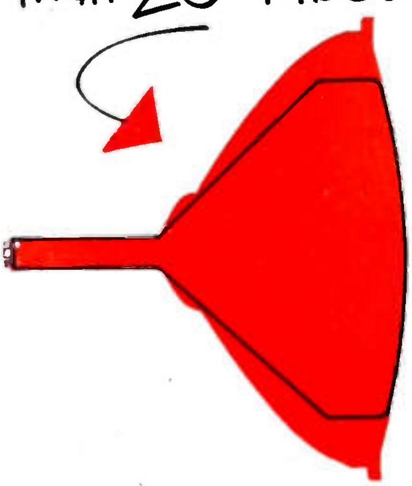
This program, forming the basis for future use of missiles requires specialized training on these specialized weapons.

Fairchild experts and equipment are playing their part in this basic tactical missile training program for the use of this effective defense weapon of the future.

ENGINE AND AIRPLANE CORPORATION
FAIRCHILD
Guided Missiles Division
WYANDANCH, L.I., N.Y.

Fairchild Aircraft Division, Hagerstown, Md.
Fairchild Engine, Farmingdale, N.Y.
Stratos Division, Bay Shore, L.I., N.Y.

Actually shorter than 20" tube!



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

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

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

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

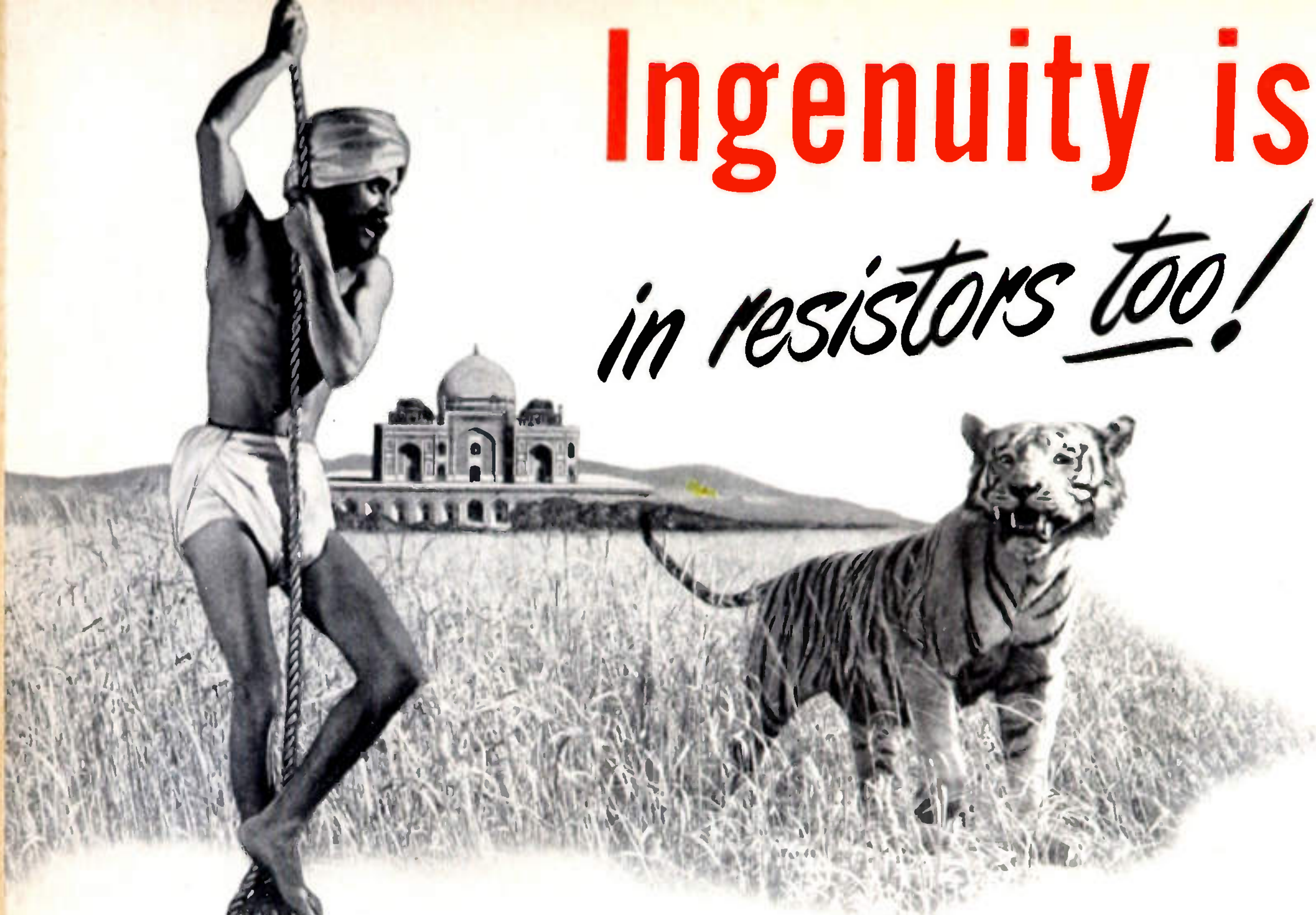
A picture actually more than 70 sq. in. larger than the center spread of a tabloid newspaper. Rectangular for minimum cabinet height and width. And actually permitting a small reduction in depth from today's 20" cabinets!

THE RAULAND CORPORATION

Perfection Through Research

4245 N. Knox Avenue, Chicago 41, Illinois

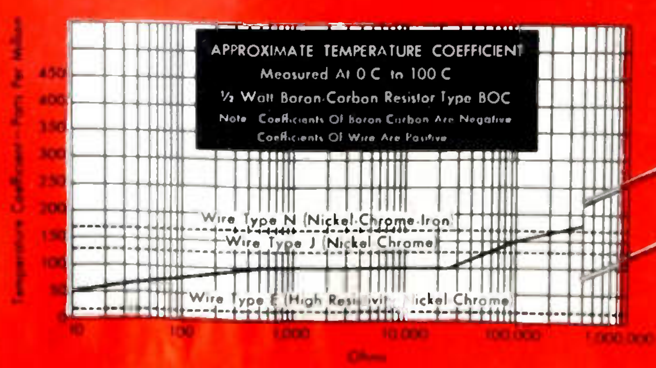
Ingenuity is *in resistors too!*



New Materials — New Techniques — New Advantages Features in 4 New IRC Resistors

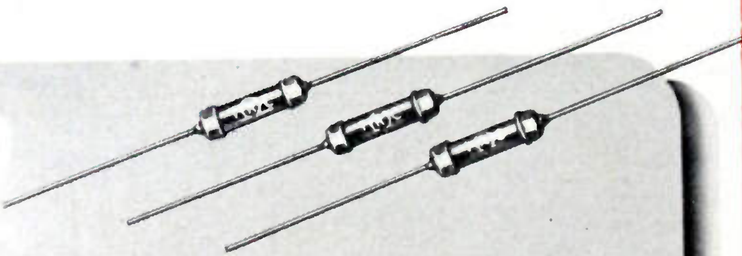
1 IRC Type BOC Boron-Carbon ½-Watt PRECISTOR Meets All Requirements of MIL-R-10509 Specification

No other non-wire-wound resistor combines the advantages of this all-new Boron-Carbon unit. Type BOC reduces the temperature coefficient of conventional deposited carbon resistors—provides high accuracy *and* long-time stability—replaces high value wire wound precisions at savings in space and cost. You'll find it adaptable to a host of critical circuitry needs—in electronics and avionics, communications, telemetering, computing and service instruments. Send for full details in Catalog Data Bulletin B-6.



Type BOC conforms to *all* requirements of MIL-R-10509. Exposed to a temperature of [illegible]5°C. for one hour, the new BOC shows a resistance change of less than .2%. High temperature operation with reliability is now possible. Voltage coefficient is less than 20 parts per million per volt. Load life is outstanding; on a 500-hour test at ambient temperature of 40°C., resistance change will not exceed 2%.

essential



2

New IRC Type DCC (Deposited Carbon) Small-Size, High-Stability Resistors

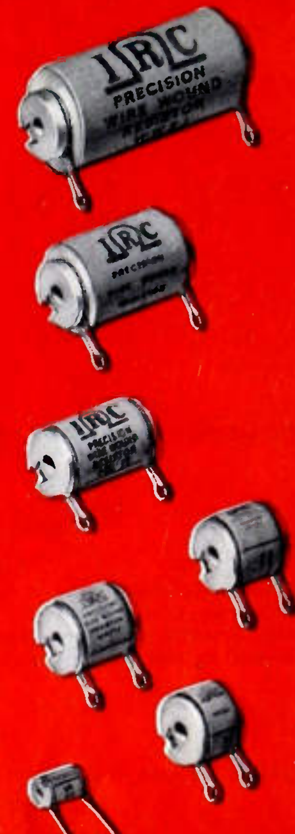
This is the latest small-size addition to IRC's famous line of deposited carbon PRECISTORS. Conservatively rated at ½ watt, it combines accuracy *and economy*—assures high stability, low voltage coefficient, and low capacitive and inductive reactance in high frequency applications. Recommended for:—Metering and voltage divider circuits requiring high stability and close tolerance—High frequency circuits demanding accuracy and stability—Other critical circuits in which characteristics of carbon compositions are unsuitable and wire-wound precisions are too large or expensive. Type DCC meets Signal Corps Specification MIL-R-10509. Complete technical data in Catalog Bulletin B-7.

3

New IRC Type FS Fuse Resistor

This completely insulated unit functions as a resistor under normal conditions and as a fuse under abnormal conditions. Small, compact, stable, it can be wired into a circuit as easily as a molded wire-wound resistor. Bulletin B-3.

4 New IRC Type WW Precision Wire Wounds Surpass JAN-R-93 Characteristic B Specifications



Here is the most reliable and stable of all wire-wound precisions . . . by unbiased test! Actually, new Type WW's far surpass JAN-R-93 Characteristic B Specifications. *New winding forms* hold more wire for higher resistance values. *New winding technique* and rigid insulation tests eliminate possibility of shorted turns or winding strains. *New type insulation* withstands humidity, assures long life, provides stability and freedom from noise. *New terminations* (except in small size WW-10) are rugged lug terminals for solder connection. Full data in Catalog Bulletin D-3.

–	Original Resist.	1st Cycle % Chge	2nd Cycle % Chge	3rd Cycle % Chge	4th Cycle % Chge	Resist. at End of 100 hrs. load	Total % Chge	% Chge from Last Temp. Cycle to End of 100 hrs. load %	Resistance Chge at End of 100 Hrs. Load only % no cycling	
1	100,010	+.04	+.04	+.05	+.05	100,050	+.04	—.01	100 040	—.02
2	100,000	+.03	+.04	+.03	+.05	100,060	+.06	+.01	100,000	0
3	100,000	+.01	+.02	+.02	+.05	100,000	0	+.05	100,050	—.02
4	100,000	+.02	0	+.02	+.02	100,000	0	—.02	100,040	—.01
5	100,010	+.03	+.04	+.04	+.05	100,000	0	—.05	100,030	—.03
6	100,000	0	+.03	+.04	+.04	100,100	+.1	+.06	99,980	0
7	100,000	+.04	+.05	+.04	+.04	100,070	+.07	+.03	100,000	0
8	100,000	+.03	+.05	+.05	+.05	100,050	+.05	0	100,000	0
9	100,000	+.04	+.03	+.05	+.04	100,010	+.01	—.03	100,050	0
10	100,000	+.02	+.02	+.02	+.04	100,010	+.01	—.03	100,000	0
11	100,000	0	+.01	+.01	+.03	100,000	0	—.03		

Tested side-by-side with competing resistors, new IRC Type WW's proved superior to all. Severe cycling and 100-hour load tests resulted in virtually zero changes in resistance. Other stringent tests proved Type WW's high mechanical strength, freedom from shorting, resistance to high humidity.

For full information on these products, or assistance in adapting them to any specific application, write IRC. Types BOC and DCC are currently available on short delivery cycles to manufacturers of military equipment only.

Boron-Carbon PRECISTORS • Power Resistors • Voltmeter Multipliers • Insulated Composition Resistors • Low Wattage Wire Wounds • Volume Controls • Voltage Dividers • Precision Wire Wounds • Deposited Carbon PRECISTORS • Ultra-HF and High Voltage Resistors • Insulated Chokes

Wherever the Circuit Says —ʌʌʌ—

INTERNATIONAL RESISTANCE COMPANY

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

Mail Coupon Today for Full Details of These New IRC Resistors

INTERNATIONAL RESISTANCE CO.,
405 N. BROAD ST., PHILADELPHIA 8, PA.

Please send me full data on the following checked items:—

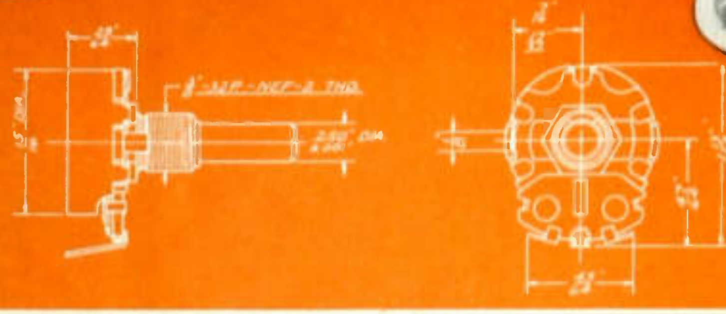
- ☐ Type BOC Boron-Carbon PRECISTORS
- ☐ Type WW Precision Wire Wounds
- ☐ Type DCC Deposited Carbon PRECISTORS
- ☐ Type FS Fuse Resistors
- ☐ Name and Address of Nearest IRC Distributor

NAME . . . .
TITLE . . . .
COMPANY . . . .
ADDRESS . . . . CITY . . . . ZONE . . . STATE . . . . .

A Complete Line

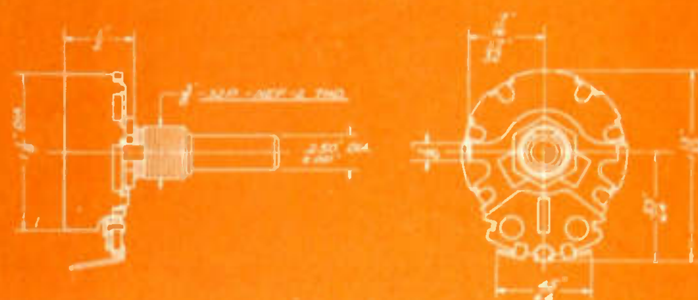
that

TYPE 45
(JAN-R-94, Type RV2)



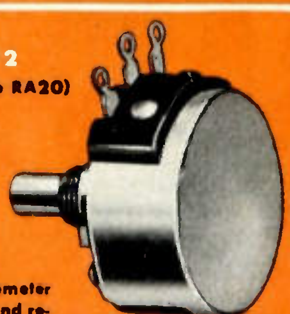
¼ watt, 15/16" diameter variable composition resistor.
Also available with other special military features not covered by JAN-R-94. Attached Switch can be supplied.

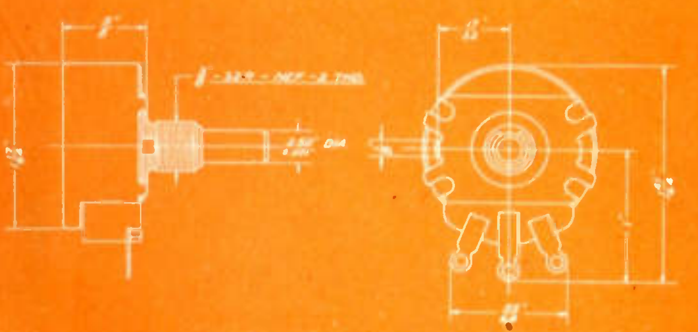
TYPE 35
(JAN-R-94, Type RV3)



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

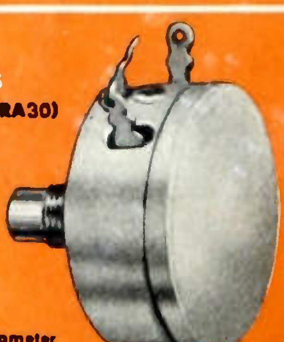
TYPE 252
(JAN-R-19, Type RA20)

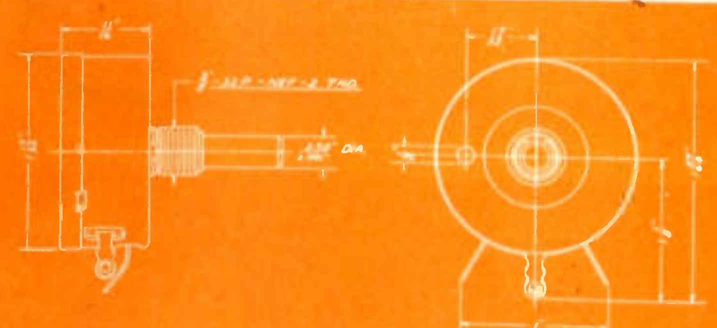




2 watt, 1 11/64" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19. Attached Switch can be supplied.

TYPE 25
(JAN-R-19, Type RA30)





4 watt, 1 17/32" diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19. Attached Switch can be supplied.

For additional information on these 7 controls, write for Data Sheet No. 160

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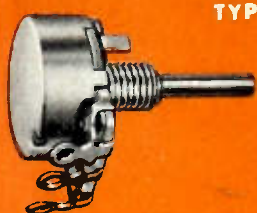
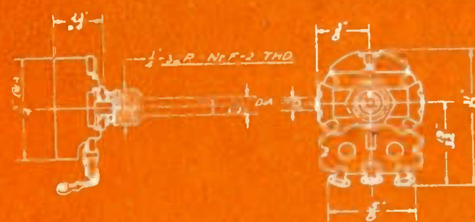
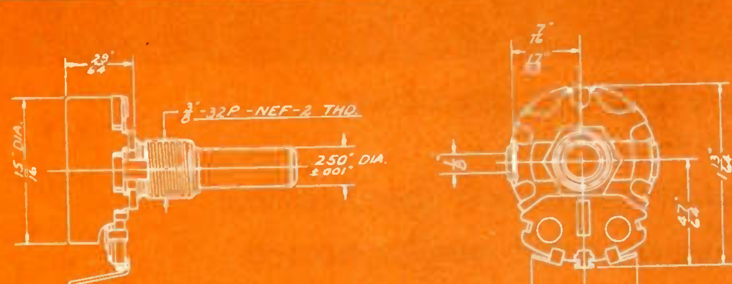




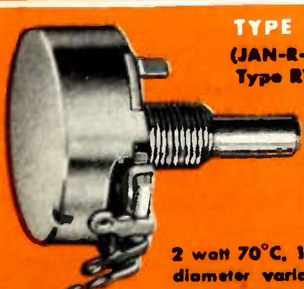
½ watt 70°C, ¾" diameter miniaturized variable composition resistor.

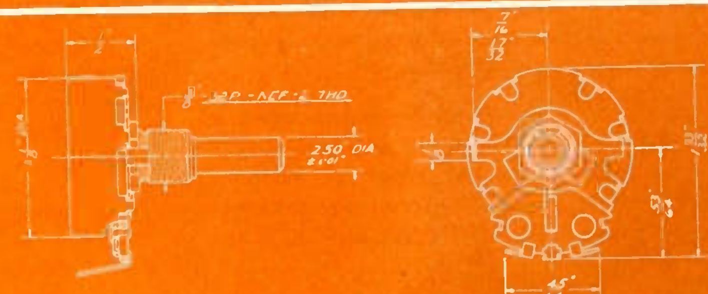
TYPE 90



1 watt 70°C, 15/16" diameter variable composition resistor. Attached Switch can be supplied.

TYPE 95 (JAN-R-94, Type RV4)





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

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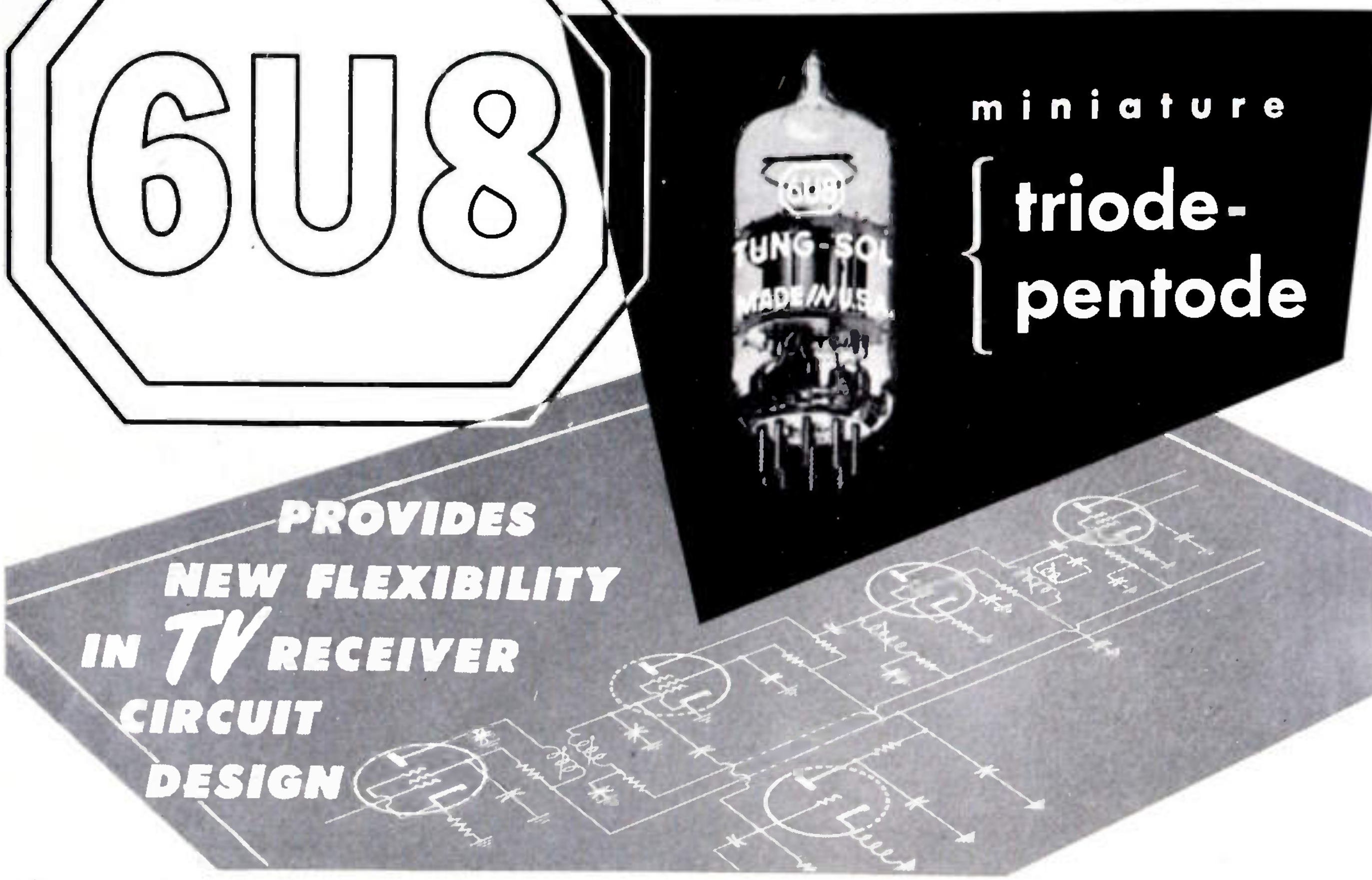
6U8



✓ Completely independent sections
✓ Versatility in circuit application
✓ Improved circuit performance

This tube has two electrically independent sections—a triode and a pentode and is intended as a local oscillator mixer for FM and TV receivers. Each section is adequately shielded, and both are capable of exceptionally good performance at the higher frequencies.

Because the two sections are completely independent, a high degree of flexibility of circuit design is available—especially valuable in TV tuner oscillator use. Performance of the 6U8 triode at low voltages is superior to that of many types previously used for this service. It has sufficient reserve emission to operate efficiently under widely varying supply voltage conditions.

The pentode provides excellent gain with low local oscillator voltage injection resulting in low oscillator radiation from TV receivers. Use of the pentode section as the mixer permits the high (40 m. c.) I. F. so desirable to reduce interference and increase stability.

The construction and characteristics of the 6U8 provide designers with extremely desirable flexibility in combining circuit functions. The pentode section of the tube may be used as an I. F. amplifier, video amplifier, sound limiter or synchronizing separator. The triode performs satisfactorily as a horizontal or vertical oscillator, or sync clipper.

Wherever there is need for a triode and a pentode in a receiver, they can be combined in the 6U8.

The TUNG-SOL engineering which has produced the 6U8 is constantly at work on a multitude of special electron tube developments for industry. Many exceptionally efficient general and special purpose tubes have resulted. Information about these and other types is available on request to TUNG-SOL Commercial Engineering Department.

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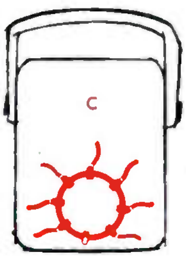

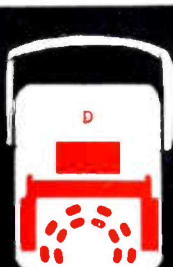

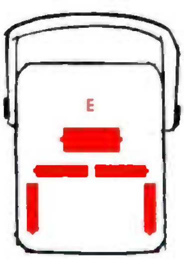

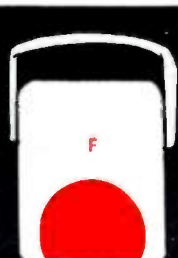

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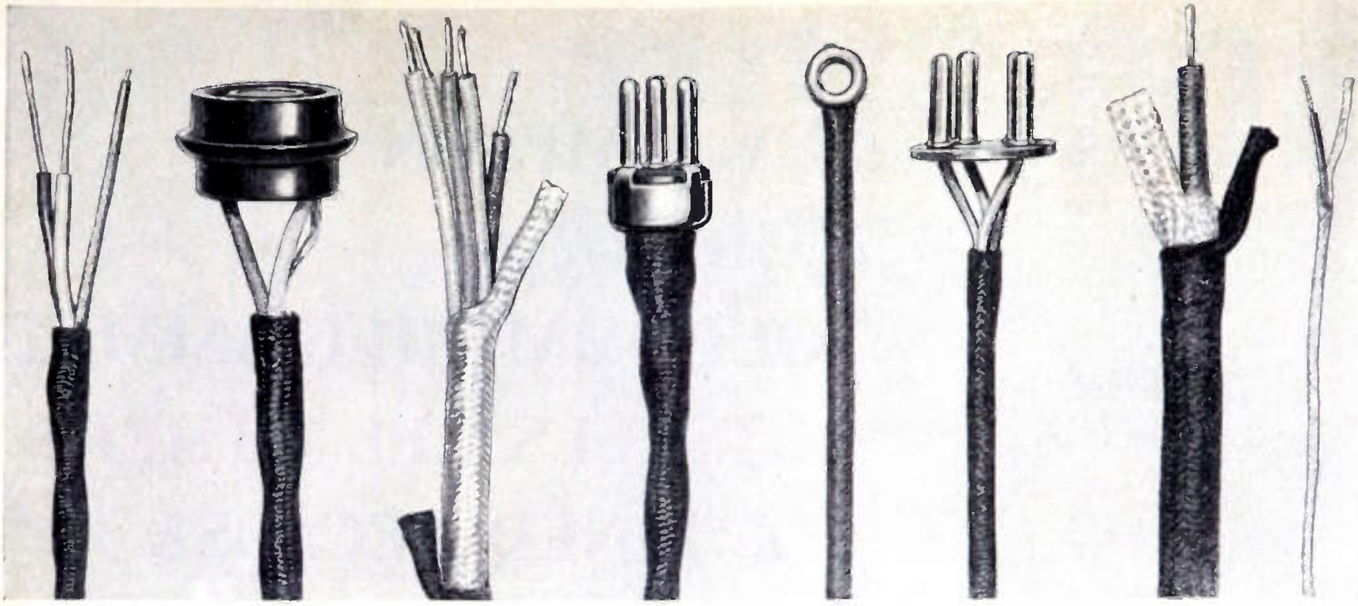

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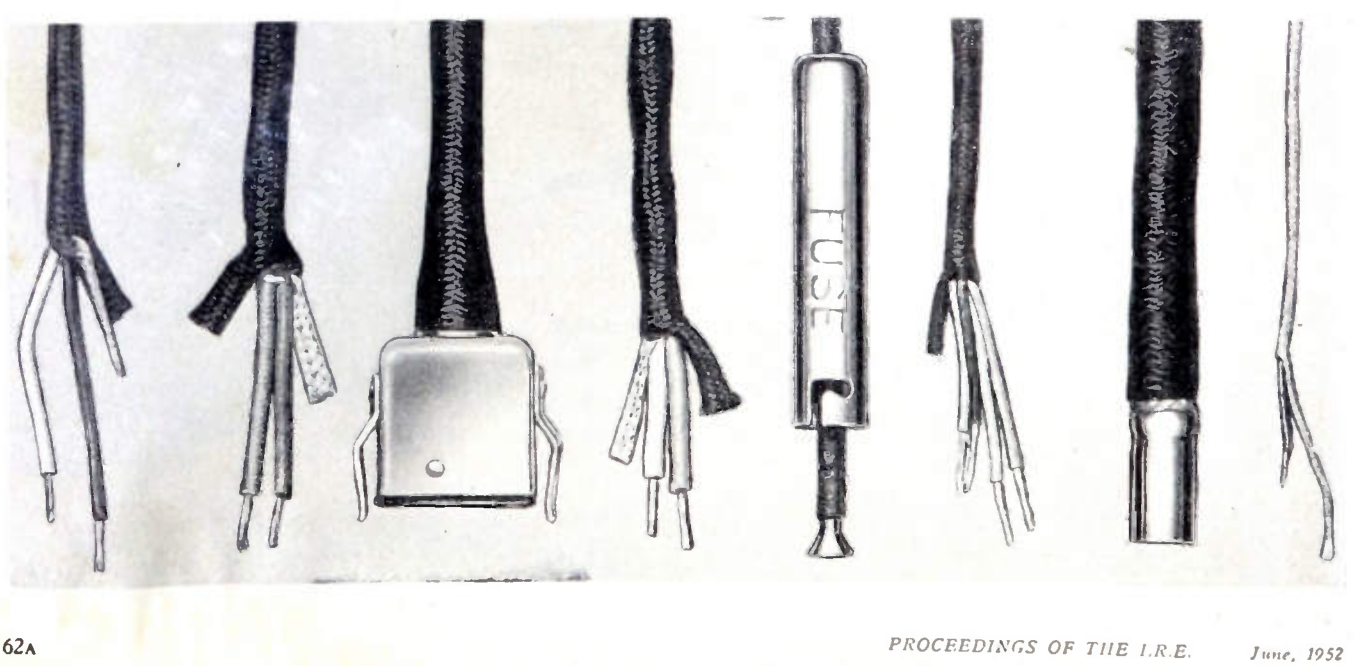

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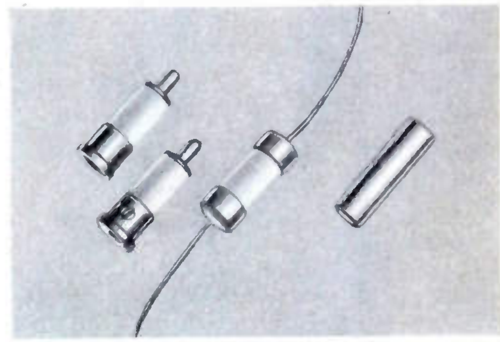

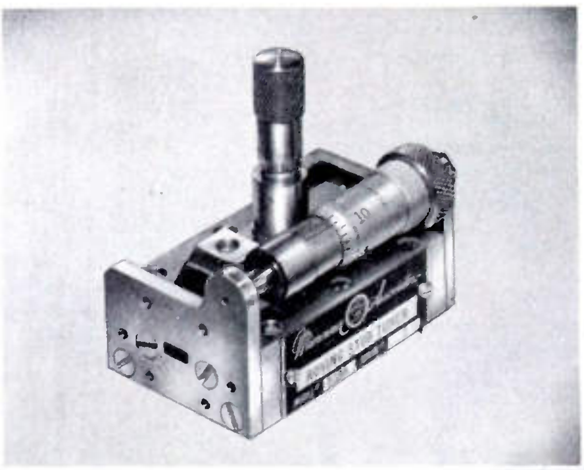

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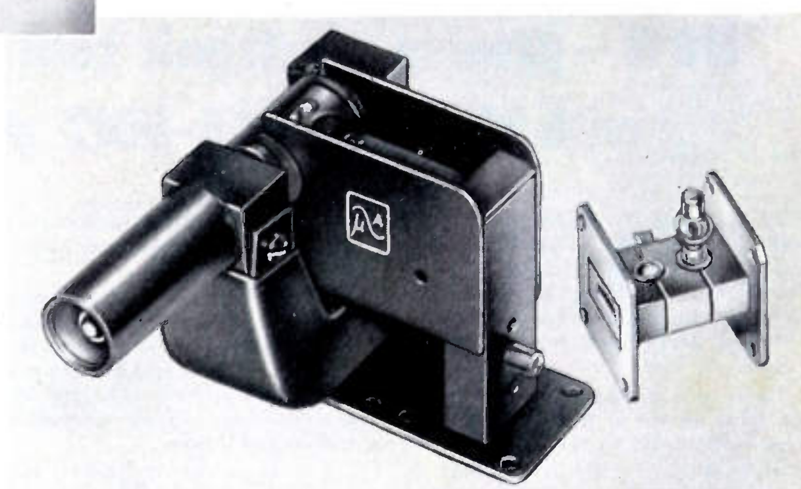

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GRID-NO. 1 (CONTROL-GRID) VOLTAGE:	
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Fault, for duration of 0.1 sec. max.	20 max. amp
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Average*	0.05 max. amp
GRID-NO. 1 CURRENT:	
Average*	0.05 max. amp
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For complete technical data on the RCA-6012, write RCA, Commercial Engineering, Section FR47, Harrison, N. J. . . . or contact your nearest RCA field office.

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GEORGE D. O'NEILL

ELECTRONIC COMPUTERS GROUP

Morton M. Astrahan, Chairman of the IRE Professional Group on Electronic Computers, was born on December 5, 1924, in Chicago, Ill.

After attending Purdue University for a year and serving in the United States Army during World War II, Dr. Astrahan attended Northwestern University. He received the B.S.E.E. degree in 1945, and was awarded a Fortesque Fellowship. He attended the California Institute of Technology, receiving the M.S.E.E. degree in 1946. He received the Ph.D. degree in electrical engineering in 1949, at Northwestern University, the subject for his thesis being Hollow Tube Dielectric Waveguides.

Dr. Astrahan joined the International Business Machines Corporation in 1949, at the Endicott Laboratory. At present, he is working on engineering planning of digital computers for IBM, in Poughkeepsie, N. Y.

Dr. Astrahan joined the Institute as a Student in 1945, transferred to Associate Member in 1950, and became a Member in 1951.

He is a member of the American Institute of Electrical Engineers, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

ELECTRON DEVICES GROUP

George D. O'Neill, Chairman of the IRE Professional Group on Electron Devices, was born in Montclair, N. J., and received the B.S. degree from the University of Michigan.

Mr. O'Neill began his professional career with the Westinghouse Lamp Company, and in 1928, he joined the Hygrade Lamp Company, in Salem, Mass., the predecessor of Sylvania Electric Products Incorporated. Early in World War II, he worked on special tubes for radar and the proximity fuse for Sylvania, in Emporium, Pa., and in 1943, he was transferred to their Central Engineering Laboratories at Long Island. He organized the solid-state section of the physics laboratories at Sylvania Center, Bayside, N. Y., and was recently made manager of the Company's engineering personnel development and education program.

Mr. O'Neill is Vice Chairman of the IRE Technical Committee on Electron Devices and a member of various other committees. He became an Associate Member of the Institute in 1928, a Senior Member in 1946, and received the IRE Fellow Award in 1949.

He is a member of the American Physical Society and the American Society for Engineering Education.

Intercommunication Among Engineers*

DONALD B. SINCLAIR

"The value of the mathematical treatment of physical phenomena is obvious. Mathematical deductions and inductions are indeed powerful instruments of analysis, and are sometimes astonishingly clear and general expressions descriptive of part of the physical universe.

"Their compressed 'shorthand' character, and the occasional difficulty of their comprehension or physical interpretation by well-qualified engineers must, however, be acknowledged. In the following address, the President of the Institute significantly emphasizes that there are various modes of communicating engineering facts, and that none of them should be neglected nor over-emphasized to the point of exclusivity. Only thus can communications engineers communicate most clearly among themselves.—*The Editor.*"

The argument is often heard that engineers cannot communicate successfully in any language but mathematics, and unfortunately, there is a good deal of truth in this popular belief. Scientists and engineers have found mathematics a universal tongue. It is concise, unambiguous, and accurate. A single equation, $e=mc^2$, has changed the course of history. But mathematics is a shorthand. Although it compresses and makes quantitative the ideas that form the stream of expanding human knowledge, the essence of these ideas can often best be presented in nonmathematical terms.

In particular, when a problem or concept is put into words, the mind can grapple with its general meaning, undeflected by preoccupation with the mechanics of detailed analysis. Equations, properly used, can facilitate communication between specialists in the same field, but they can often inhibit profound thinking, and they can certainly prevent communication with those not versed in the art.

This point seems to me most important. Not only are fields of specialization becoming more numerous, but they are becoming more refined and more difficult to understand by those who are not in direct contact with them. There is, therefore, a growing problem of intercommunication within the engineering fraternity itself. And there is a still greater one in communication with nontechnical people.

This latter problem is probably not much worse than it has always been, since professional men have been consistently accused of employing an esoteric jargon for the express purpose of shutting the general public out of their discussions. It is becoming a more serious one, however, as the engineer becomes more of a factor in society. And, make no mistake, he is indeed becoming more of a factor. The new weapons of World War II, and the developments of materials and gadgets since, have made a profound impression upon the public mind. People are interested in what we are doing, and are convinced that we can accomplish anything. The engineer's status has never been so high. The word "research" has taken on connotations of infallibility that must be dispelled lest failure to produce the expected miracles causes a revulsion that carries opinion to the other extreme.

It is vital that the limitations of research and the so-called scientific method be generally understood, and engineers should be able to do much to bring this about. To do so requires, first, that we achieve an understanding of them ourselves and, second, that we be able to communicate this understanding to others.

So far as our own understanding is concerned it is important that our thinking not be confined to special fields. Like other specialists, engineers tend to congregate in like-minded groups where common problems and ideas can be readily grasped and discussed. In familiar territory the engineer is judicial and scientific. He analyzes problems, collects and evaluates pertinent data, and decides objectively upon the proper course of action. Outside this territory, unfortunately, he all too often forgets his scientific training entirely and attempts to solve problems as intuitively as if he had never heard of the scientific method. An important underlying cause is doubtless unfamiliarity with the field in which the problem arises. To understand the other fellow's field, it is frequently essential to know the other fellow. Specialization to the extent of shutting off contact with people in other fields narrows the individual's horizons and may even inhibit his appreciation and understanding of his own specialty.

The problem of communicating with others should be our meat. We are communications people, and we should know all about it. But do we? So far as electrical devices are concerned we certainly do. We know all about telegraphs, telephones, radios, and television. We know all about antennas, vacuum tubes, and microphones. And we can analyze them in mathematics of the utmost refinement and complexity. But sometimes we have a little trouble with language.

That this is lamentable we all agree. Engineering schools worry about this problem. Professional societies are concerned about it. The Institute has established the Editor's Award specifically to encourage lucidity and the use of correct English in engineering writing. Employers are constantly alert for the engineer who expresses himself easily and well.

It is interesting to speculate as to why this problem should be so well recognized and yet remain basically unsolved. It may be that a new educational approach specifically aimed at the engineer is needed, and much thought has been and is being given to it in educational circles. Engineers, as a whole, are doers. Their job is to make useful things embodying scientific principles and discoveries. Their lives and thoughts are devoted to phenomena. English, as a means of conveying mood, emotion, impressions, and abstract matters, does not satisfy their requirements. They must set down exact descriptions of things and laws. They are in need, not so much of the color and warmth and shading of English, as of its precision and conciseness and clarity. They respect good tools and their workmanlike use. In the intellectual sphere, mathematics is recognized as such a tool. Language should be considered in the same light.

Lucid exposition is not beyond the reach of the engineer. Proper choice of words and grammatical exactness can make exposition a delight to speaker and audience alike. To speak clearly and easily to audiences of all kinds, and to be literate and understandable, challenges us all to show that communications is indeed our business.

* Based on an address presented at the President's Luncheon, 1952 IRE National Convention, New York, N. Y., March 4, 1952.

On the Dissemination of Research Information*

J. B. McCANDLESS†, MEMBER, IRE

Summary—This paper deals with the importance of obtaining and disseminating the maximum amount of technical information. The usefulness of technical briefs and periodicals is questioned and the interrelation of all fields of science is stressed. It is advocated that the inclination of selected key scientists to specialize be kept at a minimum and that their training, educationally and practically, cut across all fields of science. These scientists are then utilized to extract from a wide variety of reports and publications the maximum amount of technical data, which is then passed on to the working scientist.

RESEARCH INFORMATION as a tool for military use or application during a preparedness period, or for a general cultural scientific advance in peaceful periods, must be widely and thoroughly disseminated to be most advantageously utilized.

It has been postulated that at least part of man's superior mental ability stems from his being able to carry on intelligible communication with his fellow beings. Whether or not this is fully true might be questioned; however, the fact that man has been able to communicate with man is of such significance it is difficult to grasp a civilization of any sort without means of communication.

Since ancient times, man's curiosity, which is the forerunner of a research investigation, has compelled him to seek answers to problems posed by himself and others. If this information obtained by him is not passed on so that others may become acquainted with and curious about the problems and possible solutions, then research and its results slowly stagnates and dies out.

It is well recognized that science and research received a great impetus with the advent of the printing press and its consequent effect on people, i.e., creating a desire to be able to read, thus stimulating the thinking mechanism (sometimes, in what might be considered a wrong direction).

However, it was several generations before the manifestations of printing were felt in science. This is understandable as people did not all begin to read and write at the same time. While there is no question of the effect that printing has had on the advancement of knowledge in all fields, particularly in the sciences, it must be recognized that printing *per se* did not initiate this awakening in science; the wide dissemination of printed material was equally responsible. Thus as the population grew, transportation became less hazardous and communication relatively more rapid and widespread. These two occurrences then are the primary causes for the rapid advances in science within the last three centuries. At the same time thirst for knowledge rapidly encroached on all strata of society so peasant as well as prince craved knowledge; schools and universities came into being all over the world.

It was not until the nineteenth century however that the practical applications of this revolution in society became apparent. Then the manufacturing industries arose and completely altered society and its peoples. As these industries grew within a specialized field, it is not too surprising that their research and development investigations should be limited to this field. As a consequence the employee within a given organization has become a complete specialist in his field, and his contact with other fields is only through the technical publications he is sufficiently interested in to read; and these, generally speaking, also embrace his own field. Thus we have a picture of a vast amount of technical literature being turned out, the maximum utilization of which is certainly open to question. This is not entirely the fault of the specialist. It is also due to the unalterable fact that too many periodicals are either too specialized, in their attempt to satisfy a small professional group, or cover too broad a field. Many are so cluttered up with eye-catching advertisements that one must diligently search, as it were, between the "ads" for useful information. Technical briefs and such have served an admirable purpose in alleviating this condition somewhat and bringing to the attention of the scientist information that he would not have otherwise had. Unfortunately, however, some of these reviews are often so brief and lack such basic technical information that there is more distraction than useful information contained in them.

While the application of certain physical phenomena might appear at the time to be strictly applicable to one field or a group of allied fields, it does not follow *a priori* that this limitation is completely restricted to this field.

A pertinent point in this respect is brought out by Slater in an M.I.T. Progress Report:[1] "The study of X-ray crystal structure of alloys on the one hand, of the electrical conductivity of semiconductors on the other, may seem to the experimenters carrying them out as entirely unrelated activities. And yet the theorist, who understands that the one depends on the diffraction of X-rays by the periodicities or lack of periodicities in a crystal lattice, and that the other depends on the similar diffraction or scattering of electrons by these same lattice features, understands their essential interrelations." This interrelation between apparent diverse fields is one which has existed since the inception of modern-day physics, and men of great intellectual ability and stamina who can grasp the gist of each of several or all fields independently have indicated that the division is one of intellectual limitation and does not exist. It must also be recognized that channeling of information into a particular field is to some extent due to immediate necessity of utilization and application of this information militarily or commercially.

In a government laboratory, where a great deal of the work performed is by contractors whose results are in the form of reports which must be analyzed, one can readily perceive that dissemination of this material along with information from current technical books and periodicals becomes a problem of increasing importance and dimension.

The problem then is that of obtaining the maximum pertinent information available not only within a certain field but in all other fields, some of which may not seem to have any interrelation at all.

The solution is not simple; it calls for the education and training of a different type of research scientist. The scope of duties of such an individual may at first glance appear insurmountable. For fundamentally, what is required is an all-embracing mind—one who knows everything about everything. This of course cannot currently (and can probably never) be achieved. Still a great step can be made by keeping to a minimum the inclination of an individual to specialize. Some of the larger manufacturers and government laboratories do put their young scientists through a period of training where they spend several months within one department and then pass on to another. In some instances, however, the time spent within any one department is relatively short and all the departments are still usually within the over-all field of endeavor. Moreover, at the end of this short training period, the individual is often placed more or less permanently in one or another department, and then proceeds to forget all but the most allied fields, thus becoming a specialist. Psychologically speaking, the conjecture might be made that the specialist remains a specialist and tends to greater specialization because a new and strange field might well bring up doubts of his ability to cope completely with new problems; and having achieved some success and/or respect within his field, he hesitates to expose himself as being inexperienced in another field for fear of subjecting himself to unjust ridicule. Such actions may be interpreted as lack of interest in and general disdain for any field other than his own.

A much more positive approach is to have the individual whose basic education has not been influenced by specialization spend several years in intensive training and research investigations in possibly three or four basic fields of endeavor, preferably nonallied. This individual is particularly earmarked for a combination scientific-administrative position, which immediately establishes him as an exceptional person in an era of specialization.

His general knowledge must embrace several basic fields of endeavor so that he may recognize and analyze fundamental and important information from a variety of sources. He must be able to present this information to the working scientist and discuss its possible utilization. Furthermore, he must determine whether or not the incorporation of such information would be advantageous from the over-all management viewpoint. He should, therefore, have full

* Decimal classification: R010. Original manuscript received by the Institute, January 4, 1952.
† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

[1] J. C. Slater, M.I.T. Quarterly Progress Report, No. 1, Project NSori-07856; July 15, 1951.

knowledge of managements' plans and formulations. His position in the organization is as distinct and individual as the scientific worker himself, and like this worker, his usefulness is decreased if he is burdened with administrative details or used as a crutch to assist in general office procedure.

He becomes the funnel through which pours the vast amount of printed technical material to be analyzed and disseminated, thus leaving the scientific worker free and content in the knowledge that if there is anything new or useful in his field or another field the information will come to him promptly and concisely.

The principal advantages in utilizing such an individual may be broadly summed up as follows: First, management is assured that the most recent and novel knowledge and techniques throughout the country are not only available to them but, more important, are also being used by them; and secondly, that a minimum amount of duplication of effort exists both within their own organization and throughout the country as a whole, thus incidentally alleviating the current shortage of technical personnel.

The Plasmatron, A Continuously Controllable Gas-Discharge Developmental Tube*

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Summary—**The "plasmatron," a new type of continuously controllable gas tube, is described and its operation analyzed. This tube utilizes an independently generated gas-discharge plasma as a conductor between a hot cathode and an anode. Continuous modulation of the anode current can be effected either by variation of the conductivity or the effective cross section of the plasma. The first of these is accomplished by the modulation of the electron ionizing beam which controls the plasma density and hence its conductivity. The second method makes use of the gating action of positive-ion sheaths which surround the wires of a grid located between the anode and cathode. The plasmatron appears to have considerable promise for such applications as motor drive, direct loudspeaker drive, high-efficiency rectification and inversion, and the many other uses which require the high-current and low-voltage operation that the high-impedance vacuum tube cannot supply.**

I. Introduction

As a consequence of its space-charge-limited current the vacuum tube is a relatively high-impedance device. It is a most useful device because the current through it can be continuously controlled. On the other hand, a gas tube such as the thyratron, which normally operates with a neutralized space charge, is a low-impedance device. Unfortunately, it seems that it has been necessary to pay a large price for this low-impedance operation. The enormous advantage of continuous current control has been lost.

However, the situation is far from hopeless. In fact, if one approaches the operation of hot-cathode gas tubes in terms of the fundamental processes, it is entirely possible and practicable to have both low-impedance operation and continuous control in one tube. The plasmatron is such a tube. It can deliver, and continuously control, large currents at anode potentials of a few volts. This tube, now in the developmental stage, shows excellent promise of fulfilling the long-standing need for a tube that will continuously control large currents at low voltages. In this category are found such applications as motor control, direct loudspeaker drive, inverters, and the like.

The name "plasmatron" stems from the word *plasma*[1] which denotes an important part of a gas discharge. The plasma is a region of very high, but essentially equal, concentrations of free electrons and ions. Due to the high mobility of the plasma electrons and the absence of net space charge, the plasma is a rather good electrical conductor. It has a resistivity of the order of one ohm centimeter which places it in a class with semiconductors such as germanium. As a consequence of this high conductivity, electric fields within the plasma are small.

The principle of operation of the plasmatron lies in the use of a plasma as a conductor between a hot cathode and an anode. This plasma, however, is generated by an auxiliary discharge (which operates independently of the work circuit) and *not* by the anode current as is the case with a thyratron. It will be seen that this seemingly small difference is of fundamental importance.

There are two broadly different methods of effecting continuous control over the anode current. The first makes use of the fact that the conductivity of the plasma, and hence the anode current, depends upon the plasma density and this, in turn, upon the intensity of the auxiliary discharge which generates the plasma. The second method involves the use of a grid structure that is interposed between the cathode and the anode. This grid, by virtue of the positive-ion sheaths that surround it, gives a gating action that acts to control the effective cross-section area of the plasma conductor and hence the current-carrying capacity of the tube.

Briefly, the reason that the plasmatron grid can effect continuous control over the anode current, whereas the thyratron grid commonly cannot, lies in the fact that the ion generation in the plasmatron is achieved by an independent discharge and not by the anode current itself. It is also pertinent to point out that in a thyratron

* Decimal classification: R337.1. Original manuscript received by the Institute, November 13, 1952.

† RCA Laboratories Division, Princeton, N. J.

[1] L. B. Loeb, "Fundamental Processes of Electrical Discharges in Gases," John Wiley and Sons, New York, N. Y.; 1939.
J. D. Cobine, "Gaseous Conductors," McGraw-Hill Book Co., Inc., New York, N. Y.; 1941.
Rompe and Steenbeck, "The plasma state of gases," *Ergeb. exakt. Naturw*, vol. 18, pp. 257–376; 1939. Translated from the German by G. C. Akerlof, Mellen Inst. of Industrial Research.

the applied anode potential really serves a double function: (1) to provide electrons with enough energy to generate ions, and (2) to provide the electric field needed for pulling the required anode current through the tube. When these functions are separated, as they are in the plasmatron, it is found that the potential needed for drawing a given anode current is greatly reduced and, in addition, continuous grid control becomes possible, provided that the anode potential is insufficient to cause ionization.

It seems expedient that the first part of the paper be concerned with the building-up of a general picture of the over-all operation and characteristics of the tube. This picture will provide a convenient point of departure for treating the operation in more detail. With this approach in mind, the first mode of modulation or control will now be considered.

II. Diode Operation[2]

Fig. 1 shows a diagram of an experimental form of plasmatron and accompanying circuit operating in the diode mode. The plasmatron is seen to consist of a main anode in the shape of an inverted U; two cathodes, one referred to as the main and the other the auxiliary cathode; and a cylinder with a narrow aperture, this cylinder being designated as the garrote. Fig. 2 is a photograph

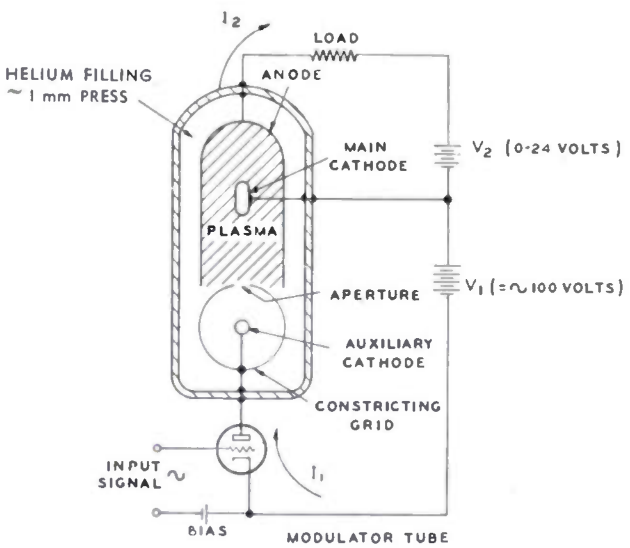


Fig. 1—Circuit arrangement for diode operation.

of a tube of the form sketched in Fig. 1. While it has been found convenient to work with experimental tubes of the form shown in Fig. 2, the principle of operation puts no definite restrictions upon the size or geometry of a tube provided the basic features are retained. The circuit is seen to consist of two loops referred to as the main loop and auxiliary loop, respectively. Most experimental tubes were filled with helium to a pressure of one millimeter of mercury.

In operation V_1 is made sufficiently large so that a discharge occurs in the auxiliary loop, current flowing between the auxiliary cathode and the upper portion of the structure. This discharge generates a plasma between the main cathode and anode, enabling large currents to flow between them for small values of V_2. In fact, as will be seen, in order to insure plasmatron operation V_2 must be held below a value which would initiate a discharge in the main loop. The garrote (generally tied to the auxiliary cathode) serves to achieve the generation of a dense plasma for small discharge currents. It increases the ionization efficiency to such an extent that several ma in the auxiliary circuit enables hundreds of ma to flow in the main or work circuit.

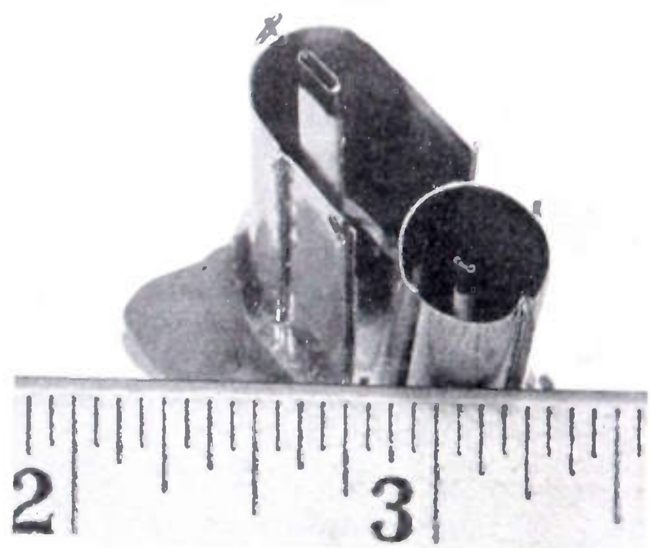


Fig. 2—Photograph of typical plasmatron diode.

In the circuit, as shown, the auxiliary discharge current can be varied by means of changes in the grid potential of the modulator tube, which in this case is a small vacuum tube such as the 6J5.

A typical volt-ampere characteristic, with the auxiliary discharge current as parameter, is shown in Fig. 3. It is seen that the anode current starts to flow when the anode is about minus one v, rises rapidly, and finally saturates at a current of over half an ampere (for $I_1=8$ ma) when the anode potential is only about 3 v. The current remains saturated along this curve until the anode potential exceeds about 25 v, the ionization potential of helium, and then rises steeply without regard to the value of the auxiliary current. In practice, the tube is always operated below this breaking point. In fact, for normal operation a load line should be fitted in between the saturation voltage and the ionization voltage. The back current for negative anode potentials is of the order of a few ma and consists of positive ion current from the plasma. For negative anode potentials greater than about 200 v, in the tube shown, a cold discharge will take place between the anode surface and cathode.

[2] E. O. Johnson, "Controllable gas diode," *Electronics*, vol. 24, pp. 107–109; May, 1951.

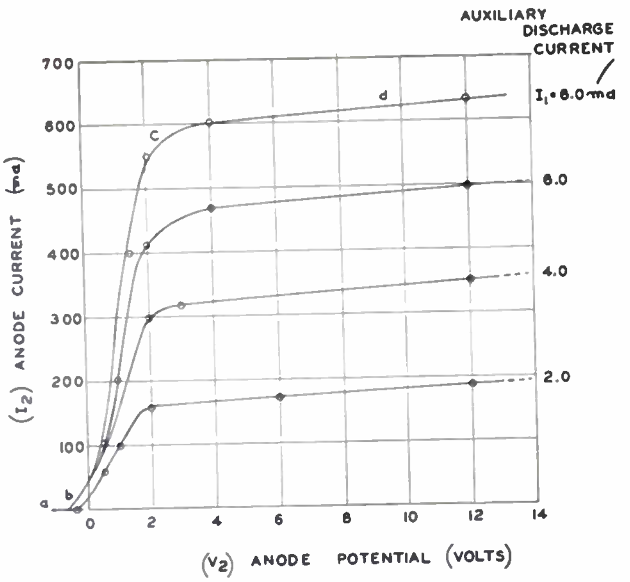


Fig. 3—Diode volt-ampere characteristic.

Fig. 4 shows the relationship between the auxiliary current and the saturated main current. The relation is seen to be a linear one until the main current exceeds about one ampere. The departure from linearity beyond here arises from the onset of emission saturation of the

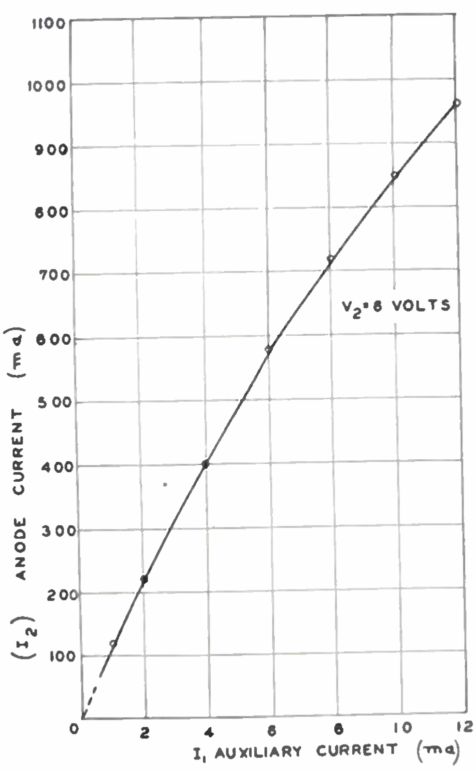


Fig. 4—Diode control characteristic.

main cathode. When the cathode is completely saturated, the anode current becomes equal to the value of this current since the plasmatron, like any other electron tube, is limited in peak anode current to the saturated cathode emission. Whereas the current gain is only about 100:1, as exemplified by the slope at low currents, improved tubes give gains of 300 or better. The power gain for this tube is about 17 db.

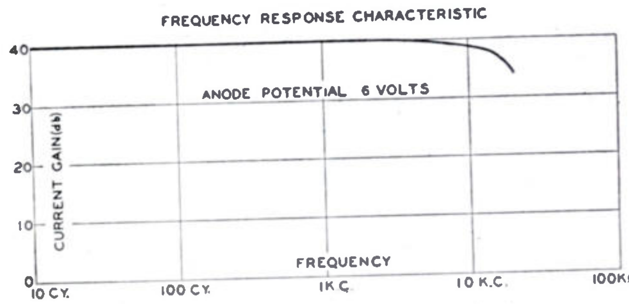


Fig. 5—Diode frequency characteristic.

The frequency response is shown in Fig. 5. The current gain is seen to be flat from dc out to about 10 kc and then fall off. This falling off with frequency arises from the length of time required for surplus ionization to diffuse away to the end micas where surface recombination of the free electrons and positive ions takes place. For helium at these gas pressures and for these tube sizes the volume recombination of the particles in the plasma is relatively small. As shown later, the unique potential distribution between cathode and anode discourages positive ions from reaching these surfaces.

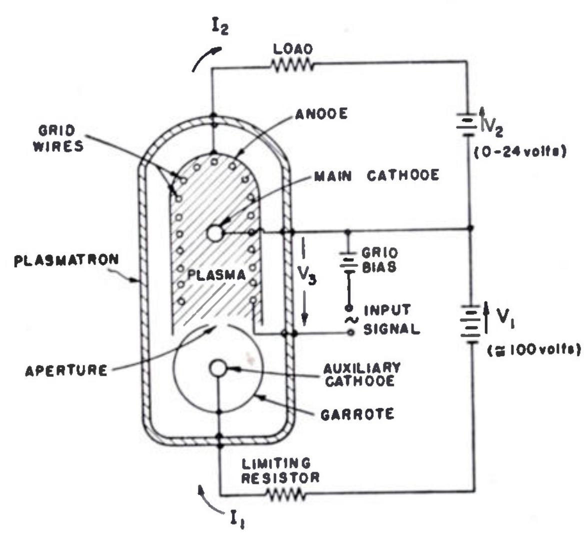


Fig. 6—Circuit arrangement for triode operation.

III. Triode Operation

The arrangement for the second mode of control employing a grid located between the cathode and anode is shown in Fig. 6. The circuit is seen to be almost identical with the diode circuit except that the modulator tube has been replaced with a limiting resistor and, in addition, a grid circuit has been added. A photograph of

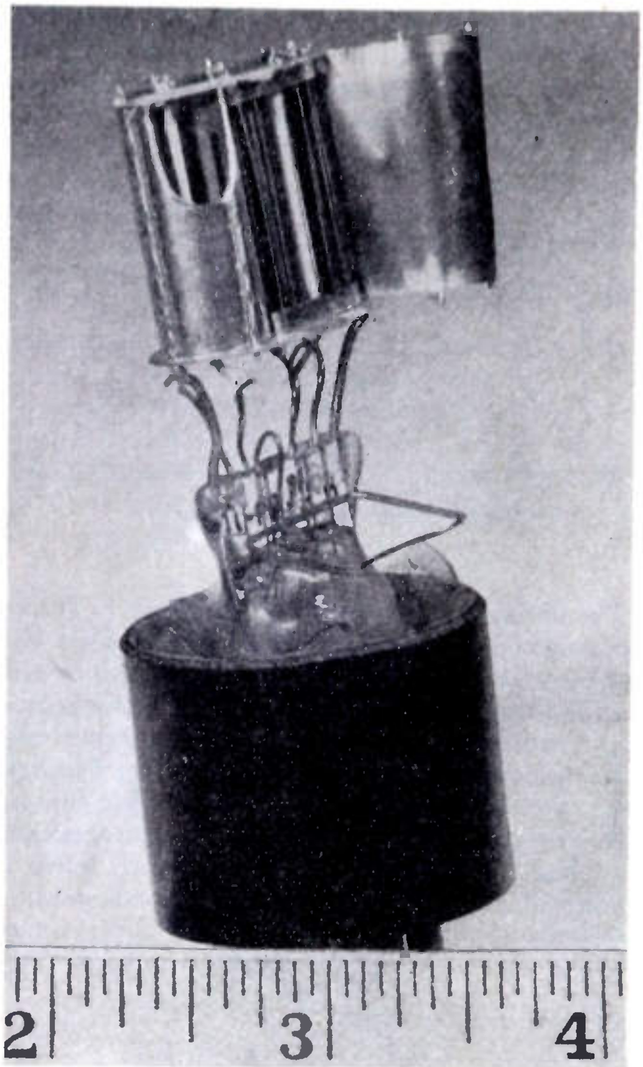


Fig. 7—Photograph of typical plasmatron triode.

a typical experimental triode plasmatron, with its anode folded back to allow a view of the grid structure, is shown in Fig. 7.

The mechanism of grid control is depicted in Fig. 8 where a section of the anode and two of the grid wires are shown. The cathode is out in the plasma towards the bottom of the figure and need not concern us. As was pointed out previously, the plasma has a high conductivity so that the electric fields within it are always small. External fields are absorbed at the plasma boundaries in a thin layer called a sheath.[1] This is illustrated in the potential diagram which is a projection of the cross section above. The sheath, which is the transition between the plasma and the grid wire, is nothing more than a region containing positive ions en route to the grid wire. It is the plasma's way of meeting its boundary conditions. One can picture the edge of the plasma as being a positive-ion emitter and the grid wire as being the collector. The thickness of the sheath is determined by the positive-ion current and the grid voltage according to the familiar 3/2 power law.[1] Since the ion current is a constant, being set by the rate at which ions diffuse out of the plasma, the sheath thickness will increase with the negative grid voltage until adjacent sheaths overlap. The negative field in the sheaths will repel the electrons which would otherwise go to the anode, and the anode current is then cut off. Intermediate anode currents result from sheaths of intermediate radii.

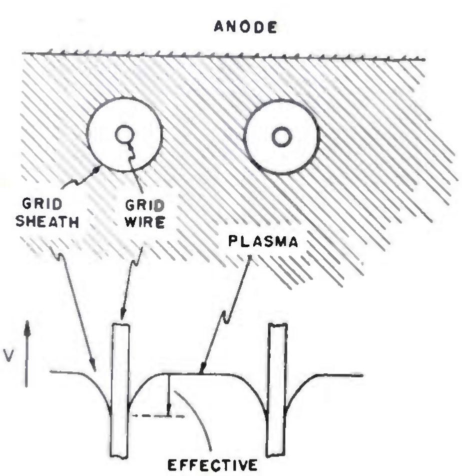


Fig. 8—Diagram illustrating grid action.

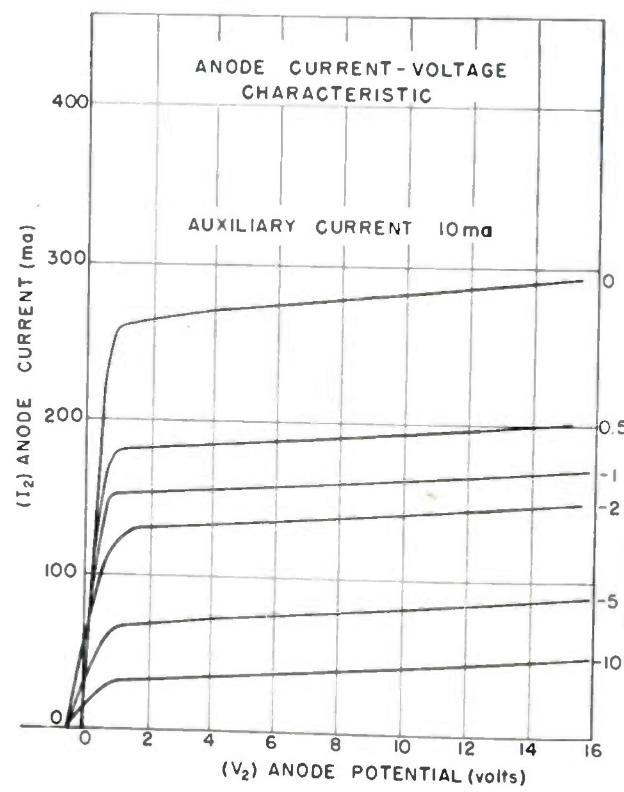


Fig. 9—Triode volt-ampere characteristic.

The volt-ampere characteristic with the grid bias as parameter is shown in Fig. 9. The auxiliary current is held constant. It is seen that this family of curves is similar to those in Fig. 3. As in the diode the load line should be placed between the saturation and the ionization voltages. However, if the grid cuts off the anode current before the anode potential exceeds the ionization potential, then the anode potential can be permitted to exceed the ionization potential without re-establishing the flow of anode current. But, just as with a thyratron, if anode current flow starts at elevated anode potentials, the grid loses control.

Fig. 10, which is a replot of the data shown in Fig. 9 for a constant anode voltage, shows that the grid characteristic is a nonlinear one. Whereas the transconductance exceeds 100,000 micromhos at low grid voltages, it drops below 10,000 for biases greater than about 6 v.

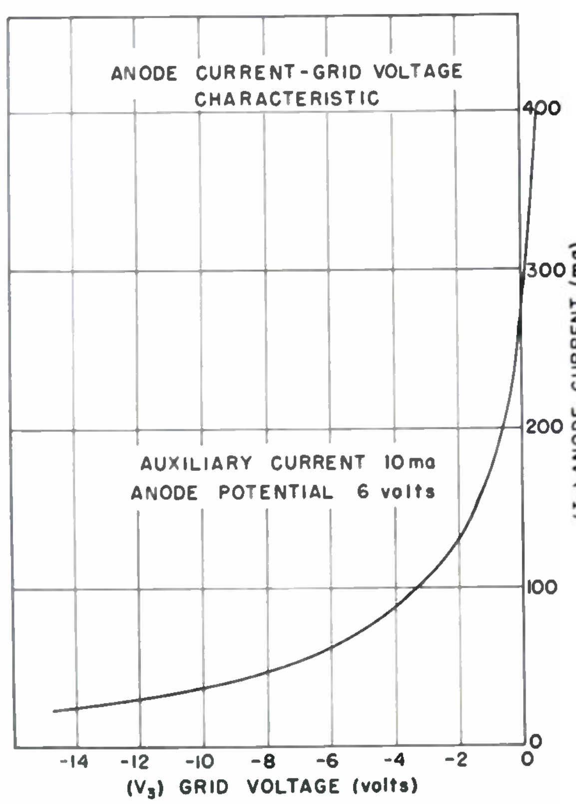


Fig. 10—Triode control characteristic.

The current-voltage characteristic of the grid is shown in Fig. 11. For negative grid potentials greater than about 1.5 v the grid current is predominately ionic, while for more positive potentials the electrons predominate. The very rapid rise in the electron current which occurs at low grid voltages is a consequence of the grid's beginning to act like an anode. In fact, if the curves were carried further, a characteristic like that of the anode would be obtained. The slope of the characteristic is seen to vary between about 300 ohms at 1-v bias to about 10,000 ohms at 10-v bias. The power gain that can be obtained with triode operation is about the same as that with diode operation.

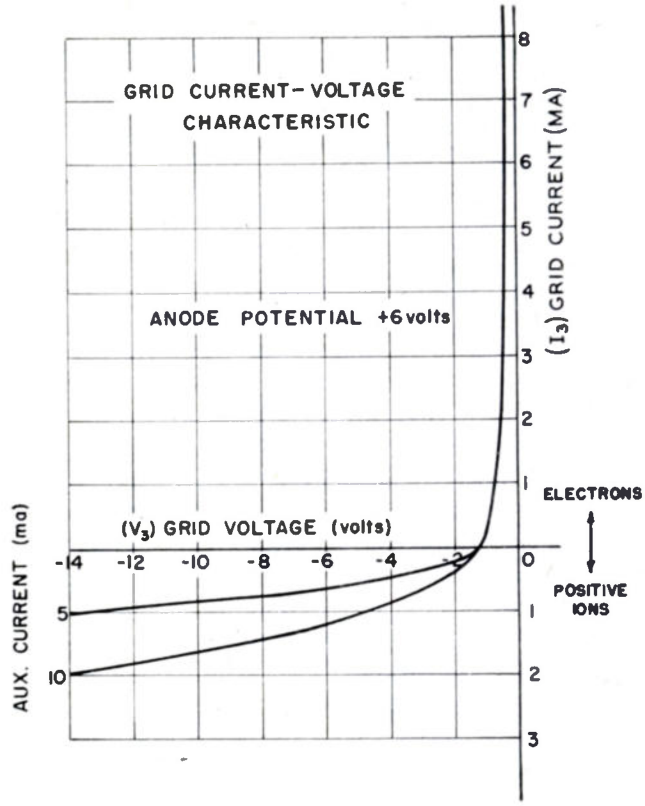


Fig. 11—Grid volt-ampere characteristic.

A typical frequency response of a grid-controlled tube is displayed in Fig. 12. Since grid modulation affects the plasma only in a small region adjacent to the grid, whereas the diode modulation requires bulk changes in the plasma, it is not surprising that the triode frequency response is a markedly better one. The wiggles in the triode response result from a complex combination of ion and electron movements which accompany the propagation and retraction of the grid sheath. The response was not studied beyond 10 mc other than observing that a tube could operate as a class C oscillator at frequencies as high as 17 mc.

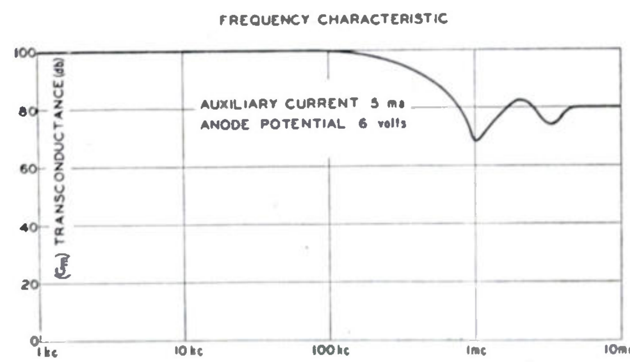


Fig. 12—Triode frequency characteristic.

IV. Plasma Characteristics[1]

The plasma in a gas discharge can often be identified with the region that shows a visible glow. Whereas the glow itself arises from excitation of the gas atoms, the excitation in many gas-discharge devices is accompanied by the dense ionization that characterizes a plasma. This is the case in the main section of the plasmatron. The density of the free electrons and positive ions which constitute the plasma in experimental or commercial gas-discharge tubes ranges from 10^9 to about 10^{11} particles per cm.[3] Since there are essentially equal densities of the oppositely charged particles, the plasma is electrically neutral. In a typical plasma the charged particles usually execute random motions conforming to a Maxwellian distribution of velocities. It is convenient to describe these motions in terms of a temperature. In the cases to be described it will be assumed that the positive-ion temperatures are about the same as the ambient-gas temperature. This seems reasonable because of the high collision frequency between the ions and neutral atoms at the pressures in the tubes to be described. The electrons, in most cases, have a much higher temperature.

The high density of the electrons along with their high mobility in a region largely devoid of space charge makes possible the conduction of large electron currents for very low applied potentials. The conductivity of a uniformly dense plasma which is electrically neutral is given by

$$\sigma = Ne\mu \text{ mhos per centimeter,} \tag{1}$$

where N is the electron density per cubic centimeter, e is the electron charge in coulombs, and μ is the electron mobility in centimeters per second per volt per centimeter. For a typical plasma $N = 5\times10^{10}$ and

$$\mu = 5 \times 10^6 \frac{\text{cm}}{\text{sec}} \Big/ \frac{\text{volt}}{\text{cm}}.$$

Thus σ is 4.0×10^{-2} mhos per cm which is comparable with the conductivity of a semiconductor like germanium.

From kinetic theory, it is found that the electron space-current density is given by

$$i_e = \tfrac{1}{4}Ne\bar{c}, \tag{2}$$

where $\bar{c}$, which is the average thermal velocity of the electrons, is given by

$$\bar{c} = 1.87 \times 10^{-8}\sqrt{\frac{T_e}{m}}. \tag{2'}$$

T_e is the electron temperature and m the electron mass. Choosing as typical values $N = 5\times10^{10}/\text{cm}^3$, $\bar{c} = 2\times10^7$ cm/sec, and $T_e = 1{,}000°$ k., it turns out that $j_e = 40$ ma/cm^2. A similar computation for the positive-ion space-current density yields (for helium) $j_p = 0.27$ ma/cm^2.

The transition between a plasma and a bounding electrode or wall occurs in a thin region of space charge which is termed a sheath. It is in the sheath that the fields from electrodes are dissipated, leaving the plasma essentially unaffected by electrode or wall potentials. If the electrode is negative with respect to the plasma, electrons are repelled from the sheath leaving only a current of positive ions flowing to the electrode. The thickness of this layer of moving positive ions adjusts itself so that the electric flux from the electrode ends on the moving charges. Accordingly, the sheath thickness d in centimeters can be related to the potential V in volts, and to the positive-ion current density j_p in amperes per square centimeter by the familiar Child-Langmuir space-charge law (for plane structures):

$$j_p = \frac{2.33 \times 10^{-6}}{\sqrt{\frac{M}{m}}} \frac{V^{3/2}}{d^2}, \tag{3}$$

where M is the ion mass and m the electron mass. If the electrode is positive with respect to the plasma, the sheath will contain a space charge of electrons. The current density j is fixed by conditions within the plasma, and is given by (2) written for the appropriate particle. The density value used in this expression must apply close to the edge of the plasma. The current value obtained by (2) is only approximately correct since no account is taken of the drift due to the weak electric fields within the plasma and at its edge. However, for purposes of simplicity this expression will be assumed to be correct.

It is now possible to consider that the behavior of a cold electrode immersed in a plasma as its potential with respect to the plasma is varied. A typical current-voltage characteristic is shown in Fig. 13. Along portion

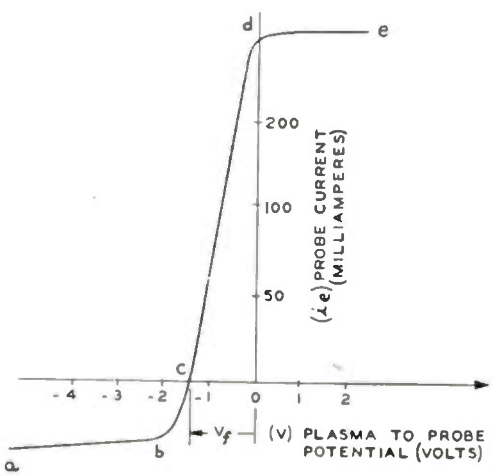


Fig. 13—Plasma cold-electrode characteristic.

ab of the characteristic all the electrons are repelled and only positive ions flow from the plasma to the electrode. The sheath thickness is given by (3) in which j_p is derived from (2). As the potential of the electrode becomes increasingly negative, the current increases slightly due to sheath expansion. Beyond point b on the way to d the electron current to the electrode increases rapidly according to the Boltzmann relation

$$i_e = A_p j_e e^{\phi V}$$

$$\phi = \frac{e}{kT_e}, \tag{4}$$

where $j_e = Ne\bar{c}/4$ is the random electron space current near the edge of the plasma, A_p is the probe area, and k is the Boltzmann constant. The electron temperature T_e can be determined from the semilog plot of i_e against V for the region bd. When point d is reached, the voltage V is zero and the electrode is at plasma potential where it collects the full electron space current. For convenience the magnitude of the positive-ion current with respect to the electron current has been greatly exaggerated. Actually, the saturated ion current, as pointed out in conjunction with (2), is smaller than the electron space current by a factor of more than a hundred. Beyond point d, as we proceed to point e, the electron current increases slowly as the electron sheath builds up, according to (3).

At point c the electron and ion currents to the electrode are equal and, since this is the point at which the electrode would "float," the corresponding plasma to electrode potential V_f is called the floating potential.

V. Cathode-Plasma Anode System

In the preceding discussion the complete circuit was neglected for purposes of simplicity. It is apparent that in a practical system a return circuit must be provided. The simplest possible system with two cold electrodes and an independently produced plasma has been treated elsewhere.[3] Let us now suppose that one of the electrodes is a hot cathode with a total emission density j_k; the other electrode is the anode. As will be seen there are two cases to consider: (1) The cathode is a relatively weak emitter such that $j_k \ll j_{0k}$, and (2) the cathode is a strong emitter such that $j_k \gg j_{0k}$. Here, j_{0k} is the random electron current density in the vicinity of the cathode; j_{0a} is the random current density in the vicinity of the anode; V_k is the plasma-cathode potential difference; V_a is the plasma-anode potential difference; V_s accounts for the potential drop through the plasma and for the contact potential difference between the cathode and anode; V_2 is the cathode-anode potential difference. These quantities are indicated in the circuit diagram of Fig. 14 and in the potential diagrams of Fig. 15.

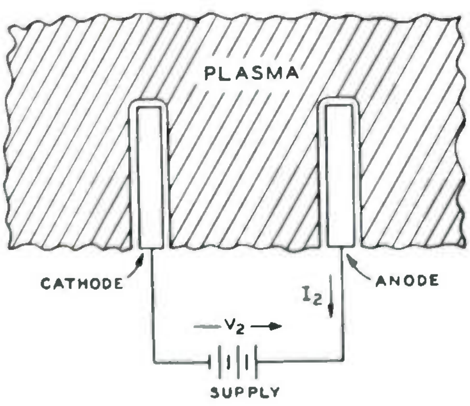


Fig. 14—Cathode-plasma-anode system.

[3] E. O. Johnson and L. Malter, "A floating double probe method for measurements in gas discharges," *Phys. Rev.*, vol. 80, pp. 58–68; October, 1950.

In the first case the potential diagram is as shown in Fig. 15(a). Neglecting the relatively small ion currents, there will be an electron current $j_k A_k$ leaving the cathode, a current $j_{0k} A_k \exp(-\phi V_k)$ entering the cathode from the plasma, and a current $j_{0a} A_a \exp(-\phi V_a)$,

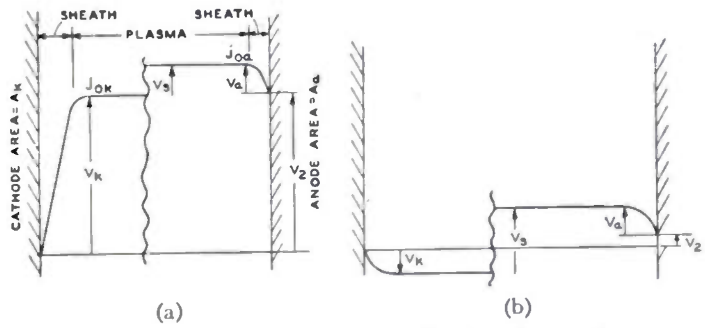


Fig. 15—Potential diagram. (a) Weak emitter case. (b) Strong emitter case.

which is equal to the electron current I_2 in the circuit, entering the anode from the plasma. ϕ is defined in (4). For equilibrium

$$j_k A_k = j_{0k} A_k \epsilon^{-\phi V_k} + j_{0a} A_a \epsilon^{-\phi V_a}.$$

From the potential diagram we see

$$V_k + V_s = V_a + V_2$$

Manipulation of the preceding equations yields

$$I_2 = \frac{A_k j_k}{1 + \left[\dfrac{A_k j_{0k}}{A_a j_{0a}} \epsilon^{\phi V_s}\right] \epsilon^{-\phi V_2}}. \tag{5}$$

While experimental studies show the essential correctness of (5), the fact that it represents a mode of operation that is not used in the plasmatron impels us to proceed at once to the second case.

With the cathode acting as a very strong emitter, a retarding field must be present at its surface to limit the net current leaving. In this case the potential distribution is as shown in Fig. 15(b). One can, as before, equate the electron current leaving the electrodes to that arriving. Thus,

$$A_k j_k \epsilon^{-\phi V_k} = A_k j_{0k} + A_a j_{0a} \epsilon^{-\phi V_a},$$

$$I_2 = A_a j_{0a} \epsilon^{-\phi V_a}, \quad \text{and} \quad V_k + V_2 + V_a = V_s.$$

The solution of these equations is

$$I_2 = \frac{A_k j_{0k}}{2}\left[\sqrt{1 + 4\left(\frac{A_a}{A_k}\,\frac{j_k}{j_{0k}}\,\frac{j_{0a}}{j_{0k}}\,\epsilon^{-\phi V_s}\right)\epsilon^{\phi V_2}} - 1\right]. \tag{6}$$

The electron sheath at the cathode is so thin compared to the usual cathode radius that the area of this sheath has been replaced with that of the cathode. Also, in this simple derivation, we have neglected any changes in the potential across the plasma as well as changes in the plasma density.[4]

[4] See Appendix.

In Fig. 16 the anode current as determined by (6) is compared with experiment. The plasmatrons used in these experiments were of about the same over-all dimensions and geometry as in Figs. 1 and 2, except that a more symmetrical (cylinder of 2-cm diameter) anode structure was employed. In addition, two 10-mil wire probes were mounted parallel with the cathode, one close to the anode and the other close to the cathode.

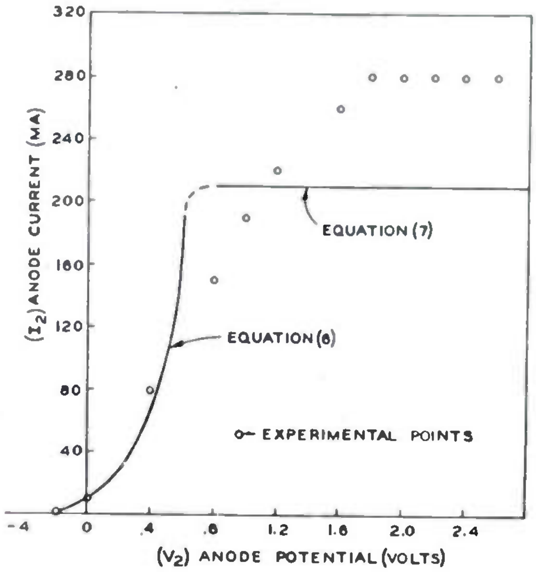


Fig. 16—Comparison of computed anode current with experiment.

The values of the space-current densities used in (6) were determined by conventional probe techniques. The electron temperature was measured by the double-probe method,[3] and was found to be close to the thermal temperature of the cathode. The potential drops across the plasma were measured by noting the change in floating potential between the two probes. This potential difference was found to be about 0.04 v, indicating a plasma resistance of about 2-ohm centimeters. This is lower than the value predicted by (1), probably because of the density gradients in the plasma.

The emission density of the 50-mil diameter oxide-coated cathode was determined by running this cathode, in conjunction with its anode, as a conventional gas diode and noting the value of anode current at which the discharge broke out of the "ball-of-fire" mode.[5] This value of anode current is approximately one-half the total field-free emission of the cathode.

If the potential V_s is given a value of 0.95 v, (6) includes the point 0.0 v, 10 ma. This value of potential seems reasonable and about what might be expected from contact potentials and a small drop in the plasma.

The curve representing (6) was terminated by the horizontal portion at the point where

$$I_2 = A_a j_{0a} \tag{7}$$

since (6) does not hold beyond here. The agreement between the two curves is fair and about as good as one might expect from this simple analysis. At least part of the deviation between the curves stems from a small cathode-coating resistance.

Whereas the positive-ion current has been neglected, it will actually amount to several ma when the anode is negative. This back current must be considered in certain applications such as in rectification where this current could result in accelerated gas clean-up as well as in premature inverse breakdown.

A comparison between the actual saturated anode current and that computed from (7), by probe measurements of plasma density, is displayed in Fig. 17. The deviation at low currents probably arises from the tendency of the positive-ion sheath surrounding the probe to overlap the electron sheath adjacent to the anode for low plasma densities. The deviation at high currents stems from the fact that the measuring probe was about 2 millimeters from the anode surface, and consequently measured a somewhat higher plasma density than that actually adjacent to the anode.[6]

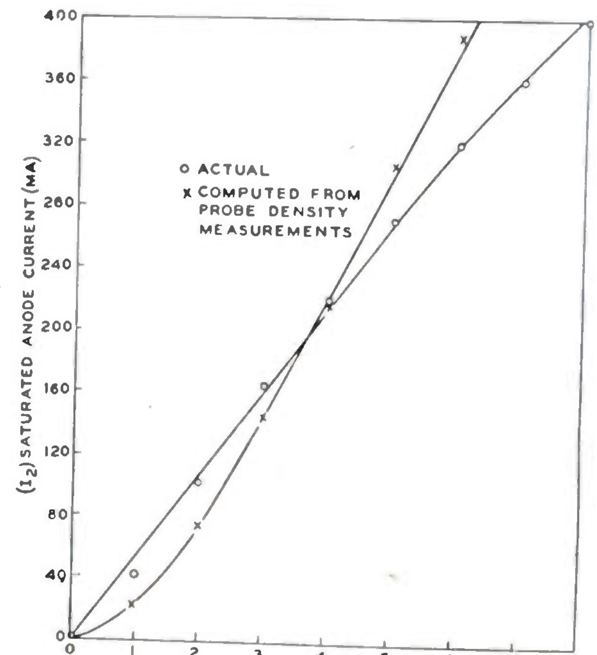


Fig. 17—Comparison of saturated anode current with its computed value.

The relation between the saturated anode current and the auxiliary current, with the cathode emission as parameter, is shown in Fig. 18. This family of curves shows how the saturated anode current is first limited at the anode, as exemplified by the linear dependence of this current on the auxiliary current and hence on the plasma density, and then saturates at a value corresponding to the emission capability of the cathode. The values of cathode emission were measured in the same manner as before. The agreement between these values

[5] L. Malter, E. O. Johnson, and W. M. Webster, "Studies of externally heated hot cathode arcs, part I—modes of the discharge," *RCA Rev.*, vol. XII, pp. 415–435; September, 1951. The "ball-of-fire" form of discharge is also described by M. J. Druyvesteyn and F. M. Penning, "Electrical discharges in gases," *Rev. Mod. Phys.*, vol. 12, p. 148; April, 1940.

[6] The increase in plasma density with decreasing radial distance from the cathode is discussed in the Appendix.

and the anode current along the horizontal portion of the curves is surprisingly good considering the usual vagaries of the emission from oxide-coated cathodes. Measurements were made rapidly as possible to prevent changes in cathode surface due to ion bombardment.

Theoretically, the departure from linearity should occur when the cathode-to-plasma potential V_k becomes zero. We can then write from (6)

$$A_k j_k = A_k j_{0k} + A_a j_{0a},$$

whence

$$\frac{A_k j_k}{I_2} = \frac{A_k j_{0k}}{A_a j_{0a}} + 1. \tag{7a}$$

Since saturation occurs at the anode (along the linearly dependent portion of Fig. 18), it follows that the total random current flow through a cylindrical shell coaxial with the cathode must be greater than through the anode sheath.[4] Then $A_k j_{0k} > A_a j_{0a}$. Substitution of this inequality into (7a) yields $A_k j_k > 2\, I_2$. That is, for normal plasmatron operation cathode emission should be at least twice maximum saturated anode current demanded.

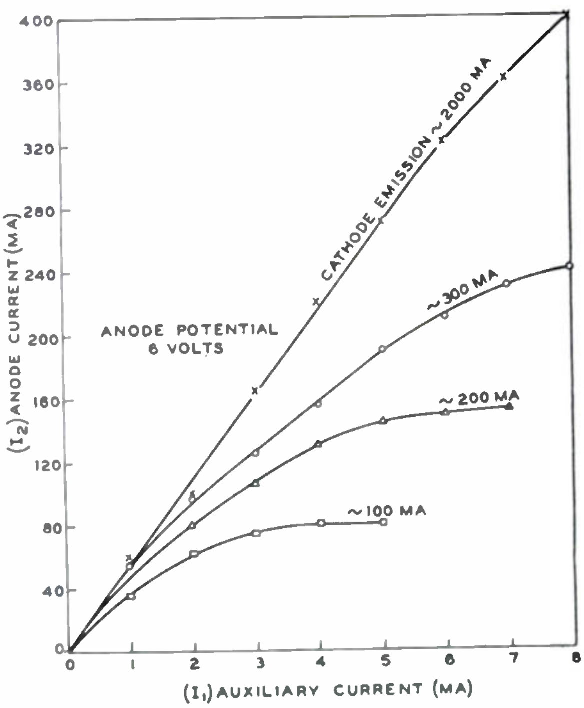


Fig. 18—Relation between anode and auxiliary discharge currents with cathode emission as parameter.

Equation (7) also indicates that the anode current should be independent of cathode area when the tube is operating with anode saturation. This was tested by employing a tube with three separate main cathodes of equal area as shown in Fig. 19. This arrangement allowed the cathodes to be used separately or in various parallel combinations. The results are tabulated in Table I (a). It is seen that the anode current is essentially unaffected by cathode area. Part (b) of the same table shows the effects which occur when the heater power is reduced to make the cathode-emission temperature limited. Here, as was expected, the anode current is seen to be equal to the sum of the cathode emissions.

TABLE I

(a) Anode Saturation

Cathode No.	Anode Current (ma)
1	300
2	240
3	200
1 and 2	280
1 and 3	260
2 and 3	240
1 and 2 and 3	260

(b) Cathode Saturation

Cathode No.	Anode Current (ma)
1	Negligible
2	14
3	56
2 and 3	68

To verify qualitatively the reality of the behavior of the plasma potential (as it has been portrayed) the arrangement of Fig. 19 was employed. This arrangement enabled the floating potential V_f of cathode No. 3, acting as a probe, to be observed as the volt-ampere characteristic of the main cathode-anode circuit was swept at a 60-cycle rate. For these tests cathode No. 3 was made a strong emitter so that its floating potential was very

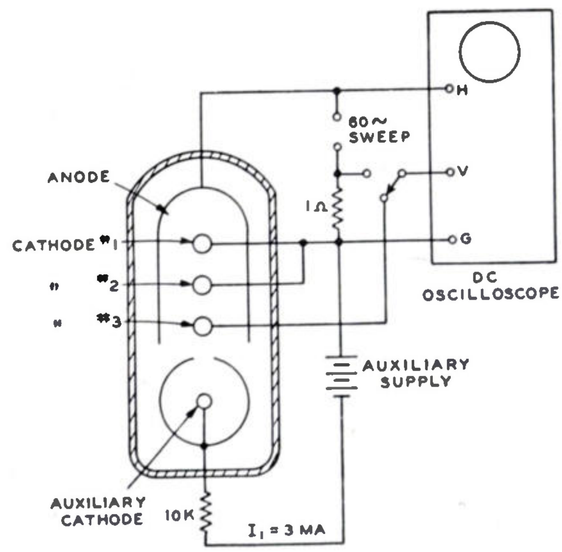
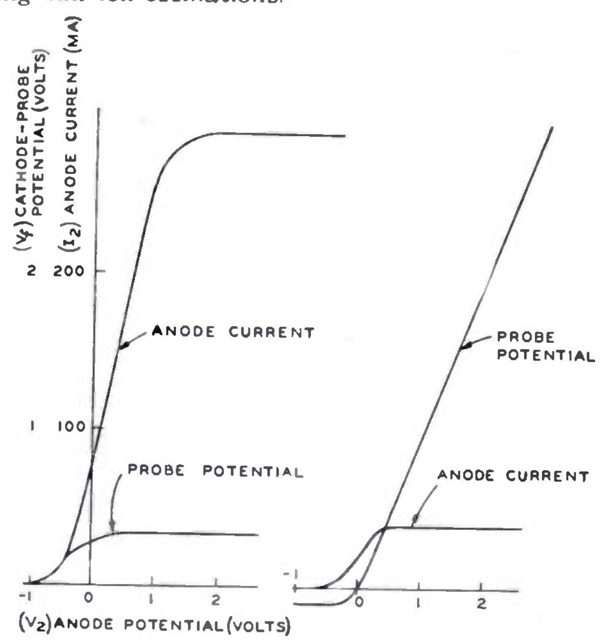


Fig. 19—Circuit for investigating the behavior of plasma potential.

close to the plasma potential. The curves in Fig. 20 (a) and (b) show the correspondence between the volt-ampere characteristics and the potential V_f which is essentially plasma potential. In the first case, where the cathode is a strong emitter, the plasma potential starts to rise with the anode current and stops rising when the anode current saturates. This behavior is in accord with that predicted by the analysis. Furthermore, the magnitude of the change in plasma potential is such as to account for the change in the current leaving the cathode. Little can be said with respect to the absolute value of the plasma potential since the contact potentials are not known. However, in other experiments made under more ideal conditions where the contact differences of potential could be measured, it was found that the plasma potential was actually about 0.5 v below that of the cathode surface. Also, the fact that the plasma electron temperature, measured in plasmas which are closely associated with copiously emitting cathodes, is generally very close to the temperature of the cathode is in itself a strong indication that such plasmas are below cathode potential. This temperature correspondence holds for oxide as well as pure metal cathodes.

In Fig. 20(b), where the cathode is a weak emitter, we see that the plasma potential moves up linearly with the anode potential once the anode current has saturated. This is in accord with the analysis which predicts that most of the tube drop occurs near the cathode when the cathode is a poor emitter.

There are conditions under which the cathode-plasma-anode system becomes unstable. Usually the only tendency towards instability that shows up in a helium-filled tube seems to occur when the system is operated with the cathode as an emitter of intermediate capabilities. Under these conditions probe measurements show that the plasma potential alternates between the distributions of Fig. 15 (a) and (b). When the instabilities give coherent oscillations, these have a period corresponding roughly to that of the Tonks-Langmuir ion oscillations.[7]



(a) (b)

Fig. 20—Relation between plasma potential and anode current. (a) Cathode a strong emitter. (b) Cathode a weak emitter.

When the gas filling is one of the heavier gases, such as xenon, then the tube is quite likely to become unstable. The reason for this is not yet clearly understood. There are indications, however, that this difficulty arises from nonsymmetrical plasma density distributions that are not appropriate to the current densities that have to be carried at the various cross sections in a cylindrical-type geometry.[4] Nonsymmetries in the plasma density are fostered by the fact that the ion generation is usually most intense in the region between the garrote and the main cathode in a tube type such as those already illustrated.

In summary it can be said that our simple picture of plasmatron behavior is in good agreement with a diversity of experimental results.

VI. Plasma Generation and Loss[8]

The auxiliary discharge is the sole agent for generating the necessary plasma since the main anode current is not allowed to ionize in normal operation. In plasmatrons of the types already described the major part of the ionization can be ascribed to the ionizing collisions that the stream of electrons emerging from the aperture in the garrote makes upon neutral atoms.

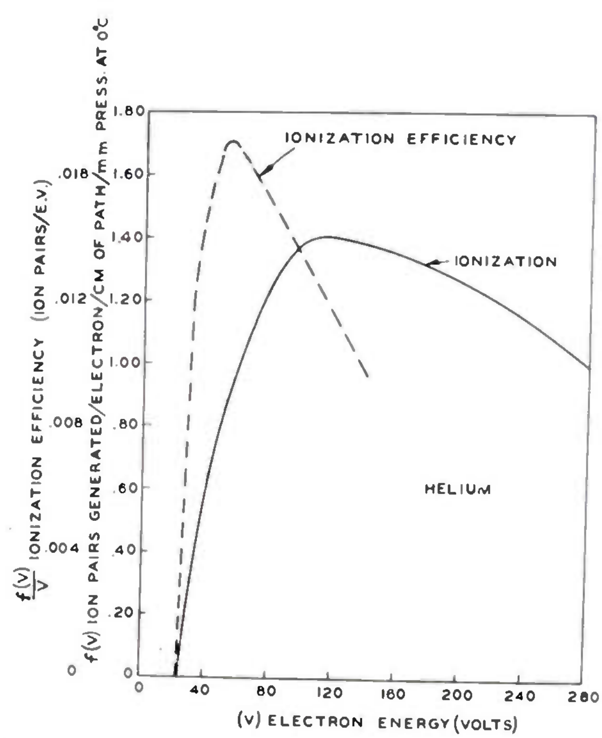


Fig. 21—Relation between ion generation and electron energy.

The dependence of this type of ionization in helium upon electron energy is shown in Fig. 21.[9] The dashed curve refers to the efficiency of ion generation as defined by the ion pairs generated per volt energy of the impinging electrons. The most copious generation occurs at $V=100$ v and the most efficient at $V=55$ v. At a gas pressure of p millimeters of mercury the differential ion current di_p arising from the ion generation, in a distance dx, by an electron current i_e is given approximately by

$$di_p = f(V) i_e \epsilon^{-x/\lambda'_e} dx.$$

Here the ionization function $f(V)$, is to a first approximation, as displayed in Fig. 21, the factor ϵ^{-x/λ'_e} is the

[7] L. Tonks and I. Langmuir, "Oscillations in ionized gases," *Phys. Rev.*, vol. 33, pp. 195–210; February, 1929.

[8] The plasma density is largely determined by the ion supply since the copious thermionic supply and high mobility of the electrons allows them to accommodate readily to the density conditions imposed by the ions. Thus the term "ion" will often be used in place of the term "plasma" in this section.

[9] M. Knoll, "Gasentladungstabellen," J. Springer, Berlin, p. 66; 1935.

fraction of the original electron current i_e still able to ionize after traveling the distance x, and λ_e' is a distance comparable to the mean free path of the electrons at the pressure p with regard to inelastic collisions.

Integration of the above equation over a distance d, large in comparison to λ_e', yields

$$i_p = f(V) i_e \lambda_e'. \tag{8}$$

Since the ionization function varies directly, and λ_e' inversely with pressure, (8) is independent of pressure. This equation represents the total ion current i_p generated in the region that the ionizing current i_e acts upon. The process by which the current i_p is lost from this region will now be investigated.

The volume recombination that takes place in the plasma is relatively negligible in the plasmatron so that the loss occurs principally by wall recombination. That is, the free electrons and ions in the plasma diffuse to the available bounding surfaces where they recombine. The travel of the free electrons and positive ions to the available surfaces occurs by an ambipolar diffusion process wherein ions and electrons leave the plasma at equal rates. This process will result in a total rate of loss of ions from the plasma as accounted for by

$$\frac{dN}{dt} = \frac{N}{\tau}, \tag{9}$$

where N is the total number of ions in the plasma and τ is the plasma decay constant, in seconds, and is described by[10]

$$\tau = \frac{\Lambda^2}{D_a}. \tag{10}$$

Here Λ is the characteristic diffusion path in centimeters and D_a is the ambipolar diffusion coefficient in square centimeters per second. The quantities N, τ, and Λ all refer to the lowest density mode of the plasma. The higher density modes have been neglected for the sake of simplicity since such an omission will not have any serious effect upon the approximate results desired in this investigation.

The application of these quantities to the plasmatron requires comment. In the direction parallel to the main cathode the usual ambipolar diffusion coefficient can be used because the total ion and electron currents arriving to the end micas are equal. However, in the cathode-anode direction the usual value of the ambipolar coefficient is no longer valid.[4] When the anode current is saturated, the electron sheath at the anode prevents any ions from leaving at this boundary. Also, the potential barrier at the cathode discourages ion loss at this point. Actually, there is some ion loss to most cathodes arising from variations of potential over the emitting surface. Except for this, loss to the end micas, the garrote and a very small amount of volume recombination, the positive ions are effectively trapped. Consequently, the values of Λ and D_a to be used in describing the plasmatron losses are generally not related to the vessel geometry or the mobility and diffusion coefficients in the manner described by Biondi and Brown.[10] This added complication presents no special problem since the effective value of τ can easily be evaluated by pulsing the auxiliary discharge and noting the rate at which the saturated main anode current decays. It is felt that the value of τ so obtained is not materially different from that which holds sway in the tube during normal operating conditions.

By introducing the electron charge e into (9), the ion current i_p lost from plasma region is found to be

$$i_p = \frac{Ne}{\tau}.$$

When (8) is introduced into this, there results

$$f(V)\lambda_e' i_e = \frac{Ne}{\tau}. \tag{11}$$

Probe studies of plasma density and the requirements upon the plasma density[4] indicate that N can be expressed in terms of the average plasma density n_{0a} adjacent to the anode, and the active tube volume v, by

$$N = \psi n_{0a} v.$$

The factor ψ, which has a value of about 2.0, is independent of the anode current and arises from the greater plasma density in the cathode region. Combined with (2), (2'), and (7), (11) becomes

$$f(V)\lambda_e' I_1 = 0.4 \times 10^{-6} \frac{j_{0a} v}{\tau}. \tag{12}$$

Here j_{0a} is expressed in ma per square centimeter, v in cubic centimeters, the electron temperature is taken as 1,000°K, and the ionizing current i_e has been replaced by its equivalent I_1, the auxiliary discharge current expressed in ma.

This equation, first of all, predicts that the anode current should be proportional to the ionization function $f(V)$. This prediction is borne out by the results shown in Fig. 22, where the crosses represent normalized values taken from Fig. 21 (as shown on page 654). The auxiliary voltage was controlled by varying the garrote bias. The larger the bias, the greater the constricting effect at the aperture,[11] and hence the larger the auxiliary discharge arc drop.[12]

Secondly, (12) indicates that the anode current should be proportional to the auxiliary current I_1. The results in Figs. 4 and 17 bear adequate testimony to the validity of this conclusion.

Thirdly, (12) suggests that the current gain, as defined by the ratio of the saturated anode current to the auxiliary current, should be directly proportional to the decay constant τ. Both the qualitative and quantitative behavior shown in Fig. 23 confirm this very nicely.

[10] M. A. Biondi and S. C. Brown, "Measurements of ambipolar diffusion in helium," *Phys. Rev.*, vol. 75, pp. 1700–1705; June, 1949.
[11] H. Fetz, "On the control of a mercury arc by means of a grid in the plasma," *Ann. der Phys.*, vol. 37, pp. 1–40; January, 1940.
[12] Probe studies show that virtually all of the arc drop appears in the aperture.

However, the unusually good quantitative results probably contain an element of fortuity which derives from the approximations introduced into the derivation of (12).

In evaluating (12) the following values were used: $A_a = 8.5$ cm², $v = 3.25$ cm³, $V = 60$ volts, and $\lambda_e' = 0.235$[13] centimeters at $p = 1$ millimeter pressure. The tubes represented in Fig. 23 were identical with the one shown in Fig. 2, except for small geometrical differences in the main cathode-anode region which altered τ. For example, tube No. 3 had positive baffles over the end micas; these prevented ion losses to the micas, and so altered τ.

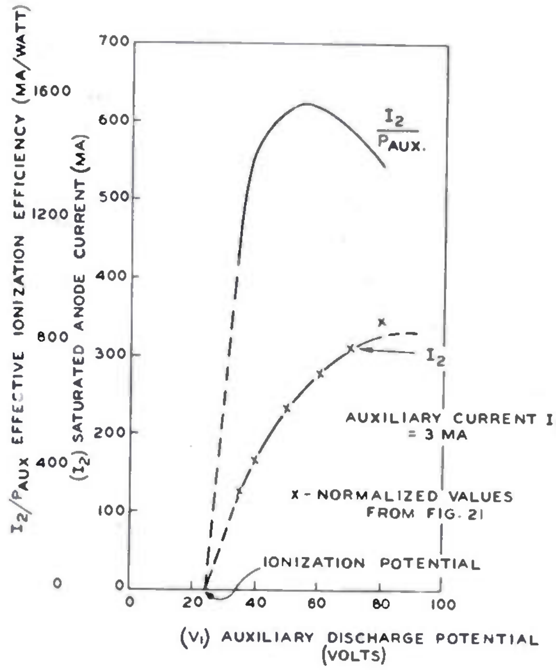


Fig. 22—Relation between saturated anode current and auxiliary discharge potential.

It is instructive to consider the effective ionization efficiency curve shown in Fig. 22. Here the peak efficiency corresponds to 1.5 amperes of anode current for each watt in the auxiliary discharge. Operation under these conditions is several times as efficient as the operation of an ordinary hot cathode diode. To illustrate this point compare the total energy consumed in tube drops for a given anode current (1) using plasmatron operation and (2) using the same tube without the auxiliary discharge. Now if the necessary auxiliary heater power of

	(1) Plasmatron	(2) Diode
Anode current	0.26 a	0.26 a
Anode voltage	2.0 v	27.0 v
Auxiliary voltage	55.0 v	—
Auxiliary current	0.003a	—
Total power in tube drops	0.69 w	7.0 w

1.0 watt (at the conservative rate of 20-ma emission per watt heating power) is added, the plasmatron shows a power loss which is only one quarter that of the diode. Had the gas been xenon, which has an ionizing potential of about 11 v, the saving would still be 50 per cent or more. Thus, the use of the plasmatron principle in rectifier circuits, where efficiency is a consideration, appears attractive.

[13] Landolt-Bornstein, "Zahlenwerte und Funktionen aus Physik und Chemie," J. Springer, Berlin; 1950.

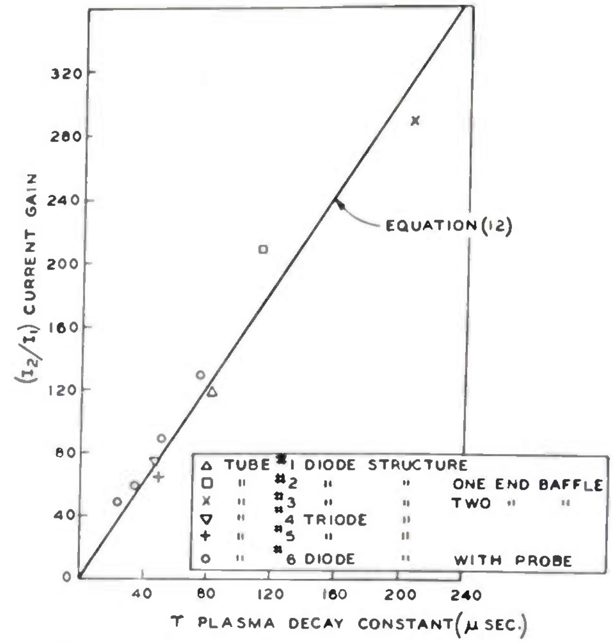


Fig. 23—Relation between current gain and plasma decay constant.

These results emphasize that the ordinary hot cathode gas diode operates at a disadvantage in that its anode potential is beset with the dual problem of pulling electrons through the tube as well as having to take care of the ionization requirements.

Any particular ion density can be put to the best use by taking advantage of the fact that for a given ion density the anode current is proportional to the anode area as shown in (7). By using various stratagems, such as finned anodes, the anode area can be increased several-fold without increasing the outside dimensions of the tube. When this is done, current gains of 300:1 are easily achieved in tubes such as those illustrated. If the previously mentioned trapping action is pushed to extremes, by extending the anode over the end micas and garrote, even greater current gain can be obtained.[14]

In dealing with the frequency response of the device in the first or diode method of operation one is concerned with the time constants of plasma build-up and decay. The plasma build-up time is a function of how fast ions can be generated by the auxiliary discharge. This process occurs so much faster than the decay time that it can be ignored. The decay of the plasma density will proceed largely by ambipolar diffusion according to the relation[10]

$$n = n_0 \epsilon^{-t/\tau} \tag{13}$$

from which (9) was derived. Here n is the ion density at a particular point in the tube at a time t, n_0 is the initial

[14] The volume recombination loss should become important when the plasma density n_{0a} at the anode has a value given approximately by $n_{0a} = 1/2\alpha\tau$ which is derived by equating the diffusion to the recombination loss. Using the value 1.7×10^{-8}, given by Biondi and Brown[10] for the volume recombination coefficient α of helium, and 100 microseconds for τ the value of n_{0a} turns out to be 3×10^{11}. This corresponds to an anode current of about an ampere in these tubes, which is approximately the current at which indications of volume recombination become evident.

ion density at this point, and τ is the decay constant previously described. As might be expected the sine-wave frequency response is inversely related to the decay constant and drops off at the expected frequency.

VII. Grid Action

In the second mode of operation the saturated anode current is controlled by varying the cross-section of the plasma in the vicinity of the anode. This is accomplished by interposing a grid between the main cathode and anode (see Fig. 7). The anode characteristic with grid bias as a parameter is plotted in Fig. 9. When the grid is at a negative potential with respect to the plasma, its wires are surrounded by positive-ion sheaths which serve as barriers through which electrons cannot flow. Thus, the anode current is made to depend upon the thickness of the grid sheaths. This control of the anode current appears as a variation of the effective anode area A_a in (7). Actually, an increasingly negative grid bias is effective in reducing the anode current in still another fashion. The sheaths act as ion sinks whose effect is to reduce the plasma density in their neighborhood. As the sheaths expand, the plasma density decreases correspondingly, resulting in a further decrease of the saturated anode current. This second control mechanism may be considered analytically as a variation of the effective random current density j_{0a} in (7). Thus, the grid exerts control over the anode current through two co-operating mechanisms: (1) the variation of the effective collection area of the anode, and (2) the variation of the current density able to flow through this area.

The relative contributions of the A_a and j_{0a} variations have been estimated in the following fashion. From a measurement of the positive-ion current to the grid when it was highly negative, a value for plasma density was obtained. Assuming that the plasma density is the same for all other values of grid bias, the sheath thickness as a function of grid bias was computed by the methods described by Malter and Webster.[16] It was further assumed that A_a, and consequently the anode current, was proportional to the region between the grid wires not closed off by adjacent sheaths. The resultant anode-current grid-voltage characteristic is plotted in Fig. 24. Here the ordinate is plotted in terms of the current at zero bias. For comparison the experimental results are plotted on the same figure. The anode current was always maintained in a saturated condition. It is seen that the experimental characteristic is very nearly the square of the theoretical characteristic (indicated by crosses). This suggests that the variation in anode current is controlled about equally by the two mechanisms, at least for tubes of the structure shown in Fig. 7, and that the effective anode area and the random electron current density are nearly linear functions of the grid-sheath radius.

[16] L. Malter and W. M. Webster, "Rapid determination of gas discharge constants from probe data," *RCA Rev.*, vol. 12, p. 191; June, 1951.

The similarity between the va output characteristics of the diode and triode operation stems from the fact that the grid action affects the total quantity $A_a j_{0a}$ of (7) just as the auxiliary discharge does. Comparison of Fig. 3 with Fig. 9 shows that a higher value of auxiliary current is required to give the same anode current in the triode as in the diode. This comes about because even at zero bias the grid wires act as ion sinks which reduce the plasma density adjacent to the anode.

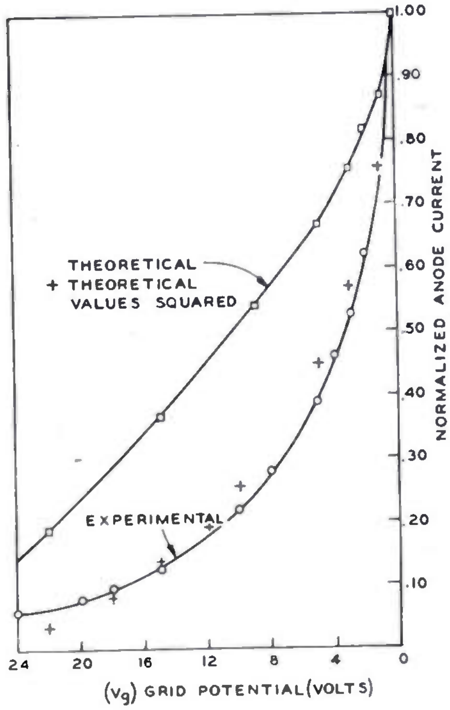


Fig. 24—Comparison between theoretical and experimental grid control characteristic.

The triode frequency response, plotted in Fig. 10, is compounded of a complex combination of ion and electron movements in the grid-anode region. The falling off of response is a result of the finite time required to alter the plasma density in the grid-anode region. For example, if the region is closed off from the main body of the plasma, the residual plasma therein soon disappears by diffusing to the grid wires. When the sheaths are then retracted, a definite length of time, primarily determined by the slow-moving ions, is required for the plasma to be restored in the grid-anode region. This build-up time requires about 5 microseconds in these tubes, whereas the cut-off of the anode current by the grid sheaths occurs in a fraction of a microsecond.

It is interesting in the light of the above to point out why the thyratron grid, in contradistinction to the plasmatron grid, loses control after the onset of a discharge with its concomitant plasma formation. The loss of grid control, in the case of the thyratron, stems from the increase in ionization that necessarily accompanies the increase in the thickness of the grid sheath.[11] As the grid sheaths reduce the span of the grid openings the anode current at first tends to decrease. This decrease, in the usual resistance-limited circuit, raises the anode potential, making possible a higher rate of ion generation and hence a more dense grid-anode plasma. As a result of this greater plasma density the effect of the original grid sheath expansion is virtually cancelled and little net effect upon the anode current ensues. At very low gas pressures the rate of ion generation is less able to keep pace with the grid-sheath expansion so that a certain amount of grid control can be achieved with grids that have small openings. At higher gas pressures, such as are normally used in commercial thyratrons, this is not usually the case. This entire matter has been studied very thoroughly by Fetz[11] and need not be elaborated upon at this point. The plasmatron escapes the difficulties of the thyratron by utilizing a plasma that is produced independently of the anode current.

VIII. General Behavior and Applications

While exhaustive life studies have not been carried out, it would seem that the plasmatron would be at least as good in this respect as a comparable gas tube such as a thyratron. The presence of the retarding field, for positive ions, at the cathode precludes the possibility of trouble from disintegration of the cathode due to ion bombardment. This should also apply to the auxiliary cathode since the arc drop of the auxiliary discharge occurs across the aperture in the garrote and, in addition, there is also a small retarding field at the auxiliary cathode. It would seem that the low operating potentials normally required would not cause prohibitive gas clean-up.

There is no reason to believe that the tube should behave any differently with respect to ambient temperatures than a thyratron filled with the same type of gas. Thus, a tube filled with a noble gas should be relatively temperature insensitive.

Since the tube is a low-voltage, high-current device, special attention must be given to the cathode. As regards over-all efficiency, the cathode-emission efficiency must be kept as high as possible and voltage drops in the cathode coating and interface must be kept to a minimum. It has been observed that such drops can add several volts to the potential at which anode current saturation occurs.

Contact differences of potential between the cathode and anode are often noticable in the plasmatron. During cathode activation the entire va characteristic can be seen to drift toward lower anode potentials. This shift of the characteristic is presumed to be due to the decrease in the anode work function which follows from the deposit of evaporated cathode material.

In its present developmental stage the tube has been used in the laboratory for direct loudspeaker drive, motor control, switching, regulation, inversion, and low drop rectification.

Conclusion

This new electronic device, a continuously controllable hot cathode gas tube, shows excellent promise for fulfilling the long-standing need for a device that will continuously control large currents at low voltages.

Appendix

It is instructive to make a simplified analysis of plasma conditions in a cylindrical geometry such as in Fig. 25. It is assumed that the charged particle densities are equal so that $n^+ = n^- = n$. This assumption seems justified on the basis of an analysis made on plane parallel structures where n^+ and n^- were not assumed equal. It cannot be expected to hold near plasma boundaries.

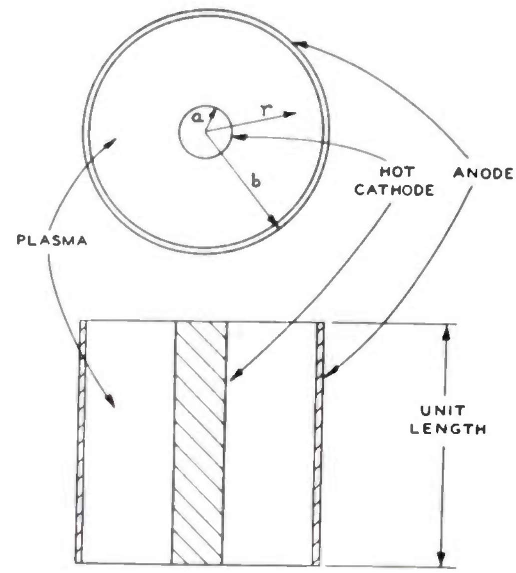


Fig. 25—Section of plasmatron diode.

The electrons enter the plasma at the cathode with a density q_a^- and leave at the anode with a flow density aq_a^-/b. Positive ions are allowed to enter either at anode or cathode with a flow density q_0^+, or to be present and have no net drift velocity in a radial direction.

In addition to the current continuity relations and the above boundary conditions, the flow equations

$$q^+ = nE\mu^+ - D^+ \frac{\partial n}{\partial r} \tag{1}$$

and

$$q^- = -nE\mu^- - D^- \frac{\partial n}{\partial r} \tag{2}$$

must be satisfied; q^+ and q^- are the respective positive ion and electron particle flows, respectively, u^+ and u^- are mobilities, D^+ and D^- the diffusion coefficients, n is the particle (plasma) density at the radius r, and E is the field.

From (1) and (2) we get

$$q^+ = -\left[\frac{D^+\mu^- + D^-\mu^+}{\mu^- + P\mu^+}\right]\frac{\partial n}{\partial r} = -D_a' \frac{\partial n}{\partial r}, \tag{3}$$

where D_a' is the modified ambipolar diffusion coefficient and $P = \pm q^-/q^+$. The positive sign refers to the case where the particle flows are in the same direction and the negative sign where the flow directions are opposed. By definition D_a' is the ambipolar diffusion coefficient D_a when $P=1$. Since $\mu^- \gg \mu^+$, D_a' will not start to differ appreciably from D_a until the absolute value of P becomes greater than 20 or so. This is an interesting result since it indicates that diffusion conditions are still "ambipolar" even when there is a relatively large disparity between the electron and positive-ion flows to boundaries. But, when the flows differ by a factor of 100 or more, as in the radial direction in the plasmatron, then conditions are far removed from ambipolar.

From (1) and (2), the continuity relations, and the boundary conditions one obtains:

$$dn = -\frac{aq_a^-}{D_a}\left[\frac{\mu^+}{\mu^-} \pm \frac{r_0 q_0^+}{aq_a^-}\right]\frac{dr}{r}. \tag{4}$$

The positive sign refers to the case where the positive ions enter at the cathode and the negative sign to the case where they enter at the anode. In the first case $r_0=a$, and in the second $r_0=b$.

If n_b is the positive-ion density adjacent to the anode, (4) becomes

$$n = n_b + \frac{aq_a^-}{D_a}\left[\frac{\mu^+}{\mu^-} \pm \frac{r_0 q_0^+}{aq_a^-}\right]\ln\frac{b}{r}. \tag{5}$$

This shows that if the positive ions enter at the cathode, remain essentially stationary (as they would for low rates of generation and loss), move in an axial direction, or enter at the anode at such a rate that the quantity in the bracket does not become negative, the plasma density will be greatest close to the cathode. If in (5), q_a^- is proportional to n_b and q_0^+/q_a^- remains approximately constant, then the density distribution function will be essentially independent of the anode current. Experiment indicates that such is the case. If q_a^- is not proportional to n_b, as would be the case if the anode current is not saturated, then (5) shows that the plasma density in the cathode region should increase with q_a^-. Also because fewer ions can run out at the anode the density in this region might be expected to increase somewhat. These conclusions are well borne out by the results of probe measurements in the anode and cathode regions of a plasmatron. Typical behavior (shown in Fig. 26), where the probe current was measured as the anode current was changed. Plasma decay studies also show that the disappearance of positive ions is markedly lessened as the anode current approaches its saturation value.

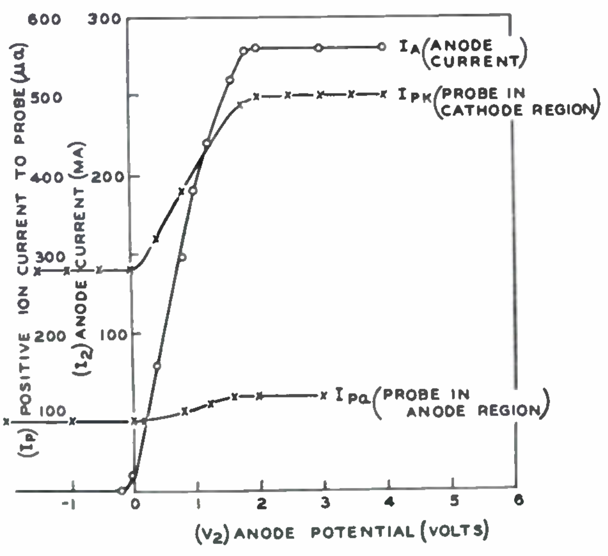


Fig. 26—Effect of anode current and potential on plasma density.

In developing the expressions to describe the plasmatron operation, it has been assumed that the current carrying capacity of the plasma is a minimum at the anode, that is, that $A_b j_b < A_r j_r < A_a j_a$. Here the A's refer to the plasma cross-section area normal to a radius and the j's to the electron space current densities at these cross sections. The conditions necessary for this situation to exist can be found from (5). Thus all will be well if

$$n_b b < nr$$

or

$$n_b b < n_b r + \frac{aq_a^-}{D_a}\left[\frac{\mu^+}{\mu^-} \pm \frac{r_0 q_0^+}{aq_a^-}\right] r \ln\frac{b}{r} \tag{6}$$

$$1 < \frac{r}{b} + \frac{aq_a^-}{bn_b D_a}\frac{\mu^+}{\mu^-} Fr\ln\frac{b}{r}$$

$$F = \left\{1 \pm \frac{r_0}{a}\frac{q_0^+}{q_a^-}\cdot\frac{\mu^-}{\mu^+}\right\}.$$

It is convenient to introduce the dimensionless factor F to account for the effect of the current ratio. The equation can be simplified further by the introduction of the relations

$$\frac{a}{b}q_a^- = q_b^-, \qquad \frac{q_b^-}{n_b} = \frac{c^-}{4}, \qquad \theta = \frac{T^-}{T^+}.$$

Here c^- is the average electron thermal velocity and the T's refer to the particle temperature. If the classical expressions for the average thermal velocities, the electron mean free path L^-, and mobilities are used, one obtains

$$1 < \frac{r}{b} + \frac{4}{\pi}\frac{\theta}{1+\theta}\frac{r}{L^-}F\ln\frac{b}{r}$$

or, approximately

$$1 < \frac{r}{b} + \frac{r}{L^-}F\ln\frac{b}{r}. \tag{7}$$

From (7) it is seen that the necessary conditions for normal plasmatron operation, wherein plasma saturation occurs at the anode and not elsewhere, can be met when (1) the ratio a/b is not too small, (2) the pressure is such that L^- is not too large, and (3) the ions are injected close to the cathode so that F tends to be large. Qualitative observations bear out these conclusions. In fact, when instabilities in operation do occur, the behavior of the plasma potential strongly suggests that the point of plasma saturation alternates between the cathode and anode regions, indicating that (7) is not being continuously satisfied.

On the Possibility of Obtaining Radar Echoes from the Sun and Planets*

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Summary—The sun is, after the moon, the simplest major astronomical body to reach by radar, and it is a more interesting object for study than the moon or the planets. In this paper, the process of reflection from the sun is investigated theoretically, together with the manner in which the echo intensity would be expected to depend on the various parameters of a radar system. It is shown that the detection of sun echoes appears technically possible, but a radio-engineering project of considerable magnitude is involved. Orders of magnitude of planetary echoes are also briefly discussed, and a reference is made to several astronomical studies which could be carried out with a very high-power radar.

I. Introduction

FOLLOWING the successful reception of radar echoes from the moon,[1-5] speculations have appeared in various places regarding the possibility of obtaining echoes from more distant objects. These speculations do not yet appear to have been put on a quantitative basis. Although Mercury, Venus, and Mars approach more closely to the earth, the sun presents the best possibility (after the moon) of returning detectable echoes because of its much greater size. Furthermore, the sun is the most interesting astronomical object for a radar investigation. Reflection would take place high in the solar atmosphere in the fully ionized gases of the corona. This region is very difficult to investigate optically. The corona is known to exhibit large spatial and temporal variations which could be studied by radar. It is also believed that streams of ionized particles of the type which cause terrestrial auroras are sometimes emitted from the sun, and these might be detected while in its vicinity.

An order of magnitude for the power required for a detectable echo from the sun can be obtained by the following simplified argument:

The angular diameter subtended by the sun at the earth is nearly the same as that of the moon ($\frac{1}{2}°$), so that both will receive about the same power flux from a terrestrial transmitting antenna. Neglecting for the moment the differing reflection coefficients of the two bodies, the decrease in the sun's echo relative to that from the moon will be due solely to the greater distance traveled by the diverging reflected radiation. Since the sun is about 400 times as distant as the moon, the sun's echo will be smaller by a factor of $(400)^2$, or 52 db. The more detailed discussion, presented below, must take several additional factors into account, such as, the partial absorption of the incident energy in the solar atmosphere, the greater size of the sun's radio disk, the necessity of overriding the sun's own radiation ("solar noise"), and the variation of these and other factors with frequency.

The evaluation of these factors is based on current knowledge of the nature of the solar atmosphere, but it should be noted that the relevant quantities are not known exactly. Because great difficulties exist in obtaining information concerning the outer corona and surrounding clouds or streams of charged particles by optical means, interest centers on the possibility of obtaining echoes from these regions. Hence, the use of a low frequency is indicated. For this reason a frequency of 30 mc will be used to illustrate points in the discussion, the behavior at other frequencies being later expressed in terms of that at 30 mc.

Following a derivation of the power requirements for echoes from the sun and those for the nearer planets, some comments are made on the practicability of building equipment of the necessary magnitude, and on uses to which a suitably powered astronomical radar system might be put.

II. Reflection from the Sun

A. The Solar Corona

The solar corona is a mass of ionized gas at high temperature, extending from the sun out to a distance of several solar radii. In the generally accepted model, the electron density decreases outwards according to the formula derived by Baumbach,[6] and later modified by Allen[7] and van de Hulst.[8] The existence of a high-electron temperature (about 10^6 °K) has been inferred from the width of the coronal emission lines, the high state of ionization required to produce these lines, and the intensity levels of the "quiet" solar noise. Spherical symmetry is often assumed for simplicity, but eclipse observations of the corona indicate that this is far from correct, the typical pattern showing a number of radial streamers. There is some evidence that these streamers are associated with prominences in the chromosphere,

* Decimal classification: R115.23×R116. Original manuscript received by the Institute, October 1, 1951.

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[1] J. Mofensen, "Radar echoes from the moon," *Electronics*, vol. 19, pp. 92–98; April, 1946.

[2] Z. Bay, "Reflection of microwaves from the moon," *Hungarica Acta Physica*, vol. 1, pp. 1–22; 1946.

[3] J. H. De Witt, Jr., and E. K. Stodola, "Detection of radio signals reflected from the moon," Proc. I.R.E., vol. 37, pp. 229–242; March, 1949.

[4] F. J. Kerr, C. A. Shain, and C. S. Higgins, "Moon echoes and penetration of the ionosphere," *Nature* (London), vol. 163, pp. 310–313; February 26, 1949.

[5] F. J. Kerr and C. A. Shain, "Moon echoes and transmission through the ionosphere," Proc. I.R.E., vol. 39, pp. 230–242; March 1951.

[6] S. Baumbach, "Strahlung Ergiebigkeit und Electronendichte der Sonnenkorona," *Astron. Nach.*, vol. 263, pp. 121–134; July 2, 1937.

[7] C. W. Allen, "Interpretation of electron densities from corona brightness," *Mon. Not. R. Astr. Soc.*, vol. 107, pp. 426–432; 1947.

[8] H. C. van de Hulst, "Zodiacal light in the solar corona," *Astrophys Jour.*, vol. 105, pp. 471–488; May, 1947.

so that the pattern presumably changes in periods of the order of days or less, while the occurrence of variations of some kind of the order of seconds and minutes is indicated by the intensity variations of solar noise from the corona.

Some, at least, of the variations are doubtless due to the ejection of prominence material from the chromosphere into or through the corona. The visible portion of a large prominence extends typically to 100,000 km from the photospheric surface. The current interpretation of several terrestrial phenomena, for example, auroras and magnetic storms, involves the interception by the earth of streams of particles (most probably neutral streams, with equal numbers of positively and negatively charged particles). It is inferred that these streams are ejected from active regions on the sun, but no direct evidence is yet available.

The coronal filaments, and, thus, presumably the main body of the corona, are known to have the same angular velocity of rotation as the photosphere, with an approximate period of rotation of 27 days.[9] This corresponds to a linear velocity of $[1.9\times a/R_0]$ km per second for a region at a distance of a/R_0 solar radii from the axis of the sun. Any internal mass motions existing in the corona must also be taken into account in the radar considerations, as they would involve a Doppler dispersion of the reflected energy. The high thermal velocities of the electrons (order of 25 km per second) will have a negligible effect in this connection, as they average to zero on a macroscopic scale. Of great importance is the possibility of high turbulent velocities in the rest of the coronal material, though current optical evidence suggests that this motion is small.

B. The Reflection Process

For simplicity, we will neglect the large irregularities known to be present in the corona. These will modify the spatial distribution of the reflected energy, but should not greatly affect the mean echo strength received at the earth.

We will, however, take into account the smaller irregularities, small compared with solar dimensions, but large compared with the wavelength of the incident radiation. In other words, the contours of equal density in the corona will be "rough" surfaces. The parameters derived by Smerd[10] for the Baumbach-Allen spherically symmetrical model will be used for the mean characteristics at any level.

Each ray entering the corona from outside will, in general, suffer deviation from its original direction, together with a certain amount of absorption attenuation. After a time, the ray will encounter an electron density sufficient to turn it away from the sun. For a ray normal to the ionization gradient, this critical density is that which makes the refractive index (μ) zero for the frequency concerned; for an oblique ray, a smaller electron density will suffice.

The relevant properties of 30-mc rays are shown in Fig. 1, for the case of a spherically symmetrical (smooth) corona. On this idealized model, the mean level of the $\mu=0$ surface for the 30-mc ordinary ray is at 1.56 solar radii. All rays except the central one are reflected away from the sun at levels outside this.

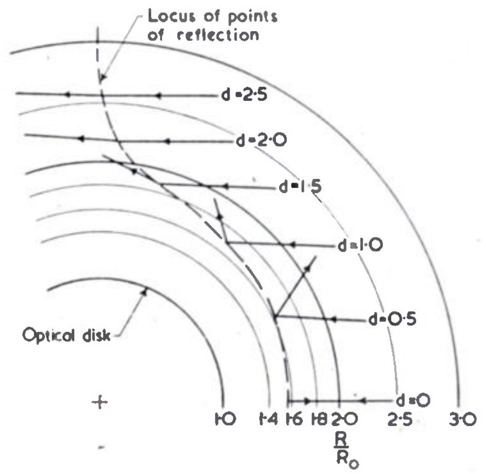

Fig. 1—Reflection from spherically symmetrical corona, 30 mc. (d=distance, measured in terms of the sun's optical radius R_0, between a ray and the ray towards the center of the sun, before they enter the sun's atmosphere).

If the reflecting surface were smooth, only the first Fresnel zone would contribute effectively; but when the reflecting surface is rough, each region will contribute something to the echo received at the earth since in each region there will be rays which will meet some $\mu=0$ surface normally. (We are neglecting the possibility of reflection from a sharp discontinuity in electron density, about which little can be said from present knowledge.) The energy scattered back in the echoing direction is greater for a rough sphere than for a smooth sphere of the same size, which scatters uniformly in all directions, because (1) the fraction of the "illuminated" area seen from any direction decreases on going away from the incident direction, and (2), in regions towards the limb, each irregularity shadows an area behind it, reducing the effective illuminated area capable of reflection to directions away from the incident direction.

To form an estimate of the directivity of the radiation reflected from the corona, we can use figures for the optical directivity of the planets derived from the variation of their brightness with astronomical phase. The result obtained for Venus, a planet with an opaque atmosphere, is 3.3. Corresponding figures for radio reflection would be less if the irregularities were small, owing to the smaller apparent roughness of a surface for the longer wavelength. In the case of the sun (also a gaseous type of reflector) the irregularities are presumably larger than those of Venus. Furthermore, the central strip of the sun is rendered more important in the reflection process by the Doppler spread because of rota-

[9] M. Waldmeier, "Polarkarten der Sonnenkorona," *Zeit. für Astrophysik*, vol. 27, pp. 24–41; April, 1950.

[10] S. F. Smerd, "A radio-frequency representation of the solar atmosphere," *Proc. IEE* (London), part III, vol. 97, pp. 447–452; November, 1950.

tional velocities. Thus, a not unreasonable estimate of radio-frequency directivity for the sun is 4.

We must now consider the effects of the sun's refracting and absorbing atmosphere. Refraction would reduce the effective reflecting area, by deviating rays away from the reflecting surface, while absorption would lower the echo intensity by reducing the total amount of energy scattered. For the accuracy required, we can separate the effects of refraction, absorption, and directivity, and consider each singularly for the whole sun.

In addition to the decrease of curvature of the actual reflecting surface (by a factor of 1.56, see Fig. 1), there is a further effect associated with the gradual bending of the rays. A ray which enters the solar atmosphere at a distance of, for example, 0.5 solar radii from the central ray, has its turning point at about 0.65 solar radii distant. This is equivalent to a decrease of effective radius by a factor of about 1.3. The net effect is an increase of the sun's effective radius over that of the photosphere by 1.56/1.3 to the value 1.2 R_0.

The absorption loss for central and oblique rays can be obtained from Smerd's results for the optical depth to the reflecting region,[11] giving a loss for the double passage varying from 4.3 db, for $d=0$ to 1.5 db, for $d=1.5$. Making rough allowance for the directional characteristics of the reflected energy, we will adopt 3.5 db the mean loss of all rays contributing to the echo.

After deducing the fraction of the incident energy which is reflected back to the earth, we must consider how the echo energy is dispersed in frequency by the Doppler effect. The solar rotation will produce a shift of $[375\times a/R_0]$ cps at 30 mc for a point on the sun's disk at a projected distance of a/R_0 solar radii from the axis of rotation. The energy spectrum arising from this effect will be symmetrical about a frequency corresponding to the central strip of the sun's disk, this frequency being displaced from the transmitted frequency by a slowly varying shift which depends on the earth's orbital and diurnal motions. The spectrum can be computed approximately, by using the ray trajectories of Fig. 1 to deduce the levels at which various portions of the incident beam will be reflected, and, hence the corresponding frequency shift. Making allowance for the variation of directivity over the various portions of the disk, it appears that at 30 mc about 55 per cent of the echo energy will be contained in a band of 500 cps, and 90 per cent in 1,000 cps.

Optical evidence suggests that the turbulent velocities within the corona are smaller than the rotational velocity, so that they should not produce appreciable additional dispersion of energy. Some of the energy will be reflected from material traveling with prominence velocities, and consequently will suffer quite large frequency shifts (for example, nearly 200 kc for a velocity of 1,000 km per second), but it is to be expected that only a negligible fraction will be dispersed in this way.

C. Echo Cross Section of Sun

We have seen that the incident energy contributing to the echo is that which would pass normally through an area of $\pi(1.2\times R_0)^2$ at the sun's distance. To a first approximation, and making the additional assumption that the pulse length exceeds the spread in delay time of the sun echo (7 seconds), the echo cross section, σ, will be the product of

A, the equivalent area, $\pi(1.2\times R_0)^2$,

$\beta(\leq 1)$, the receiver acceptance factor, the proportion of the dispersed reflected energy which is contained within the receiver bandwidth,

D, the directivity in the back-scattering direction of that portion of the scattered radiation which is accepted by the receiver,

α, the mean ratio of emergent to incident power over all rays, considering only the effect of absorption,

$$\sigma = A\beta D\alpha.$$

Thus for a 500 cps bandwidth, for which $\beta=0.55$,

$$\begin{aligned}\sigma &= \pi(1.2\times R_0)^2\times 0.55\times 4\times 0.45\\ &= 1.5\pi R_0^2.\end{aligned}$$

III. Noise Background

Sun echoes will be received against a background of solar and galactic noise, which will be so large at the frequencies considered that receiver noise will be of little importance provided the noise factor is reasonably good. Solar noise can be expected to rise at some times during disturbed periods to values of many times the sun-echo level. With the indicating and integrating system discussed later, it may be possible to see small changes in even a highly variable noise level, but no quantitative data are available about this. We will consider in detail the case of the "quiet" sun.

Smerd's results[11] for the 30-mc optical depth, which were used above to obtain a figure for the absorption, led to an apparent temperature, T_a, for the quiet sun of 900,000°K.[12] The noise power available from a 30-mc plane-polarized antenna of gain 3,000 directed towards the sun would then be

$$P_s = \frac{kT_a}{\lambda^2}\,\Omega_s\,\frac{G\lambda^2}{4\pi}\text{ watt (cps)}^{-1},$$

where

k = Boltzmann's constant (joule/deg.)
Ω_s = solid angle subtended by Sun's optical disk (steradian);

therefore,

$$P_s = 2.0\times 10^{-19}\text{ watt (cps)}^{-1}.$$

[11] S. F. Smerd, "Radio-frequency radiation from the quiet sun," *Aust. Jour. Sci. Res., Ser. A*, vol. 3, pp. 34–59; March, 1950.

[12] Temperature referred to sun's optical disk.

The antenna temperatures to be expected from galactic noise have been inferred from published results of observations made in and near the frequency range concerned.[13] The estimated variation throughout the year of 30-mc galactic noise in the direction of the sun is shown in Fig. 2 for an antenna of gain 3,000. The range of temperatures is about 12,000 to 100,000°K., the highest value occurring in December. The median

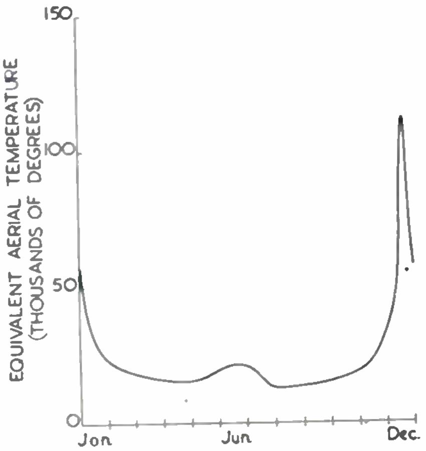

Fig. 2—Estimated 30-mc antenna temperatures due to galactic noise for an antenna directed towards the sun on the days shown.

value, which will be little affected by change of antenna gain, is 17,500°K., equivalent to an available power at the receiver of

$$P_G = kT = 2.4 \times 10^{-19} \text{ watt (cps)}^{-1}.$$

The total power from the quiet sun and galactic noise (median value) will then be

$$P_N = 4.4 \times 10^{-19} \text{ watt (cps)}^{-1}.$$

The constancy of this level at 30 mc is not known, but some special observations have been made by Bolton[14] at 100 mc, under conditions in which about one-tenth of the antenna noise power was galactic, and the rest solar. He finds that on the average day, described as "quiet," the noise level shows variations of up to 5 to 10 per cent with significant times of the order of 2 to 5 seconds. On exceptional days, the variations are less than 1 to 2 per cent for considerable periods.

In addition to solar and cosmic noise, trouble is likely to be experienced under some conditions from atmospherics and station interference, although use of a frequency above 25 mc would minimize these effects.

IV. Detection of Sun Echoes

A. Propagation between Sun and Earth's Surface

Electron densities in interplanetary space are normally so small that negligible absorption should take place there. Similarly absorption in the terrestrial ionosphere at high angles and at the high frequencies considered should be negligible.

B. Echo-to-Noise Ratio

In more definite terms, let us consider the following system characteristics, taking up later the effects of varying some of the parameters.

Wavelength	$\lambda = 10$ meters
Radiated power	$P_T = 250$ kw
Pulse length	$\tau > 7$ seconds
Antenna power gain (isotropic)	$G = 3{,}000$
Receiver bandwidth	$B = 500$ cps
Post-detector integration time	$t = 10$ seconds
Echo cross section of sun	$\sigma = 1.5\pi R_0^2$
Solar radius (optical)	$R_0 = 6.95 \times 10^8$ meters
Solar distance (mean)	$S = 1.49 \times 10^{11}$ meters.

Then the echo power intercepted by the receiving antenna (available power) would be

$$P_R = \frac{P_T G_T}{4\pi S^2} \sigma \frac{1}{4\pi S^2} \frac{G_R \lambda^2}{4\pi}$$

$$= 5.2 \times 10^{-16} \text{ w.}$$

Therefore, the echo-to-noise ratio before the detector becomes

$$\frac{P_R}{P_N} = 2.4$$

$$= +3.7 \text{ db.}$$

The possibility of observing echoes whose strength is near the noise level is limited by the statistical fluctuations of the noise. These fluctuations can be reduced by integrating after the detector over a period of time containing many noise impulses. By integration over n noise pulses, the minimum detectable signal $P_{\min}$ can be reduced by a factor $\sqrt{n}$. Thus for an IF bandwidth of 500 cps and an integration time of 10 seconds, half of which is used to receive echo energy, the detectability of weak echoes will be increased by a factor

$$M = \left(\frac{Bt}{2}\right)^{1/2}$$

$$= 50$$

$$= 17 \text{ db.}$$

Thus under the conditions considered, the echo power would exceed minimum detectable echo by about 21 db.

C. Choice of Optimum Frequency

A large number of factors enter into the choice of the optimum frequency, but some of these can be combined into a single curve of sensitivity against frequency.

As the frequency increases, the echo cross section of the sun decreases, for two reasons. The radius of the coronal level from which $\mu = 0$ reflection can be expected, decreases; also, the absorption of energy in traveling

[13] C. A. Shain, "Galactic radiation at 18.3 mc," *Aust. Jour. Sci. Res., Ser. A.*, vol. 4, pp. 258–267; September, 1951.

[14] J. G. Bolton, Division of Radiophysics, CSIRO (private communication).

to the reflecting level and out again increases rapidly. Both these effects can be calculated from Smerd's data.[11]

The two components of the noise background, solar and galactic, vary with frequency. The apparent temperature of the quiet sun is found to increase slowly over the frequency range concerned, and, in addition, for an antenna of constant area the radiation flux for a given apparent temperature increases with the square of the frequency. The galactic noise level varies in the opposite direction, the power accepted by the antenna being approximately proportional to $f^{-2.5}$ in the "cold" parts of the sky. For an antenna of the size which has a gain of 3,000 at 30 mc, which was considered above, solar noise would be predominant above 30 mc, and galactic noise below 30 mc.

Again, as the frequency increases, the Doppler spread due to solar rotation increases, though this is partially compensated by the decrease in apparent size of the sun. An increased Doppler spread requires a greater bandwidth in order to accept the same proportion of the echoed energy. This leads to a reduction in sensitivity by increasing the noise.

Coming to equipment factors, it is reasonable to consider the effect of frequency variation in terms of an antenna of constant area. The power sensitivity of a radar system in such a case is proportional to f^2. The difficulty and cost of producing a given transmitter power varies little up to 30 mc, but then increases steeply with frequency. A rough estimate can be made of the relative amounts of power that can be produced at various frequencies for a given cost.

TABLE I

EFFECT OF FREQUENCY ON VARIOUS FACTORS WHICH INFLUENCE THE SYSTEM SENSITIVITY ON SUN ECHOES, RELATIVE TO THE SENSITIVITY AT 30 MC

Frequency (mc)	Transmitter antenna gain (db)	Transmitter power for given cost (db)	Sun effective area (db)	Sun absorption (db)	Noise (db)	Doppler spread (db)	Relative system sensitivity (db)
10	−9.5	0	+3.4	+3.6	−9.3	+1.5	−10
20	−3.5	0	+1.2	+2.1	−2.2	+0.6	− 2
30	0	0	0	0	0	0	0
40	+2.5	− 5	−0.8	−2.6	−0.6	−0.4	− 7
60	+6.0	− 8	−1.7	−9.0	−4.0	−1.1	−18
80	+8.5	−12	−2.3	−11.5	−6.7	−1.5	−26

The effect of frequency variation on all the above factors is given in Table I. The figures in some of the columns can be accepted as reliable, but others are somewhat speculative. The over-all effect is plotted in Fig. 3, in terms of the performance at 30 mc., which has been considered in detail. Two curves are given, the dotted line for constant transmitter power and the full line a rough estimate for a transmitter of constant cost. The curves indicate that, under the assumed conditions, there is an optimum sensitivity for sun echoes at 25 to 30 mc.

Three additional factors must be considered, namely, ionospheric transmission, atmospherics, and station interference. The frequency used must be sufficiently high to penetrate the ionosphere without difficulty. These factors set a minimum frequency of about 25 mc. At the same time, the interest in searching for echoes from clouds or streams of electrons near the sun and in interplanetary space suggests the use of the lowest suitable frequency. The over-all conclusion then is that 25 to 30 mc appears to be the optimum frequency range for sun echo work.

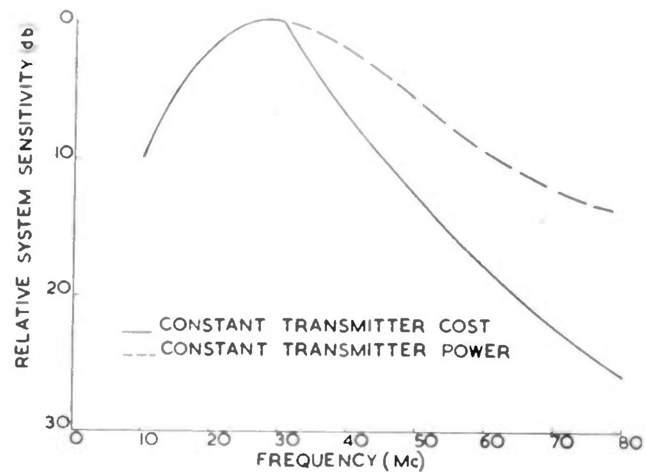

Fig. 3—Variation with frequency of sensitivity for sun echoes, relative to 30 mc, for constant antenna gain.

D. Effect of Changing Other System Constants

The increase of echo-to-noise ratio corresponding to an increase of antenna gain G will be less than the normal G^2 because the noise from the sun will rise at the same time. Since the solar noise comprises only a portion of the antenna noise, while the increase of transmitting gain will be fully effective, the curve of system sensitivity against gain will be between the G and G^2 curves, as shown in Fig. 4.

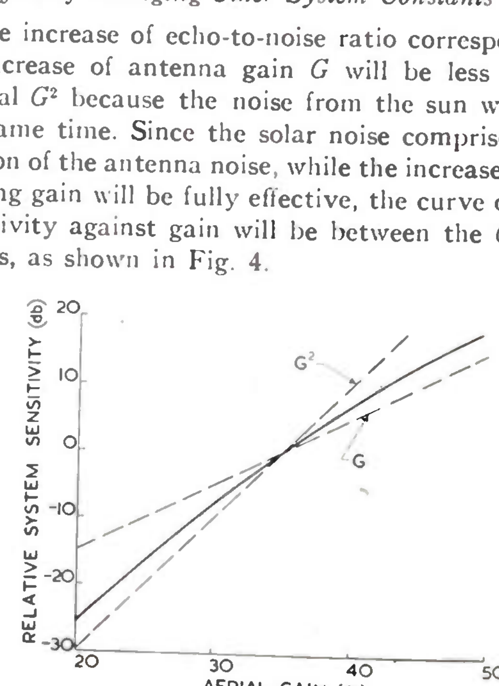

Fig. 4—Variation with antenna gain of 30-mc sensitivity with sun echoes, relative to gain = 3,000.

Variation of bandwidth (B) affects the proportion of the reflected energy which is accepted by the receiver (β), the noise background, and the integration gain. The number of noise pulses contained in the integration period varies as $B^{1/2}$ for a given integration time, so that the sensitivity is proportional to $\beta/B^{1/2}$. This quantity is found to increase with B up to about 400 cps as the effective area of the sun increases. It is then approximately constant up to 1,000 cps, eventually decreasing when the bandwidth exceeds the Doppler spread of the reflected energy.

The sensitivity increases as the square root of the integration time, t, provided the pulse length exceeds $(t+7)$ seconds. (7 seconds is the echo delay time corresponding to the sun's 30-mc radius.)

V. Echoes from Planets

Echo intensities to be expected from planets can be calculated quite simply from the standard radar equation if it is assumed that absorption does not take place in possible planetary ionospheres. Although Venus, Mars, and Mercury come closer to the earth than the sun at times, they are much smaller, and, therefore, should produce weaker echoes. To give an idea of the order of magnitudes, a number of figures are given in Table II for the estimated echo strength relative to the minimum detectable echo, under various conditions.

TABLE II

Examples of 30-mc Echo Intensities from Planets, Using the Equipment Characteristics of Subsection IV B

Planet	Distance at closest approach (meters)	Diameter (meters)	IF bandwidth (cps)	Integration time (seconds)	30-mc echo strength relative to minimum detectable echo (db)
Venus	4.1×10^{10}	12.2×10^{6}	500	10	+2
			50	100	+12
			6	270	+18
Mercury	9.2×10^{10}	4.9×10^{6}	50	615	−6
			1	615	+2
Mars	7.8×10^{10}	6.8×10^{6}	96	520	−3
Jupiter	63×10^{10}	143×10^{6}	4,800	4,200	−16

These are for a 30-mc radar system with the same characteristics as were used in the sun-echo discussion. The reflection coefficient of the surface has been taken as 0.15 in each case, and the directivity as 5. These values were used in considering reflection from the moon,[5] and can be used as a first approximation for the planets. Post-detector integration for the times noted is assumed. The last figure quoted for each planet gives the greatest possible sensitivity for that particular planet, using the minimum bandwidth (as limited by the planet's velocity of rotation) and assuming the energy can be integrated over the whole travel time corresponding to the planet's distance.

The variation with frequency is quite different for planets than that derived for the sun, chiefly because the reflection coefficient is not expected to vanish as the frequency goes up, and also it would not now be necessary to override solar noise. The use of a much higher frequency is favored in this case, the choice depending largely on equipment.

VI. Practicability

A. Transmitter and Antenna

We will now consider the practicability of realizing the equipment parameters which have been found to be necessary for the detection of sun echoes. The critical items are of course the transmitter and the antenna. A power of 250 kw has been introduced above as a reasonable maximum on present techniques, leading to a requirement for antenna gain of 3,000. At 30 mc, this corresponds to an aperture of 200 to 250 meters square and a beamwidth of 3°. Such an antenna could be built in the form of an array of dipoles or as an array of unit antennas of intermediate size. It would require facilities for directing the beam towards the sun for some fraction, at least, of the day and of the year.

A system of this kind is within the range of present techniques, but it is obvious that a project of considerable magnitude is involved.

B. Receiving and Integrating Equipment

On the receiving side, the sensitivity requirement is approximately that of the best receiver practice today. In achieving high sensitivity, use would have to be made of integrating techniques, together with the Dicke comparison system, to reduce the effect of variations in the solar noise background and in receiver gain. The most sensitive arrangement would be to switch rapidly between two slightly displaced frequencies, one containing the sun echo together with solar and galactic noise, the other solar and galactic noise only. A displacement of a few kc from the echo frequency would get outside the sun-echo Doppler spectrum, and produce a negligible change in the noise and in antenna characteristics. By this method it should be possible to observe sun echoes in the presence of relatively large solar noise variations.

VII. Some Possible Astronomical Investigations

By far, the most interesting object for study with a high-power astronomical radar would be the sun. The mere obtaining of echoes on the threshold of visibility would not have much scientific value, as little could be learned from such echoes. It would be essential to build equipment capable of giving echoes with about 20 db in hand so that the fine structure of the echo could be studied. Since antenna resolution would be poor, this study would be limited to the fine structure in depth, unless one could detect coronal streamers or outgoing corpuscular streams subtending a large angle. The chances of detecting such objects would be greatest when near the sun's limb, for then they would present the greatest projected area.

In addition to obtaining some information on the shape of the corona and its time variations, data on rotational velocities and internal motions in the corona could be derived from the Doppler spectrum of the reflected energy. Also, the distance to the reflecting region of the corona might be measured, with about the accuracy obtained in the best optical determinations of the distance of the sun from the earth.

Planetary echoes would have less interest. The rotation period of Venus, which is not at present known, might be determined, and approximate reflection coefficients of the various planets obtained. Accurate distance measurements could, however, only be made with a further large increase in equipment capabilities beyond those considered in this paper. The moon's distance, incidentally, could be determined more precisely than it is yet known from astronomical measurements, with such a powerful equipment. This would be a measurement to the nearest point of the moon, and its main interest would be as a determination of the tidal bulge which very probably exists in the region permanently turned towards the earth.

A Novel Type of Monoscope*

S. T. SMITH†, ASSOCIATE MEMBER, IRE

Summary—**A cathode-ray monoscope tube has been developed to provide output closely resembling the directivity pattern of a radar antenna. The variation of secondary electron emission from aluminum with angle of incidence is used as the basis for design of the monoscope target. The conical aluminum target is so shaped that variation of the target current with beam deflection corresponds to the variation of received radar signal with angular displacement of the radar antenna.**

Introduction

In a radar application at the Naval Research Laboratory an electronic function generator was required that would provide electrical output signals proportional to an over-all radar antenna directivity function in both the horizontal and vertical planes. Input voltages to the device were to be proportional to the angular displacement of the antenna in azimuth and elevation. An accuracy of ±8 per cent was required as well as reliability of operation and no critical adjustments.

A cathode-ray tube with an electron gun is well suited for this application. The electron beam deflected horizontally or vertically resembles a radar antenna beam deflected in azimuth or elevation, and the electron target thus corresponds to the radar target. The problem, therefore, is one of designing the target for the electron beam so that the current to the target will vary with the position of the beam on the target in the same manner as the amplitude of received radar echoes varies when the antenna is directed towards a radar target.

Design of the Monoscope

Shown in Fig. 1 is the desired output characteristic of the monoscope. It is the antenna response curve of combined radar transmitter and receiver and represents the relative voltage amplitude of return echoes as measured in the receiver output when the antenna is directed to either side of a small radar target.

The net current to any electrode bombarded by an electron beam is the difference between the secondary escape current and the beam current. The electrode current, or in this case, the target current to the monoscope, can be caused to vary with position of the beam by controlling the effective secondary escape current at each point on the target surface. After several preliminary experiments it was concluded that the most practical way of controlling the secondary escape current at each point on the target surface was to shape the target in such a way as to control the angle of incidence of the electron beam and thus vary the secondary emission ratio.

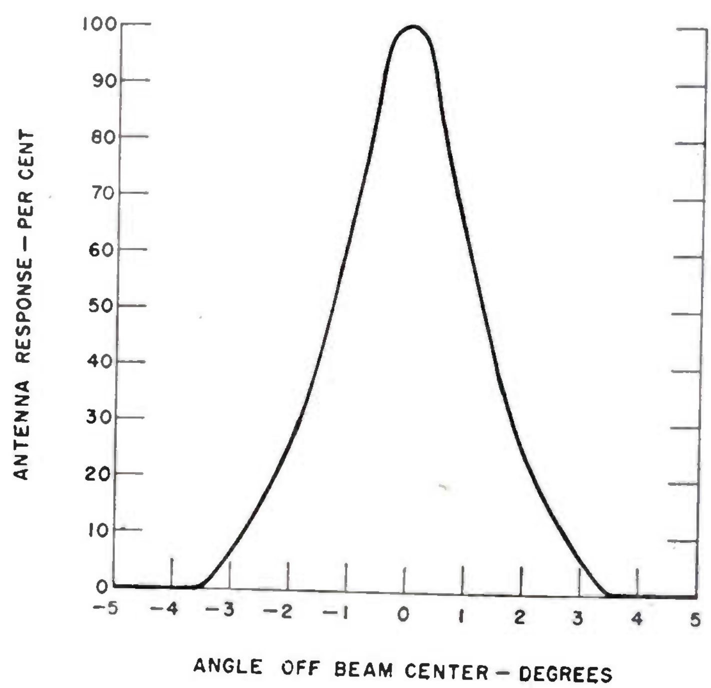

Fig. 1—Antenna response curve.

Müller[1] and others[2] have shown that the secondary electron emission of metals increases as the angle of incidence of the primary beam approaches grazing incidence. Müller's measurements indicate that the relative change in secondary emission with angle of inci-

* Decimal classification: R138.1. Original manuscript received by the Institute, March 20, 1951; revised manuscript received, January 2, 1952.

† Naval Research Laboratory, Washington, D. C.

[1] H. O. Müller, "Die Abhängigkeit der Sekundärelektronenemission einiger Metalle von Einfallswinkel des Primären Kathodenstrahls," *Zeit. für Phys.*, vol. 104, pp. 475–486; March, 1937.

[2] R. Kollath, "Sekundärelektronenemission fester Körper," *Phys. Zeit.*, vol. 38, pp. 202–224; April, 1937.

dence is greatest for metals of least density and that the relative change is approximately constant for all bombarding energies to 4,000 volts. The relative change in secondary emission for metals is little affected by temperature. Of the metals that were studied by Müller aluminum produces the greatest relative change in secondary emission. The curve of Fig. 2 is based on measurements made by Müller using an aluminum target at a bombarding energy of 2,500 volts.

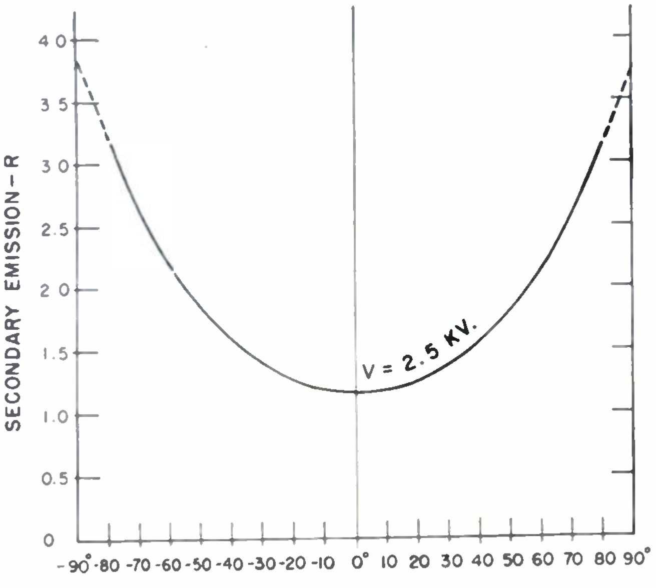

Fig. 2—Variation of secondary-emission ratio with angle of incidence for aluminum (from Müller).

Explanations of the increase in secondary emission with angle of incidence are based on the assumption that at larger angles of incidence the path of the primary electrons through the metal is nearer the surface than at normal incidence. Therefore, since the mean distance to the surface is shorter, more of the secondaries liberated within the metal are able to escape. The process is made more complicated by surface contamination and by dispersion of the primary electron beam as it proceeds through the metal. Assuming an exponential absorption with distance to the surface, Bruining[3] has shown that the variation of secondary-emission ratio is given quite closely by

$$R_\phi = R_0 e^{p(1-\cos\phi)},$$

where R_ϕ is the secondary-emission ratio at angle of incidence ϕ, R_0 is the secondary-emission ratio at normal incidence, and p is a coefficient proportional to the absorption of the material and the mean depth of penetration of the primaries.

Once the variation of secondary emission with angle of incidence is known (Fig. 2), it is a straightforward procedure to design the shape of the target so as to cause the target current to vary with position on the target surface according to the antenna directivity function shown in Fig. 1. The diagram of Fig. 3 illustrates the arrangement of electron gun and target assembly to accomplish the desired results. The target is pointed in the center to make the angle of incidence, the secondary-emission ratio, and the target current a maximum value.

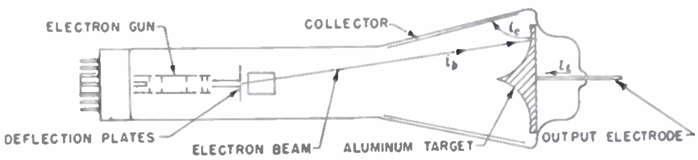

Fig. 3—Diagram of the angle of incidence monoscope.

At progressively larger distances from the center the target is shaped to cause the target current to decrease according to the function shown in Fig. 1.

Near the edges of the target it was required that the target current be zero. This was accomplished by shaping the target for normal incidence of the beam and by lowering the voltage on the collector electrode so that the secondary current was just equal to the beam current. In determining the shape of the rest of the target it was assumed that the same fraction of secondaries was collected at all points on the target surface.

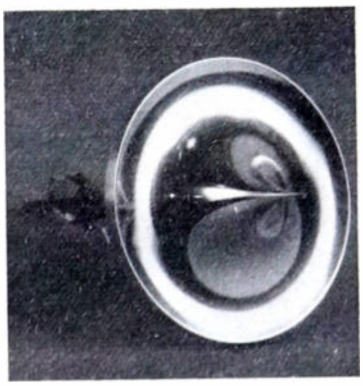

Fig. 4—The angle of incidence target.

After the curve of target shape was obtained by using point-by-point calculations, a template was made and the targets were formed on a lathe from commercial aluminum. The active surface of the aluminum was polished and cleaned by standard cleaning processes. Fig. 4 is a photograph of a completed target.

Fig. 5—The angle of incidence monoscope.

A completed tube is illustrated in Fig. 5. It is 18 inches long and employs a type 5CP electron gun. It is

[3] H. Bruining, "Secondary electron emission, part II, absorption of secondary electrons," *Physica*, vol. V, pp. 901–912; December, 1938.

believed that by proper design the over-all size of the tube could be reduced considerably. Surrounding the target is a glass shoulder on which phosphor has been deposited for observation and focusing of the electron beam.

The monoscope characteristic was indicated on an oscilloscope by connecting the output of a sweep generator to the horizontal deflection plates of both the monoscope and the oscilloscope, while the amplified output of the monoscope target electrode was applied to the vertical deflection plates of the oscilloscope. The beam current of the monoscope was pulsed at a high rate so that the output could be amplified without the use of a dc amplifier. The output characteristic of the monoscope (Fig. 6) was taken in this manner and represents the envelope of the target current pulses as the electron beam is deflected diametrically across the target surface. Fig. 7 represents the output envelope with a high collecting field. In this case all secondary electrons from the target are collected.

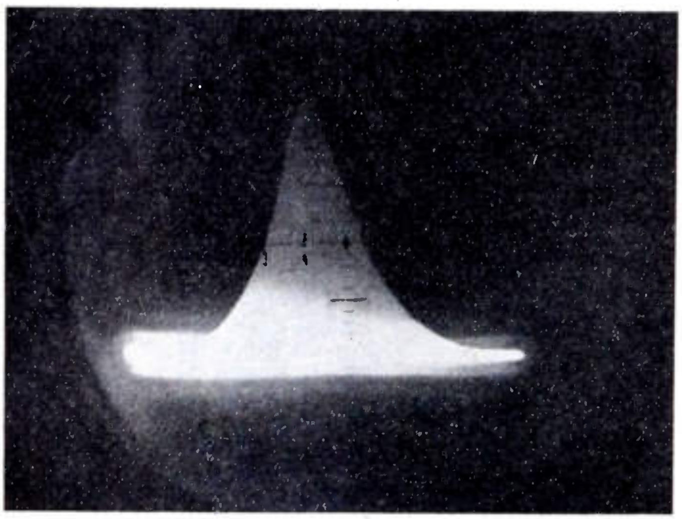

Fig. 6—Output characteristic with low collecting field.

A comparison of the output characteristics shown in Fig. 6 with the desired antenna characteristic shown in Fig. 1 indicates that the agreement is within ± 5 per cent.

Since the target is made from commercial aluminum stock with possible impurities, and since a surface layer of aluminum oxide is present, it might be suspected that the secondary-emission characteristics of the surface would change from tube to tube and with age, but such does not appear to be the case. Six targets have been made from stock selected at random, and each exhibited secondary-emission characteristics with angle of incidence similar to those obtained by Müller. One tube has been operated intermittently for three years with no noticeable adverse effects. There has been no evidence of contamination of the target surface by evaporation of barium from the cathode. However, no quantitative measurements of tube life have been taken, and no conclusive statements can be made.

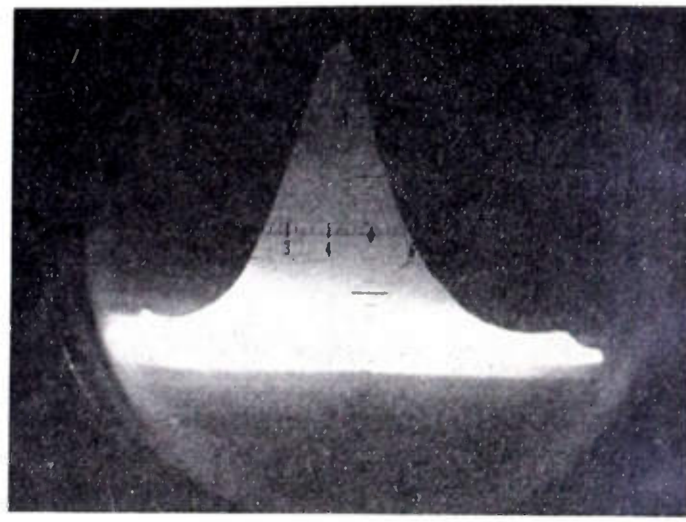

Fig. 7—Output characteristic with high collecting field.

It is interesting to compare the magnitude of output current with that of the conventional television monoscope. In the television monoscope the secondary emission ratio is controlled by varying the amount of carbon deposited on aluminum foil by a half-tone printing process. Thus at one-kv energy the secondary-emission ratio can be varied between the limits of approximately 0.5 for carbon areas to approximately 2 for aluminum areas, or a difference of 1.5. In the case of the angle of incidence target, when all secondaries are collected, the secondary-emission ratio can be controlled between the limits of 3.8 and 1.2, a difference of 2.6. Therefore, for the same beam current the angle of incidence target provides more output than the carbon-on-aluminum target.

Conclusions

The variation of secondary-emission ratio with angle of incidence for aluminum at 2,500-volts energy is a characteristic sufficiently reliable to be used as the basis for the design of a monoscope. The stability of operation of experimental monoscopes favors the use of the angle of incidence effect for special monoscopes and similar applications.

Acknowledgment

Dr. A. V. Haeff and P. L. Sheretz of the Naval Research Laboratory contributed basic ideas for the Target Simulator Monoscope, and M. A. Chambers, D. E. Rudert, and W. M. Campbell were largely responsible for the construction and processing of the tubes. Their co-operation and assistance are very much appreciated.

RF Phase Control in Pulsed Magnetrons*

E. E. DAVID, JR.†, ASSOCIATE, I.R.E.

Summary—This paper describes the behavior of a magnetron oscillator started in the presence of an externally applied rf exciting signal whose frequency is not greatly different from the unperturbed steady-state frequency of the magnetron.

Two points of view are presented: First, quasi-steady-state starting is assumed, and a differential equation representing the system within this limitation is derived. Solutions obtained specify the phase of the oscillator as a function of time subsequent to starting. Second, the inhomogeneous Van der Pol equation is used to describe the system. The oscillator is represented as a parallel RLC circuit shunted by a negative, nonlinear conductance. Approximate analytical and differential analyzer solutions of this equation are used to investigate the frequency and phase transients during starting and distortion of the build-up envelope by the exciting signal. The initial conditions for both equations are established in terms of the exciting signal-to-preoscillation noise ratio.

Results of the two analyses are essentially in agreement; details of the solutions are different since reactive beam loading has been neglected in the latter case. The phase transient initiated during starting may have a duration which is long compared to the build-up time of the rf voltage on the magnetron anode. Preoscillation conditions which fix the initial phase play an important part in determining the phase for a considerable time subsequent to starting. Other influencing factors include the oscillator tuning relative to the injected frequency; the value of a coefficient directly related to the injected power and the natural frequency of the oscillator cavity, and inversely to the power output and coupling Q; and design parameters determining the rf rise time.

Introduction

In some applications of electronic oscillators it is desirable to establish coherence between the generated oscillations and another signal whose frequency is not greatly different. It is often proposed to achieve coherence by injecting energy from this source directly into the oscillator circuit. If the injected signal satisfies certain amplitude and frequency conditions, nonlinear characteristics of the oscillator cause it to synchronize and, therefore, become coherent with the injected signal.

Several investigators have examined this phenomenon both experimentally and theoretically.[1-7] Their results have shown that the power output, frequency, and phase of an oscillator may be uniquely determined if a locking signal of known power, frequency, and phase is injected, and if the oscillator's operation is known uniquely as a function of its load impedance when its dc conditions are specified.

This important conclusion may be deduced immediately from the approximate differential equation of the system,

$$\frac{d\phi}{dt} + S \sin \phi = (\omega_1 - \omega'), \tag{1}$$

where S is a parameter directly related to the locking power and inversely related to the oscillator output power and Q, ω' is the unperturbed oscillator frequency, ω_1 is the perturbing frequency, and ϕ is the phase *difference* between the injected signal and the generated oscillation. If a stable lock-in occurs, $d\phi/dt$ will be zero and the steady-state solution to the equation is merely

$$\phi = \sin^{-1}\left(\frac{\omega_1 - \omega'}{S}\right). \tag{2}$$

Thus the locking phase, ϕ, is determined by locking signal power, frequency, and the steady-state properties of the oscillator itself. The complete solution to (1) also shows that the transient "pull-in" of the phase to its steady-state value (2)[8] when the locking signal is suddenly applied to the steady-state oscillator is not instantaneous, but occurs approximately exponentially with a time constant of $1/S$.

When $(\omega_1 - \omega'/S) > 1$, the solution to (1) shows an entirely different character. Specifically, synchronization is not accomplished and $d\phi/dt$ varies periodically with time instead of settling monotonically to zero. This periodic variation of ϕ corresponds to a frequency modulation of the oscillator which may be calculated in detail from (1).

Thus the effects of an injected rf signal on a steady-state oscillator are well known. Many applications, however, require that the oscillator operate intermittently with a very short duty cycle. The most important example of this kind is the microwave magnetron. It is particularly adaptable to applications where rf pulses of very short duration are needed. In such cases coherence from pulse to pulse may be a desirable condition. This can be achieved if a cw signal of similar frequency is used to phase or synchronize the oscillator during its operation. Hence, we would like to know the dynamics of the synchronization during and subsequent to rf build-up of the oscillator, when a phasing signal

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† Bell Telephone Laboratories, Murray Hill, N. J.

1 E. E. David, Jr., "Locking Phenomena in Microwave Oscillators," Technical Report No. 63, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; 1948.

2 B. Van der Pol, "Forced oscillations in a circuit with non-linear resistance," *Phil. Mag.*, vol. 7-3, p. 65; 1927.

3 R. Adler, "A study of locking phenomena in oscillators," Proc. I.R.E., vol. 34, pp. 351–357; 1946.

4 R. D. Huntoon and A. Weiss, "Synchronization of oscillations," Proc. I.R.E., vol. 35, pp. 1415–1423; 1947.

5 E. V. Appleton, "The automatic synchronization of triode oscillators," *Proc. Camb. Phil. Soc.*, vol. 21, pp. 231–242; 1922–23.

6 E. E. David, Jr., "RF Phase Control in Microwave Oscillators," Technical Report No. 100, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; 1949.

7 J. C. Slater, "Microwave Electronics," McGraw-Hill Book Co., Inc., New York, N. Y.; 1950.

8 See Appendix II for derivation.

has been previously applied for an arbitrarily long time. Eventually, the oscillator will assume the steady-state indicated by (2). The steady-state theory, however, is not adequate to describe the pulsed case. In particular, the effects of the starting transient must be accounted for.

Our objective shall be to find a differential equation which describes the system. The solution to such an equation will specify the locking phase as a function of time over the interval of interest. This solution for large values of time should reduce to the one previously discussed. A second but equally important objective will be to set up the initial conditions for the equation in terms of the oscillator's preoscillation noise. Finally, we shall endeavor to establish the most important factors determining pulse-to-pulse coherence.

Two different but complementary approaches to the problem have been made: (1) The starting of the magnetron is considered a quasi-steady-state process. In this case, the electronic conditions during starting are given by relations determined empirically from measurements on the steady-state magnetron. (2) The magnetron is approximated by a parallel LRC circuit shunted by a negative, nonlinear conductance. The voltage across such an arrangement may be described by a differential equation, one form of which is known as Van der Pol's equation. When such an equation is set equal to a sinusoidal driving term, it specifies the perturbed operation of the oscillator. The electronic conditions are determined by observations of magnetron rf envelopes during starting. Before presenting these analyses, let us consider the preoscillation state of the magnetron and derive the initial condition which is applicable to either equation to be discussed.

I. Preoscillation Conditions in the Magnetron

The initial conditions on the starting equation may be established if we examine the state of the oscillator at the first instant of starting. Upon application of a step-function voltage to the plate of a magnetron, an electronic charge immediately begins to fill the interception space. Initially, the coupling between the incoherent space charge and the resonant structure is small. The increase of this coupling, with time, is accompanied by build-up of noise voltage on the anode. When the necessary conditions are established, the rf voltage begins to build up. These oscillations spring from the voltage already present on the plate, in this case, the preoscillation noise. If there is an external sinusoidal signal impressed upon the magnetron, the initial oscillations start from the vector sum of this and the noise voltage. If the character of the preoscillation noise can be deduced, this sum may be used to establish the required initial conditions.

During the preoscillation time of the magnetron, the incoherent space charge is inducing noise voltage on the anode. The spectrum of this voltage has an amplitude distribution that is determined by the sinusoidal cavity response. The situation is analogous to the case of a tuned circuit being driven by a current whose frequency spectrum is much wider than the bandwidth of the tank. The voltage-response waveform to such a disturbance is practically independent of the exciting waveform, and is almost completely defined by the tuned-circuit selectivity and sensitivity. In our case, a wide-band noise current is driving a high-Q resonant structure. Certain conclusions, therefore, may be drawn as to the character of the resulting voltage on the magnetron vanes.

First, the noise-voltage bandwidth is that determined by the cavity Q. Approximately, therefore, the voltage is sinusoidal over any few rf cycles. The average frequency of these sinusoids over a large number of rf cycles is the cavity resonant frequency; however, this signal is frequency-modulated in a random manner as indicated by the noise sidebands. Over a large number of cycles, then, its phase is likewise random in the interval 0 to 2π. Similarly, the noise envelope varies slowly in a random manner. It can be shown that the probability distribution of this envelope is approximately[9]

$$P(e_n)de_n = \frac{2e_n}{\overline{e^2_n}} e^{-e^2_n/\overline{e^2_n}} de_n, \tag{3}$$

where e_n is the noise-envelope amplitude and $\overline{e^2_n}$ is the mean-squared noise-envelope amplitude. The discussion above is equivalent to the statement that the fine-grained noise structure is sinusoidal while the coarse-grained structure is random.

It is seen, therefore, that when no external voltage is impressed, the starting phase is completely random. In the presence of an external signal, the starting phase may be determined statistically from the relative signal and noise powers.

II. Statistical Properties of the Phase of a Sine Wave Plus Random Noise

It has been postulated that the magnetron preoscillation noise approximates a sine wave of random phase and statistical amplitude over a small number of rf cycles. We are interested in the phase of the resultant magnetron voltage during a very short interval at the beginning of the starting period. This voltage, therefore, may be found by considering the sum of two vectors: one of constant phase and fixed amplitude, the other of random phase and statistical amplitude. The statistical phase of the resultant may be deduced therefrom.

Now in an actual case of starting, the "mean-noise frequency" may differ appreciably from the synchronizing frequency. Fortunately, this difference is small in percentage and, since the noise phase is random, does not affect the validity of our representation. The randomness fixes the phase during any short period, and

[9] S. Goldman, "Frequency Analysis, Modulation and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., p. 336; 1948.

the phase shift due to the frequency difference is negligible during this interval.

Consider, then, the vector relationship shown in Fig. 1, where N is the noise-voltage amplitude of phase ξ, C is the locking signal amplitude, and A is the resultant of phase ϕ. Now A may be expressed as

$$Ae^{j\phi} = C + N\cos\xi + jN\sin\xi, \tag{4}$$

whose phase is

$$\phi = \tan^{-1}\left(\frac{N\sin\xi}{C + N\cos\xi}\right) = \tan^{-1}\left(\frac{R\sin\xi}{1 + R\cos\xi}\right), \tag{5}$$

where $R = N/C$. Let us first find the statistical properties of ϕ from those of ξ, considering R merely as a parameter. The probability of R may be superimposed on this solution to obtain the desired result. Equation (5) shows that the initial part of our solution falls into two cases: Case I, $0 \leqq R \leqq 1$; and Case II, $1 < R < \infty$.

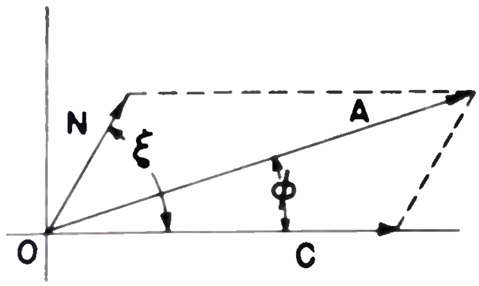

Fig. 1—Vector relationship of ϕ to the noise angle ξ.

Case I

If $R<1$, ϕ always lies in either the first or fourth quadrant for any value of ξ. More specifically, ϕ is a double-valued function of ξ, as seen from (5). For any finite sector, $d\phi$, there are two corresponding sectors $d\xi_1$ and $d\xi_2$. This is shown in Fig. 2. The probability of ϕ

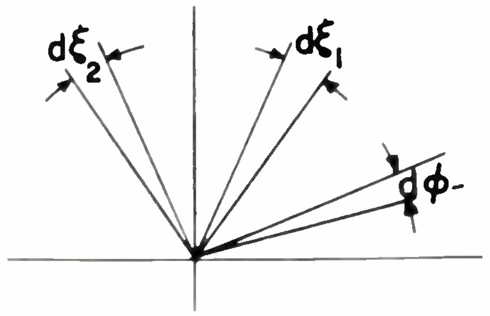

Fig. 2—Finite sector $d\phi$ with its corresponding sectors $d\xi$ and ξ_2.

lying in the interval $d\phi$ is the sum of the probabilities of ξ lying in the intervals $d\xi_1$ and $d\xi_2$. That is,

$$P(\phi, R)d\phi = P(\xi)[\,|d\xi_1| + |d\xi_2|\,], \tag{6}$$

where $P(\xi) = 1/2\pi$. Differentiating (5) with respect to ξ, we obtain

$$P(\phi, R)d\phi = \frac{1}{2\pi}\left[\left|\frac{1 + R^2 + 2R\cos\xi_1}{R\cos\xi_1 + R^2}\right| + \left|\frac{1 + R^2 + 2R\cos\xi_2}{R\cos\xi_2 + R^2}\right|\right]d\phi. \tag{7}$$

In order to have $P(\phi, R)$ as a function of ϕ and R only, it is necessary to find an expression for $R\cos\xi$. By employing a little trigonometric manipulation, (5) will yield

$$R\cos\xi = -\sin^2\phi \pm \cos\phi\sqrt{R^2 - \sin^2\phi}. \tag{8}$$

Combining (7) and (8), there results, after simplification,

$$P(\phi, R)d\phi = \frac{1}{2\pi}\left[\left|1 + \frac{\cos\phi}{\sqrt{R^2 - \sin^2\phi}}\right| + \left|1 - \frac{\cos\phi}{\sqrt{R^2 - \sin^2\phi}}\right|\right]d\phi. \tag{9}$$

This is the required expression.

Equation (5) yields another useful result. If $d\phi/d\xi$ is equated to zero, one obtains

$$\cos\xi = -R, \tag{10}$$

which, when substituted into (5), yields

$$\tan(\phi_{\max}) = \frac{R}{\sqrt{1 - R^2}} \tag{11}$$

or

$$\sin(\phi_{\max}) = R.$$

That is, for any ratio of signal-to-noise greater than 1, ϕ has a maximum value given by (11). Conversely, for any ϕ, the minimum allowable R is given by (11). Therefore, the validity conditions on (9) may be stated in two ways, both of which must be satisfied simultaneously:

$$\begin{gathered} \sin\phi \geqq R \geqq 1 \\ 0 \geqq \phi \geqq \sin^{-1} R. \end{gathered} \tag{12}$$

Case II

When $R>1$, ϕ becomes a single-valued function of ξ. Therefore, the probability of ϕ lying in the interval $d\phi$ is the same as the probability of ξ being in $d\xi$.

$$P(\phi, R)d\phi = P(\xi)d\xi. \tag{13}$$

This expression may be evaluated in a manner exactly analogous to that used in Case I. There results

$$P(\phi, R)d\phi = \frac{1}{2\pi}\left[1 + \frac{\cos\phi}{\sqrt{R^2 - \sin^2\phi}}\right]d\phi. \tag{14}$$

The validity conditions on this expression are simply

$$\begin{gathered} 1 < R < \infty \\ 0 \leqq \phi \leqq 2\pi. \end{gathered} \tag{15}$$

The probability densities, $P(\phi, R)$, for Cases I and II are shown in Fig. 3. These curves are plotted for positive values of ϕ only since they are even functions of that variable. Note that the probability of ϕ falling in the interval $\Delta\phi$ is

$$\int_{\phi}^{\phi+\Delta\phi} P(\phi, R)d\phi,$$

an interval which proves to be finite regardless of the infinite portions of the densities for $R \leqq 1$. In fact, it is easily shown that

$$\int_{0}^{\sin^{-1} R} P(\phi, R)d\phi = 1/2 \quad \text{for} \quad R \leqq 1$$

and

$$\int_{0}^{\pi} P(\phi, R)d\phi = 1/2 \quad \text{for} \quad R > 1,$$

a result which merely states that for any value R, ϕ will certainly lie in the range given by (12) or (15).

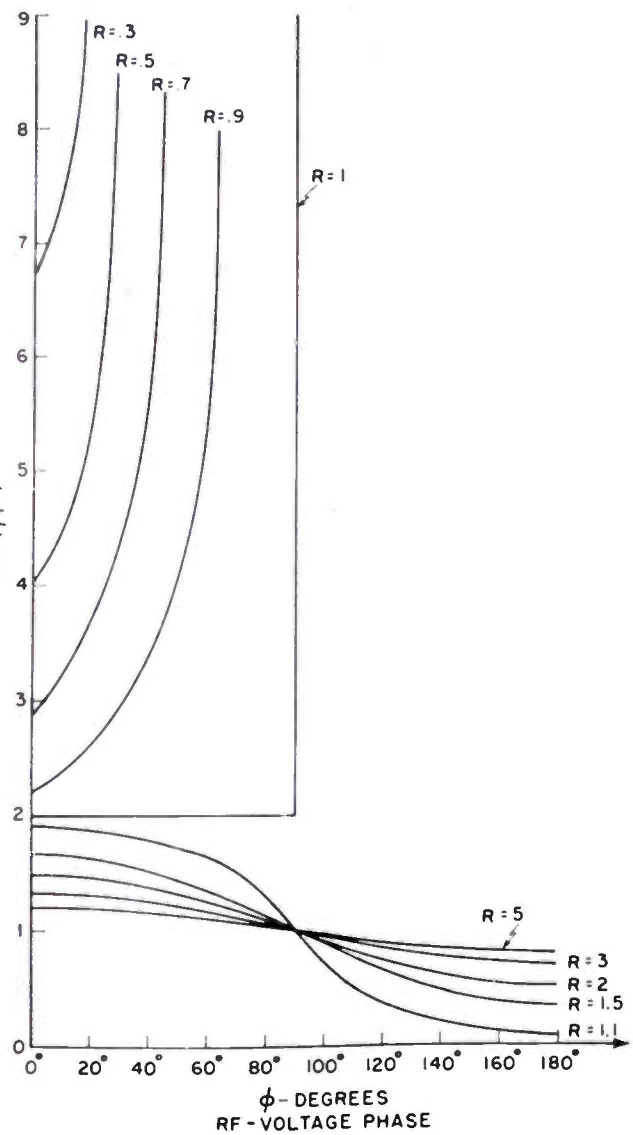

Fig. 3—Probability density of rf voltage phase for constant values of noise amplitude.

It is now necessary to superimpose the noise-envelope distribution on our present solution. More specifically, this distribution may be written as a function of the variable R by consulting (3).

$$P(R)dR = 2R/\overline{R^2}e^{-R^2/\overline{R^2}}dR, \tag{16}$$

where $\overline{R^2}$ is the noise-to-signal power ratio. Now it may be seen that the probability of ϕ is the sum of all the $P(\phi, R)$ for all admissible R, multiplied by the weighting factor $P(R)$. Therefore, in general,

$$P(\phi) = \int_{0}^{\infty} P(\phi, R)P(R)dR. \tag{17}$$

This integral may be evaluated by substituting (9), (14), and (16), letting $y = R^2 - \sin 2\phi$, and noting that

$$\int_{0}^{y} e^{-y^2/\overline{R^2}}dy = 2\sqrt{\frac{\overline{R^2}}{\pi}} \operatorname{erf}\left(\frac{y}{\sqrt{\overline{R^2}}}\right),$$

where erf $(y/\sqrt{\overline{R^2}})$ is the well-known error function. The expression for the probability density is then

$$\begin{aligned} P(\phi) = 1/\pi\overline{R^2}\Big\{ \Big[&\overline{R^2}/2(e^{-\sin^2\phi/\overline{R^2}} - e^{-1/\overline{R^2}}) \\ &+ \sqrt{\pi\overline{R^2}}/2 \cos\phi e^{-\sin^2\phi/\overline{R^2}} \operatorname{erf}(\cos\phi/\sqrt{\overline{R^2}}) \Big] \\ &+ \Big[\overline{R^2}/2(e^{-\sin^2\phi/\overline{R^2}} - e^{-1/\overline{R^2}}) \\ &- \sqrt{\pi\overline{R^2}}/2 \cos\phi e^{-\sin^2\phi/\overline{R^2}} \operatorname{erf}(\cos\phi/\sqrt{\overline{R^2}}) \Big] \\ &+ \overline{R^2}/2e^{-1/\overline{R^2}} + \sqrt{\pi\overline{R^2}}/2 \cos\phi e^{-\sin^2\phi/\overline{R^2}} \\ &\cdot [1 - \operatorname{erf}(1/\sqrt{\overline{R^2}})]\Big\} \end{aligned} \tag{18}$$

for $0 \leqq \phi \leqq \pi/2$, where the limitation on ϕ may be deduced from (12). For $\pi/2 \leqq \phi \leqq \pi$, we have

$$\begin{aligned} P(\phi) = 1/\pi\overline{R^2}\Big\{&\overline{R^2}/2e^{-1/\overline{R^2}} + \sqrt{\pi\overline{R^2}}/2 \cos\phi e^{-\sin^2\phi/\overline{R^2}} \\ &\cdot [1 - \operatorname{erf} 1/\sqrt{\overline{R^2}}]\Big\}. \end{aligned} \tag{19}$$

Fig. 4 shows these probability densities plotted from (18) and (19) for various values of noise-to-signal ratio.

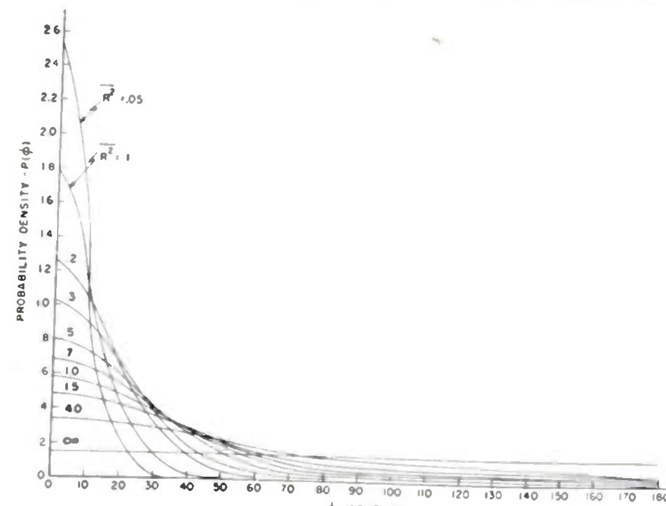

Fig. 4—Probability density of phase of sine wave plus noise with noise-to-signal ratio as a parameter.

Fig. 5 shows

$$P(0, \phi) = \int_0^{\phi} P(\phi) d\phi,$$

the probability of the initial phase falling in interval $0-\phi$. These curves were obtained from those of Fig. 4 by graphical integration and are, of course, also even functions of ϕ. We have established, therefore, the probability of initial phase as a function of noise-to-signal ratio. With this information available, we will now proceed to set up and solve the differential equations for magnetron starting in the presence of an externally applied sinusoidal signal.

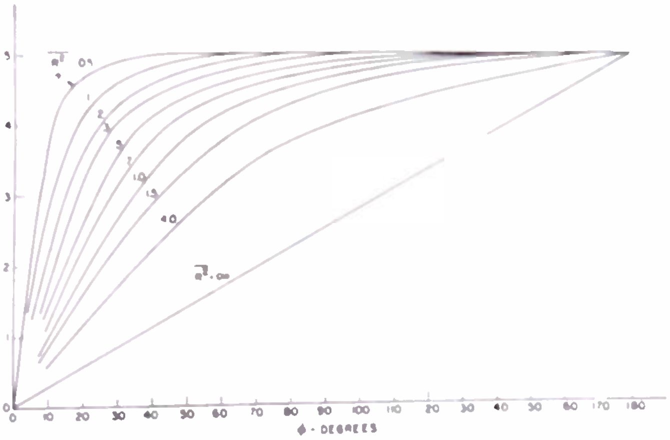

Fig. 5—The probability $P(0, \phi) = \int_0^{\phi} P(\phi)d\phi$ with noise-to-signal ratio as a parameter.

III. Unperturbed Starting and the Accompanying Variations in Instantaneous Frequency

It has been found experimentally[10] that the electronic behavior of the steady-state magnetron may be described approximately by the two relations

$$g = \frac{E/R}{V_{rf}} - \frac{1}{R} \tag{20}$$

$$b = \beta - g \tan \alpha, \tag{21}$$

where E, R, β, and α are constants which change with dc conditions, and V_{rf} is the amplitude of the rf voltage. If the starting of a magnetron is considered as a quasi-steady-state process, that is, a succession of steady-states, then (20) and (21), together with the oscillator operating equations, may be used to write a differential equation of starting. Such an analysis is reproduced in Appendix I. The result indicates that the rf envelope during starting may be represented as

$$v_{rf} = V_{RF_0}(1 - e^{-kt}), \tag{22}$$

where V_{RF_0}, the steady-state value of v_{rf}, and k, the reciprocal of the build-up time constant, are evaluated as functions of the design and steady-state properties of the magnetron. It has been found experimentally that at small rf voltage (22) is not a good approximation of true conditions. During the beginning of the build-up, the rf voltage increases exponentially, a condition impossible unless g is constant. Equation (20), therefore, is valid only after the initial instant of starting. If a locking signal is present, however, quite a large rf voltage may already exist at this first instant. Under these conditions, it is probable that the build-up follows (22) closely, even at the beginning. Of course, if the locking signal is quite small, the exponential behavior will doubtless be present. In either case, it is desirable to rewrite (22) as

$$v_{rf} = V_{RF_0}(1 - \eta e^{-kt}), \tag{23}$$

where η is a constant whose value lies between 0.7 and 1. Equation (23) states that at $t=0$, the time at which the build-up has progressed far enough for (20) to be valid, the rf voltage is finite and has the value

$$v_{rf_i} = (1 - \eta)V_{RF_0}. \tag{24}$$

Subsequent to this time, the build-up continues exactly as expressed by (22). Actual observations on magnetron rf build-up show η to be 0.8 or greater in most cases. The experimental evidences show the above theory to be a good approximation so long as the rf load is not badly mismatched.[11]

Equation (23) expresses the form of the rf build-up envelope; however, nothing has been said about the frequency and phase during the transient period. This information is readily obtained by use of (20) and (21):

$$g = \frac{E/R}{V_{rf}} - \frac{1}{R} = \frac{E/R}{V_{RF_0}(1 - \eta e^{-kt})} - \frac{1}{R}$$

or

$$g = g_0 + \frac{E\eta}{RV_{RF_0}} \frac{e^{-kt}}{1 - \eta e^{-kt}}, \tag{25}$$

where g_0 is the steady-state electronic conductance. Then from (21),

$$b = b_0 - \frac{E\eta}{RV_{RF_0}} \tan \alpha \left(\frac{e^{-kt}}{1 - \eta e^{-kt}}\right), \tag{26}$$

where b_0 is the steady-state electronic susceptance. Using the magnetron operating equation, as derived in Appendix I, we have

$$2\left(\frac{\omega - \omega'}{\omega_0}\right) = -\frac{E\eta}{\omega_0 CRV_{RF_0}} \tan \alpha \left(\frac{e^{-kt}}{1 - \eta e^{-kt}}\right), \tag{27}$$

where $\omega' = 2b_0/C - \omega_0 B/2Q_{ext} + \omega_0$ = steady-state operating frequency with zero-locking signal.

Further simplification yields

$$\omega = \omega' - \left(\frac{E\eta}{2CRV_{RF_0}} \tan \alpha\right)\left(\frac{e^{-kt}}{1 - \eta e^{-kt}}\right). \tag{28}$$

[10] J. C. Slater, "Microwave electronics," *Rev. Mod. Phys.*, vol. 18, pp. 489–495; 1946.

[11] W. Rotman, "A Study of Transient Phenomena in Magnetrons," unpublished notes, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; March 18, 1948.

This result shows that the frequency of the magnetron during the initial part of the build-up may be remote from the steady-state value ω'. Rotman's measurements on steady-state and transient $g-V_{rf}$ relations, when used to evaluate (28) for $t=0$, indicate that the frequency difference may be as much as 20 mc. Further experimental evidence is necessary, however, before an absolute evaluation is made.

IV. The Quasi-Steady-State Starting Equation

With this understanding of magnetron starting, we may now derive an equation expressing the phase of the magnetron during build-up when an external, sinusoidal signal is impressed. The synchronizing signal is considered small enough so that the fundamental nature of the build-up is not disturbed. It is shown in Appendix II that the load susceptance of a synchronized oscillator is approximately

$$B' = B - 2|\rho|\sin\phi,$$

where B is the passive susceptance, ϕ is the locking phase, and

$$|\rho| = |Y_{LP\rho_0}| = |Y_{LP}|\left|\frac{V_1}{V_i}\right|,$$

where Y_{LP} is the passive load admittance and V_1/V_i is the ratio of locking voltage to magnetron voltage at the magnetron reference plane. During starting, then, the reflection factor becomes

$$|\rho| = |Y_{LP}|\left|\frac{V_1}{V_{RF_0}(1-\eta e^{-kt})}\right|$$
$$= |Y_{LP}|\left|\frac{V_1}{V_{RF_0}}\right|\left(\frac{1}{1-\eta e^{-kt}}\right). \tag{29}$$

This expression is a good approximation so long as V_1/V_{RF_0} is considerably less than the quantity $(1-\eta)$. Now V_1 is the total voltage at the reference plane because of the locking signal. This voltage includes that caused by the wave incident on the magnetron cavity and that reflected from it. During the build-up, therefore, V_1 may be also a function of time. The nature of the variation depends on the locking frequency, the steady-state oscillator load, and the variation of electronic admittance with time. An exact evaluation of this effect is difficult and, for our purposes, unnecessary, since we are interested in qualitative rather than quantitative results at the moment. It is easy to see, however, how changes in V_1 will affect $|\rho|$ as a function of time. If V_1 is considered constant, (29) shows that $|\rho|$ is large at $t=0$ and decreases exponentially to its steady-state value as t increases. Actually, V_1 may start at some small value and increase with time, so that the variation shown by (29) is exaggerated in magnitude. That is, the initial value of $|\rho|$ is not as large as indicated, but it may, nevertheless, exceed the steady-state $|\rho|$ considerably. An assumption of constant V_1, therefore, will not change the fundamental nature of the solution although it will exaggerate the effect of the locking signal. In our solution, (29) will be used to represent the reflection factor as a function of time. We should expect, then, that synchronization will be indicated earlier in time than is actually the case. In this respect, the solution will be an optimistic one.

With this approximation in mind, we may write down the differential equation of starting. Again using (27) and substituting the expression for B', we have

$$2\left(\frac{\omega-\omega'}{\omega_0}\right)+\frac{E\eta}{\omega_0 CRV_{RF_0}}\left(\frac{e^{-kt}}{1-\eta e^{-kt}}\right)$$
$$-2\frac{|\rho_0|}{Q_{\text{ext}}}\left(\frac{\sin\phi}{1-\eta e^{-kt}}\right)=0, \tag{30}$$

where $|\rho_0|$ is the steady-state reflection factor. Now if ω_1 is the frequency of the synchronizing signal,

$$\frac{d\phi}{dt}=\omega_1-\omega. \tag{31}$$

If this is substituted into (30), there results

$$\frac{d\phi}{}+\left(\frac{S}{-\eta e^{-kt}}\right)\sin\phi = M+\frac{Ne^{-kt}}{1-\eta e^{-kt}}, \tag{32}$$

where $S=\omega_0|\rho_0|Q_{\text{ext}}$, $M=\omega_1-\omega'$, and $N=E\eta/2CRV_{RF_0}$ $\tan\alpha$. This is the differential equation of starting which was desired. We need only solve the equation in order to find the phase ϕ as a function of time with the steady-state properties of the system as parameters.

The steady-state portion of the solution is the same as that discussed in the Introduction, for if t is allowed to approach infinity, (32) becomes identical with (1).

Unfortunately, (32) is nonlinear in the dependent variable ϕ, so that an explicit analytical solution is not to be expected. The nature of the solution may be deduced by considering the equation for small values of ϕ only, so that $\sin\phi$ may be replaced by ϕ. Such a study is made in the following section. If more information is desired, machine methods are available which give a complete solution for particular values of the coefficients. Results of this type are discussed in section VI.

V. An Approximate Solution for Small Locking Angles

An analytical expression for the phase as a function of time may be obtained from (32), provided we are interested in the solution for small ϕ only. If this is the case, $\sin\phi$ may be replaced by ϕ, and (32) becomes an ordinary linear differential equation. It may be solved by use of an integrating factor $e^{\int P dt}$, where P is the coefficient of the term in ϕ. In this case

$$e^{\int P dt} = e^{St}(1-\eta e^{-kt})^{S/k}.$$

If the linear equation is multiplied by this factor and integrated, there results

$$e^{St}(1-\eta e^{-kt})^{S/k}\phi = \int \left[M e^{St}(1-\eta e^{-kt})^{S/k} + N e^{(S-k)t}(1-\eta e^{-kt})^{(S/k)-1}\right] dt. \quad (33)$$

The indicated integrations are not difficult to carry out if the factors $(1-\eta e^{-kt})^{S/k}$ and $(1-\eta e^{-kt})^{(S/k)-1}$ are expanded in series by the binomial expansion and integrated term by term. Then ϕ may be written

$$\phi = \frac{1}{(1-\eta e^{-kt})^{S/k}}\left[\frac{M}{S} + M\sum_{m=1}^{\infty}(-1)^m \frac{\frac{S}{k}\left(\frac{S}{k}-1\right)\cdots\left(\frac{S}{k}+1-m\right)}{km!\left(\frac{S}{k}-m\right)}\eta^m e^{-mkt} + \frac{N}{k\left(\frac{S}{k}-1\right)}e^{-kt} + N\sum_{m=1}^{\infty}(-1)^m\frac{\left(\frac{S}{k}-1\right)\left(\frac{S}{k}-2\right)\cdots\left(\frac{S}{k}-m\right)}{km!\left(\frac{S}{k}-m-1\right)}\eta^m e^{-mkt} + \phi' e^{-St}\right], \quad (34)$$

where ϕ' is a constant of integration. If the factor

$$(1-\eta e^{-kt})^{S/k}$$

is expanded in a series and divided into the bracketed series, and ϕ' evaluated in terms of the initial angle $\phi_0 (\phi = \phi_0$ when $t=0)$, there results

$$\phi = \frac{1}{(1-\eta e^{-kt})^{S/k}}\left\{\frac{M}{S} + \frac{N-\eta M}{k\left(\frac{S}{k}-1\right)}e^{-kt} + \left[(1-\eta)^{S/k}\phi_0 - \frac{M}{S} - \frac{N-\eta M}{k\left(\frac{S}{k}-1\right)} - (1-\eta)^{S/k}\left(\frac{N-\eta M}{\eta}\right)\sum_{m=1}^{\infty}(-1)^m\frac{m!\eta^{m+1}}{k\left(\frac{S}{k}-1\right)\cdots\left(\frac{S}{k}-m-1\right)}\right]e^{-St}\right\} + \frac{N-\eta M}{\eta}\sum_{m=1}^{\infty}(-1)^m\frac{m!\eta^{m+1}e^{-(m+1)kt}}{k\left(\frac{S}{k}-1\right)\left(\frac{S}{k}-2\right)\cdots\left(\frac{S}{k}-m-1\right)}. \quad (35)$$

Interpretation of this result is facilitated if the constants involved are evaluated within an order of magnitude. Although sparse experimental evidence is available for this purpose, some information may be obtained from data published by the M.I.T. Radiation Laboratory.[12] The majority of these data describe the operation of a 3,000-mc magnetron. Therefore, the numbers deduced from them are believed representative of such oscillators.

The constant S is defined as $S=\omega_0|\rho_0|/Q_{ext}$. Its value ranges approximately from 10^6 to 5×10^7 for typical S-band magnetrons. Note that it is directly proportional to the square-root of the synchronizing power and inversely related to the product of external Q and square-root of power output. We may say, therefore, that its value indicates in a quantitative manner the effect of the locking signal.

The time-constant of the rf voltage build-up is $1/k$. Direct observations by means of high-speed oscillographs indicate that a value of about 10^{-8} sec is typical. This constant is a function of the oscillator resonant frequency, loaded Q, and electronic characteristics.[13] Now η and N may be related by means of the expressions derived in Appendix I. There it is shown that

$$N = k\eta \tan\alpha = \eta \times 10^{-8}\tan\alpha. \quad (36)$$

The value of $\tan\alpha$ ranges approximately between the limits 0.1 and 1, and, as previously stated, η lies between 0.7 and 1.

Thus it is seen that the constants k, N, and η are intimately related to the electronics of the oscillator, while S and M reflect the character of the locking signal. The latter, therefore, are the primary operating parameters of the system, while the former are related to the design. Of course, this separation is not absolute since all are functions of design and operating conditions, but it does indicate in a qualitative manner the most important factors determining the constants. With

[12] G. B. Collins, "Microwave Magnetrons," Radiation Laboratory Series No. 6, McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

[13] See Appendix I.

these concepts in mind, we may now examine our approximate solution.

In almost all practical cases of synchronization, the frequencies ω_1 and ω' are made as nearly equal as possible, and the phenomenon is used merely as a "phasing" device. Let us first consider $M=\omega_1-\omega'=0$ and also assume that the preoscillation noise is small, so that the initial angle, ϕ_0, is zero. The final phase, of course, will also be zero, since $\phi_{ss}=M/S=0$. Fig. 6 shows the form of the phase transient under these conditions for various values of S. Note that the phase is disturbed violently during the magnetron voltage build-up.

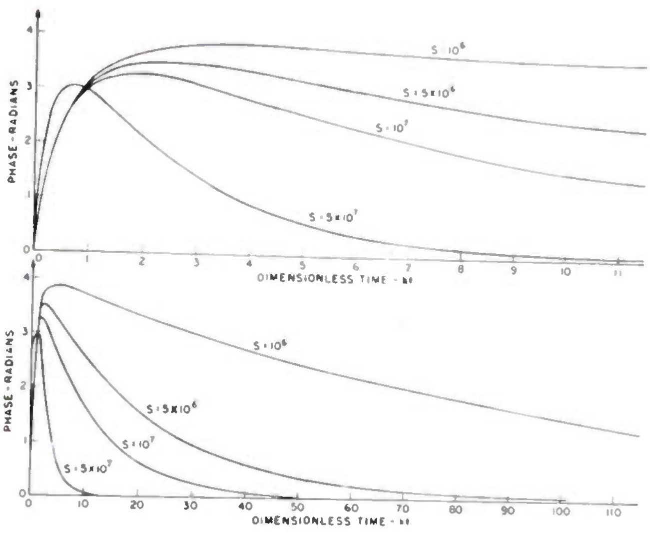

Fig. 6—Phase transient accompanying the starting of a synchronized magnetron with locking power as a parameter for $M=0$, $N=2\times10^7$, $k=10^8$, $\phi_0=0$, and $\eta=0.8$.

This effect results from frequency modulation of the tube during the starting period as described earlier for the unsynchronized case. It is interesting to note that

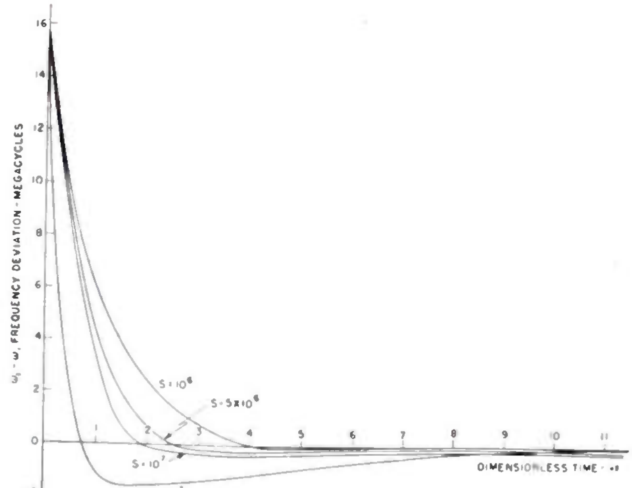

Fig. 7—Frequency transient accompanying the starting of a synchronized magnetron for $M=0$, $N=2\times10^7$, $k=10^8$, $\phi_0=0$, and $\eta=0.8$.

this electronic frequency pushing is present regardless of the locking signal. After this initial perturbation, the phase approaches its steady-state value exponentially with a time constant $1/S$. The values of η and N[14] do not greatly affect the form of the transient; however, the magnitude of the disturbance increases with both, as can be seen by noting that the initial frequency difference is $d\phi/dt|_{t=0}$ and

$$\left.\frac{d\phi}{dt}\right|_{t=0}=\omega_1-\omega\Big|_{t=0}=\frac{N}{1-\eta}. \tag{37}$$

The instantaneous frequency difference, $d\phi/dt$, is shown in Fig. 7 for the same values of S as in Fig. 6. Here the frequency-modulation effect is quite apparent.

Preoscillation noise present in the magnetron will cause the initial angle, ϕ_0, to be different from zero.

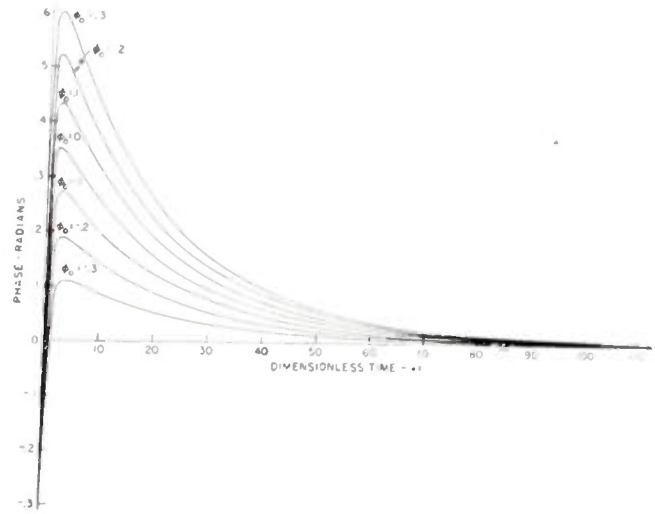

Fig. 8—Effect of initial conditions on phase transient for $M=0$, $N=2\times10^7$, $k=10^8$, $\eta=0.8$, and $S=5\times10^6$.

How this condition affects the phase transient is shown in Fig. 8. Note that since the duration of the transient is much greater than that of the magnetron voltage build-up, the consequences of the preoscillation noise are evident for a comparatively long period. Hence, the state of the oscillator at the first instant of starting is of paramount importance if a high degree of phase coherence from pulse to pulse is desired. Another consideration of equal importance is, of course, that of variations in the dc pulse voltage. The effect of such variations is to make ω' and M, and therefore the steady-state phase, M/S, different from zero. Figs. 9 and 10 show

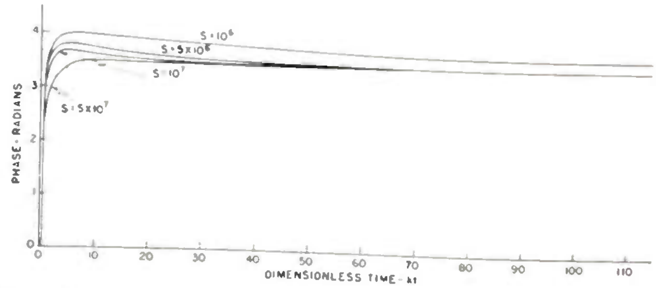

Fig. 9—Phase transient accompanying the starting of a synchronized magnetron with locking power as a parameter for $M/S=0.348$, $N=2\times10^7$, $k=10^8$, $\phi_0=0$, and $\eta=0.8$.

[14] The values $N=2\times10^7$ and $\eta=0.8$ were computed from data on a Raytheon QK-61 magnetron. The corresponding value for S is 5×10^6 if $|\rho|=0.1$.

the phase transient for $M/S = 0.348$ and -0.348, respectively. The actual magnitude of the steady-state incoherence (M/S) resulting from a particular voltage change may be computed exactly from (1).

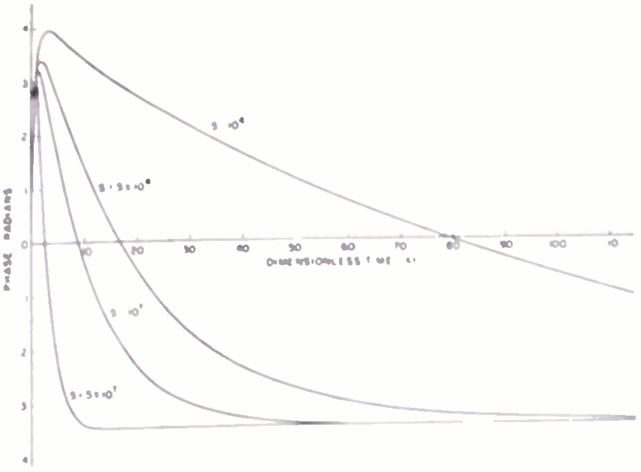

Fig. 10—Phase transient accompanying the starting of a synchronized magnetron with locking power as a parameter for $M/S = -0.348$, $N = 2\times10^7$, $k = 10^8$, $\phi_0 = 0$, and $\eta = 0.8$.

It may be seen quite clearly from Figs. 9 and 10 that the operation of a pulsed synchronized magnetron is different for positive and negative values of the frequency difference, $\omega_1 - \omega'$. This characteristic has been noted very distinctly in experimental circuits.

We have now examined the solution of (32) for small ϕ. A more general solution, obtained by machine methods, will be presented in the following section.

VI. Differential Analyzer Solutions for Large Locking Angles

The use of analogue computers is one of the most powerful methods presently available for obtaining the solution to nonlinear differential equations. An electronic differential analyzer has been designed and constructed by Macnee of the Research Laboratory of Electronics at M.I.T.[15] It is through the use of this machine, and with Macnee's assistance, that the solutions presented in this section were obtained.

Sources of error in these solutions are twofold. First, calibration errors may be as much as 10 per cent. Second, an important part of the transient occurs within a time comparable with the time constant of the analyzer itself. The initial portion of these solutions, therefore, is considerably in error.

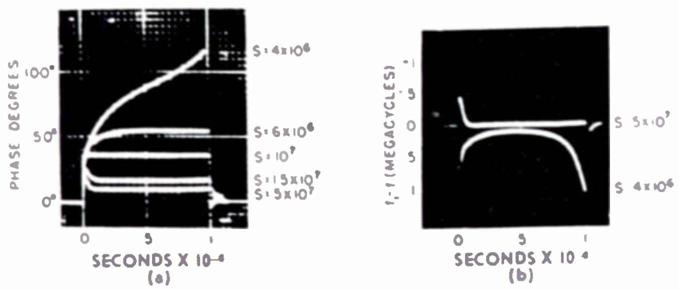

Fig. 11—Phase and frequency for $M = 5$ megaradians, showing the effect of changes in synchronizing power.

[15] A. B. Macnee, "An Electronic Differential Analyzer," Technical Report No. 90, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.: 1949.

Fig. 11(a) shows the phase ϕ as a function of time for values of S when the frequency difference, $M = \omega_1 - \omega'$, is 5 megaradians per second. Note that when the quantity M/S becomes greater than one, there is no longer sufficient synchronizing power to maintain locking and the phase becomes a continuously increasing function. The frequency for this condition, along with that for the locked case, is shown in Fig. 11(b). The periodic variations of frequency are typical of the unsynchronized oscillator when perturbed by an external signal. Figs. 12(a) and 12(b) show phase and frequency for the same conditions, except that $M = \omega_1 - \omega'$ is negative.

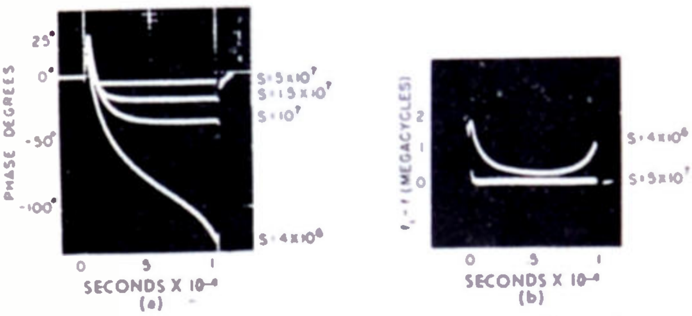

Fig. 12—Phase and frequency for $M = -5$ megaradians, showing the effect of changes in synchronizing power.

How variations in $|\omega_1 - \omega'|$ affect the phase and frequency with S constant is shown in Figs. 13(a) and 13(b). The similarity to Figs. 11 and 12 should be noted.

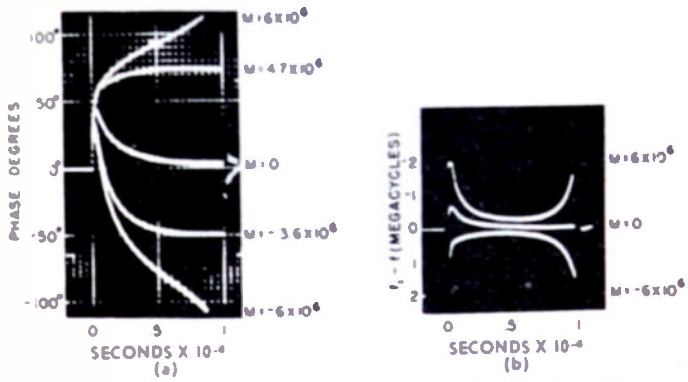

Fig. 13—Phase and frequency for $S = 5\times10^6$, showing effect of changes in frequency difference $\omega_1 - \omega$.

The effect of initial angles $-60°$, $0°$, and $60°$ when M/S is 0.95 is depicted in Figs. 14(a) and 14(b). It is seen that a large positive or negative initial angle con-

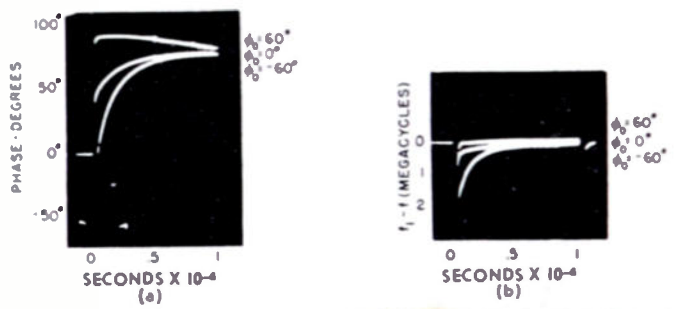

Fig. 14—Phase and frequency for $M/S = 0.95$, showing the effect of changes in the initial condition.

siderably increases the magnitude and duration of the transient. This effect is shown more graphically in Figs. 15 and 16, where $\phi_0 = \pm 180°$. Under such conditions, the transient becomes quite intense and synchronization does not occur until almost a microsecond has elapsed. Since in some applications the pulse length is considerably shorter than the duration of the transient, it is again seen that preoscillation noise can give rise to large incoherencies. Of course, the probability of such conditions may be reduced to insignificance by the use of large synchronizing power.

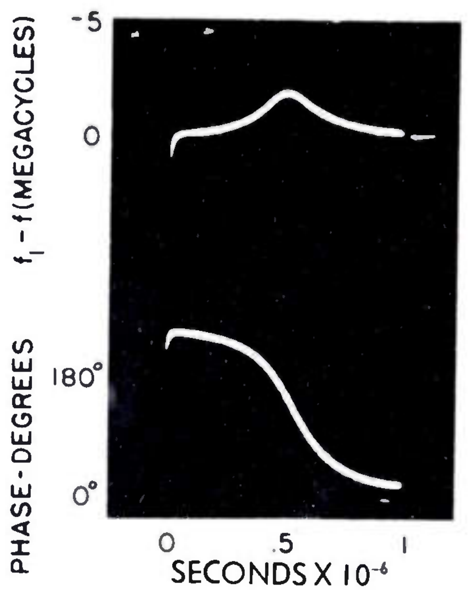

Fig. 15—Phase and frequency for $M/S = 0.4$, showing the effect of large positive initial angle.

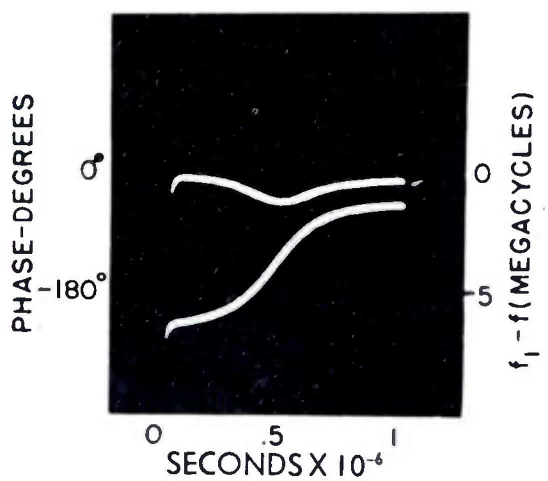

Fig. 16—Phase and frequency for $M/S = 0.4$, showing the effect of large negative initial angle.

Thus the exact solution to the starting equation shows the same general character as does the approximate solution introduced in Section V. Both solutions emphasize the importance of the preoscillation state when pulse-to-pulse coherence is desired. Furthermore, the preoscillation noise-to-locking signal ratio is the important parameter determining this condition. Finally, the other important factors are oscillator tuning relative to the injected frequency, value of the coefficient S, and the values of the design paramenters N and η.

VII. Phase Control as Described by the Van der Pol Equation

A single-mode oscillator may be properly represented by a parallel RLC combination shunted by a negative conductance and susceptance, which are functions of the terminal voltage. The differential equation describing such a circuit is known as Van der Pol's equation, and its solution shows a transient build-up, followed by steady-state sinusoidal oscillations. If the circuit is acted upon by a sinusoidal current source, a driving term is added to the equation. Its solution will now show the same steady-state synchronization behavior that has been discussed previously. The conditions for synchronization can be found as a function of the ratio of the injected current to the oscillation amplitude, and the frequency difference between the oscillator and injected signal. This calculation has been made by Van der Pol,[2] and is in agreement with other literature in the field. The transient conditions existing during the oscillator build-up, however, have never been studied exhaustively. While the general form of the build-up envelope is well known, such things as the instantaneous phase and frequency and the distortion of the envelope by the injected signal have not been examined. The latter two aspects, moreover, are of fundamental interest in our discussion of transient behavior.

In the previous treatment, a shape for the build-up envelope was assumed and the instantaneous phase calculated on a quasi-steady-state basis. The validity of this analysis may be more firmly established if the solution to the Van der Pol equation leads to similar results.

VIII. The Van der Pol Equation with a Driving Current

The form of Van der Pol equation with which this section concerns itself is easily derived from the equivalent circuit shown in Fig. 17. It is to be noted that reactive loading effects of the electronic space charge have been neglected for the sake of simplicity. The presence of this loading serves only to exaggerate the frequency pushing exhibited during the transient build-up. We are justified, therefore, in ignoring this factor so long as we are concerned only with the nature of the solution and not its quantitative details.

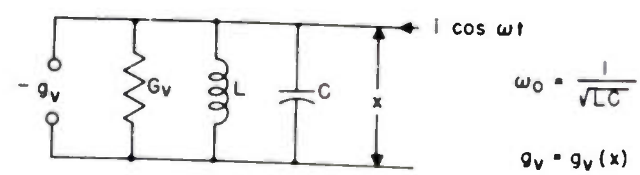

Fig. 17—The single-mode equivalent circuit of an oscillator, neglecting reactive loading by the electronic space charge.

Returning to Fig. 17, if we write the nodal equation for the voltage x and differentiate with respect to time, we find

$$\frac{d^2x}{dt^2} + \mu\left(G_V - g_V - x\frac{dg_V}{dx}\right)\frac{dx}{dt} + \omega_0^2 x = -E\omega^2 \sin \omega t, \quad (38)$$

where $\mu = 1/C$, $\omega_0^2 = 1\sqrt{LC}$, and $E = i/\omega C$. The function relationship, $g_V = g_V(x)$, must be evaluated so that the equation may be reduced to one in x and t only.

IX. Relation of Van der Pol's Equation to Actual Conditions in the Magnetron

From experimental observations, the form of the magnetron rf voltage build-up is well known. The voltage initially shows an exponential growth, a behavior which reveals that the electronic conductance is constant during this period. That such is the case may be seen by considering (1) of Appendix I. This equation may be written as[16]

$$t = \int \frac{dA/A}{\frac{\omega_0}{2}\left(\frac{g}{\omega_0 C} - \frac{1}{Q_L}\right)}$$

If g is constant, the equation integrates directly, giving

$$A = e^{\omega_0/2(g/\omega_0 C - 1/Q_L)t}. \quad (39)$$

Hence, if $g/\omega_0 C > 1/Q_L$, the voltage increases exponentially.

Near the end of the build-up, the voltage "saturates" and approaches its final value asymptotically. It is shown in Appendix I that such behavior is characteristic when the electronic conductance is an inverse function of the voltage.

The region intermediate between these two extremes is one of transition in which the voltage is described equally well by either boundary equation. Hence, we would like to find an expression for g_V which closely approximates a constant at small voltage, decreases at larger amplitudes, and has a smooth transition in between. Fig. 18 shows a possible form of this function as

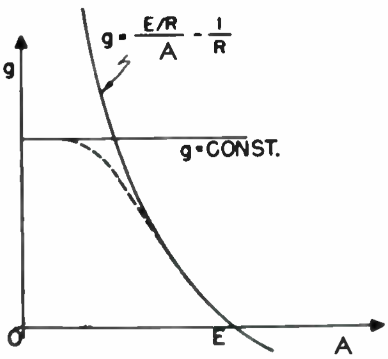

Fig. 18—Dotted curve shows a possible form of $g(A)$ as interpolated from the limiting loci.

interpolated from the limiting conditions. One recognizes that this curve may be closely represented by the square-law function as shown in Fig. 19. There it has been assumed that

$$g = G_V + 1 - \frac{A^2}{4}. \quad (40)$$

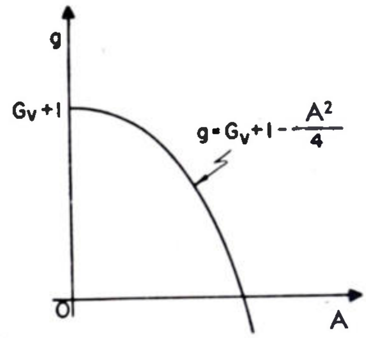

Fig. 19—Square-law approximation to the locus shown in Fig. 18.

As a function of the voltage x, (40) may be written as

$$g_V = G_V + 1 - \frac{x^2}{3}. \quad (41)$$

The details of this transformation will be derived in Section X. Now, substituting (41) into (38), we arrive at

$$\frac{d^2x}{dt^2} + \mu(x^2 - 1)\frac{dx}{dt} + \omega_0^2 x = -E\omega^2 \sin \omega t. \quad (42)$$

This is the form of Van der Pol's equation with which we shall be concerned. Its form shows clearly that it represents a second-order system with nonlinear damping, driven by a sinusoidal forcing function.

X. An Approximate Analytical Solution to the Homogeneous Equation

If the driving current is reduced to zero, (42) becomes

$$\frac{d^2x}{dt^2} + \mu(x^2 - 1)\frac{dx}{dt} + \omega_0^2 x = 0. \quad (43)$$

One may obtain an approximate solution to this homogeneous equation by assuming $x = A \cos(\omega_0 t - \phi)$, where A and ϕ are both functions of time.

Van der Pol's original solutions were made under a similar assumption. His results, however, are not in the form most useful for this analysis. Hence, the purpose of this redundant presentation will be to formulate $A(t)$ and $\phi(t)$ in convenient terms.

When one substitutes the assumed solution into (43), he obtains

$$\left\{\frac{d^2A}{dt^2} - A\left(\omega_0 - \frac{d\phi}{dt}\right) + \mu\left(\frac{3A^2 - 4}{4}\right)\frac{dA}{dt} + A\omega_0^2\right\} \cos(\omega_0 t - \phi)$$

[16] Note that A is the envelope function of x. In other words, we have assumed $x \sim A \cos(\omega t - \phi)$, where $A = A(t)$ and $\phi = \phi(t)$.

$$
-\left\{2\left(\omega-\frac{d\phi}{dt}\right)\frac{dA}{dt}-A\,\frac{d^2\phi}{dt^2}
+\mu\left(\frac{A^3-4A}{4}\right)\left(\omega-\frac{d\phi}{dt}\right)\right\}\sin\,(\omega_0t-\phi)
+\mu\,\frac{A^2}{4}\,\frac{dA}{dt}\cos\,3(\omega_0t-\phi)
-\mu\,\frac{A^3}{4}\left(\omega-\frac{d\phi}{dt}\right)\sin\,3(\omega_0t-\phi)=0. \tag{44}
$$

Now if the coefficients in this equation are slowly varying compared to the sines and cosines, each coefficient must be identically zero for the equation to be true at all instants of time. This condition implies that during the transient build-up, the amplitude and phase do not change appreciably during one rf cycle. For instance, in *S*-band magnetrons, which have typical starting times of about 10^{-7} seconds, containing 300 rf cycles, the assumption is quite well satisfied.

Let us, then, set the coefficient of $\sin\,(\omega_0t-\phi)$ equal to zero and neglect the second derivative, $d^2\phi/dt^2$, by the previous assumption

$$
\frac{dA}{dt}=-\mu\,\frac{A^3-4A}{8}\,. \tag{45}
$$

Equation (44) may be integrated directly after solving for dt. The solution is

$$
A=\frac{2}{\sqrt{1+\dfrac{4-A_0{}^2}{A_0{}^2}\,e^{-\mu t}}}\,, \tag{46}
$$

$$
\omega=\left[\omega_0{}^2+\frac{\mu^2}{2}\left(\frac{4-A_0{}^2}{A_0{}^2}\right)e^{-\mu t}\,\frac{\left\{2-\left(\dfrac{4-A_0{}^2}{A_0{}^2}\right)e^{-\mu t}\right\}}{\left\{1+\left(\dfrac{4-A_0{}^2}{A_0{}^2}\right)e^{-\mu t}\right\}^2}\right]^{1/2}, \tag{50}
$$

where A_0 is the initial amplitude. This result is shown in Fig. 20 for various values of A_0, plotted against the dimensionless time μt. It is to be noted that the initial amplitude, A_0, is the critical factor fixing the time of starting, while μ enters only as a scale factor. Also, so long as the assumption $f_0 \gg \mu$ is valid, the value of ω_0 does not enter.

Consider (41) which expresses g_V as a function of x. Rewriting this equation, we have

$$
g_Vx=G_Vx+x-\frac{x^3}{3}\,. \tag{47}
$$

If the assumed solution for x be substituted, there results

$$
g_VA\cos\,(\omega_0t-\phi)=G_VA\cos\,(\omega_0t-\phi)+A\cos\,(\omega_0t-\phi)
-\frac{A^3}{3}\left[\tfrac{3}{4}\cos\,(\omega_0t-\phi)-\tfrac{1}{4}\cos\,3(\omega_0t-\phi)\right]. \tag{48}
$$

The triple-frequency term may be neglected since the circuit is sharply tuned to the frequency ω_0. Under this assumption g_V reverts to our former g and

$$
g=G_V+1-\frac{A^2}{4}\,. \tag{49}
$$

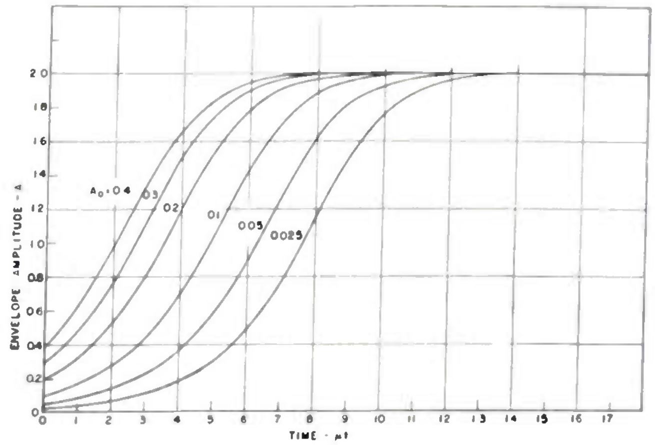

Fig. 20—Approximate envelopes of starting transient.

Equation (46) shows that the steady-state amplitude, A, is equal to 2. Then the steady-state g is just G_V and the total conductance shunting the circuit is (G_V-g) or zero. This result gives a quantitative check on our solution for A and shows the process by which (41) was arrived at from (40).

Returning to (44) and putting the coefficient of the cosine equal to zero while neglecting the factor d^2A/dt^2, we find

where $\omega=\omega_0-d\phi/dt$ and (46) has been substituted. The instantaneous frequency is shown in Fig. 21 for various

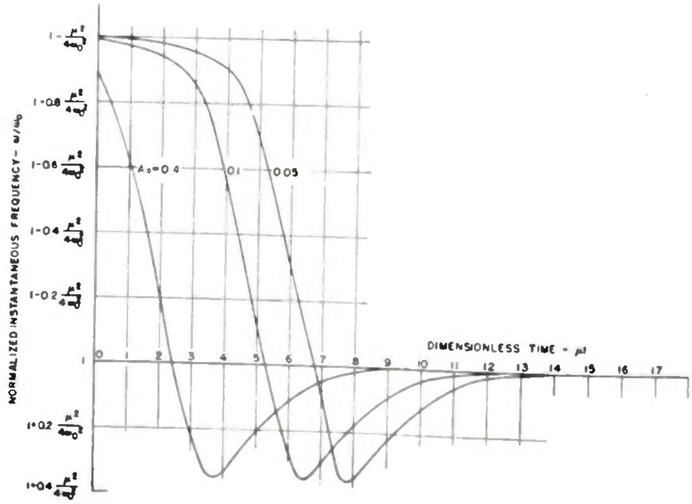

Fig. 21—Instantaneous frequency during starting.

values of A_0, where the approximation $(1+x)^{1/2}=1+x/2$ when $x\ll 1$ has been made in (50). Like the build-up loci of Fig. 20, a change of A_0 merely shifts the time scale and does not alter the form of the curves. Hence, the frequency and build-up for any A_0 may be found from those shown by an appropriate shift on the time axis. It is of interest to note that the quantity $\omega-\omega_0$ changes sign during the transient and that this change is coincident with the inflection point of amplitude build-up.

XI. An Approximate Analytical Solution to the Inhomogeneous Equation

If one attempts to obtain an analytic solution to (42), he is eventually confronted with insoluble nonlinear differential equations in A and ϕ. Some information may be obtained, however, by making the assumption that the driving current is small enough so that our former expression for $A(t)$, with a suitable value of A_0 inserted, is still representative of the build-up envelope. Let us, then, make a solution for the inhomogeneous equation under this assumption.

Now (42) may be written as

$$\frac{d^2x}{dt^2}+\mu(x^2-1)\frac{dx}{dt}+\omega_0^2x$$
$$=-E\omega^2[\sin\phi\cos(\omega t-\phi)+\cos\phi\sin(\omega t-\phi)]. \quad (51)$$

If we assume the solution $x=A\cos(\omega t-\phi)$ and substitute in (51), we find it may be written in a form analogous to (44) in which the coefficient of $\cos(\omega t-\phi)$ is

$$\frac{d^2A}{dt^2}-A\left(\omega-\frac{d\phi}{dt}\right)^2+\mu\left(\frac{3A^2-4}{4}\right)\frac{dA}{dt}$$
$$+A\omega_0^2+E\omega^2\sin\phi.$$

As before, this coefficient may be set equal to zero if it is slowly varying. Neglecting the term d^2A/dt^2, we can obtain the following:

$$\frac{d\phi}{dt}=\omega+\sqrt{\omega^2+\mu\left(\frac{3A^2-4}{4A}\right)\frac{dA}{dt}+(\omega_0^2-\omega^2)+\frac{E\omega^2}{A}\sin\phi}.$$

From the known nature of the solution, $d\phi/dt\ll\omega$; hence we may expand the one-half-power term in a power series and discard all but the first two terms. Then, $d\phi/dt$ becomes

$$\frac{d\phi}{dt}=-\frac{\mu}{2\omega}\frac{(3A^2-4)}{4A}\frac{dA}{dt}-\frac{\omega_0^2-\omega^2}{2\omega}-\frac{E\omega}{2A}\sin\phi. \quad (52)$$

We may substitute (46) for A, and thus reduce our equation to one in ϕ and t alone. We have

$$\frac{d\phi}{dt}+\frac{E\omega}{4}\left[1+\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]^{1/2}\sin\phi=\omega-\omega_0$$
$$-\frac{\mu^2}{4\omega}\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\frac{\left[2-\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]}{\left[1+\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]^2}, \quad (53)$$

where we have used the approximation $(\omega^2-\omega_0^2)/2\omega\approx\omega-\omega_0$.

We should note here that when $e^{-\mu t}$ becomes quite small, (53) reduces to

$$\frac{d\phi}{dt}+\frac{E\omega}{4}\sin\phi=\omega-\omega_0. \quad (54)$$

Equation (54) is the same as (1). Hence, the solution to (53), after the effects of the starting transient die out, is exactly similar to that found previously for the steady state.

Examination shows that (53) is very much like the equation derived for quasi-steady-state build-up (32). Both contain a term expressing the effect of frequency pushing during starting. These terms are the functions of time located on the right-hand side of the equalities, and both show time variations exactly similar to the frequency variations during unsynchronized starting, as expressed by (50) and (28). These variations are different in detail, for we have neglected reactive beam loading in the transient analysis. The important conclusion to be drawn is that both analyses show this frequency disturbance to be present regardless of the locking signal. In addition, both equations show that the effect of the locking signal is enhanced during starting by a factor which is, in magnitude, merely the inverse of the rf voltage envelope.

An analytic solution for (53) may be obtained if one considers the special case, $\omega-\omega_0=0$; (53) then becomes

$$\frac{d\phi}{dt}=-\frac{E\omega}{4}\left[1+\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]^{1/2}\sin\phi$$
$$-\frac{\mu^2}{4\omega_0}\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\frac{\left[2-\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]}{\left[1+\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]^2}. \quad (55)$$

Values of the constants in this equation corresponding to those used previously are approximately as follows:

$$S\to\frac{E\omega}{4}=5\times10^6$$
$$k\to\mu=10^8$$
$$\omega'\to\omega_0=2\pi\times3\times10^9$$
$$1-\eta\to A_0/2=0.2.$$

Now, the second term on the right in (55) has its minimum value at $t=0$ and increases rapidly thereafter, being nearly zero at the end of the starting transient. The minimum value is approximately $-\mu^2/4\omega_0\sim-1.33\times10^6$ [assuming that $(4-A_0^2/A_0^2)\gg1$]. Even if this term remained at its minimum value during the entire starting period, it would correspond to a phase change of only $-1.33\times10^6\times10^{-7}\approx-1°$. Hence, we may safely rewrite (55) as

$$\frac{d\phi}{dt}=-\frac{E\omega_0}{4}\left[1+\left(\frac{4-A_0^2}{A_0^2}\right)e^{-\mu t}\right]^{1/2}\sin\phi. \quad (56)$$

By separating the variables and integrating, we obtain

$$\ln \tan \frac{\phi}{2} = -\frac{E\omega_0}{4} \int \left[1 + \left(\frac{4 - A_0^2}{A_0^2}\right) e^{-\mu t}\right]^{1/2} dt + C. \quad (57)$$

The integral may be evaluated by making the substitution

$$y^2 = 1 + \left(\frac{4 - A_0^2}{A_0^2}\right) e^{-\mu t}.$$

After utilizing the initial condition, $\phi = \phi_0$ at $t = 0$, the final result is

$$\ln \tan \frac{\phi}{2} = \ln \tan \frac{\phi_0}{2} + \frac{E\omega}{2\mu} \left\{ (1 + \chi e^{-\mu t})^{1/2} - (1 + \chi)^{1/2} + \frac{1}{2} \ln \left[\frac{(1 + \chi e^{-\mu t})^{1/2} - 1}{(1 + \chi e^{-\mu t})^{1/2} + 1} \frac{(1 + \chi)^{1/2} + 1}{(1 + \chi)^{1/2} - 1} \right] \right\}, \quad (58)$$

where $\chi = 4 - A_0^2/A_0^2$. The phase, ϕ, is plotted as a function of time in Fig. 22 for two values of χ. It is seen that after the envelope build-up goes to completion, the phase decreases exponentially for small ϕ. The time-constant of this exponential is, as before, the inverse of the coefficient of the sine term in the differential equation. This fact can be confirmed by considering (58) for $e^{-\mu t} \ll 1$.

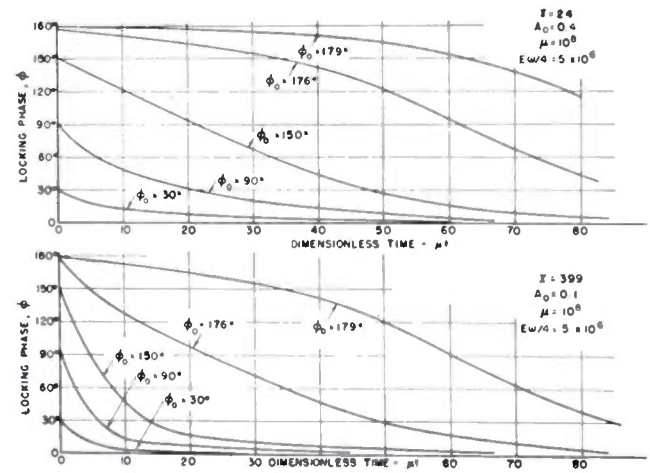

Fig. 22—Phase transient during starting.

For large initial angles, it is noted that the duration of the transient is greatly prolonged. Hence, we may again conclude that the conditions at the initial instant of starting are of prime importance in determining pulse-to-pulse coherence.

Before discussing the "goodness of phasing," we must choose a criterion. Let us define the parameter U, where

$$U = \frac{1}{N_P} \sum_{n=1}^{N_P} \int_0^{\tau_n} [\phi_n(t) - \phi_{ss}]^2 dt. \quad (59)$$

This definition is arbitrary, but reasonable. We have taken the square of the integrated phase deviation from the steady-state value and averaged it over a large number, N_P, of pulses, each of duration τ_n.

Now, U is a function of the static parameters of the oscillator, the ratio of the steady-state amplitude to the locking signal, the initial amplitude A_0, and the ratio of preoscillation noise to synchronizing power. For a given oscillator, the static parameters are important only in relation to other oscillators. Hence, we may exclude this factor in the sequel. The ratio of steady-state to locking signal amplitude determines the time-constant of the phase transient resulting from the phase deviations which exist when the rf envelope *first reaches its final value*. This time-constant decreases with increasing locking power. Equation (58) and Fig. 22 show that phase deviation existing when the envelope has reached its final value *increases* with A_0 for a *given* initial phase deviation. On the other hand, the initial phase deviations *decrease* with increasing locking power. Consequently, the behavior of U as a function of locking power is somewhat complicated and will depend rather strongly on the individual case. We can, however, draw some generalized conclusions.

For a given oscillator, the value of U will, for the most part, decrease with increasing locking power. This decrease will not be uniform, however. It will be most rapid for small locking signals (small A_0) and large locking signals (small initial phase deviations and time-constant). More important is the possibility that U will increase or decrease rather slowly in the region of intermediate locking power. In other words, there can exist a situation in which increase of locking power does not improve and may even impair phasing.

XII. Differential Analyzer Solution to the Driven Van der Pol Equation and Distortion of Build-Up Envelope by Driving Signal

The driven Van der Pol equation (42) may be solved by the electronic differential analyzer. Solutions obtained from this computer can be used to determine the distortion of the build-up envelope by the driving signal. We may examine the solution for given values of the parameters μ, ω_0, ω, and $E\omega/2A_f$. It has been seen previously that when $E\omega/2A_f = 0$ and $f_0 \gg \mu$, the form of the build-up envelope is independent of f_0, and μ enters merely as a scale factor. When $E\omega/2A_f$ becomes finite, but not large, this condition will still be closely valid. For the solutions to be presented, the ratio $f_0/\mu \sim 3.2$; hence, our conclusions as to distortion of the envelope will hold in general, so long as $f_0 \gg \mu$. For these solutions we have chosen the following values:

$$\omega_0 = 2 \times 10^9 \text{ radians/sec}$$

$$\mu = 10^8 \text{ sec}^{-1}$$

$$\omega_0 - \omega = 0,$$

and the synchronizing parameter, $E\omega/2A_f$, is variable over the range 0 to 1.9×10^7, which provides a maximum

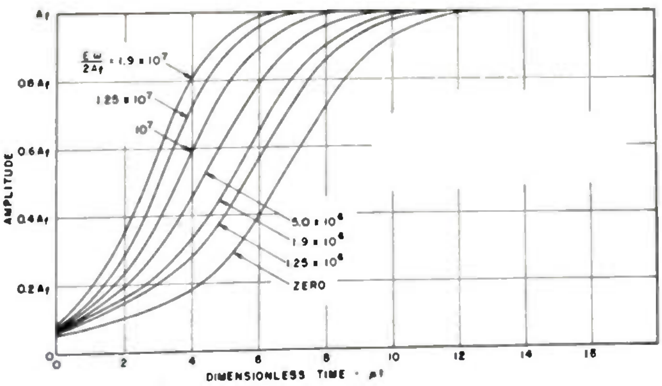

Fig. 23—Distortion of build-up envelope by driving current as a function of the locking parameter.

locking band of 3 mc. Fig. 23 shows build-up envelopes for several values of the parameter. These curves were taken from computer solutions, such as those shown in Fig. 24. Comparison of Figs. 20 and 23 shows that the build-up curves with and without the driving signal are quite similar. Hence, our assumption that distortion of the envelope by the driving signal may be accounted for by utilizing a suitable value of A_0 in (46) seems well justified. Thus, the material in Section XI, which was derived under this assumption, may be considered reliable. Furthermore, the results of Section XI agree well with those found from the quasi-steady-state analysis. Since the solutions are mutually compatible, we may, with confidence, apply the conclusions drawn from them to the actual situation.

Conclusion

The effects of the starting transient of a magnetron on its synchronization by a continuous signal have been calculated. These computations have been carried out from two different points of view, and their results are compatible. Briefly, the phase transient initiated during the starting may have a duration much longer than that of the rf envelope transient. Preoscillation conditions, therefore, which fix the initial phase, play an important part in determining the phase for a considerable time

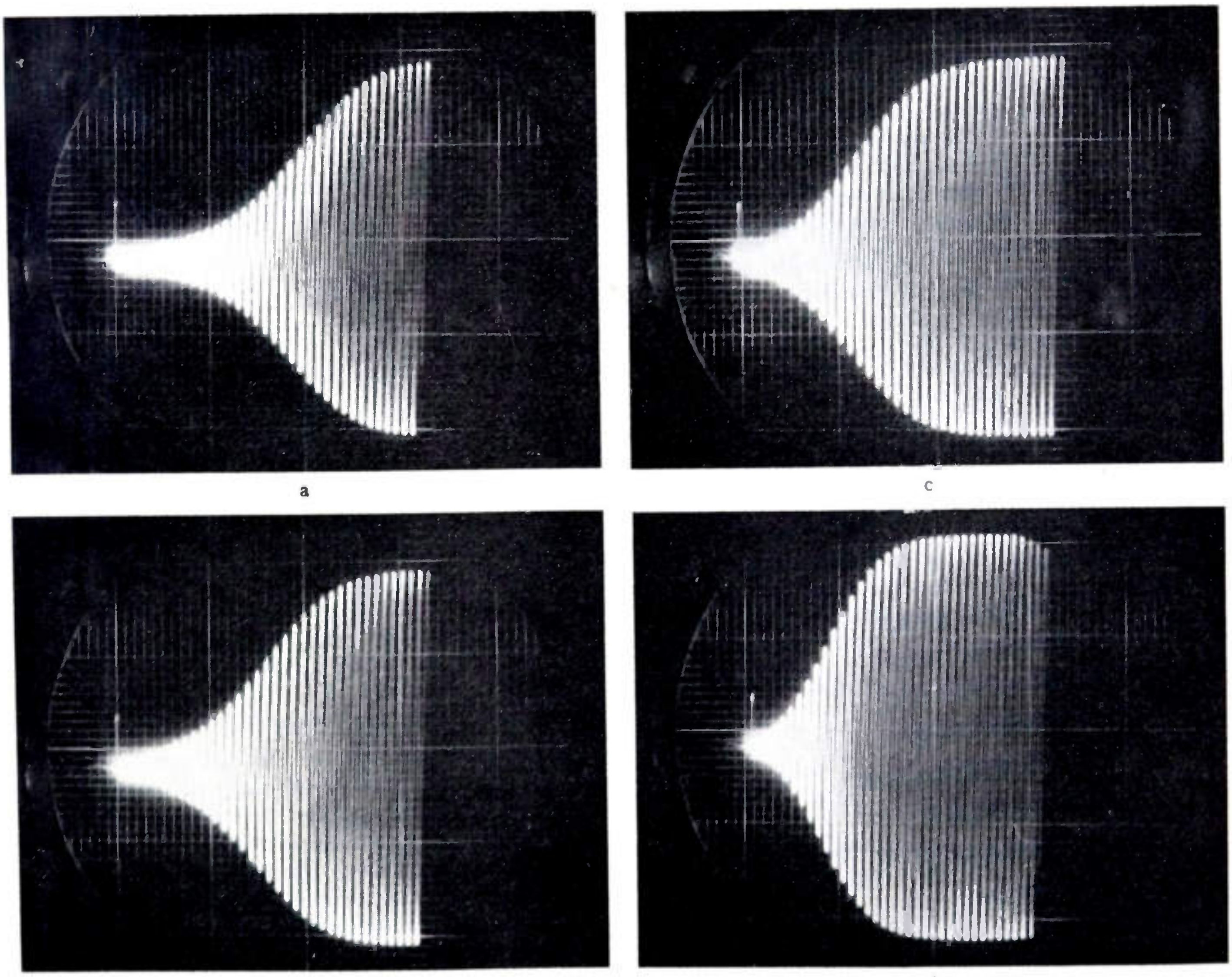

Fig. 24—Differential analysis solutions to the inhomogeneous Van der Pol (42) show distortion of the build-up envelope by the synchronizing signal: (a) $E\omega/2A_f=0$, (b) $E\omega/2A_f=1.9\times10^6$, (c) $E\omega/2A_f=10^7$, and (d) $E\omega/2A_f=1.9\times10^7$.

subsequent to starting. Furthermore, the statistics of the initial phase depend upon the preoscillation noise-to-locking signal ratio. This, then, is the important parameter fixing pulse-to-pulse coherence. Other factors influencing the performance are: oscillator tuning relative to the injected frequency, value of a coefficient which is directly related to the injected power and the natural frequency and inversely related to the power output and external Q, and design parameters determining the rf rise time.

Appendix I

The equivalent circuit usually used to characterize a magnetron is shown in Fig. 25. Here the normalized load admittance, $G+jB$, is coupled to the operating mode, represented by R, C, and L. The factor K_C accounts for the transformer action of the coupling loop or iris, and $g+jb$ is the nonlinear admittance which characterizes the electronic discharge. If the magnetron is operating in the steady state, conservation of energy requires that the total admittance shunting any pair of terminals be zero. From this condition, we find the operating equations

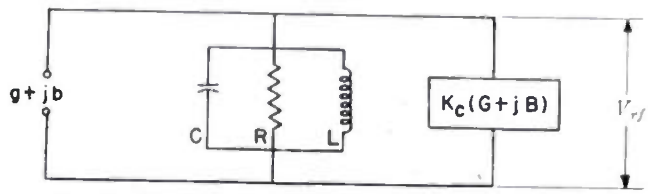

Fig. 25—Single-mode equivalent circuit of a magnetron.

$$\frac{g}{\omega_0 C} = \frac{1}{Q_0} + \frac{G}{Q_{ext}}$$

$$\frac{b}{\omega_0 C} = \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) + \frac{B}{Q_{ext}}, \tag{I-1}$$

where $\omega_0 = 1/\sqrt{LC}$ = resonant mode frequency, $Q_0 = R/\omega_0 L = R\omega_0 C$ = quality factor of the mode, and $Q_{ext} = \omega_0 C/K_C = 1/K_C\omega_0 L$ = loading effect of matched load. If the magnetron is in a transient condition, another term which accounts for energy storage in the mode must be added to the equations above. Specifically, the energy stored is

$$W = CV_{rf}^2 \tag{I-2}$$

and the rate of energy storage is

$$\frac{dW}{dt} = 2CV_{rf}\frac{dV_{rf}}{dt}. \tag{I-3}$$

At any instant this rate may be represented as a conductance shunting the circuit of Fig. 25. The magnitude of this shunting effect is

$$G_S V_{RF}^2 = 2CV_{rf}\frac{dV_{rf}}{dt}$$

or

$$G_S = \frac{2C}{V_{rf}}\frac{dV_{rf}}{dt} \tag{I-4}$$

so that (I-1) becomes

$$\frac{g}{\omega_0 C} = \frac{1}{Q_0} + \frac{G}{Q_{ext}} + \frac{2}{\omega_0}\frac{1}{V_{rf}}\frac{dV_{rf}}{dt} \tag{I-5a}$$

$$\frac{b}{\omega_0 C} = \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) + \frac{B}{Q_{ext}}. \tag{I-5b}$$

Now if g is known as a function of V_{rf}, then (I-5a) may be solved to find V_{rf} as a function of time. Assuming the empirical relation

$$g = \frac{E/R}{V_{rf}} - \frac{1}{R}, \tag{I-6}$$

we find

$$V_{rf} = \frac{E}{RC\omega_0}\frac{1}{\left(\frac{1}{RC\omega_0} + \frac{1}{Q_L}\right)}\left\{1 - e^{-\omega_0/2(1/RC\omega_0 + 1/Q_L)t}\right\}$$

$$V_{rf} = V_{RF_0}(1 - e^{-kt}, \tag{I-7}$$

where

$$V_{RF_0} = \frac{E}{RC\omega_0}\frac{1}{\left(\frac{1}{RC\omega_0} + \frac{1}{Q_L}\right)},$$

$$k = \frac{\omega_0}{2}\left(\frac{1}{RC\omega_0} + \frac{1}{Q_L}\right), \quad \text{and}$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{G}{Q_{ext}}.$$

Now in the derivation of (32) we found the coefficient N was

$$N = \frac{E\eta}{2CRV_{RF_0}}\tan\alpha. \tag{I-8}$$

If we substitute our expression for V_{RF_0}, we find

$$N = \frac{\omega_0}{2}\left(\frac{1}{RC\omega_0} + \frac{1}{Q_L}\right)\eta\tan\alpha = k\eta\tan\alpha. \tag{I-9}$$

Appendix II

It is possible to examine the equivalent load of a synchronized oscillator in an analytical manner. The resulting expression, together with the oscillator operating equations, allows the steady-state and transient behavior of the oscillator phase to be calculated. Assumptions involved in such a procedure make the result an approximation which is quite good when the synchronizing signal is small. This type of analysis has been applied to both triode and microwave oscillators. Portions of this particular solution are adapted from the work of Slater.

Consider a microwave oscillator whose voltage and current at its reference plane are $Ve^{j\omega t}$ and $ie^{j\omega t}$. Let there be an externally injected signal whose voltage and current at the oscillator reference plane are $V_1e^{j\omega_1 t}$ and $i_1e^{j\omega_1 t}$. If ω_1 and ω are not greatly different, the admittance presented to the oscillator at its plane of reference may be written

$$Y_L = \frac{ie^{j\omega t} + i_1e^{j\omega_1 t}}{Ve^{j\omega t} + V_1e^{j\omega_1 t}} \tag{II-1}$$

or

$$Y_L = \frac{i}{V}\left[\frac{1 + \dfrac{i_1}{i}e^{j(\omega_1-\omega)t}}{1 + \dfrac{V_1}{V}e^{j(\omega_1-\omega)t}}\right]$$

$$= (G + jB)\left[\frac{1 + \dfrac{i_1}{i}e^{j(\omega_1-\omega)t}}{1 + \dfrac{V_1}{V}e^{j(\omega_1-\omega)t}}\right], \tag{II-2}$$

where $G+jB$ is the passive load admittance. If we carry out the indicated division and neglect all except the first two terms, (II-2) becomes

$$Y_L \cong G + jB + 2\rho e^{j(\omega_1-\omega)t}, \tag{II-3}$$

where $\rho=1/2(i_1/i-V_1/V)(G+jB)$ and the approximation is good only when $V_1/V \ll 1$. Now ρ is known as the reflection factor, and may be written

$$\rho = \frac{Y_{LP}}{2}\left(\frac{i_1}{i_i+i_r} - \frac{V_1}{V_i+V_r}\right)$$

$$= \frac{Y_{LP}}{2}\left(-\frac{\rho_S}{1-\rho_L} - \frac{\rho_S}{1+\rho_L}\right)$$

or

$$\rho = \frac{Y_{LP}\rho_S}{2}\left(\frac{2}{1-\rho_L{}^2}\right) \cong -Y_{LP}\rho_S, \tag{II-4}$$

where we have separated the oscillator current and voltage into their incident and reflected components: $\rho_S = V_1/V_i = -i_1/i_1$ = reflection coefficient resulting from the locking signal, and $\rho_L = V_r/V_i = -i_r/i_i$ = reflection coefficient resulting from the passive load. Furthermore, the approximation in (II-4) assumes that $|\rho_L|$ is less than about 0.3. Then the load admittance may be written approximately as

$$Y_L \cong G + jB - 2|\rho|e^{j[(\omega_1-\omega)t+\Omega]}$$
$$= G + jB - 2|\rho|e^{j\phi}, \tag{II-5}$$

where Ω is the phase of the reflection factor. Note that Ω is the sum of the oscillator, locking signal, and load phases if the approximation in (II-4) is valid. In any case, the reflection factor phase is a linear function of the oscillator phase. Now we are particularly interested in the condition $\omega_1=\omega$, that is, when the oscillator is synchronized. Incorporating this condition in (II-5) and substituting the resulting expression into the oscillator operating equations, we have, after separation of the real and imaginary parts,

$$\frac{g}{\omega_0 C} = \frac{1}{Q_0} + \frac{G}{Q_{ext}} - \frac{2|\rho|}{Q_{ext}}\cos\phi \tag{II-6a}$$

$$\frac{b}{\omega_0 C} \cong 2\left(\frac{\omega_1-\omega_0}{\omega_0}\right) + \frac{B}{Q_{ext}} - \frac{2|\rho|}{Q_{ext}}\sin\phi. \tag{II-6b}$$

Equation (II-6b) is concerned with the oscillator frequency and phase, while (II-a) specifies its power output. The former may be made to read

$$\frac{\omega-\omega'}{\omega_0} = \frac{|\rho|}{Q_{ext}}\sin\phi, \tag{II-7}$$

where $\omega' = b/2C + \omega_0 - \omega_0/2Q_{ext}$ and is the frequency of oscillation of the absence of the locking signal. Then in the synchronized condition we have

$$\phi = \sin^{-1}\left[\frac{Q_{ext}(\omega_1-\omega')}{|\rho|\omega_0}\right]. \tag{II-8}$$

Now, (II-7) may be written as a differential equation by allowing ω and ω_1 to differ and noting that

$$\frac{d\phi}{dt} = \omega_1 - \omega. \tag{II-9}$$

This permits us to write

$$\frac{d\phi}{dt} + \frac{|\rho|\omega_0}{Q_{ext}}\sin\phi = (\omega_1 - \omega'). \tag{II-10}$$

This equation may be solved directly for ϕ as a function of time.

A Family of Designs for Rapid Scanning Radar Antennas*

R. F. RINEHART†

Summary—**A family of parallel-plate radar lenses is derived from an optical result of Luneberg. These lenses theoretically produce a beam with a straight wave front from a point source of microwave energy located at any point on a certain circle. A constant refractive index greater than one in the planar portion of the lens permits an inversely proportional smaller feed-circle radius. This provides the possibility of reducing feed-circle size to a point where rapid rotation of the source is feasible, producing rapid scan by the output beam.**

INTRODUCTION

IN AN EARLIER PAPER[1] a theorem of Luneberg's in lens optics was utilized to derive a theoretical shape for a perfectly focussing radar lens which constituted a promising solution to the problem of rapid scanning. Tests have since been conducted on a model of the lens at a wavelength of 3 cm with good results.

From a practical standpoint the lens mentioned above has the disadvantage of requiring that the diameter of the circle on which the antenna feed must be rotated to produce scanning be undesirably large. The present paper develops a system of lenses possessing theoretically perfect optics and requiring smaller feed circles than those in footnote reference 1.

THE LUNEBERG LENS AND ITS CURVED SURFACE ANALOGUE

Luneberg[2] has shown that light from a point source P (Fig. 1) in a plane medium of refractive index 1 can be focussed into a parallel beam by a disk with a variable refractive index which is a function only of the distance r from the center of the disk.

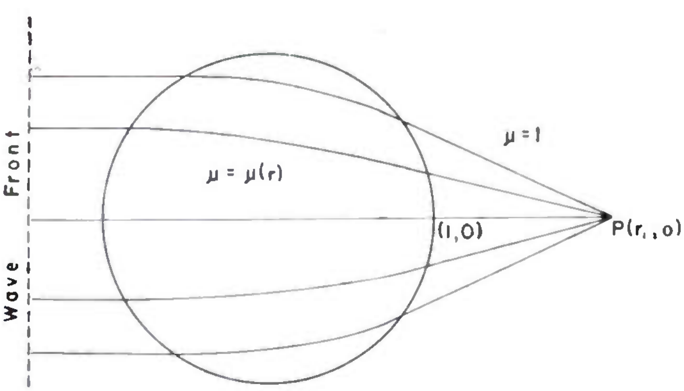

Fig. 1—The Luneberg collimating lens (Plane Case). The index of refraction is unity outside the unit circle and is an appropriately chosen function $\mu(r)$ inside the unit circle for a given choice of point source P.

Taking the radius of the disk to be unity, the index $\mu(r)$ is determined by

$$\mu = e^{w(r, r_1)}, \tag{1}$$

where the function $w(r, r_1)$ is determined by

$$w(r, r_1) = 1/\pi \int_r^1 (\arcsin t/r_1)/\sqrt{t^2 - r^2}\, dt. \tag{2}$$

When a source of electromagnetic energy moves around the circle $r = r_1$, the emergent parallel beam scans through 360 degrees.

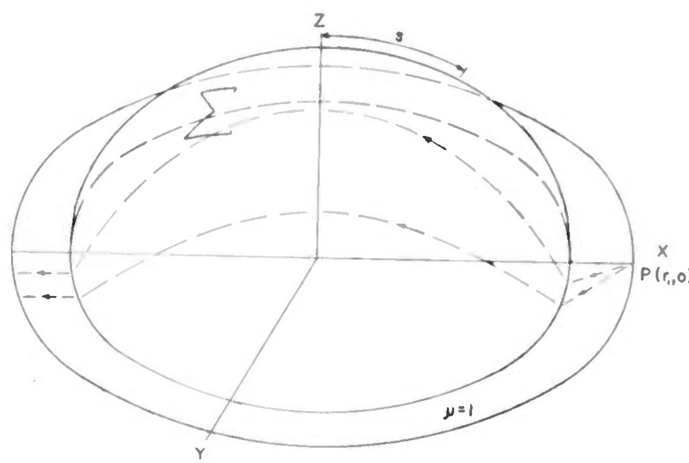

Fig. 2—The curved surface lens analogue of the Luneberg lens. With a constant refractive index of one, this surface has the same optical properties as the corresponding Luneberg planar lens. A parallel conducting plate waveguide with this surface as mean surface theoretically possesses the same focussing characteristics.

In footnote reference 1 the case $r_1 = 1$ (i.e. P on the disk circumference) was shown to be optically equivalent to an appropriate curved surface of revolution with a *unit index of refraction*; henceforth, this surface would serve as mean surface for a focussing parallel-plate radar lens. In that case $\mu(r)$ turns out to be $\sqrt{2 - r^2}$ and the equation of the generating curve of the corresponding surface, $s = 1/2(r + \sin^{-1} r)$, where s is the arc length of the generating curve measured from the axis of revolution.

The methods employed in footnote reference 1 can also be applied to the more general Luneberg theorem where $r \neq 1$. We seek a surface of revolution Σ such that the hat-shaped surface consisting of Σ and the annular region of the XY plane between $r = 1$ and $r = r_1$ (see Fig. 2) will duplicate, with a refractive index of unity, the optical properties of the planar Luneberg system of Fig. 1. Let s be the arc length, measured from the axis of revolution; then the generating curve $r = r(s)$ of the desired surface Σ is determined by the equations

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† Case Institute of Technology, Cleveland, Ohio.
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[2] R. K. Luneberg, "Mathematical Theory of Optics," Brown University Lecture Notes, Providence, R. I., pp. 208–213; 1944.

$$ds/r = du/u \qquad (3)$$
$$r = \mu u,$$

where r and μ are related by the Luneberg equations (1) and (2).

For each choice of $r_1 \geqq 1$, (1), (2), and (3) determine a surface which will yield a perfect lens with feed circle $r = r_1$. Equations (1), (2), and (3) are not solvable in terms of elementary functions for the relation between r and s, except for the case $r_1 = 1$. For any numerical choice of r_1, approximation methods can be employed to obtain the curve $r = r(s)$.

A Modification of the Luneberg Analogue Surface to Reduce Feed-Circle Size

The surface defined in the preceding section possesses the disadvantage of requiring the relatively large feed circle $r = r_1$. However, it is possible to construct an equivalent electro-optical system consisting of the same curved surface Σ, and an attached annular ring of *outer radius* unity and with a *constant* nonunit refractive index. This amounts to "turning the hat-brim in" on Fig. 2, as in Fig. 3.

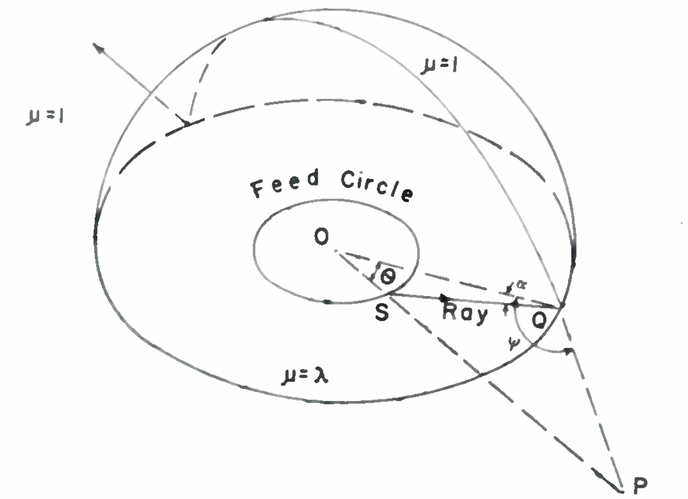

Fig. 3—An equivalent surface to that of Fig. 2, but with the "brim" turned in. The index of refraction in the planar part of the surface is a constant greater than one.

To this end we seek to determine for a given $r_1 > 1$ a feed-circle radius $r_0 < 1$ and a constant refractive index λ, such that a ray emanating from a point S on that feed circle will be refracted into the curved surface Σ at Q in the same direction as is a ray from the corresponding point P on the circle $r = r_1$ to Q. Accordingly, the constants r_0 and λ must satisfy the following relationships (Fig. 4):

$$\lambda \sin \alpha = \sin \psi \qquad \text{(Snell's Law)}, \qquad (4)$$

$$r_0 \sin (\theta + \alpha) = \sin \alpha \text{ (Law of sines, triangle } OQS), \quad (5)$$

$$r_1 \sin (\psi + \theta) = \sin \psi \text{ (Law of sines, triangle } OPQ). \quad (6)$$

By elimination of ψ and α from (4), (5), and (6) and imposing the requirement that the resulting relation in r_0, r_1, λ, and θ be an identity in θ, one finds that

$$\lambda = 1 \quad \text{or} \quad \lambda = r_1.$$

The case $\lambda = 1$ leads to $r_0 = r_1 = 1$, which is the case discussed in footnote reference 1. The case $\lambda = r_1$ implies that $r_0 = 1/r_1$. It is readily verified that these necessary conditions are also sufficient, by noting that the relations $\lambda = r_1 = 1/r_0$ together with (4) and (6) insure the holding of (5).

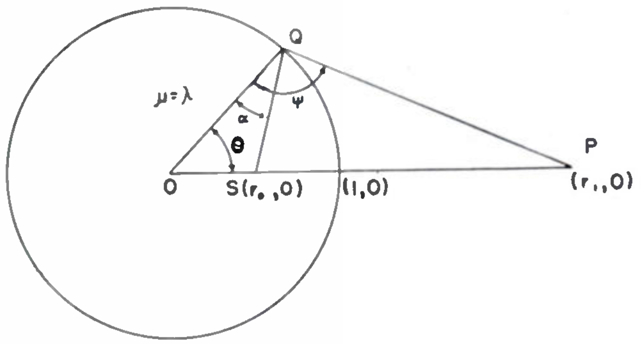

Fig. 4—Partial plan view of Fig. 3.

Thus for any choice of feed-circle radius, r_0, there exists a corresponding focussing lens system consisting of the annular ring with radii r_0 and 1, with constant refractive index $1/r_0$ attached to the surface Σ determined by (1), (2), and (3). The only limitation on feed-circle size is the size of the refractive index λ which may be practicable, taking into consideration the discontinuity in refractive index at $r = 1$.

The Angle of Bend at the Junction of Annulus and the Curved Surface

The lens system developed here has an abrupt bend at $r = 1$, and toroidal-bend or reflector-plate methods would need to be employed to eliminate reflections. Particular difficulty would attend a bend in excess of 90 degrees. It is therefore desirable to know at what angle Σ meets the planar section of the lens system.

Theorem: The surface Σ has a vertical tangent plane at $r = 1$.

Proof: We wish to compute ds/dr at $r = 1$. From (3),

$$ds/du = r/u \frac{du}{dr} = \mu \frac{du}{dr} = \mu d/dr(r/\mu) = 1 - (r/\mu)d\mu/dr$$
$$= 1 - (r/\mu)e^{w(r,r_1)}(dw(r, r_1)/dr).$$

At $r = 1$, $(r/\mu)e^{w(r,r_1)} = 1$, and we need to compute dw/dr.

Now from (2) with $0 < r < 1$,

$$w(1 - \Delta r, r_1) - w(1, r_1)$$
$$= 1/\pi \int_{1-\Delta r}^{1} (\arcsin t/r_1 dt)/\sqrt{t^2 - (1 - \Delta r)^2}$$

since $w(1, r_1) = 0$, because $\mu]_{r=1} = 1 = e^{w(1,r_1)}$.

Setting $1 - \Delta r = m$, we have

$$w(m, r_1) = 1/\pi \int_{m}^{1} (\arcsin t/r_1 dt/\sqrt{t^2 - m^2}.$$

Now since $0 < \arcsin t/r_1 > \arcsin m/r_1 > 0$,

$$w(m, r_1) > [(\arcsin m/r_1)/\pi] \int_m^1 dt/\sqrt{t^2 - m^2}$$
$$> [(\arcsin m/r_1)/\pi]$$
$$\cdot 1/2 \ln [(1 + \sqrt{1 - m^2})/m],$$
$$w(m, r_1)/(1 - m) > [(\arcsin m/r_1)/2\pi] \quad (7)$$
$$\cdot [1/(1 - m)]$$
$$\cdot [\ln (1 + \sqrt{1 - m^2})/m].$$

The first factor on the right is greater than zero for $m > 0$. By application of L'Hospital's rule to the product of the last two factors, it is readily determined that as $m \to 1$ from below, the product $\to +\infty$. Hence by (7) lim $[w(m, r_1)/(1-m)] = +\infty$, and therefore ds/dr is infinite at $r = 1$, and the generating curve of the surface has a vertical tangent at $r = 1$.

Since the surface has a vertical tangent plane at $r = 1$, the bend needed at $r = 1$ in a waveguide constructed about this mean surface is no more severe, namely 90 degrees, than in the previously given solutions. If the discontinuity in refractive index at $r = 1$ is not a practical limitation, then this solution provides a neat way of achieving a small feed-circle radius in a radar lens with theoretically perfect optics.

Space-Charge Waves in an Accelerated Electron Stream for Amplification of Microwave Signals*

PING KING TIEN†, ASSOCIATE, IRE AND LESTER M. FIELD†, FELLOW, IRE

Summary—Space-charge waves in an accelerated or decelerated electron stream are studied in this paper using a one-dimensional small-signal approximation. Exact solutions are given for an idealized stream of infinite extent, accelerated or decelerated uniformly through a space where dc space-charge effects are assumed to be neutralized by positive ions.[1] It is argued that the results obtained can be applied to streams of small diameter by use of an appropriate reduction factor on plasma frequency even in the absence of positive ions.

The solution obtained indicates that space-charge waves on a decelerated stream grow in amplitude, and can thus provide a method for amplifying microwave signals. The amplifying mechanism which is discussed is one involving an electron stream which has a single value of velocity in any transverse plane, but which is decelerated and accelerated alternately by a set of properly spaced parallel electrodes. Three amplifiers of this type have been constructed and tested and the theoretical gain expression has been verified.

PART I

General Theory of the Space-Charge Waves in an Accelerated Electron Stream

A. Introduction and Assumptions

IN A RECENT letter,[2] a mechanism for the amplification of microwave signals by use of space-charge waves was announced. The principle involved came to the authors' attention during a study of space-charge waves in a single-velocity electron stream, accelerated or decelerated by a dc electric field.

A theory of space-charge waves in a constant potential drift space was published early in 1939 by Hahn[1,3] and Ramo.[4,5] In their theory, space-charge waves are discussed in terms of a traveling wave whose amplitude varies periodically to form a standing wave in space. The propagation constant of the traveling wave is equal to the signal angular frequency divided by the stream velocity, and the periodic variation of the standing-wave envelope is similarly related to the plasma frequency, which in turn depends upon the dc electron-charge density of the stream.

The general problem to be solved in this paper is that of space-charge waves in a long and slender stream of small diameter, accelerated or decelerated by electrodes which produce some arbitrary dc potential distribution along the stream. The stream is enclosed in a space of high vacuum and is well-focused by an essentially infinite axial magnetic field.

Instead of analyzing such a problem directly, the problem will be simplified by treating an idealized stream of infinite extent with the following assumptions:

(*a*) The stream is considered uniform in a plane transverse to the beam-flow direction. Sufficient positive ions are assumed to neutralize dc effects of space charges, or alternatively, the dc potential distribution along the idealized stream in the direction of beam flow is assumed to be exactly the same as that of an actual small diameter stream by imagining a set of fictitious permeable-membrane plane electrodes inserted across the stream parallel to each other whose purpose is to hold the potentials at the desired levels. (Both ions and membranes may be considered at the same time if desired.)

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† Electronics Research Laboratory, Stanford University, Calif.

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[4] S. Ramo, Proc. I.R.E., vol. 27, pp. 757–763; December, 1939.

[5] S. Ramo, *Phys. Rev.*, vol. 56, pp. 276–283; August, 1939.

(*b*) Motions of electrons are limited to the axial direction by assuming an infinite magnetic focusing field.

(*c*) The stream is modulated with a signal small as compared with the comparable dc quantities.

(*d*) The dc velocity of the stream is assumed to be small as compared with the velocity of light, so that the effect of relativity can be neglected.

(*e*) Emission velocities are neglected and electrons are assumed to start from the cathode with zero dc velocity.

The theory is thus reduced to a one-dimensional small-signal analysis which, it is believed, will give most of the important information that is needed for an actual stream of small diameter. Actually, the idealized stream of infinite transverse extent with ions or membranes has essentially the same dc conditions as those of an actual small-diameter stream without ions. The ac effect of space charge, however, is reduced in a stream of small diameter as compared with the effect in a stream of infinite extent. This factor will be taken into account by reducing the value of the plasma frequency by an appropriate factor which will be described in detail in Section F of Part I.

In a stream of small diameter, the dc potential distribution curve along the beam-flow direction can be adjusted to an arbitrary form by properly shaping the electrodes. (Consider a Pierce gun as an example.) The dc effect of space charge as studied by Wiltshire[6] can, in general, be neglected here except in the vicinity of the cathode. It is evident that different potential distributions produce space-charge waves of different forms. It would seem that in such diode-type problems the Llewellyn-Peterson[7] equations might be directly applied. Unfortunately, the Llewellyn[8] equations, which are built on the basis that the potential distribution is determined from space charge in the stream for a stream of infinite transverse extent with no ions, do not give the right solutions in the case considered here where the dc effect of space charge is unimportant and the form of dc potential distribution is mainly dependent upon the form of accelerating electrodes. In Section D of Part I it will be shown that the Llewellyn-Peterson equations may be adapted to fit the problem here by subdividing the whole space into small regions.

Two methods are given in Section D of Part I which solve the idealized stream described above. The first is particularly suitable for the case where the dc potential distribution curve is determined from an electrolytic tank and is difficult to put in analytical form.

B. The Space-Charge Wave Matrix

In a one-dimensional analysis the total ac current of a stream, i.e. the sum of ac convection current and ac displacement current, is equal to zero. The Llewellyn-Peterson[7] equations may then be expressed in the form

$$\begin{aligned} q(z) &= (E^*q_s + F^*v_s)e^{-\beta} \\ v(z) &= (H^*q_s + I^*v_s)e^{-\beta}, \end{aligned} \tag{1}$$

where z is the co-ordinate in the axial direction. The subscript s denotes the starting plane of the region considered. $q(z)$ and $v(z)$ are, respectively, the ac convection current density and the ac electron velocity at the z-plane; q_s and v_s are the same quantities at the s-plane. With ω the angular signal frequency $\beta = i\omega T$, T the dc transit time between the s-plane and z-plane and $i=(-1)^{1/2}$. E^*, F^*, H^* and I^* are the Llewellyn coefficients and are[7] functions of the dc voltages and dc current densities at both ends of the space. We will similarly express the space-charge waves on a stream in the form

$$\begin{aligned} q(z) &= (Eq_s + Fv_s)e^{-\beta} \\ v(z) &= (Hq_s + Iv_s)e^{-\beta}; \end{aligned} \tag{2a}$$

except that in the case here E, F, H and I, called *space-charge wave coefficients* in this paper, are functions of the dc potential distribution in the space as well as the dc end conditions. This is because in our case the dc potential distribution in the space is not determined by the dc effect of the space charges of the stream, but is mainly dependent upon the form of the accelerating electrodes. To simplify mathematical manipulation, matrix notation will be used throughout this paper. Expression (2a) can be represented in the matrix form

$$\begin{Vmatrix} q(z) \\ v(z) \end{Vmatrix} = \begin{Vmatrix} E & F \\ H & I \end{Vmatrix} e^{-\beta} \begin{Vmatrix} q_s \\ v_s \end{Vmatrix}. \tag{2b}$$

The square matrix in (2b) is called the *space-charge wave matrix*. The factor $e^{i\omega t}$ is omitted for convenience in all expressions of this paper. The factor $e^{-\beta}$, together with the omitted factor $e^{i\omega t}$ in (2b), showed the traveling wave of the signal frequency, and the space-charge wave coefficients E, F, H, and I describe the amplitude variations of the waves.

C. Matrix Expression of the Space-Charge Waves in a Constant Potential Drift Space

A brief review of the space-charge waves in a constant potential drift space is necessary for later explanations. The space-charge wave matrix of the drift space of a velocity-modulated tube has the form[4,9]

$$\begin{Vmatrix} \cos(hz) & ig\sin(hz) \\ \dfrac{i}{g}\sin(hz) & \cos(hz) \end{Vmatrix} e^{-\beta}, \tag{3}$$

where

$$h = \omega_1/v_0, \qquad \omega_1^2 = \rho_0 e/\epsilon m_0,$$
$$g = \omega\rho_0/\omega_1, \qquad \beta = i\omega z/v_0;$$

[6] A. S. Wiltshire, Technical Note No. 71, Telecommunication Research Establishment, Ministry of Supply, England.

[7] F. B. Llewellyn and L. C. Peterson, Proc. I.R.E., vol. 32, pp. 144–166; March, 1944.

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[9] L. T. Zitelli, "Space Charge Effects in Modulated Electron Beams," Ph.D. Dissertation, Stanford University, Calif.

e and m_0 are respectively the electron charge and electron mass, v_0 is the dc velocity of the beam, ρ_0 is the dc charge density, z is the distance measured from the entering plane, ω_1 is known as the angular plasma frequency, and ϵ is the dielectric constant of the vacuum.

Taking $-i/g$ as the characteristic impedance of the system and defining v/q as impedance, expression (3) has exactly the same form as the matrix of the ideal conventional transmission-line equations. The Smith chart could therefore be used in computations.

D. General Theory of the Space-Charge Waves in an Accelerated Electron Stream of Infinite Extent

Two methods are developed here to solve for space-charge waves in an accelerating stream, namely, the *method of partition of space* and the *method of electronic equations.* Both are perfectly general and may be applied to a space having any dc potential distribution.

1. Method of Partition of Space

This method consists of dividing the accelerating space into a large number of small cells and considering each cell as a separate diode. The Llewellyn equations can be applied to each diode. In order to fit the given dc potential distribution, the dc end conditions for each cell are evaluated from the given dc potential distribution. The problem is then reduced to that of a number of diodes connected in cascade. The equations for each

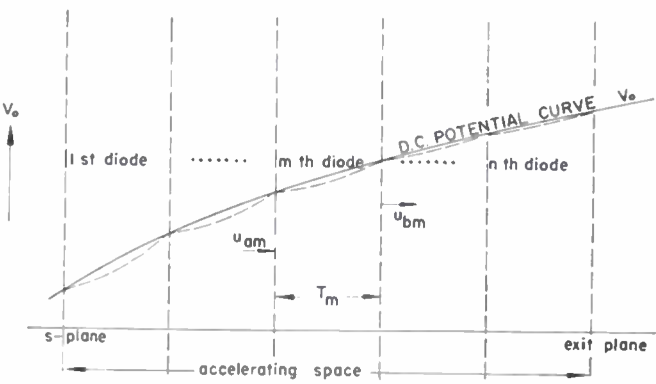

Fig. 1—Method of partition of space. The actual potential distribution curve shown as a solid line is approximated by the dotted line in the method of analysis.

diode can be expressed in matrix form by the Llewellyn equations. The product of all matrices gives the equation for the space-charge waves in the space. Referring to Fig. 1, the entire accelerating space is assumed to be divided into n diodes, and then

$$\begin{Vmatrix} q(z) \\ v(z) \end{Vmatrix} = \begin{Vmatrix} E_n & F_n \\ H_n & I_n \end{Vmatrix} \times \begin{Vmatrix} E_{n-1} & F_{n-1} \\ H_{n-1} & I_{n-1} \end{Vmatrix} \times \cdots \times \begin{Vmatrix} E_m & F_m \\ H_m & I_m \end{Vmatrix} \times \cdots \times \begin{Vmatrix} E_1 & F_1 \\ H_1 & I_1 \end{Vmatrix} e^{-\beta z} \begin{Vmatrix} q_s \\ v_s \end{Vmatrix}, \tag{4}$$

where β is again the total dc transit angle multiplied by the imaginary unit i.

E_m, F_m, H_m, and I_m for the mth diode can be obtained from the Llewellyn coefficients[7] after eliminating the space-charge factor ζ by the following equation, in which $\eta = e/m_0$:

$$(\eta/\epsilon)q_0 = (u_a + u_b)2\zeta/T^2. \tag{5}$$

They are

$$E_m = 1 - \frac{\eta T_m^2 q_0}{2\epsilon u_{bm}}$$

$$F_m = i\omega T_m q_0/u_{bm}$$

$$H_m = i\frac{\eta}{\epsilon}\frac{T_m}{2\omega}\left(1 + \frac{u_{am}}{u_{bm}}\right) - i\left(\frac{\eta}{\epsilon}\right)^2 \frac{T_m^3}{4\omega}\frac{q_0}{b_{bm}}$$

$$I_m = \frac{u_{am}}{u_{bm}} - \frac{\eta T_m^2 q_0}{2\epsilon u_{bm}}, \tag{6}$$

where, according to the Llewellyn-Peterson[7] notation, u_{am} and u_{bm} are respectively the dc electron velocity at the entering and the exit plane of the mth diode, $q_0(=I_d$ in reference number 7) is the dc electron current density, and T_m, the dc transit time for the mth diode. All those quantities are evaluated from the given dc potential distribution in the space.

2. Method of Electronic Equations

The following electronic quantities are assumed:

Electron velocity	$v_1 = v_0 + ve^{i\omega t}$
Charge density	$\rho_1 = \rho_0^+ + \rho_0^+ + \rho^+ e^{i\omega t}$
Convection current density	$q_1 = q_0 + qe^{i\omega t}$
Potential	$V_1 = V_0 + Ve^{i\omega t}$,

where all quantities except q_0 are functions of z, and z is the distance measured from some reference plane denoted as the o-plane; v_0, ρ_0, q_0 and V_0 are dc quantities, v, q, ρ, and V are ac quantities, ρ_0^+ is the dc charge density of the positive ions, and ρ_0 is that of the electrons. Through the use of Poisson's equation, the force equation, the continuity equation, and the current equation, after neglecting higher-order small quantities and by using the substitutions $q = Qe^{-\beta}$ and $\beta = i\omega\int(dz/v_0)$, the following electronic equation[10] may be obtained:

$$\frac{d^2Q}{dz^2} + \frac{dQ}{dz}\frac{3dv_0}{dz} + Q\frac{\eta q_0}{\epsilon v_0^3} = 0 \tag{7}$$

$$v = \frac{-iv_0^2}{\omega q_0}\frac{dQ}{dz}e^{-\beta}. \tag{8}$$

The signs of the quantities e and ρ_0 for electrons are negative. When electrons move in the direction of posi-

[10] A similar equation derived by L. D. Smullin, M.I.T. and P. Parzen, Federal Telecommunication Laboratory, Inc. (unpublished).

tive z, q_0 is negative. When e is negative $\eta = e/m_0$ is also.

It is interesting to note from (8) that the amplitude of the velocity modulation goes to zero when that of the current modulation is a maximum, and it reaches its maximum when the amplitude of the current modulation has its maximum rate of change. The detailed solution will be given in a particular case.

E. Space-Charge Wave Coefficients of a Uniformly Accelerated Electron Stream

The solutions are given below for the special case in which the dc potential distribution is a straight line, that is, the electron stream is accelerated uniformly. This case is shown in Fig. 2. The potential gradient in

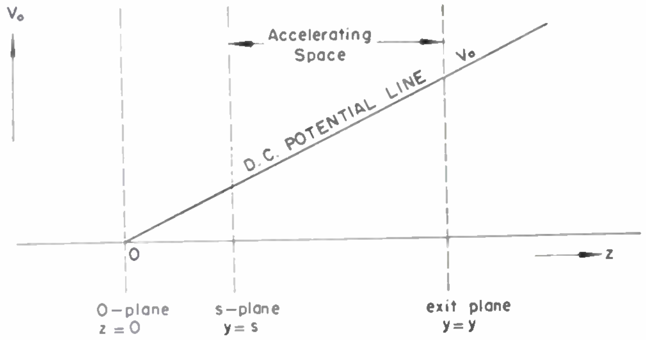

Fig. 2—The accelerating space with a linear potential distribution.

the space is constant and is equal to E_0. The o-plane, which will be taken as the origin of the axial co-ordinate z, is the plane on which the dc potential versus distance curve of the accelerating space indicates zero potential when extrapolated. Define k as $v_0 = kz^{1/2}$.

Assume the electron stream enters the accelerating space at the s-plane. Define a dimensionless quantity which is always positive as

$$x = \left(\frac{-4\eta \mid q_0 \mid z^{1/2}}{\epsilon k^3} \right)^2. \tag{9}$$

The space-charge wave coefficients E, F, H and I[11] are plotted in Figs. 3, 4, 5, and 6 as a function of x over the range x equal to one to 5,000, with the initial plane at $x = 1$. The space-charge wave matrix with the initial plane at any other position, say $x = a$, can be computed from the curves by

$$\begin{Vmatrix} F_{az} & F_{az} \\ H_{az} & I_{az} \end{Vmatrix} = \begin{Vmatrix} E_{1a} & F_{1a} \\ H_{1a} & I_{1a} \end{Vmatrix}^{-1} \times \begin{Vmatrix} E_{1z} & F_{1z} \\ H_{1z} & I_{1z} \end{Vmatrix}. \tag{10}$$

Where the double subscript is used, the first letter denotes position of initial plane and second letter denotes position of interest. E_{1z}, F_{1z}, H_{1z}, and I_{1z} and E_{1a}, F_{1a}, H_{1a}, and I_{1a} are space-charge wave coefficients with the initial plane at $x = 1$, and can be found from the curves.

From these curves it is noted that accelerated streams have space-charge waves with an amplitude which varies in a shrinking oscillatory form. This fact has been utilized to reduce the noise content in an electron beam.

[11] Appendix II.

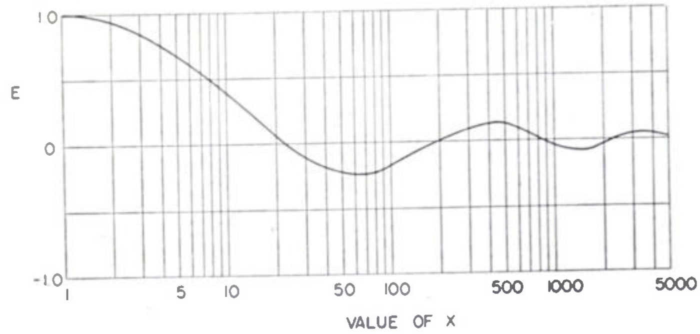

Fig. 3—Space-charge wave coefficient E for a uniformly accelerated electron stream of infinite extent, with the initial plane at $x = 1$.

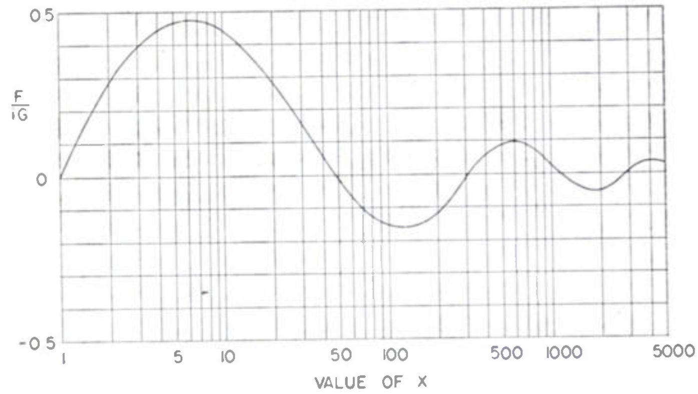

Fig. 4—Space-charge wave coefficient F for a uniformly accelerated electron stream of infinite extent, with the initial plane at $x = 1$.

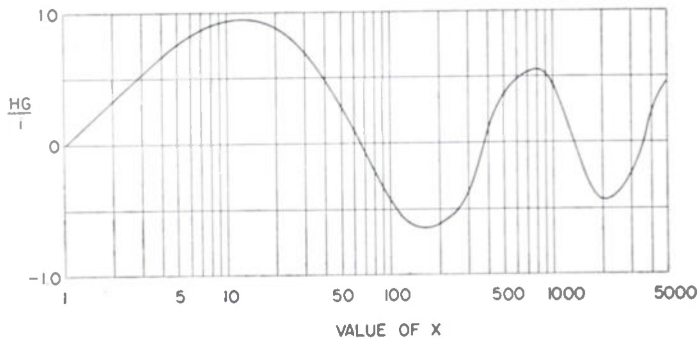

Fig. 5—Space-charge wave coefficient H for a uniformly accelerated electron stream of infinite extent, with the initial plane at $x = 1$.

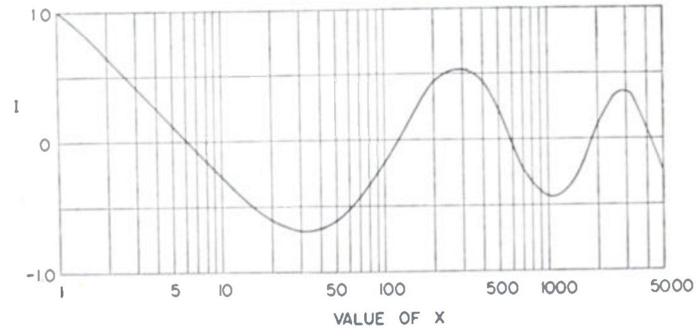

Fig. 6—Space-charge wave coefficient I for a uniformly accelerated electron stream of infinite extent, with the initial plane at $x = 1$.

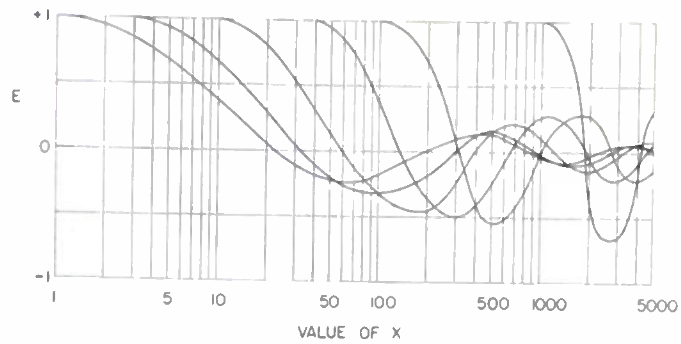

Fig. 7—Space-charge wave coefficient E for a uniformly accelerated electron stream of infinite extent, with the initial planes at $x=1$, 3, 10, 40, 100, and 1,000.

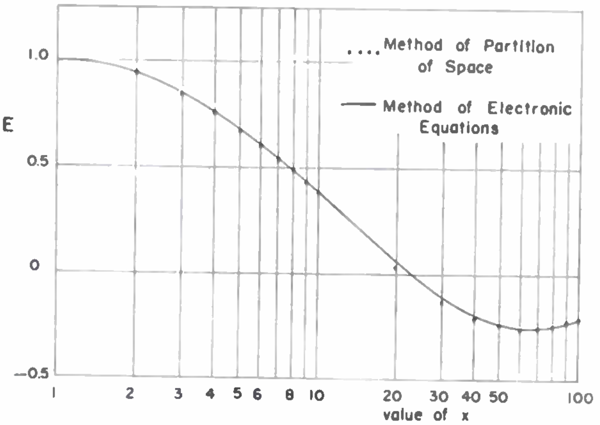

Fig. 8—The results computed by the method of partition of space are compared with those computed by the method of electronic equations.

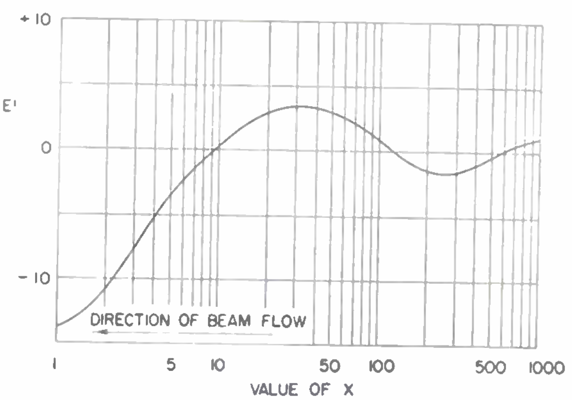

Fig. 9—Space-charge wave coefficient E' for a uniformly decelerated electron stream of infinite extent, with the initial plane at $x=1{,}000$.

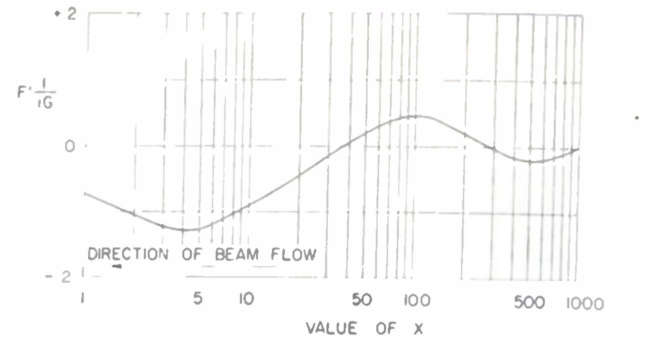

Fig. 10—Space-charge wave coefficient F' for a uniformly decelerated electron stream of infinite extent, with the initial plane at $x=1{,}000$.

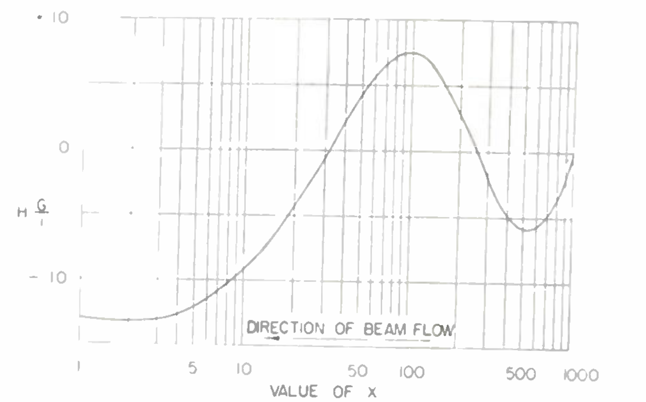

Fig. 11—Space-charge wave coefficient H' for a uniformly decelerated electron stream of infinite extent, with the initial plane at $x=1{,}000$.

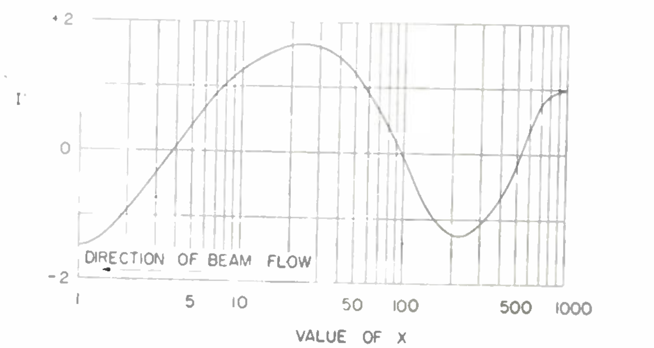

Fig. 12—Space-charge wave coefficient I' for a uniformly decelerated electron stream of infinite extent, with the initial plane at $x=1{,}000$.

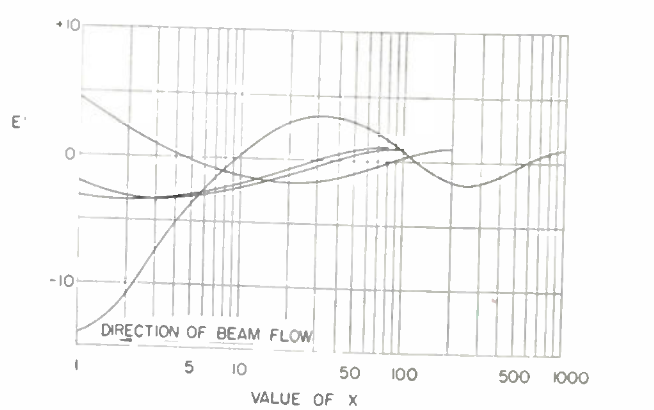

Fig. 13—Space-charge wave coefficient E' for a uniformly decelerated electron stream of infinite extent with the initial planes at $x=1{,}000$, 200, 100, and 80.

A stream which has been accelerated in a space could have been decelerated by simply reversing the direction of flow. In order to avoid confusion, the space-charge wave coefficients for a decelerated stream are denoted by E', F', H', and I', E, F, H and I being used to refer to an accelerated stream. It can be shown that

$$\begin{Vmatrix} E' & F' \\ H' & I' \end{Vmatrix} = \begin{Vmatrix} E & F \\ H & I \end{Vmatrix}^{-1}. \tag{11}$$

E', F', H' and I'[11] are plotted in Figs. 9, 10, 11, and 12 with the initial plane at $x=1{,}000$. The space-charge wave coefficients with the initial plane at any other position can be computed by the same expression (10).

The prediction of space-charge waves with amplitude in a growing oscillatory form for a continuously decelerated stream is an important result of this analysis.

F. A Correction Factor for a Stream of Small Diameter Surrounded by a Wall

The results derived by Ramo[4] for the space-charge wave of a stream of infinite extent and that of a stream of finite diameter surrounded by a metallic drift tube wall may be compared for the case of a constant potential drift space. One finds there that the solution for the stream of finite diameter can be obtained from that of the stream of infinite extent by merely multiplying the value of angular plasma frequency ω_1 by a reduction factor

$$\left(\frac{1}{1+\Gamma^2/\gamma_0^2}\right)^{1/2}. \tag{12}$$

This factor neglects higher-order variations across the beam cross section, and in general will be true only for a stream of small diameter. The correction factor (12) depends upon the beam and tube diameters, the dc velocity of the stream, and the applied signal frequency.

It is believed that the same correction factor can be used for accelerated and decelerated streams, except that in these cases the dc velocity along the stream varies and the correction should be made point by point.

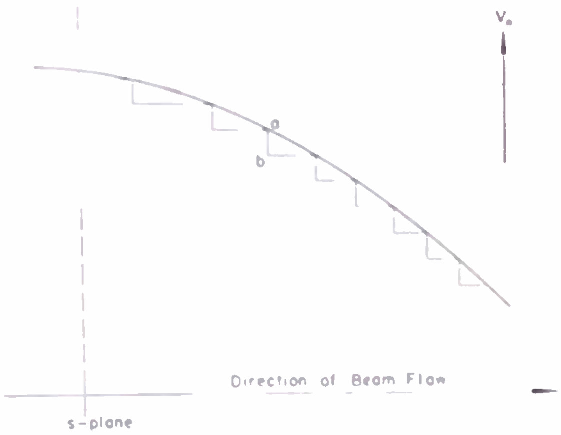

Fig. 14—The dc potential distribution versus distance curve approximated by a step line.

The possibility of applying the same correction factor for accelerated and decelerated streams arises from the fact that the space-charge waves will not be noticeably changed if the continuous dc potential distribution curve is replaced by a step function as shown in Fig. 14.

PART II

Amplification of Microwave Signals by Use of Decelerated Electron Stream

A. An Amplifying Mechanism for Microwave Signals

Space-charge waves of increasing amplitude in a decelerated stream can provide a method of amplifying microwave signals. It is observed from Fig. 9 that when an electron stream, excited with pure current modulation at the position of $x=100$, is decelerated through a space of linear potential distribution to the position of $x=3$, the amplitude of current modulation is increased by a factor of about three.

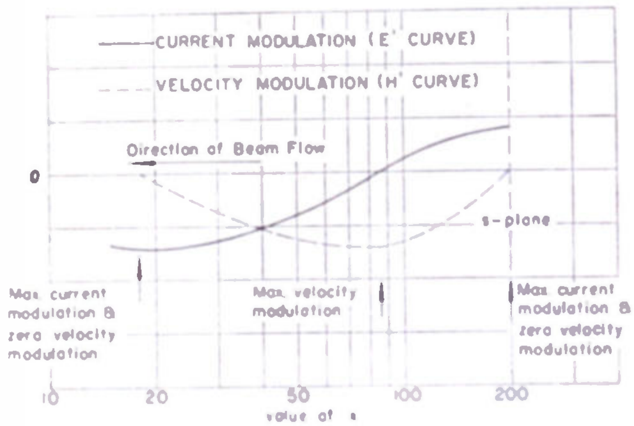

Fig. 15—The amplitude variation of current and velocity modulations of a uniformly decelerated electron stream excited with current modulation at $x=200$.

One form of structure by which a signal can be amplified is shown in Fig. 16. This structure is composed of identical sections, each consisting of a decelerating

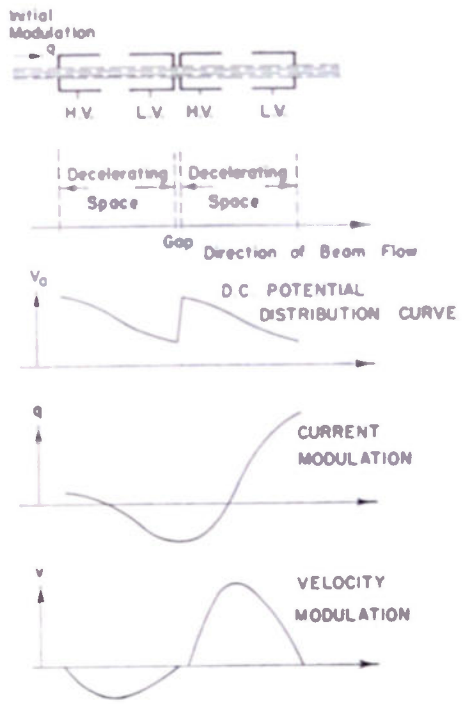

Fig. 16—The repeated structure of the new amplifying mechanism.

space and a gap. The length of the decelerating space is so designed that the stream of electrons which enters one end of the space with maximum current modulation (Fig. 15) is brought back to maximum current modulation again at the other end of the space, after being amplified during the period of deceleration. A voltage jump is then made to bring the stream back to the original high potential through a gap. Current modulation would remain continuous across the gap, if it is short compared to a space-charge quarter wavelength, in accordance with the Llewellyn-Peterson equations.[7] The same process is repeated in the next section.

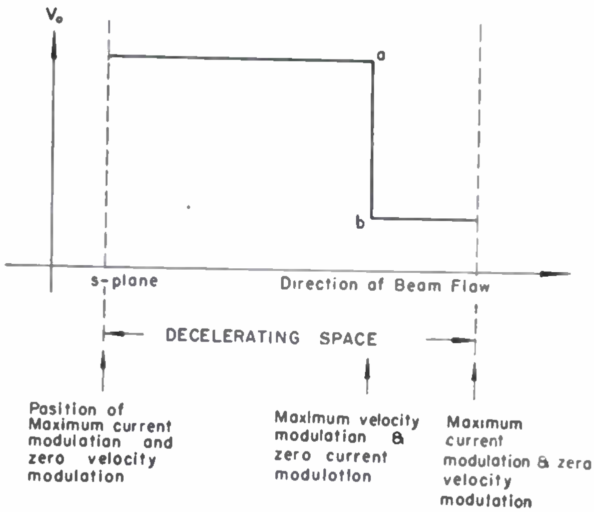

Fig. 17—The form of the dc potential-distribution curve which achieves maximum amplification with a given deceleration.

Now the form of the dc potential distribution in the decelerating space will be considered. By studying various forms of decelerating potential distribution curves, it can be shown that the most effective dc potential distribution curve for amplification is in the form of Fig. 17. The corresponding structure is shown in Fig. 18. In this case, the signal power gain per section can be de-

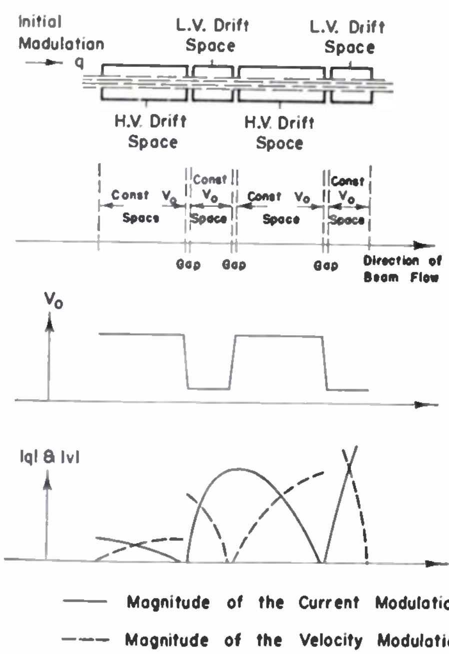

Fig. 18—The structure of the single stream space-charge wave amplifier.

rived directly from kinematic theory and solutions of space-charge waves in a constant potential drift space.

Suppose that the structure starts with current modulation q_1. This current modulation is transformed into velocity modulation v_1 after coasting along the high-potential drift space. The length of the high-potential drift space is designed to be a quarter, or an odd multiple of a quarter, of the effective plasma wavelength. According to expression (3) in section C of Part I, the stream reaches the decelerating gap with velocity modulation

$$v_1 = \frac{i}{g_1} q_1 = \frac{i\omega_1 v_{01} F_1}{\omega q_0} q_1, \tag{13}$$

where the subscript 1 denotes quantities before the deceleration. The subscript 11 will be used to denote quantities after deceleration; ω_1 is the plasma angular frequency at the high potential drift space and F_1 is the associated correction factor due to the effects of the metallic wall and the finite beam diameter of the stream.

v_1 is then amplified through a deceleration at the gap in accordance with straight forward kinematics[12]

$$v_{11}/v_1 = v_{01}/v_{011}. \tag{14}$$

After coasting along the low potential drift tube, designed to be a quarter or an odd multiple of a quarter of the plasma wavelength long, the velocity modulation v_{011} takes on the form of current modulation again, according to (3)

$$q_{11} = i g_{11} v_{11} = i \frac{\omega q_0}{\omega_{11} v_{011} F_{11}} v_{11}. \tag{15}$$

The final result for power gain per section from expressions (13), (14) and (15) is thus

$$\text{power gain} = 20 \log_{10} \left(\frac{V_{01}}{V_{011}}\right)^{3/4} \left(\frac{F_1}{F_{11}}\right) \text{db}, \tag{16}$$

where V_{01}, F_1 and V_{011}, F_{11} are respectively the dc potential and the plasma frequency correction factor, associated with the drift tube before and after deceleration.

The amplification from (16) seems to be insensitive to signal frequency although the correction factors F_1 and F_{11} change somewhat with the signal frequency.

Either a cavity or helix can be used to modulate the electron stream and to extract the amplified signal output. The helix is preferred for wide-band operation.

This kind of amplifier (Fig. 19) may have many advantages. It is not frequency sensitive as compared with klystrons. No circuit structure is necessary between the input and output ends as in traveling-wave tubes. Difficulties from mixing of beams in a double-beam tube are avoided since only a single stream is used.

It is believed that the same structure can be used to construct an oscillator when a proper feedback system is provided.

Detailed design procedures for this kind of amplifier have been obtained, but will not be included in this paper.

[12] Appendix I.

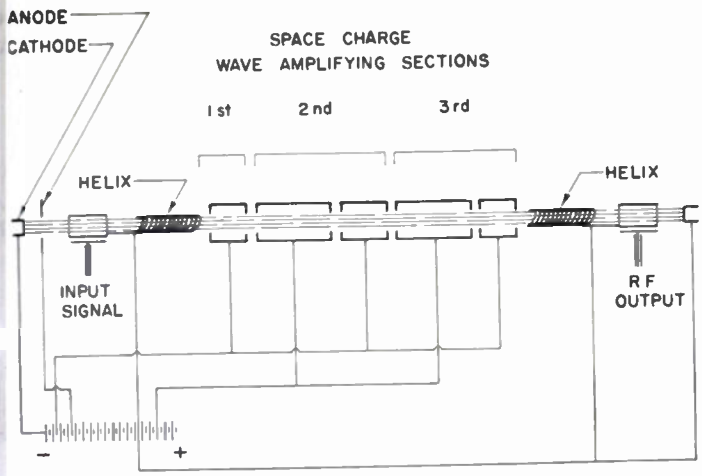

Fig. 19—A sketch of the single stream space-charge wave amplifier.

B. Experimental Results

Three amplifiers of this kind have been constructed and tested. They are shown in Figs. 20, 21, and 22. A beam of 0.055-inch diameter was used in all tubes. Tube No. 1 has six cylinders. With the first three cylinders connected at the same potential as a high-potential drift tube, and with the other three cylinders connected together as a low-potential drift tube, it was operated as a typical space-charge wave amplifier of one space-charge wave section. With a sudden deceleration from 1,900 to 65 volts, 24 db of increase in gain over that with no deceleration was observed at a beam current of 0.6 ma and a frequency of 3,000 mc.

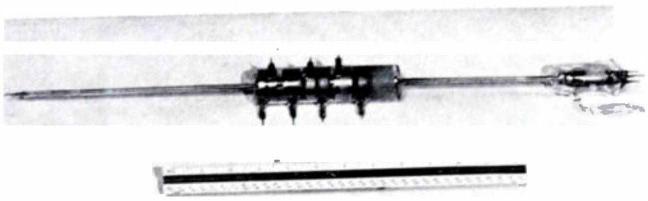

Fig. 20—Single stream space-charge wave amplifier No. 1.

Fig. 21—Single stream space-charge wave amplifier No. 2.

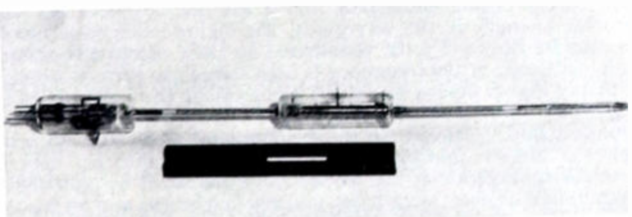

Fig. 22—Single stream space-charge wave amplifier No. 3.

When the tube is connected as two long decelerating sections, a total gain of 17 db has been observed at the same beam current and operating frequency.

Tube No. 2 has one long and one short cylinder. It was operated with a sudden deceleration from 860 to 70 volts. A 10-db increase in gain due to the space-charge wave section was measured.

Tube No. 3 has one long cylinder of 5.10-cm length and two short cylinders of 0.86-cm length. The tube was operated as an amplifier with two space-charge wave sections. The helix was operated at 680 volts. The first section was operated with a deceleration from 680 volts of helix potential to 66 volts and the second section was operated with a deceleration from 500 to 68 volts. Total gain measured due to the space-charge wave sections was 20 db at 2,891 mc with a beam current of 0.50 ma. The gain from the helix was 9 db and the over-all gain of the tube was 29 db under the operating conditions described above.

In all of these experiments the gain expression (16) was verified closely.

Acknowledgment

The work described was carried out under the sponsorship of the Office of Naval Research, the United States Army Signal Corps and the U. S. Air Force.

Appendix I

When a stream is decelerated (Fig. 16) or accelerated through a gap of small spacing, current modulation remains unchanged. From conservation of energy

$$e(V_{0a} - V_{0b}) = \tfrac{1}{2}m_0(v_{0b} + v_b)^2 - \tfrac{1}{2}m_0(v_{0a} + v_a)^2$$

or

$$\tfrac{1}{2}m_0(v_{0b}{}^2 - v_{0a}{}^2) = \tfrac{1}{2}m_0(v_{0b} + v_b)^2 - \tfrac{1}{2}m_0(v_{0a} + v_a)^2.$$

Assuming ac quantities v_a and v_b are small compared with dc quantities v_{0a} and v_{0b}, and neglecting second order small quantities, we have

$$v_b/v_a = v_{0a}/v_{0b}.$$

Appendix II

$$E = \left\{\frac{J_1(y)N_2(s) - N_1(y)J_2(s)}{J_1(s)N_2(s) - N_1(s)J_2(s)}\right\}\frac{s}{y}$$

$$F = i2G\left\{\frac{J_1(y)N_1(s) - N_1(y)J_1(s)}{J_1(s)N_2(s) - N_1(s)J_2(s)}\right\}\frac{1}{y}$$

$$H = \frac{i}{2G}\,\frac{J_2(y)N_2(s) - N_2(y)J_2(s)}{J_1(s)N_2(s) - N_1(s)J_2(s)}\,s$$

$$I = \frac{N_2(y)J_1(s) - J_2(y)N_1(s)}{J_1(s)N_2(s) - N_1(s)J_2(s)},$$

where $G = 2\omega q_0/k^2$, $y = 2x^{1/4}$, and $y = s$ at the initial plane, J_1 and N_1 are respectively Bessel functions of the first and the second kind of the first order, and J_2 and N_2 are Bessel functions of the second order.

Expressions for E', F', H', and I' have the same forms as E, F, H, and I, except with s and y interchanged.

Microwave Coupling by Large Apertures*

SEYMOUR B. COHN†, SENIOR MEMBER, IRE

Summary—**In this paper a frequency-correction factor is proposed for Bethe's small-aperture coupling relation for a transverse diaphragm in a waveguide. Experimental data on many shapes and sizes of apertures have shown this factor to be highly accurate up to and somewhat above the resonant frequency of each aperture. Also included are approximate formulas for the resonant Q and the resonant length of an aperture, and for the effect of wall thickness.**

Introduction

BECAUSE OF THE importance of apertures in many different microwave devices, considerable theoretical and experimental work has been done in the past on this type of circuit element. For example, circular and rectangular transverse irises in waveguides have been analyzed thoroughly by various investigators.[1,2] Apertures of shapes other than these have not thus far had accurate solutions that take full account of the proximity of the aperture to the waveguide walls, or that are valid near the resonant frequency of the aperture. Bethe has worked out a general approach to the problem, however, that can be used with an aperture of any shape in an infinitely thin conducting wall between any two regions if the aperture is small compared to the wavelength and to the distance to the nearest sharp bend of the wall or any other perturbation.[3-5] Specifically, the resonant frequency of the aperture should not be less than three times the operating frequency if good accuracy is to be obtained.

In many microwave devices, such as filters, antennas and broadbanding elements, apertures are used at or near resonance, and, hence, Bethe's coupling formulas are not applicable. In this article a frequency-correction factor will be added to Bethe's coupling formula for an aperture in a transverse diaphragm in a rectangular waveguide. Also, an approximate correction for wall thickness will be presented. Although coupling between other regions has not been investigated, the same principles should be applicable.

The High-Frequency Response of a Large Aperture

The effect of an infinitely thin perforated transverse diaphragm on the fundamental mode of a waveguide may be computed from an equivalent circuit in which the diaphragm is represented by a two-terminal impedance shunted across a two-conductor transmission line that is assumed to carry only the fundamental mode of the waveguide. This impedance is essentially lossless, and therefore must be of the form specified by Foster's reactance theorem.[6]

Foster's theorem, which holds for any lossless, nonactive, linear, two-terminal network, may be expressed in the following mathematical form:

$$X(f) = -\frac{1}{B(f)} = \sum_{m=0}^{n} \frac{p_m f}{1 - f^2/f_m^2} + qf - \frac{r}{f}$$

$$0 < f_0 < f_1 < \cdots < f_n, \tag{1}$$

where $X(f)$ is the reactance and $B(f)$ the susceptance of the network, f is the frequency, f_m the frequencies of the poles of $X(f)$, and p_m, q and r are positive, real constants. For a lumped-constant network, n is finite; while for a distributed constant network, such as a diaphragm in a waveguide, n is infinite, and an essential singularity of the function occurs at $f = \infty$.

The reactance function for a small aperture in an infinitely thin conducting diaphragm in a transverse plane of a rectangular waveguide (TE_{10} mode) is [3,5]

$$X = -\frac{1}{B} = \frac{4\pi M Z_0}{ab\lambda_g}, \tag{2}$$

where M is the magnetic polarizability of the aperture, Z_0 the characteristic impedance of the waveguide, a and b the width and height of the waveguide cross section, and λ_g the guide wavelength. Formulas for M are

* Decimal classification: R142×R310. Original manuscript received by the Institute, May 5, 1951; revised manuscript received December 12, 1951. The work described in this paper was performed as a part of a program of research for the Squier Signal Laboratory, U. S. Army Signal Corps, under contract no. W36-039-sc-38246.

† Sperry Gyroscope Company, Great Neck, N. Y.

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4 H. A. Bethe, "Theory of diffraction by small holes," *Phys. Rev.*, vol. 66, pp. 163–182; October 1 and 15, 1944.

5 C. G. Montgomery, R. H. Dicke, and E. M. Purcell, "Principles of Microwave Circuits," Radiation Laboratory Series, vol. 8, McGraw-Hill Book Co., Inc., N. Y., pp. 176–179; 1948.

6 The diaphragm will in general couple the fundamental-mode equivalent transmission line to an infinite number of higher-mode equivalent transmission lines. If the frequency is such that the higher modes do not propagate, and if the diaphragm is far from any other obstacle in the waveguide, the higher-mode transmission lines may be assumed to be terminated by their reactive characteristic impedances. If the frequency is high enough to permit propagation of any higher modes, and if the higher-mode equivalent lines are dissipatively terminated, the diaphragm will appear effectively as a dissipative shunt element on the fundamental-mode line. In order to have a lossless impedance to which Foster's theorem may be rigorously applied, all of the higher mode lines will be assumed to be terminated by pure reactances, and the diaphragm and waveguide walls will be assumed to be of infinite conductivity. Clearly, however, if the frequency is below the cut-off frequency of the first higher mode excited by the aperture, and if the diaphragm is far from any other discontinuity, the effect of the higher-mode terminations will be completely negligible, and the manner of terminating these lines will be immaterial.

known for circular and elliptical apertures,[3,5] and accurate measured values are available for the shapes of Fig. 1.[7] Since Z_0 is proportional to λ_g/λ, it follows that $X \propto f$, and hence (2), assumes the aperture to have the reactance of a constant inductance. Comparison with (1) shows that the small-aperture theory assumes q finite, $r=0$, and $p_m=0$, for all values of n. This of course is not actually the case, since an aperture has an unlimited number of resonances; however, it is a very good approximation for $f \ll f_0$.

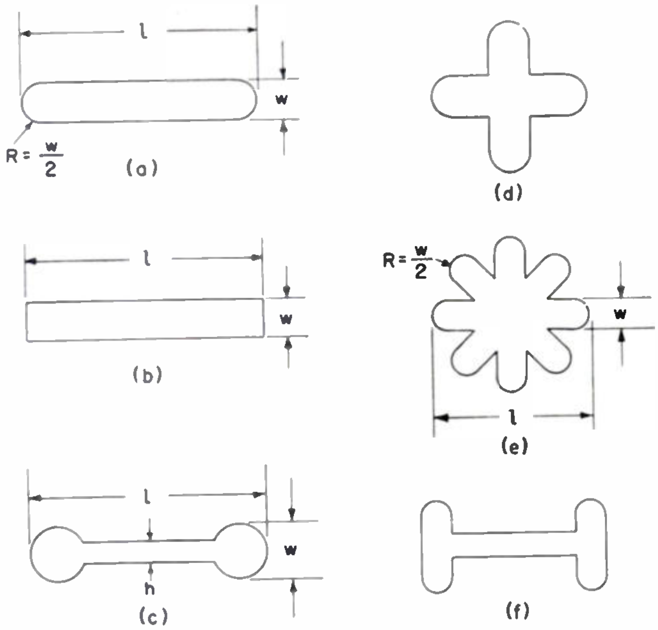

Fig. 1—Additional apertures for which accurate magnetic-polarizability data are available.

Now consider the reactance function that is one step more complicated than a straight-line function, but that reduces to (2) for $f \to 0$:

$$\frac{X}{Z_0} = -\frac{Y_0}{B} = \frac{4\pi M}{ab\lambda_g(1 - f^2/f_0^2)}. \tag{3}$$

Another simple special case is that for which the poles occur at $f=(2m+1)f_0$ and zeros at $f=2mf_0$, where m takes on all integral values from $-\infty$ to ∞:

$$\frac{X}{Z_0} = \frac{4\pi M}{ab\lambda_g}\left(\frac{2f_0}{\pi f} \tan \frac{\pi f}{2f_0}\right). \tag{4}$$

It can be seen that an infinite number of other reactance functions that conform with (1) and that reduce to (2) are possible. Nevertheless, it has been found in the experimental investigation of many aperture shapes described later in this article that the single-pole relation of (3) gives very good agreement in all cases with the measured values, and is in most cases superior to (4). Thus, (3) is, in general, recommended up to and somewhat beyond the first resonance of the aperture.

The Resonant Frequency of an Aperture

In order to use (3), it is necessary to know the resonant frequency of the aperture. This may be determined experimentally, or by certain empirical relations that will now be given.

The following empirical relationship for the resonant length l of a rectangular aperture of width w, centered in a transverse plane of a rectangular waveguide, was originally suggested by Slater:[8]

$$l = \frac{\lambda_0}{2}\sqrt{1 + \left(\frac{2aw}{b\lambda_{g0}}\right)^2}. \tag{5}$$

Equation (5) is generally accurate to within a few per cent.

For w/l small, (5) reduces to $l \approx \lambda_0/2$. Hence, the resonant frequency of a narrow rectangular aperture is approximately equal to the cutoff frequency of a waveguide having the same cross-sectional shape as the aperture. This correspondence has been found to be valid for several other aperture shapes whose width-to-length ratio is small, and it is, therefore, suggested as a general approximation. For example, consider the rounded slot of Fig. 1(a). If the cross section of the corresponding waveguide is narrow enough so that the electric field is negligible in the semicircular ends, the cutoff frequency should be the same as that of a rectangular waveguide having the same height w and the same cross-sectional area. Thus, the resonant length of a narrow rounded slot is

$$l = \frac{\lambda_0}{2} + 0.273w. \tag{6}$$

This was found to agree within one per cent for five apertures having w/l under 0.11, and within 5 per cent for $w/l=0.2$.

Approximate formulas for the cutoff wavelength of a number of other waveguide cross sections have been or can be obtained. For example, a theoretically computed graph for a "dumbbell" cross section has been given by Altar,[9] and for a "ridged" cross section by the author.[10] If theoretical or experimental data are not available for a particular aperture shape, a rough guess can usually be made that will yield good results if the aperture is not used too near resonance. For example, in the case of apertures (d) and (e) of Fig. 1, (6) may be used with fair accuracy if w/l is small.

[7] S. B. Cohn, "Determination of aperture parameters by electrolytic-tank measurements," Proc. I.R.E., vol. 39, pp. 1416–1421; November, 1951.

[8] J. C. Slater, "Microwave Transmission," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 185–187; 1942.

[9] W. Altar, *Research Report SR-287*, Westinghouse Electric and Mfg. Co.; 1945.

[10] S. B. Cohn, "Properties of ridge waveguide," Proc. I.RE., vol. 35, p. 783; August, 1947.

The Resonant Q of an Iris

The Q of a resonant iris loaded by matched waveguide terminations may be defined by

$$Q = \frac{f_0}{2} \frac{d(B/Y_0)}{df}\bigg|_{f=f_0}. \qquad (7)$$

Substitution of (3) in (7) gives

$$Q = \frac{ab\lambda_{g0}}{4\pi M}. \qquad (8)$$

Because of the effect of the higher resonant frequencies that were neglected in the derivation of (8), and because of the finite thickness of a practical diaphragm, the actual Q of an aperture would be somewhat greater than the value predicted by (8).

Effect of Wall Thickness

The attenuation α_0 introduced by an aperture in an infinitely thin transverse conducting diaphragm in a waveguide is given by

$$\alpha_0 = 10 \log_{10} \left[\left(\frac{B}{2Y_0} \right)^2 + 1 \right] \text{db}. \qquad (9)$$

If the diaphragm has finite thickness and if the frequency is below resonance, the attenuation will be greater than α_0. For very large thickness it is obvious that the increase in attenuation will approach assymptotically the attenuation of the principal mode in a waveguide whose cross section is the shape and size of the aperture, and whose length is equal to the thickness of the diaphragm.[1] Hence the following formula for the added attenuation α_t is suggested:

$$\alpha_t = \frac{54.6tA}{\lambda_c} \sqrt{1 - \left(\frac{\lambda_c}{\lambda}\right)^2} \text{ db}. \qquad (10)$$

This is the attenuation formula for a waveguide with an additional factor, A. This factor is a function of the thickness t, and approaches unity for t large; λ_c is the cutoff wavelength of the waveguide. The total attenuation of the diaphragm is

$$\alpha = \alpha_0 + \alpha_t. \qquad (11)$$

The experimental investigation described in the next section showed that in all cases tested, A is approximately three for $t<0.02l$, and decreases slowly with increasing t. For the smaller values of w/l, A tends to be somewhat greater than three, while for the larger values of w/l, A tends to be somewhat less than three. In the case of a circular aperture, A appears to be only slightly greater than one.[1]

The effect of thickness on resonant frequency has been found to be small for aperture a of Fig. 1. For aperture c, however, this effect is large.

Experimental Data

The eight apertures listed in Table I were tested in 2.840×1.340 inch ID waveguide. In every case the maximum dimension of the aperture was parallel to the transverse magnetic field. The diaphragm thickness ranged from 0.003 to 0.250 inch for the rounded slots, and 0.003 to 0.125 inch for the other apertures. The measured quantity was the attenuation α for α greater

TABLE I

List of Tested Apertures

No.	Shape	l	w/l	h/l
1	Rounded slot	1.565 inches	0.2	
2	Rounded slot	1.570	0.1	
3	Rounded slot	1.785	0.2	
4	Dumbbell	1.328	0.4	0.1
5	Dumbbell	1.410	0.3	0.1
6	Rosette	1.340	0.2	
7	Rosette	1.340	0.1	
8	Rosette	0.940	0.1	

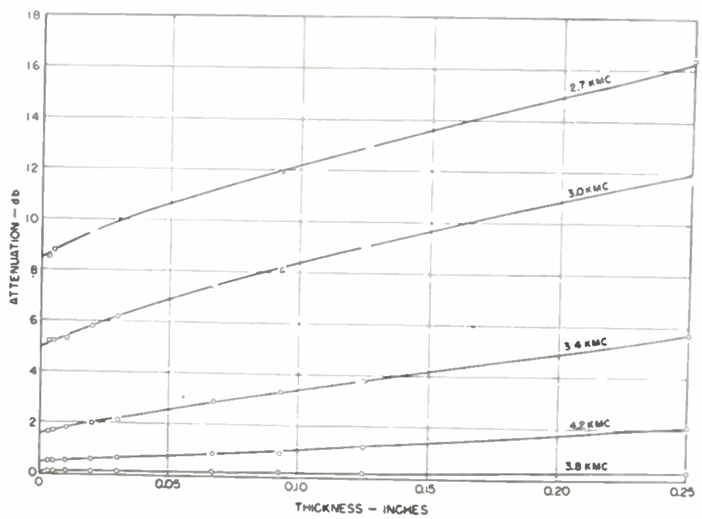

Fig. 2—Attenuation versus thickness for aperture number 2 of Table I.

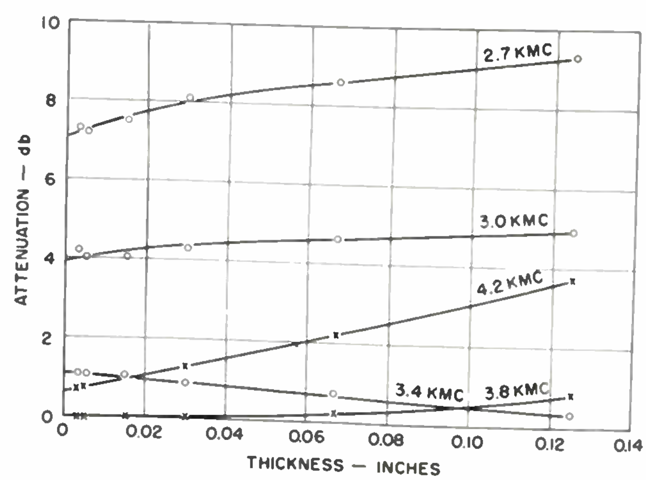

Fig. 3—Attenuation versus thickness for aperture number 4 of Table I

han 4.8 db, and the voltage standing-wave ratio S or smaller attenuations. S is convertible to α by

$$\alpha = 10 \log_{10}\left(\frac{S+1}{4S}\right)^2 \text{db.} \tag{12}$$

Γhe curves of attenuation versus thickness of Figs. 2 ınd 3 are typical.

The shunt-susceptance curves of Figs. 4 and 5 were :omputed, as follows, from the α and S values for zero thickness obtained by extrapolation of the graphical data:

$$\left|\frac{B}{Y_0}\right| = \frac{S-1}{\sqrt{S}} = 2\sqrt{10^{\alpha/10} - 1}. \tag{13}$$

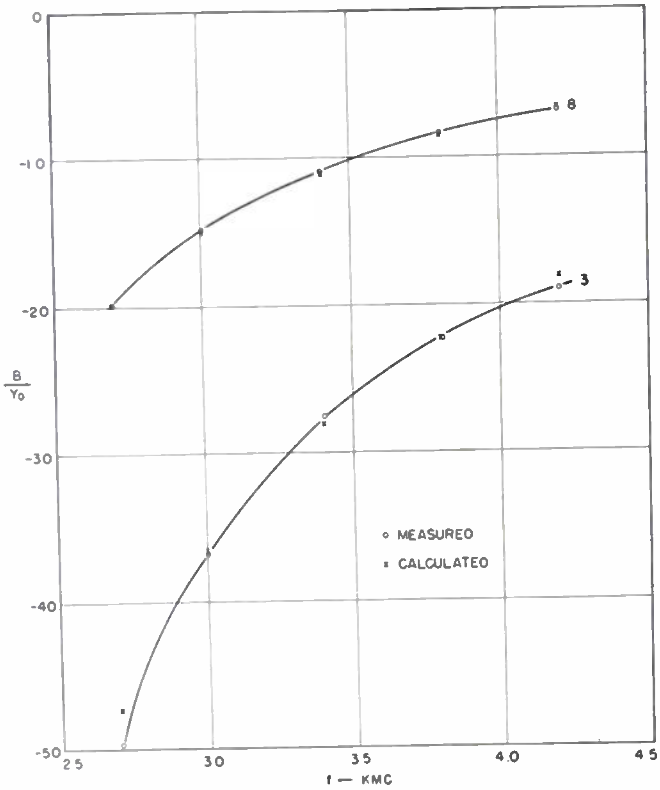

Fig. 4—Normalized susceptance of apertures numbers 3 and 8 in zero-thickness diaphragms.

In Fig. 4, points calculated with the aid of (3) are also shown, and the agreement with the measured points is

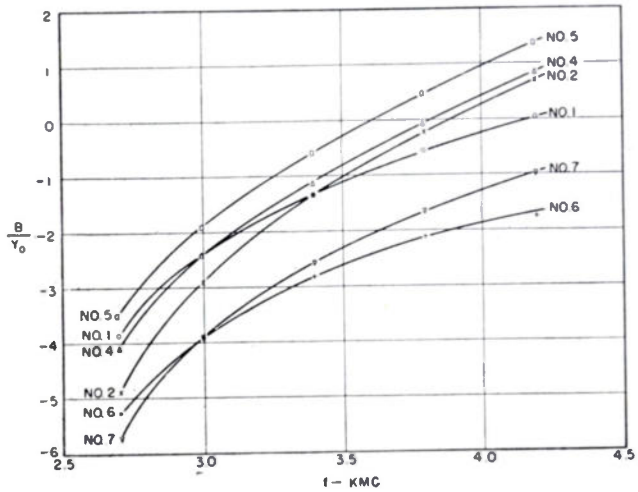

Fig. 5—Normalized susceptance of apertures numbers 1, 2, 4, 5, 6, and 7 in zero-thickness diaphragms.

seen to be very good. The calculated and measured points for all eight apertures are listed in Table II, and in every case a close correspondence occurs. The magnetic polarizabilities for the various apertures were obtained from graphs in the literature.[7] The resonant wavelengths for numbers 1, 2, 4, and 5 were obtained from Fig. 5, for numbers 3 and 8 from (6), and for numbers 6 and 7 by trial to give the best match between the measured and computed values.

TABLE II

f(mc)	Aperture 1		Aperture 2		Aperture 3		Aperture 4	
	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)
2,700	−3.94	−3.864	−4.59	−4.864	−47.3	−49.62	−4.126	−4.064
3,000	−2.62	−2.420	−2.88	−2.892	−36.5	−36.74	−2.528	−2.427
3,400	−1.49	−1.353	−1.36	−1.33	−28.0	−27.54	−1.122	−1.122
3,800	−0.67	−0.577	−0.254	−0.253	−22.3	−22.34	−0.084	−0.091
4,200	0	+0.0089	+0.628	+0.680	−18.2	−18.24	−0.754	+0.821

f(mc)	Aperture 5		Aperture 6		Aperture 7		Aperture 8	
	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)	$\frac{B}{Y_0}$ (Calc)	$\frac{B}{Y_0}$ (Meas)
2,700	−3.483	−3.496	−5.253	−5.256	−5.787	−5.700	−20.0	−19.93
3,000	−1.951	−1.91	−3.900	−3.900	−4.112	−3.892	−15.1	−14.80
3,400	−0.574	−0.578	−2.806	−2.818	−2.737	−2.580	−11.22	−10.90
3,800	+0.472	+0.476	−2.059	−2.104	−1.746	−1.685	−8.58	−8.294
4,200	+1.312	+1.339	−1.491	−1.776	−1.000	−0.998	−6.61	−6.890

Axially Symmetric Electron Beam and Magnetic-Field Systems*

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Summary—**The dynamics of hollow or solid beams in axial magnetic fields are analyzed, assuming electrons do not cross each other radially. The focusing properties are due to the angular velocity, which depends on magnetic flux linkages in the beam and at the cathode. The position of the cathode in the magnetic field is thus of vital importance.**

Radial oscillations about an equilibrium radius are always stable in the presence of a magnetic field, and can be made stable even without a magnetic field.

Design formulas are presented for two cathode arrangements; in the first the cathode is in a uniform magnetic field, while in the second it is inside a magnetically shielded structure. Special cases of each of these arrangements are considered.

Limited experimental results confirm most of the theory presented, and indicate the possibility of focusing without a magnetic field over the beam.

Introduction

THIS PAPER is a study of the theory and design methods for high-density axially symmetric electron beams. Particular attention is paid to hollow beams, where the charge density is an arbitrary function of radius between two radii and zero elsewhere.

Recently, microwave-tube developments indicate advantages of hollow beams for certain cases; for example, a coaxial waveguiding structure has been suggested for the traveling-wave tube.[1] Hollow beams increase the efficiency of high-power microwave tubes because the relatively useless core of electrons is removed.

Although hollow beams are not yet extensively used, their future importance is anticipated. Magnetic focusing as commonly used with solid beams, though successful, is by no means efficient. Much power is wasted in producing magnetic fields more intense than would be necessary were proper design methods used. For these reasons it is felt that a better understanding of a magnetic focusing of dense beams is desirable.

While certain instances of magnetic focusing have been treated in the past, notably by Brillouin,[2,3] it is only recently that any careful analysis of this subject has been done. Wang's treatment of the solid beam[4] shows the role of the cathode position in the magnetic field. His significant analysis forms the basis for much that is presented here. The equilibrium conditions derived by Samuel[5] for certain hollow beams are special cases of the ones developed here.

The purpose of the present investigation is to extend the work of Wang and Samuel in somewhat more general form so that practicable design methods making efficient use of the magnetic field may be arrived at.

Following an analysis of the motion of electrons in long axially symmetric systems the significance of cathode and other end conditions are considered. This leads to a discussion of several special cases for which explicit design formulas are developed. The paper concludes with a brief description of some experimental results.

Theory of Infinitely Long Beams in Magnetic Fields

At present we consider only steady-state conditions in infinitely long electron-beam systems. Beam production and end conditions are to be considered later. Complete axial symmetry is specified as well as longitudinal uniformity of the beam and electrodes. The magnetic field is axially symmetric, and in the part of the beam considered here, it is uniform and in the z direction. The object is to produce a beam with arbitrary radial charge distribution between two coaxial electrodes. This is shown in Fig. 1, which also defines the cylindrical co-ordinate system. This general representation may be used for hollow or solid beams, and either one of the electrodes may be removed if desired.

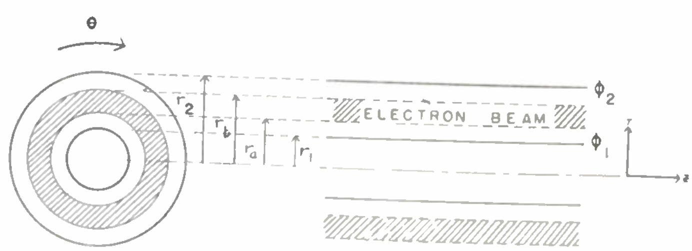

Fig. 1—Hollow beam in equilibrium, drift tubes, and co-ordinate system.

Only beams in which every electron has the same axial velocity are considered. This special mode of operation is considered because it is of greatest practical interest in microwave tubes and is mathematically simple. Relativistic effects are neglected as are thermal emission velocities from a zero potential cathode.

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† Electrical Engineering Dept., University of Minnesota, Minneapolis, Minn.

[1] L. M. Field, "Some slow-wave structures for traveling-wave tubes," PROC. I.R.E., vol. 37, pp. 34–39; January, 1949.

[2] L. Brillouin, "A theorem of Larmor and its importance for electrons in magnetic fields," *Phys. Rev.*, series 2, vol. 67, pp. 260–266; April, 1945.

[3] L. M. Field, "High current electron guns," *Rev. Mod. Phys.*, vol. 18, pp. 353–361; July, 1946.

[4] C. C. Wang, "Electron beams in axially symmetrical electric and magnetic fields," PROC. I.R.E., vol. 38, pp. 135–147; February, 1950.

[5] A. L. Samuel, "On the theory of axially symmetric electron beams in an axial magnetic field," PROC. I.R.E., vol. 37, pp. 1252–1258; November, 1949.

Conservation of energy supplies the following basic equation:

$$\dot{r}^2 + r^2\dot{\theta}^2 + \dot{z}^2 = -2\frac{e}{m}\phi, \tag{1}$$

where the dot notation represents total time derivatives; e is the electron charge, -16×10^{-20} coulomb; m is the electron mass, 9.1×10^{-31} kg; and ϕ is the electric potential at the general point measured from the cathode. All units are mks rationalized.

The solution r as a function of time t is determined as an integral of (1) after it is reduced to an equation in r alone. Setting $\dot{z}$ constant has eliminated one variable, leaving $\dot{\theta}$ and ϕ to be reduced to functions of r.

Conservation of the generalized angular monentum p_θ, given by

$$p_\theta = mr^2\dot{\theta} + eA_\theta r \tag{2}$$

where A_θ is the only component of the vector magnetic potential, leads to Busch's theorem[6]

$$\dot{\theta} = \frac{e/m}{2\pi r^2}(\psi_c - \psi). \tag{3}$$

Here ψ and ψ_c are the magnetic fluxes linking the circles at the electron position and at the cathode where the same electron started, as shown in Fig. 2.

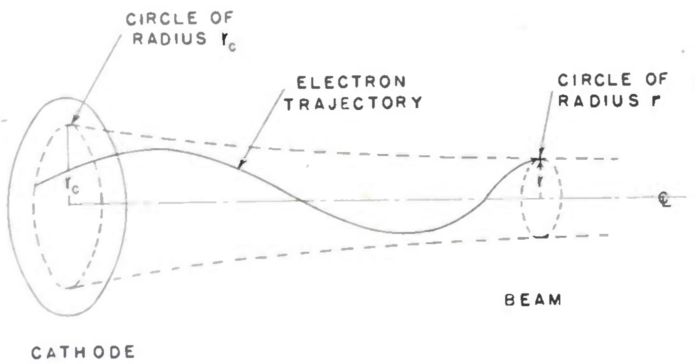

Fig. 2—The two circles linking the fluxes ψ and ψ_c.

Equation 3 expresses the angular velocity as a function of position only, and in a uniform magnetic field as a function of r only. The field need be uniform, however, only in the region of the electron's present position though it must be axially symmetric everywhere.

To reduce ϕ to a function of r we assume that electrons do not cross each other radially in their travel in the tube. Electrons move in cylindrical shells, possibly varying in radius, but never intersecting each other. Each shell encloses a constant amount of charge which may be considered to lie along the axis. Each electron, therefore, moves in a logarithmic potential field determined by its relative position in the beam. This potential is given by

$$\phi = \alpha \ln\frac{r}{r_0} + \gamma, \tag{4}$$

where r_0 is the radius at which electrons of that shell experience no radial force, their equilibrium radius. The constants α and γ are functions of r_0 but not of r or t, and are determined from the average potential $\bar{\phi}$ at r_0; α is a coefficient giving the magnitude and direction of the radial electric field.

If we now assume the magnetic field uniform and substitute (3) and (4) into (1), we obtain

$$\dot{r}^2 = -2\frac{e}{m}\alpha\ln\frac{r}{r_0} - 2\frac{e}{m}(\gamma - V) - \frac{\Omega^2}{r^2} - \omega_H{}^2r^2 + 2\omega_H\Omega, \tag{5}$$

where

$$\begin{aligned} \dot{z}^2 &= -2\frac{e}{m}V \\ \psi &= \pi r^2 B \\ \omega_H &= \frac{e}{m}\frac{B}{2} \\ \Omega &= \frac{e}{m}\frac{\psi_c}{2\pi}. \end{aligned} \tag{6}$$

This is the differential equation of radial motion for any electron in the beam.

The right-hand side of (5) is negative for large and small values of r regardless of the sign of α, but only the positive region is accessible to the electron. The magnetic field, by limiting the energy of radial motion, prevents electrons from traveling arbitrarily far from or close to the axis.

If the radial acceleration derived from (5) is equated to zero at the equilibrium radius r_0 and the equation solved for ψ_c, we obtain

$$\psi_c = \pi r_0{}^2 B\sqrt{1 + K\alpha}, \tag{7}$$

where $K = e/m/\omega_H{}^2r_0{}^2$, a negative number. The quantity $\sqrt{1+K\alpha}$ is less than or greater than one, depending on the sign of α. For real values of ψ_c, $K\alpha$ must be greater than -1. Presently we shall see that this condition guarantees that the radial motion shall be periodic (i.e., stable).

Stability requires that the equivalent potential energy (total energy less $(m\dot{r}^2/2)$) be a minimum at r_0. The condition that $\partial\ddot{r}/\partial r$ be negative at r_0 leads to the inequality

$$K\alpha > -2. \tag{8}$$

Since ψ_c must be real, the above inequality is always satisfied. An unstable radial oscillation of the beam in the presence of a magnetic field is not possible. Any

[6] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., New York, N. Y., chap. 4, p. 35; 1949.

general radial charge distribution is a possible stable configuration so long as the cathode flux condition (7) and the stability condition are satisfied.

The solution to (5) may be written (see Appendix II)

$$\frac{R}{R_0} = 1 + \frac{x^2}{2} + x\sqrt{1 + \frac{x^2}{4}} \cos \beta t, \tag{9}$$

where R and R_0 are the square of r and r_0, respectively, x is an amplitude factor defined in the appendix, and the frequency β is

$$\beta = -2\omega_H\sqrt{1 + \frac{K\alpha}{2}}. \tag{10}$$

This solution is valid only for small oscillations about the equilibrum radius.

It is apparent that as the normalized oscillation amplitude increases the center of oscillation is displaced outward as shown in Fig. 3.

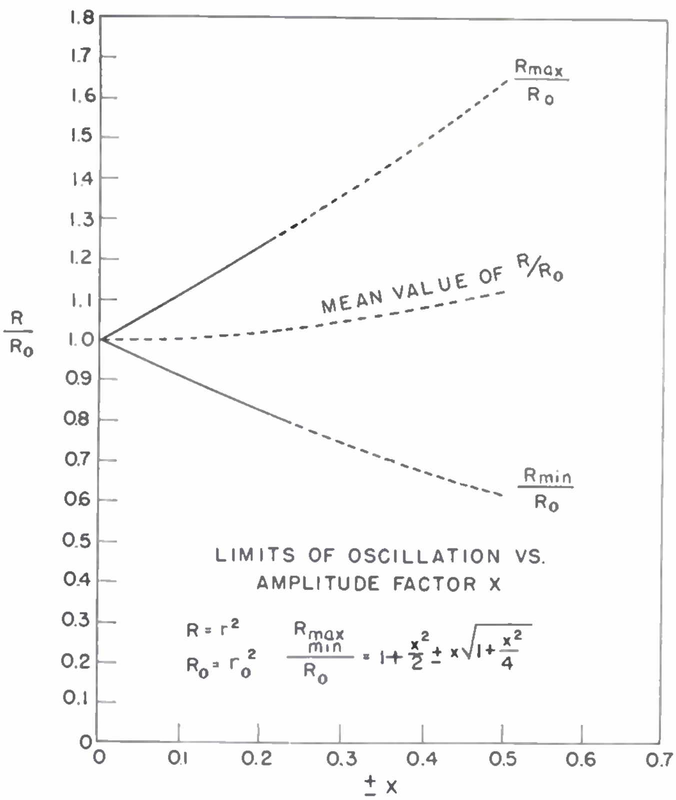

Fig. 3—Limits of radial oscillation showing outward displacement as amplitude increases.

The form of the frequency confirms the previously derived stability condition. Fig. 4 indicates the character of β as a function of flux density B. When α is positive and the electric force is radially outward, the oscillation about a given r_0 is not stable below a certain minimum flux density. But when α is negative and the electric force is inward, β remains real even for zero flux density. We shall return to discuss focusing without a magnetic field later in this paper.

The assumption that electrons do not cross radially requires that β be the same for all shells, hence independent of r_0. Unless α is proportional to r_0^2, this is not the case; but even so we shall hope the error is not a serious one. The importance of this factor must be shown by experiment, though it is clearly immaterial for a beam in equilibrium.

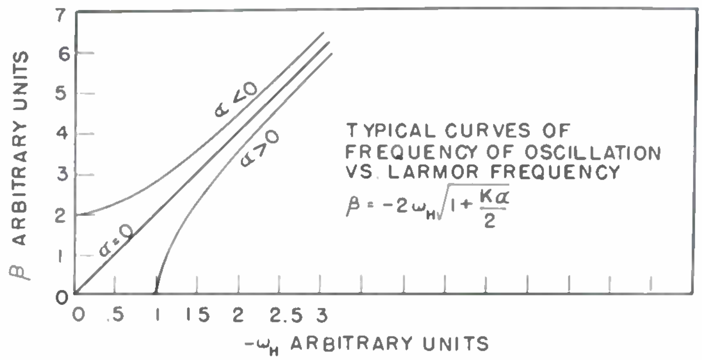

Fig. 4—Oscillation frequency as a function of magnetic field showing stability for $\alpha<0$ and $B=0$.

If the amplitude of oscillation has a zero value entirely inside or outside the charge ring, the beam fluctuates in diameter as a unit; but if the zero value occurs in the charge region, the beam pulsates in thickness, the outside expanding while the inside contracts.

By substituting (7) and (9) into (3), the average angular velocity may be found to be

$$\bar{\dot{\theta}} = \omega_H(\sqrt{1 + K\alpha} - 1). \tag{11}$$

Since $\bar{\dot{\theta}}$ has the same sign as α and is independent of the amplitude factor x, the focusing is due to the average angular velocity which, in turn, depends on the relative values of electric and magnetic fields. This rotation is just the Larmor precession or drift; the radial oscillations are part of the cycloidal or trochoidal paths whose amplitudes are determined by the particles' total energy.

At r_0 the total kinetic energy of an electron is $-e\gamma$ of which $-eV$ is due to axial motion. The difference $-e(\gamma - V)$ is the energy due to angular and radial velocities. This quantity must be positive and at least enough for the angular velocity. Any excess over this minimum represents a radial motion.

The design procedure consists of setting $\gamma - V$ equal to its minimum value, i.e., setting $x^2=0$. This optimum value is

$$(\gamma - V)_{\text{opt}} = \frac{(\sqrt{1 + K\alpha} - 1)^2}{-2K}, \tag{12}$$

from which the drift-tube potentials can be found as outlined in Appendix I.

Special Cases

The cases to be considered here are determined from a consideration of some of the end conditions in these systems.

An electron gun is used to supply and control the beam current. Because of its desirable qualities as a high current-density, low-interception gun the Pierce type of electron gun[7] is used. But the design of Pierce guns does not include the effects of magnetic fields which will therefore have to be nullified.

The magnetic field at the cathode cannot be entirely eliminated because of the cathode-flux condition (7). If magnetic focusing is to be obtained, the flux ψ must differ from the flux ψ_c. Electrons must cross field lines somewhere to acquire the angular velocity necessary for magnetic focusing.

Two ways of nullifying the effect of the magnetic field and still maintain the necessary flux linkage are available. The magnetic field lines may be made parallel to the electron paths in the gun so that there is no rotation of the electrons. The field crossing takes place outside the gun. Inside, the field lines are normal to the cathode surface, thus eliminating uncertainties as to the emission and noise properties of the cathode.[8]

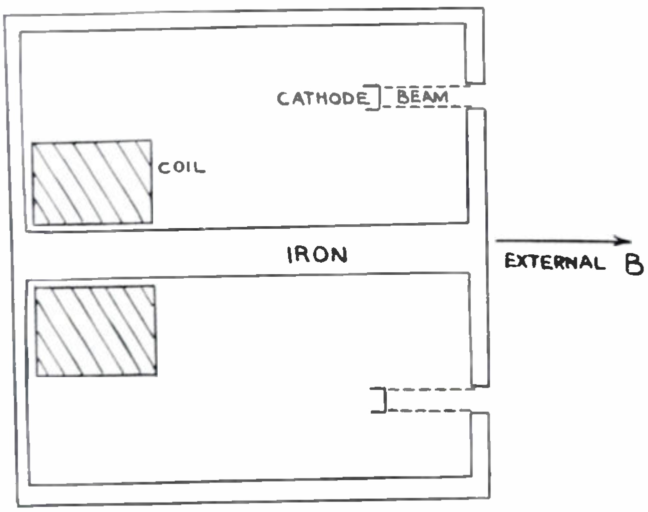

Fig. 5—A shielded electron gun used in division II.

Alternatively, the cathode may be placed inside a magnetically shielded structure, such as that shown in Fig. 5. Here all the flux ψ_c passes through the central pole piece and is the same for every electron in the beam. The electron gun is in a field-free space inside this structure.

These two arrangements determine the broad divisions of the special cases to be discussed.

[7] J. R. Pierce, *loc. cit.*, chapt. 10, pp. 168–186.
[8] R. Q. Twiss, "On the Steady State and Noise Properties of the Magnetron," Ph.D. Thesis, Physics Dept., M.I.T., Cambridge, Mass., Chap. 4; January, 1949.

Division I. Uniform Field in Gun Region

Under this heading we assume the field at the cathode is uniform of value B_c and, moreover, that for a shell of equilibrium radius r_0 the cathode flux is given by

$$\psi_c = \pi r_0^2 B_c. \tag{13}$$

Substituting this expression into (7), solving for α, and then determining the charge density from Appendix I, we find that

$$\rho = -\frac{\epsilon}{r_0}\frac{\partial\alpha}{\partial r_0} = -\frac{e}{m}\frac{\epsilon}{2}(B_c^2 - B^2). \tag{14}$$

In this case the charge density is uniform and the assumption of noncrossing can be valid. Since the attainable charge density depends on the difference $B^2 - B_c^2$, which we call the effective flux density B_{eff}^2, it is clear that the greatest economy of magnetic field is obtained when B_c is zero.

Equation 14 may be solved for B_{eff}^2 and expressed in terms of beam voltage and current, resulting in the general design equation

$$B_{\text{eff}}^2 = \frac{2}{\dfrac{e}{m}\epsilon\pi\sqrt{-2\dfrac{e}{m}}(r_b^2 - r_a^2)}\frac{I}{\sqrt{V}}$$

$$= \frac{6.9 \times 10^{-7}}{r_b^2 - r_a^2}\frac{I}{\sqrt{V}}. \tag{15}$$

The values of $\gamma - V$ obtained from this uniform charge density differ from the optimum values expressed in (12). The best design is obtained by equating them at the inner edge of the beam and letting the outer electrons oscillate. This procedure leads to an expression for the potential of the inner electrode.

$$\phi_1 = V + \frac{e}{m}\frac{r_a^2}{4}B^2\left[(1 - g_1^2)\ln\frac{r_a}{r_1} + \frac{3}{2}g_1^2 - g_1 - \frac{1}{2}\right], \tag{16}$$

where

$$g_1 = \sqrt{1 + K\alpha} = \frac{B_c}{B}.$$

The potential difference between electrodes is given by

$$\phi_2 - \phi_1 = -\frac{e}{m}\frac{r_a^2}{4}B_{\text{eff}}^2\left[\left(\ln\frac{r_2 r_a}{r_b r_1} - \frac{1}{2}\right) + \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)\right]. \tag{17}$$

This difference is positive, giving an outward electric force which aggravates the effects of space charge and increases the required focusing magnetic field.

When $B_c \neq 0$, the beam oscillates everywhere except at the inner edge. The amplitude of the oscillation is given by the amplitude factor x derived from the discrepancy between $\gamma - V$ and $(\gamma - V)_{\text{opt}}$.

$$x^2 = 4g_1 \frac{(1-g_1)}{(1+g_1^2)}\left(1 - \frac{r_a^2}{r_0^2}\right). \tag{18}$$

When $B_c = 0$, however, g_1 is zero and x vanishes, which means that the entire beam can now be in equilibrium. If the electron beam is a solid one, the above equations simplify considerably. The potential of the outer electrode is then

$$\phi_2 = V - \frac{e}{m}\frac{r_b^2}{4}B^2\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right) \tag{19}$$

if the field at the cathode is zero. It is worth noting that zero magnetic field at the cathode not only yields the most efficient and economical operation for the solid beam, but is the only condition which will allow the beam to remain in equilibrium with no oscillations.

Division II. Shielded Guns

In this division the quantity ψ_c is the same for all electrons. As before, we solve (7) for α and then find the charge density ρ.

$$\rho = \frac{e}{m}\frac{\epsilon}{2}\left(\frac{\psi_c^2}{\pi^2 r_0^4} + B^2\right). \tag{20}$$

In contrast to the previous case the cathode-flux linkage now increases the attainable charge density. By integrating this expression for ρ over the cross section of the beam and solving for the magnetic field quantities, we reproduce exactly (15) if we define

$$B_{\text{eff}}^2 = \left(\frac{\psi_c}{\pi r_a r_b}\right)^2 + B^2. \tag{21}$$

The term $\psi_c/\pi r_a r_b$ has the form of a geometric mean flux density at the cathode, which suggests the substitution

$$\frac{\psi_c}{\pi r_a r_b} = g_2 B. \tag{22}$$

Comparison of the actual and optimum values of $\gamma - V$ shows that they can be equated if

$$\phi_1 = V + \frac{e}{m}\frac{r_a^2}{4}B^2\left[\ln\frac{r_a}{r_1}\left(1 - \frac{r_b^2}{r_a^2}g_2^2\right) - \frac{1}{2}\left(g_2\frac{r_b}{r_a} - 1\right)^2\right] \tag{23}$$

and

$$\phi_2 - \phi_1 = \frac{e}{m}\frac{r_a^2}{4}B^2\left[g_2^2\frac{r_b^2}{r_a^2}\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right) + g_2^2\left(\ln\frac{r_2}{r_b} - \frac{1}{2}\right) - \left(\ln\frac{r_a}{r_1} - \frac{1}{2}\right) - \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)\right]. \tag{24}$$

The entire beam can be in equilibrium if these conditions are satisfied.

These general results lead to several particular arrangements of interest. For example, it may be desirable to keep both drift tubes at the same potential. It is necessary only to adjust the value of g_2 to make the difference $\phi_2 - \phi_1 = 0$ in (24). This value of g_2 is

$$g_2 = \left[\frac{\left(\ln\frac{r_a}{r_1} - \frac{1}{2}\right) + \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)}{\frac{r_b^2}{r_a^2}\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right) + \left(\ln\frac{r_2}{r_b} - \frac{1}{2}\right)}\right]. \tag{25}$$

Another arrangement may require that the inner drift tube be eliminated. The electric field, hence α, then vanishes, at the inner edge of the beam and

$$g_2 = \frac{r_a}{r_b}, \qquad \psi_c = \pi r_a^2 B. \tag{26}$$

Here the innermost electrons do not cross field lines and need not rotate as they experience no electric force. The outer electrons rotate only enough to balance the space-charge forces. This arrangement is considerably more efficient than the one discussed in the previous division. It is the one previously treated by Samuel,[6] Brillouin,[2] and Field,[3] although the latter two discuss a slightly different cathode arrangement.

As mentioned previously, it is possible to eliminate entirely the magnetic field over the uniform part of the beam. The design equations are derived from those developed above for the shielded gun by letting $B = 0$. The effective magnetic field remains finite because of the cathode-flux linkage. In a beam focused in this way the charge is heavily concentrated on the inside, as may be seen from (20).

The focusing action in this case is analogous to a ball spinning around the sides of a bowl in a gravitational field. The rotation acquired as the electrons leave the magnetic field structure around the cathode produces a centrifugal force. This force and the space-charge force are balanced by the inward electric force due to the negative potential difference $\phi_2 - \phi_1$.

The practical significance of this type of focusing is quite evident. Besides saving the power to produce magnetic fields, the coils and power supplies are eliminated. The tube becomes accessible for rf connections and mechanical alignment is no longer critical.

Table I brings together for convenient reference the design formulas developed previously:

TABLE I

DESIGN FORMULAS

For all Cases $B_{eff}^2 = \dfrac{6.9 \times 10^{-7}}{r_b^2 - r_a^2} \dfrac{I}{\sqrt{V}}.$

I. $\psi_c = \pi r_0^2 B_c$

General

$$B_{eff}^2 = B^2 - B_c^2 = B^2(1 - g_1^2) \qquad B_c = g_1 B.$$

$$\phi_1 = V + \frac{e}{m}\frac{r_a^2}{4}B^2\left[(1 - g_1^2)\ln\frac{r_a}{r_1} + \frac{3}{2}g_1^2 - g_1 - \frac{1}{2}\right]$$

$$\phi_2 - \phi_1 = -\frac{e}{m}\frac{r_a^2}{4}B_{eff}^2\left[\left(\ln\frac{r_2 r_a}{r_1 r_b} - \frac{1}{2}\right) + \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)\right]$$

$$x^2 = 4\frac{g_1(1 - g_1)}{(1 + g_1^2)}\left(1 - \frac{r_a^2}{r_0^2}\right).$$

$B_c = 0$

$$B_{eff}^2 = B^2$$

$$\phi_1 = V + \frac{e}{m}\frac{r_a^2}{4}B^2\left(\ln\frac{r_a}{r_1} - \frac{1}{2}\right)$$

$$\phi_2 - \phi_1 = -\frac{e}{m}\frac{r_a^2}{4}B^2\left[\left(\ln\frac{r_2 r_a}{r_1 r_b} - \frac{1}{2}\right) + \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)\right].$$

Solid Beam

$B_c = 0$

$$\phi_2 = V - \frac{e}{m}B^2\frac{r_b^2}{4}\left(\mathrm{on}\frac{r_2}{r_b} + \frac{1}{2}\right).$$

II. ψ_c *independent of* r_0

General

$$B_{eff}^2 = B^2 + \left(\frac{\psi_c}{\pi r_a r_b}\right)^2 \qquad \frac{\psi_c}{\pi r_a r_b} = g_2 B$$

$$\phi_1 = V + \frac{e}{m}\frac{B^2}{4}r_a^2\left[\ln\frac{r_a}{r_1}\left(1 - \frac{r_b^2}{r_a^2}g_2^2\right) - \frac{1}{2}\left(g_2\frac{r_b}{r_a} - 1\right)^2\right]$$

$$\phi_2 - \phi_1 = \frac{e}{m}\frac{B^2}{4}r_a^2\left[g_2^2\frac{r_b^2}{r_a^2}\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right) + g_2^2\left(\ln\frac{r_2}{r_b} - \frac{1}{2}\right) - \left(\ln\frac{r_a}{r_1} - \frac{1}{2}\right) - \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)\right].$$

$\phi_2 = \phi_1$

$$g = \left[\frac{\left(\ln\frac{r_a}{r_1} - \frac{1}{2}\right) + \frac{r_b^2}{r_a^2}\left(\ln\frac{r_2}{r_b} + \frac{1}{2}\right)}{\frac{r_b^2}{r_a^2}\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right) + \left(\ln\frac{r_2}{r_b} - \frac{1}{2}\right)}\right].$$

Inner Electrode Absent

$$g = \frac{r_a}{r_b} \qquad B_{eff}^2 = B^2\left(1 + \frac{r_a^2}{r_b^2}\right)$$

$$\phi_2 = V + \frac{e}{m}B^2\frac{r_a^2}{4}\left[1 + \ln\frac{r_2}{r_b}\left(\frac{r_a^2}{r_b^2} - \frac{r_b^2}{r_a^2}\right) - \frac{1}{2}\left(\frac{r_a^2}{r_b^2} + \frac{r_b^2}{r_a^2}\right)\right].$$

$B = 0$

$$B_{eff}^2 = \left(\frac{\psi_c}{\pi r_a r_b}\right)^2$$

$$\phi_1 = V - \frac{e}{4m}\left(\frac{\psi_c}{\pi r_a}\right)^2\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right)$$

$$\phi_2 - \phi_1 = \frac{e}{4m}\left(\frac{\psi_c}{\pi r_a}\right)^2\left[\left(\ln\frac{r_a}{r_1} + \frac{1}{2}\right) + \frac{r_a^2}{r_b^2}\left(\ln\frac{r_2}{r_b} - \frac{1}{2}\right)\right].$$

Experimental Results

A series of experiments based on some of the special cases just discussed were carried out. A few of these results are presented here.

The design parameters for the beam system were:

Inner radius $r_a = 1.13$ cm
Outer radius $r_b = 1.53$ cm
Beam voltage $V = 1{,}000$ volts
Beam current $I = 500$ ma
Drift-tube length 7 inches

The magnetic fields at the cathode and over the rest of the beam were independently controlled by currents I_K and I_B, respectively. In the tube using the shielded structure of Fig. 5 the cathode-flux linkage ψ_c was controlled by the current I_G.

The first tube was operated with the cathode in the uniform magnetic field assumed for Division I discussed above. Figs. 6 and 7 show the results of varying the radial electric field and the main magnetic field, respectively. In these figures the following notation is used:

Collector current I_c
Inner drift-tube current I_1
Outer drift-tube current I_2
Gun anode current I_a
Inner drift-tube voltage ϕ_1
Outer drift-tube voltage ϕ_2.

The small indicated currents are averages for a pulsed system with a duty ratio of 0.01.

The broken vertical lines in these figures represent the theoretically optimum values of the variable parameter.

The second tube employed a shielded electron gun, and was used to check the possibility of focusing the beam with no magnetic field outside the gun. One of the results of this test is shown in Fig. 8. The maximum of collector current indicates that the expected process does indeed take place. Quantitative checks of the

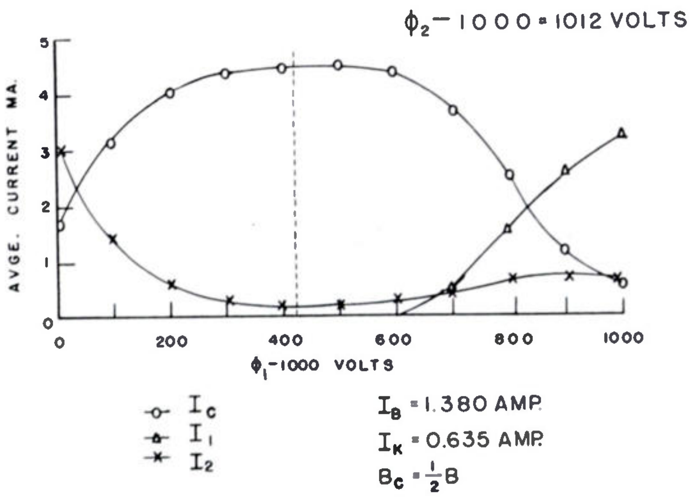

Fig. 6—Beam focusing as a function of radial electric field. Cathode in a uniform magnetic field.

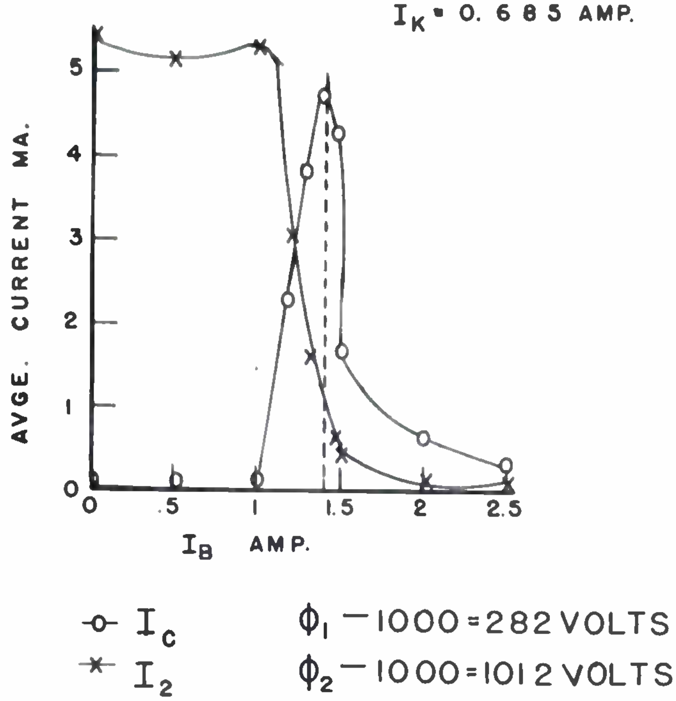

Fig. 7—Beam focusing as a function of magnetic field. Cathode in a uniform magnetic field.

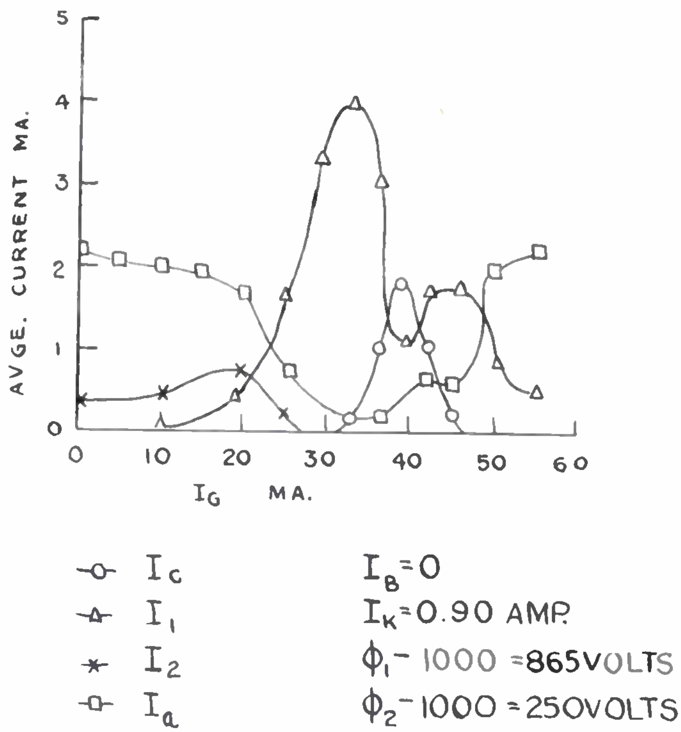

Fig. 8—Beam focusing as a function of cathode-flux linkage in a shielded gun and with no magnetic field over the beam.

theory were difficult to obtain because of poor shielding of the gun and the consequent effect on the behavior of the gun itself. The optimum drift-tube voltages, not shown here, were in rather close agreement with the design values.

Most of the discrepancies observed between the design values of beam current and the values obtained appeared to be due to the inadequacy of the transition region where electrons cross magnetic field lines. All of the experiments indicated that once the electrons could be made to enter the drift space the focusing processes were much as the theory predicted.

Conclusions

The tests described here were not as clean-cut as is desirable since many of the factors affecting the performance could not be separated. It does appear, however, that the qualitative aspects of the theory are very well borne out by the experiments and that the quantitative aspects are confirmed reasonably well. In every case where the experiments were not successful the cause of failure could be attributed with fairness to the transition region.

A particular advantage of the analysis as presented is the expression of magnetic fields in terms of flux linkages. This integrated quantity is easy to calculate or measure by mutual inductance methods, particularly in axially symmetric systems.

Little light has been shed on the validity of the assumption of noncrossing, but apparently for design purposes this assumption is relatively safe. The design formulas presented here are correct and useful as far as they go, but it should be remembered that they contain no information about the design of the transition region through which electrons pass between the gun and the main part of the beam.

Previous success in designing solid beam systems is due largely to the fortuitous circumstance that the proper focusing condition is obtained with the cathode outside the magnetic field. In hollow beams, where both inner and outer radii must be controlled, conditions are more stringent. Here the transition region problem cannot be avoided as it is generally necessary to have some flux link the cathode.

It appears possible to focus beams in any of the special arrangements considered. Noteworthy among these is the case where magnetic flux is required in the cathode region only. Because of its practical significance in economy of field and simplification of the entire system, this focusing system may well be a good reason in itself for using hollow beams in place of solid ones.

Before this method can be applied in general, however, more development work is required. Fringing of flux in the cathode region may be greatly reduced by placing the magnetic gap well forward of the gun anode itself. For smaller tubes it might be feasible to close the magnetic circuit linking the electron gun, and to activate this circuit, outside the vacuum envelope.

Acknowledgment

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Appendix I

The Potential in a Uniform Beam

In the uniform hollow beam shown in Fig. 1 the charge density ρ may be represented in the power series form

$$\rho(r) = \sum_{n=0}^{\infty} \rho_n r^n, \quad n = 0, 1, 2, \cdots \quad \text{for } r_a \leqq r \leqq r_b$$

$$0 \text{ for } r < r_a \text{ and } r > r_b. \tag{26}$$

Note that only ρ_0 has the dimensions of charge density. The evaluation of the average potential $\bar{\phi}$ consists of integrating Poisson's equation,

$$\frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial\bar{\phi}}{\partial r}\right) = -\frac{\rho}{\epsilon}, \tag{27}$$

where ϵ is the permittivity of free space, and of matching the solution to the boundary conditions at r_1 and r_2. To do this we follow Wang[4] by letting

$$\bar{\phi} = \phi_s + \phi_L, \tag{28}$$

where ϕ_s is a solution of Poisson's equation and ϕ_L is a solution of Laplace's equation.

The potential ϕ_s due to space charge alone is evaluated first. From (27), integrating twice from r_a outward,

$$\phi_s = -\int_{r_a}^{r}\left[\frac{1}{\epsilon r}\int_{r_a}^{r}\rho r dr\right]dr, \tag{29}$$

where we have set $\phi_s = 0$ inside r_a.

Before evaluating ϕ_L we shall have to calculate the contribution of ϕ_s at the outer electrode, i.e., at r_2. At the outer edge of the beam where $r = r_b$,

$$\phi_{s_b} = -\int_{r_a}^{r_b}\left[\frac{1}{\epsilon r}\int_{r_a}^{r}\rho r dr\right]dr. \tag{30}$$

Outside the electron beam, the potential ϕ_s is represented by the familiar logarithmic expression for a line charge along the axis.

$$\phi_s = -\frac{Q}{2\pi\epsilon}\ln\frac{r}{r_b} + \phi_{s_b}, \quad r > r_b, \tag{31}$$

where Q is the total charge per unit length in the beam.

$$Q = 2\pi\int_{r_a}^{r_b}\rho r dr. \tag{32}$$

The potential ϕ_s at r_2 is thus

$$\phi_{s_2} = -\frac{1}{\epsilon}\int_{r_a}^{r_b}\left[\rho r \ln\frac{r_2}{r_b} + \frac{1}{r}\int_{r_a}^{r}\rho r dr\right]dr. \tag{33}$$

Since $\phi_2 = \phi_{L_2} + \phi_{s_2}$, we have

$$\phi_{L_2} = \phi_2 + \frac{1}{\epsilon}\int_{r_a}^{r_b}\left[\rho r \ln\frac{r_2}{r_b} + \frac{1}{r}\int_{r_a}^{r}\rho r dr\right]dr; \tag{34}$$

and $\phi_{L_1} = \phi_1$ since we have set $\phi_{s_1} = 0$. The solution for ϕ_L is well known:

$$\phi_L = \frac{\phi_{L_2} - \phi_{L_1}}{\ln r_2/r_1}\ln\frac{r}{r_1} + \phi_{L_1}. \tag{35}$$

With the appropriate substitutions we obtain, for $r_a \leqq r \leqq r_b$,

$$\bar{\phi} = a\ln\frac{r}{r_1} + \phi_1 - \frac{1}{\epsilon}\int_{r_a}^{r}\left[\frac{1}{r}\int_{r_a}^{r}\rho r dr\right]dr, \tag{36}$$

where

$$a = \frac{\phi_2 - \phi_1 + \frac{1}{\epsilon}\int_{r_a}^{r_b}\left[\rho r \ln\frac{r_2}{r_b} + \frac{1}{r}\int_{r_a}^{r}\rho r dr\right]dr}{\ln r_2/r_1}. \tag{37}$$

Equation (36) may be further simplified by integrating the last term by parts, which gives us

$$\bar{\phi} = a\ln\frac{r}{r_1} + \phi_1 - \frac{1}{\epsilon}\ln\frac{r}{r_a}\int_{r_a}^{r}\rho r dr + \frac{1}{\epsilon}\int_{r_a}^{r}\rho r \ln\frac{r}{r_a}dr. \tag{38}$$

Equation (38) is the required expression for the average potential at any radius in the charge region of the beam. If we substitute the power-series expansion of ρ and perform the indicated integrations, we arrive at a general explicit form for $\bar{\phi}$, from which the corresponding expressions for several special cases may be extracted quite easily.

$$\bar{\phi} = a\ln\frac{r}{r_1} + \phi_1 - \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{(n+2)^2}(r^{n+2} - r_a^{n+2}) + \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{n+2}r_a^{n+2}\ln\frac{r}{r_a} \tag{39}$$

and

$$a = \frac{\phi_2 - \phi_1 + \frac{Q}{2\pi\epsilon}\ln\frac{r_2}{r_b} + \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{(n+2)^2}(r_b^{n+2} - r_a^{n+2}) - \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n r_a^{n+2}}{n+2}\ln\frac{r_b}{r_a}}{\ln r_2/r_1}. \tag{40}$$

Evaluation of α and γ

When the beam is in equilibrium, i.e., when $r = r_0$, the expression derived above for the potential $\bar{\phi}$ and the radial field derived from it must agree with the potential and field derived from $\phi = \alpha \ln r/r_0 + \gamma$.

It is clear that γ is just the potential $\bar{\phi}$ evaluated at r_0. Thus,

$$\gamma = a\ln\frac{r_0}{r_1} + \phi_1 - \frac{1}{\epsilon}\ln\frac{r_0}{r_a}\int_{r_a}^{r_0}\rho r dr + \frac{1}{\epsilon}\int_{r_a}^{r_0}\rho r \ln\frac{r}{r_a}dr \tag{41}$$

and

$$\gamma = a\ln\frac{r_0}{r_1} + \phi_1 - \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{(n+2)^2}(r_0^{n+2} - r_a^{n+2}) + \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{n+2}r_a^{n+2}\ln\frac{r_0}{r_a}. \tag{42}$$

The condition on the radial eletric field at r_0 demands that

$$\frac{\alpha}{r_0} = \left[\frac{\partial\phi}{\partial r}\right]_{r=r_0}, \tag{43}$$

from which

$$\alpha = a - \frac{1}{\epsilon}\int_{r_a}^{r_0} \rho r dr \tag{44}$$

and

$$\alpha = a - \frac{1}{\epsilon}\sum_{n=0}^{\infty} \frac{\rho_n}{n+2}(r_0^{n+2} - r_a^{n+2}). \tag{45}$$

The second term on the right of these last two equations represents the charge per unit length between r_a and r_0. Evidently α must increase with r_0 whether it is positive or negative, and at the inner edge of the beam $\alpha_a = a$.

From (44), we have

$$\rho(r_0) = -\frac{\epsilon}{r_0}\frac{\partial\alpha}{\partial r_0}, \tag{46}$$

an equation used in the development of design formulas.

Once the values of α and γ are determined from the design equation, the requisite electrode voltages may be determined from (41) or (42) and (44) or (45).

At $r_0 = r_a$ the integrals in (41) are zero, and we obtain

$$\phi_1 = \gamma_a - \alpha_a \ln\frac{r_a}{r_1} \tag{47}$$

since a is just the value of α at r_a, i.e., $a = \alpha_a$. Equation (40) may be solved for $\phi_2 - \phi_1$,

$$\phi_2 - \phi_1 = \alpha_a \ln\frac{r_2}{r_1} - \left[\frac{Q}{2\pi\epsilon}\ln\frac{r_2}{r_b} + \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{(n+2)^2}(r_b^{n+2} - r_a^{n+2}) - \frac{1}{\epsilon}\sum_{n=0}^{\infty}\frac{\rho_n}{n+2} r_a^{n+2}\ln\frac{r_b}{r_a}\right]. \tag{48}$$

and by differentiation

$$\ddot r = -\frac{e}{m}\frac{\alpha}{r} + \frac{\Omega^2}{r^3} - \omega_H^2 r. \tag{50}$$

Let

$$r^2 = R \quad\text{and}\quad r_0^2 = R_0. \tag{51}$$

Then

$$\dot R = 2r\dot r \quad\text{and}\quad \ddot R = 2r\ddot r + 2\dot r^2. \tag{52}$$

Substituting (49) and (50) into (52), we obtain

$$\ddot R = \left[-2\frac{e}{m}\alpha - 4\frac{e}{m}(\gamma - V) + 4\Omega\omega_H\right] - 2\frac{e}{m}\alpha\ln\frac{R}{R_0} - 4\omega_H^2 R. \tag{53}$$

For values of R close to R_0 we may substitute

$$\ln\frac{R}{R_0} \approx \frac{R}{R_0} - 1, \tag{54}$$

which reduces (53) to the linear form

$$\ddot R = a_0 + a_1 R \tag{55}$$

where

$$a_0 = \left[-4\frac{e}{m}(\gamma - V) + 4\Omega\omega_H\right] = 4\omega_H^2 r_0^2\left[\sqrt{1+K\alpha} - K(\gamma - V)\right]$$

and

$$a_1 = -\left(4\omega_H^2 + 2\frac{e}{m}\frac{\alpha}{r_0^2}\right) = -4\omega_H^2\left(1 + \frac{K\alpha}{2}\right).$$

We multiply (55) by $2\dot R$ and integrate once to obtain

$$\dot R^2 = 2a_0 R + a_1 R^2 + a_2, \tag{56}$$

where

$$a_2 = -4\omega_H^2 R_0^2\left(1 + \frac{K\alpha}{2}\right),$$

obtained by substituting (49) into (56) and setting $r = r_0$. The integral of (56), neglecting the additive phase constant, may be written

$$\frac{R}{R_0} = \left[\frac{\sqrt{1+K\alpha} - K(\gamma - V)}{1 + \frac{K\alpha}{2}}\right] + \left\{\left[\frac{\sqrt{1+K\alpha} - K(\gamma - V)}{1 + \frac{K\alpha}{2}}\right]^2 - 1\right\}^{1/2}\cos\left(-2\omega_H\sqrt{1 + \frac{K\alpha}{2}}\,t\right). \tag{57}$$

Appendix II

Solution of the Differential Equation of Motion

From (5), expressing the radial motion of the electron, we have

$$\dot r^2 = -2\frac{e}{m}\alpha\ln\frac{r}{r_0} - 2\frac{e}{m}(\gamma - V) - \frac{\Omega^2}{r^2} - \omega_H^2 r^2 + 2\Omega\omega_H \tag{49}$$

If we define

$$\left[\frac{\sqrt{1+K\alpha} - K(\gamma - V)}{1 + \frac{K\alpha}{2}}\right] = 1 + \frac{x^2}{2}, \tag{58}$$

$$-2\omega_H\sqrt{1 + \frac{K\alpha}{2}} = \beta. \tag{59}$$

Equation (57) may be written

$$\frac{R}{R_0} = 1 + \frac{x^2}{2} + x\sqrt{1 + \frac{x^2}{4}}\cos\beta t. \tag{60}$$

The Iterated Network and its Application to Differentiators*

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Summary—By induction on the Cayley-Hamilton theorem, a convenient and compact expression for the transmission matrix of an n-tuply iterated, 4-terminal passive network is obtained.

The resulting expression is used for the analysis of the behavior of an iterated RC differentiating network. It is shown that such networks are of advantage in applications requring a very short effective time constant. The effective time constant of an n-tuply iterated structure is shown to be $2\tau/n(n+1)$, where $\tau=RC$ is the time constant of the unit structure.

THE PARTICULAR PROBLEM with which this paper is concerned is the analysis of a compound differentiating circuit formed by the iteration of a simple RC differentiator. Such an iterated circuit is of advantage since the effective time constant of the assembly is much less than that of the unit element. It is of particular importance in, for example, pulse-width discriminators where a very short time constant may be required.

In solving this problem, however, we shall also solve the general problem of the iteration of any passive linear 4-terminal network. We shall obtain a compact and convenient expression for the transmission matrix[1,2] of any interated structure in terms of n, the number of units included, and of the transmission matrix of the unit element. The resultant matrix can be used in the usual way for combining with other elements or with the application of the appropriate boundary conditions for obtaining the steady state or transient response.

The manipulation of the general matrix will be illustrated by the analysis of the iterated differentiator. In application this circuit is terminated in an open circuit across which the output voltage is obtained, and a pulse is applied to the input. In our analysis, we shall use a unit step wave instead of a pulse. The utility of the circuit as a differentiator is then measured by the rate at which the output voltage decays from the initial value. We shall define the effective time constant as that corresponding to the rate of decay immediately after the application of a step voltage.

A compound differentiator does not decay along a simple exponential curve. The instantaneous time constant is not constant. To obtain a measure of this deviation, we shall show the exact responses of the structures compound of one, two, and three units. We shall see that the responses of these are very nearly identical when reduced to a common scale.

THE GENERAL ITERATED NETWORK

Let us consider as a unit a 4-terminal passive linear network. We shall not assume that it is lossless. We wish to obtain a convenient representation for the compound network formed by connecting n such units together.

If we consider the transmission matrix, M, of the unit, then the transmission matrix of n-tuply iterated structure is the nth power of M. We wish, then, to express M^n conveniently in terms of M and of n. Such an expression can be obtained by induction using the Cayley-Hamilton theorem.

Let the transmission matrix of the unit network be

$$M = \begin{Vmatrix} A & jB \\ jC & D \end{Vmatrix}, \tag{1}$$

where A, B, C, and D are real (if lossless) or complex functions of frequency.

Since the unit is passive, M is unitary, that is,

$$AD + BC = 1. \tag{2}$$

The characteristic equation of M is obtained by introducing the variable, λ, and setting

$$\begin{vmatrix} A-\lambda & jB \\ jC & D-\lambda \end{vmatrix} = 0, \tag{3}$$

where the left-hand side is a determinant.

Expanding (3) and using (2), the characteristic equation of M is

$$\lambda^2 - 2\nu\lambda + 1 = 0, \tag{4}$$

where we have introduced the parameter

$$\nu = (A+D)/2. \tag{5}$$

The solutions of (4) for λ are the "characteristic values" of M.

The Cayley-Hamilton theorem states that, if the matrix M be substituted for λ, in its characteristic equation (4) a valid matrix equation is obtained. It is true, therefore, that

$$M^2 = 2VM - I. \tag{6}$$

Now the Tschebysheff[3] polynomials of the second kind and orders 1 and 2 are

$$U_1(\nu) = 1 \tag{7}$$

$$U_2(\nu) = 2\nu. \tag{8}$$

Subsequent orders are related by the recursion formula

$$U_{n+1}(\nu) = 2\nu U_n(\nu) - U_{n-1}(\nu). \tag{9}$$

Substituting (7) and (8) in (6),

$$M^2 = U_2(\nu)M - U_1(\nu)I. \tag{10}$$

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Let us assume that

$$M^n = U_n(\nu)M - U_{n-1}(\nu)I \tag{11}$$

is true up to and including a given n. It is true, certainly, for $n=2$. Multiplying (11) by M and using (10), (8), and then (9),

$$\begin{aligned} M^{n+1} &= (2\nu U_n - U_{n-1})M - U_n I \\ &= U_{n+1}(\nu)M - U_n(\nu)I \end{aligned} \tag{12}$$

Equation (11), is therefore, by induction, a general expression for the n'th power of M.

It is a compact and convenient expression for the n'th power of any 2-square unitary matrix (1) using the Tschebyscheff polynomials of the second kind in the variable ν defined by (5).

RC Differentiating Networks

We shall now consider the network obtained by iterating the usual RC differentiator. By applying the boundary condition to the general expression found for an iterated structure, we can obtain the response of this network to an arbitrary input. To study its behavior as a differentiator, we shall consider its response to a unit step input. We shall determine the instantaneous rate of decay of the response immediately after the application of voltage, measuring this decay as an "effective time constant." Finally, since no single time constant exactly describes the response, we shall plot the exact decay shape for doubly and triply iterated structures and compare the response with the ideal exponential curve. We shall show that the approximation of the response of the iterated structure as a single time constant is, in fact, sufficiently accurate for nearly all purposes. Let us now consider an iterated differentiating RC network. The basic differentiating unit is that shown in Fig. 1.

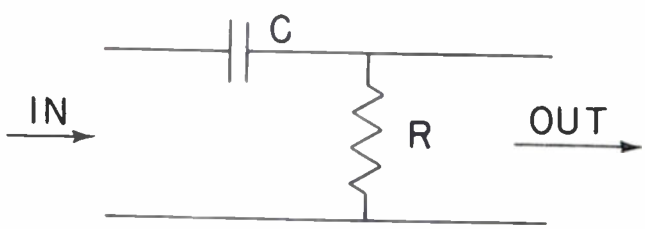

Fig. 1

The Unit RC Structure

The transmission matrix of this unit is[1]

$$M = \begin{Vmatrix} (1 + 1/p\tau) & R/p\tau \\ 1/R & 1 \end{Vmatrix}, \tag{13}$$

where $p=jw$, the complex frequency, and $\tau = RC$, the time constant. Then, from (5)

$$\nu = 1 + 1/2p\tau, \tag{14}$$

and the transmission matrix of the n-tuply iterated structure is given by (11).

Suppose we terminate the iterated structure in an open circuit. The boundary condition is that the output current is identically zero. If $E_i(p)$ is the input voltage as a function of p, and $E_0(p)$ the output, then[1] using the top left-hand term of the matrix M^n as expanded according to (11):

$$E_0(p) = \frac{E_i(p)}{(1 + 1/p\tau)U_n(\nu) - U_{n-1}(\nu)}. \tag{15}$$

Using (9), we can write this

$$E_0(p) = \frac{E_i(p)}{U_{n+1}(\nu) - U_n(\nu)}. \tag{16}$$

To analyze the behavior of this circuit as a differentiator, let us apply a unit step-voltage. The time domain input is

$$\begin{aligned} E_i(t) &= 0 \quad \text{if} \quad t < 0 \\ &= 1 \quad \text{if} \quad t > 0. \end{aligned} \tag{17}$$

The Fourier transform[4,5] of this is

$$E_i(p) = 1/p. \tag{18}$$

The time-domain response to the step input is, then, the inverse Fourier transform of

$$E_0(p) = \frac{1/p}{U_{n+1}(\nu) - U_n(\nu)}. \tag{19}$$

This can be evaluated by the usual method[4,5] for any specific case.

We have, however, defined the "effective time constant," τ_n, as that operative immediately after the application of voltage.

$$1/\tau_n = -\lim_{t\to 0+} \frac{dE_0(t)}{dt}. \tag{20}$$

It is convenient to remove the discontinuity in $E(t)$ by subtracting the unit step from it.

$$E_0'(t) = E_0(t) - E_i(t). \tag{21}$$

This does not alter (20)

$$1/\tau_n = -\lim_{t\to 0+} \frac{dE_0'(t)}{dt}. \tag{22}$$

We can now apply the limiting and differentiating theorems of the Fourier transform, as follows:

$$\begin{aligned} 1/\tau_n &= -\lim_{t\to 0+} \frac{d}{dt} F^{-1}\{E_0'(p)\} \\ &= -\lim_{t\to 0+} F^{-1}\{pE_0'(p)\} \\ &= -\lim_{p\to +\infty} p^2 E_0'(p), \end{aligned} \tag{23}$$

where $F^{-1}\{-\}$ means the inverse Fourier transform.

Substituting (19) and (18) in (21) and (23),

$$\frac{1}{\tau_n} = -\lim_{p\to+\infty} p\left(\frac{1}{U_{n+1}(\nu) - U_n(\nu)} - 1\right). \tag{24}$$

Now, as $p \to \infty$, $\nu \to 1$ by (14). The Tschebyscheff polynomial of the second kind can be expanded as a terminating series about this point.

$$U_n(\nu) = n + \frac{2}{3!} n(n^2 - 1^2)(\nu - 1) + \frac{2^2}{5!} n(n^2 - 1^2)(n^2 - 2^2)(\nu - 1)^2 + \cdots . \quad (25)$$

Using this expansion in (24) and retaining only the terms of the first order in $(\nu - 1)$,

$$1/\tau_n = \frac{n(n+1)}{2\tau}. \quad (26)$$

Hence the effective time constant, τ_n, of a network formed by the n-tuple iteration of a differentiator of time constant τ is τ reduced by the factor $2/n(n+1)$

$$\tau_n = (2/n(n+1))\tau. \quad (27)$$

This effective time constant cannot precisely describe the behavior of the multiple structure. That it is an accurate approximation is shown by the curves of Fig. 2. These have been computed by the usual methods of transient analysis. They show shape of response to a unit step input of single, double, and triple cases. To normalize the time base, we have, in each case, used t/τ_n as the variable. The solid line, being the single unit case, represents an exactly exponential case. The dashed line (– – –), representing the double case, and the dash-dot line (–·–), the triple case, are seen to deviate only slightly from this. The approximation in terms of τ_n is sufficiently accurate for most purposes.

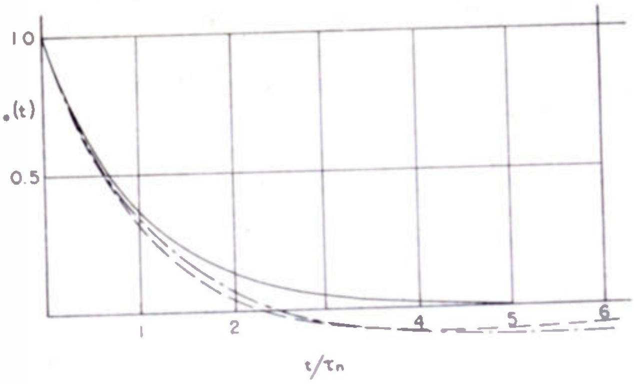

Fig. 2

Conclusion

We have obtained a compact and convenient general expression for the transmission matrix of any n-tuply iterated 4-terminal passive linear network in terms of matrix of the unit network.

We have applied this to the iteration of an RC differentiator. We have shown that such a compound differentiator is useful in that it has a very short effective time constant compared to that of the unit structure. In fact, if $RC = \tau$, then τ_n, the effective time constant of the n-tuply iterated differentiator, is $2\tau/n(n+1)$. For even small values of n, this is a large reduction from τ.

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The Approximation with Rational Functions of Prescribed Magnitude and Phase Characteristics*

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Summary—**A successive-approximations method is applied to the selection of network functions having desired magnitude and phase variation with frequency. The first approximation, the first set of pole and zero locations, can be selected on the basis of known solutions to similar problems or through use of a set of curves. In succeeding approximations the pole and zero locations are adjusted to decrease the deviation of the earlier approximations from the desired characteristics. The process adjusts the magnitude and phase characteristics simultaneously. Its flexibility permits accommodation of practical constraints not possible with other methods.**

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† Bell Telephone Laboratories, Murray Hill, N. J. On leave from the Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge 39, Mass.

A. Introduction

THE APPROXIMATION PROBLEM is the problem of selecting a rational function, identifiable as the response function of a realizable network, which approximates prescribed magnitude and phase characteristics within tolerable limits. The method of its solution presented here is one of successive approximations, and it has a number of features of practical value. Constraints on the form of the function to be obtained can be imposed easily. As a result, one can accommodate practical design constraints not possible with other methods. One chooses the complexity of the approximating function (the number of poles and zeros) at the outset, and successive adjustments always leave the number of poles and zeros unchanged. One can approximate to magnitude and phase characteristics simultaneously (not over the whole frequency spectrum with minimum-phase functions, of course). He can obtain functions which fit the prescribed characteristics more closely over some regions of frequency at the expense of poorer fits in other regions.

The method of selection employed consists of a preliminary choice of pole and zero positions followed by successive adjustments of these positions. The first approximation can be made on the basis of known solutions to similar problems which are roughly suitable to the given problem, or it can be made purely on the basis of a set of curves. One locates poles and zeros to conform to any constraints which are prescribed. Ordinarily, the preliminary function chosen, or first estiate, only roughly approximates the desired characteristics. The successive adjustments, specified in terms of shifts in pole and zero locations, are made to reduce the deviation of the trial characteristics from the desired characteristics. At the beginning of the adjustment procedure, simple calculations, based on curves presented, effect rapid improvement in the approximation. In many practical cases the simple adjustment procedure is sufficient to obtain a satisfactory approximation. Final adjustments, if they are necessary, are made on the basis of a much more powerful, but also more complicated, algebraic technique.

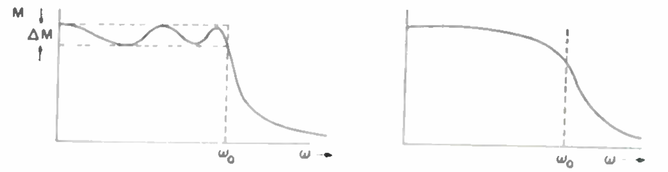
B. The First Approximation

Any response function of a lumped-element network, being a rational function of frequency, can be expressed as

$$F(\lambda) = \frac{a_0 + a_1\lambda + \cdots + a_m\lambda^m}{b_0 + b_1\lambda + \cdots + b_n\lambda^n} = \frac{h(\lambda - \lambda_{01})(\lambda - \lambda_{02}) \cdots (\lambda - \lambda_{0m})}{(\lambda - \lambda_{p1})(\lambda - \lambda_{p2}) \cdots (\lambda - \lambda_{pn})}. \tag{1}$$

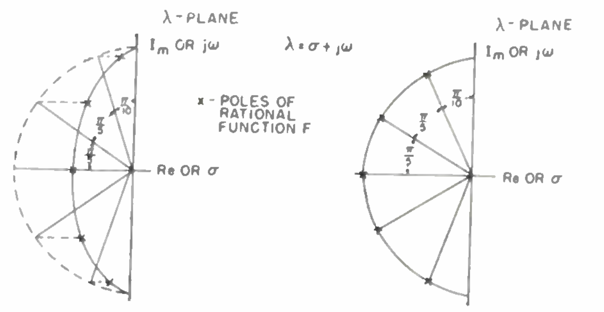
The quantities $\lambda_{01}, \lambda_{02} \cdots \lambda_{0m}$ are the zeros of the function and $\lambda_{p1}, \lambda_{p2} \cdots \lambda_{pn}$ are the poles. The response function is apparently specified within a constant multiplier by specification of the poles and zeros of the function, and it is the pole and zero locations on which attention is directed in the following.

When a designer uses the solution of a similar problem in the selection of the first trial positions of the poles and zeros, wide experience and ingenuity lead to a better first estimate and effect a saving of labor in the subsequent adjustment procedure. The well-known Tschebyscheff and Butterworth approximations[1,2,3] to the ideal low-pass filter characteristic are typical examples of approximations which suggest suitable first estimates for related problems. The characteristics obtained with these functions are sketched in Fig. 1. The maps of pole locations in the complex plane for the two types of functions (of fifth order) are given in Fig. 2. In both of these types the function has no internal zeros. The geometric figure on which the poles shown in Fig. 2(a) lie is an ellipse. The cut-off frequency is only slightly less than the value of ω at the intersection of the ellipse and the axis of imaginaries. Simple relationships exist between the level and tolerance of $|F|$ of Fig. 1(a) and the size of the minor axis of the ellipse of Fig. 2(a). For present purposes, it is sufficient to state that as the minor axis of the ellipse is decreased (the poles brought nearer the imaginary axis always along lines parallel to the real axis), the level of F in the pass band increases as does the tolerance. Increasing the number of poles[4] increases the level of $|F|$ for a fixed ellipse and the number of cycles of variation of $|F|$ in the pass band while it decreases the tolerance.



(a) TSCHEBYSCHEFF CHARACTERISTIC (b) BUTTERWORTH CHARACTERISTIC

Fig. 1—The Tschebyscheff and Butterworth approximations for low-pass filters.



(a) TSCHEBYSCHEFF MAP (b) BUTTERWORTH MAP

Fig. 2—Maps of poles for Tschebyscheff and Butterworth approximations.

$$F = \frac{1}{(\lambda - \lambda_{p1})(\lambda - \lambda_{p2}) \cdots (\lambda - \lambda_{p5})}.$$

From Fig. 2 it is apparent that the Butterworth approximation is a degenerate form of the Tschebyscheff approximation in which the ellipse becomes a circle. One observes from Figs. 1 and 2 that small shifts in the position of the critical frequencies of a function cause marked changes in the characteristics of the function. This fact suggests the study of the relationship between changes in pole positions and changes in the function characteristics. This study and a method of successive approximations arising from it constitute the results given in this paper.

[1] S. Darlington, "Synthesis of reactance four poles," *Jour. Math. Phys.*, vol. 18, pp. 257–353; 1935.

[2] S. Butterworth, "On the theory of filter amplifiers," *Exp. Wireless and Wireless Eng.*, vol. 7, pp. 536–41; 1930.

[3] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Co., Inc., New York, N. Y., Chapt. 9; 1948.

[4] For odd numbers of poles, one always lies on the negative real axis; for even numbers of poles, none lie on the real axis.

The analogy between a two-dimensional field and the logarithm of a rational function has been exploited in connection with the approximation problem.[5,6] The selection of poles and zeros of a rational function whose magnitude has a certain variation along the real-frequency axis is equivalent to the selection of charged positive and negative filaments which generate a corresponding potential along a line which can be identified with the real-frequency axis. Physical intuition based on the analogy is helpful in selecting the first approximation in the method presently described.

As explained in the next section, one employs normalized curves to compute the magnitude and phase characteristics of the successive trial functions. These curves may also be employed as a guide in the selection of the first approximation.

C. Computation of Phase and Magnitude Characteristics for Specified Pole and Zero Positions

Normalized curves are used for the computation of the characteristics of the successive trials. The same curves may be used as a guide in the choice of the first approximation. The process of normalization is easily understood by considering the logarithm of F [F is defined in (1)].

$$\ln F = \ln |F| + j \operatorname{Arg} F \tag{2}$$

$$\ln |F| = \ln h + \sum_{i=1}^{m} \ln |\lambda - \lambda_{0i}| - \sum_{i=1}^{n} \ln |\lambda - \lambda_{pi}| \tag{3}$$

$$\operatorname{Arg} F = \sum_{i=1}^{m} \operatorname{Arg} (\lambda - \lambda_{0i}) - \sum_{i=1}^{n} \operatorname{Arg} (\lambda - \lambda_{pi}). \tag{4}$$

The consideration of the logarithm of F is convenient in that it makes the contribution of each pole and zero separately evident. One observes that the sums in (3) and also in (4) are really made up of only two kinds of terms, terms corresponding to conjugate complex critical frequencies (poles or zeros) and terms corresponding to real critical frequencies. The distinction between zeros and poles is merely the distinction between plus and minus signs associated with the components in the sum.

For the complex critical frequencies, one may write the following from (3) and (4)[7]

$$\ln |F_c| = \ln |\lambda - \lambda_c| |\lambda - \bar{\lambda}_c| \tag{5}$$

and

$$\operatorname{Arg} F_c = \operatorname{Arg} (\lambda - \lambda_c)(\lambda - \bar{\lambda}_c), \tag{6}$$

in which

$$\lambda_c = \sigma_c + j\omega_c. \tag{7}$$

If one divides F_c by ω_c^2 and replaces λ by $j\omega$, as he is ordinarily interested only in real frequencies, (5) and (6) may be written

$$\ln \frac{|F_c|}{\omega_c^2} = \ln \left| \frac{j\omega}{\omega_c} - \frac{\sigma_c}{\omega_c} - j \right| \left| \frac{j\omega}{\omega_c} - \frac{\sigma_c}{\omega_c} + j \right| \tag{8}$$

and

$$\operatorname{Arg} F_c = \operatorname{Arg} \left(\frac{j\omega}{\omega_c} - \frac{\sigma_c}{\omega_c} - j \right) \left(\frac{j\omega}{\omega_c} - \frac{\sigma_c}{\omega_c} + j \right). \tag{9}$$

Clearly $\ln |F_c|$ and $\operatorname{Arg} F_c$ are easily found as functions of ω for any λ_c if one has simply plots of $\ln |F_c| / \omega_c^2$ and $\operatorname{Arg} F_c/\omega_c^2$ as functions of ω/ω_c for appropriate values of the parameter σ_c/ω_c. Fig. 3 gives sketches of families of curves corresponding to (8) and (9).[8] An accurate set of curves is given in R.L.E. Report No. 145[9] corresponding to Fig 3 as well as sets of curves corresponding to Figs. 4, 7, 8, and 9. Through use of the curves one can readily obtain plots of $\ln |F_c|$ and $\operatorname{Arg} F_c$ as functions of ω for any complex value of λ_c. Fig. 3 is instruc-

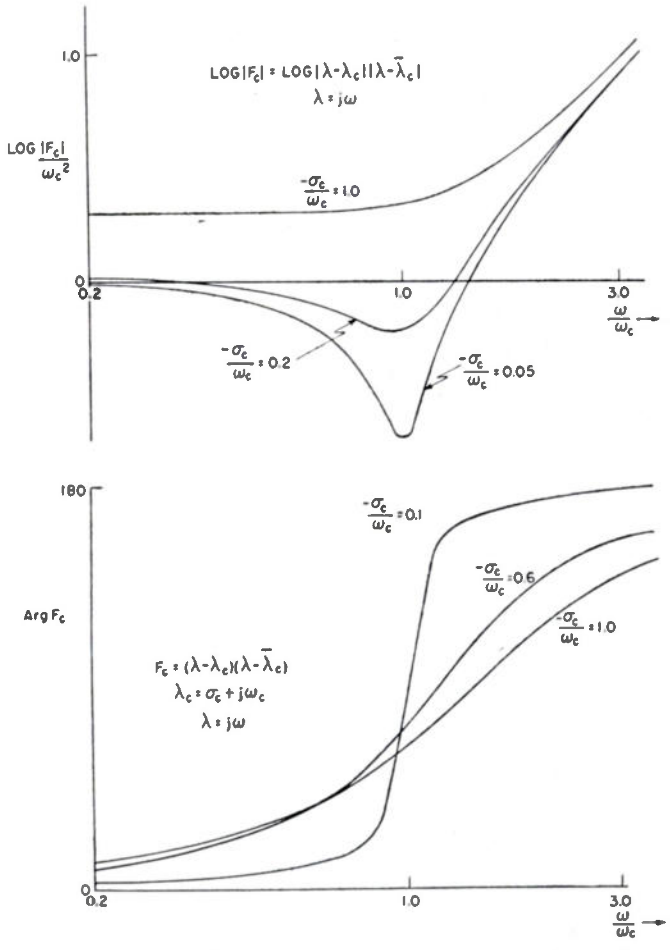


Fig. 3—Log $|F_c/\omega_c^2|$ and Arg F_c sketched as functions of ω/ω_c for a range of values of $-\sigma_c/\omega_c$.

[5] W. W. Hansen and O. C. Lundstrom, "Experimental determination of impedance functions by use of an electrolytic tank," Proc. I.R.E., vol. 33, pp. 528–534; 1945.

[6] J. G. Linvill, "An Experimental Approach to the Approximation Problem for Driving-Point and Transfer Functions," S. M. Thesis, M.I.T., Cambridge, Mass.; 1945.

[7] In the following, F_c will always represent the product of a pair of factors with complex conjugate zeros, λ_c and $\bar{\lambda}_c$.

[8] All logarithms in the curves are to the base 10 and all arguments are given in degrees. In the text, log refers to the base 10 and ln refers to the base ϵ; Arg is the angle in radians unless specifically stated otherwise.

[9] J. G. Linvill, "The Selection of Network Functions to Approximate Prescribed Frequency Characteristics," Technical Report No. 145, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; March 14, 1950.

tive in connection with the approximation problem. It is apparent that critical frequencies close to the imaginary axis in the complex frequency plane cause much more violent changes in the magnitude of the function for nearby real frequencies than do critical frequencies removed a considerable distance from the imaginary axis. Accordingly, if one is choosing a function which is to exhibit abrupt changes in magnitude at some range of frequencies, the suggestion is definite that critical frequencies with small displacement from the imaginary axis be placed near the range of $j\omega$ where the change is desired. From Fig. 3, also, it is clear that such critical frequencies cause rapid phase shift at the same time they cause an abrupt change in magnitude. This situation is a manifestation of the implicit relationship between magnitude and phase.

The second kind of terms mentioned in connection with (3) and (4) is that representing a critical frequency falling on the real axis of the complex frequency plane.[10]

$$\ln |F_r| = \ln |\lambda - \sigma_r|, \tag{10}$$

where σ_r is real.

$$\operatorname{Arg} F_r = \operatorname{Arg}(\lambda - \sigma_r). \tag{11}$$

If one divides F_r by $-\sigma_r$ and sets $\lambda = j\omega$, the result is

$$\ln \left| \frac{F_r}{\sigma_r} \right| = \ln \left| \frac{j\omega}{-\sigma_r} + 1 \right| \tag{12}$$

and

$$\operatorname{Arg} \frac{F_r}{-\sigma_r} \ (= \operatorname{Arg} F_r \text{ if } \sigma_r \text{ is negative})$$

$$= \operatorname{Arg}\left(\frac{j\omega}{-\sigma_r} + 1\right). \tag{13}$$

The sketch of Fig. 4 indicates the form of variation of $\ln|F_r|$ and Arg F_r for any real critical frequency, σ_r. The sketches of Figs. 3 and 4 are given for critical frequencies in the left half-plane. The same curves are useful for critical frequencies in the right half-plane since the logarithm of the magnitude of F_c or F_r is unchanged if the sign of σ_c or σ_r is changed, and only the sign of the argument of F_c or F_r changes when the sign of σ_c or σ_r is changed.

D. An Illustrative Example

To illustrate the choice of the first approximation, a function of the form

$$F = \frac{1}{\lambda^3 + a_2\lambda^2 + a_1\lambda + a_0}$$

$$= \frac{1}{(\lambda - \sigma_1)(\lambda - \sigma_2 - j\omega_2)(\lambda - \sigma_2 + j\omega_2)} \tag{14}$$

will be chosen. The characteristics to be approximated

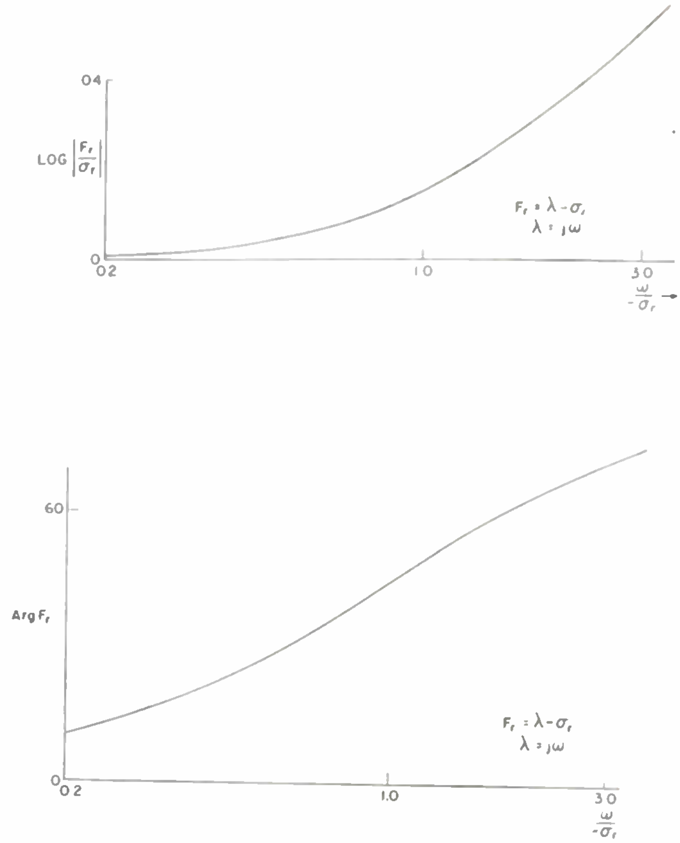


Fig. 4—Log $|F_r/\sigma_r|$ and Arg F_r sketched as functions of $\omega/-\sigma_r$

are described in Fig. 5. It will be required that the approximating function have three poles, have approximately a linear phase shift over the range $0 \leq \omega \leq 1$, and have a logarithm of magnitude which decreases linearly (approximately) with frequency such that $|F|$ at $\omega = 1$ is half of $|F|$ at $\omega = 0$. The magnitude of the argument of F at $\omega = 1$ is immaterial so long as its variation with frequency is linear. The level of $\log_{10}|F|$ is not specified. A large number of functions of the form represented in (14) will have characteristics of approximately the nature indicated in Fig. 5. Some of them approximate the magnitude characteristic well and the phase characteristic badly; some approximate the phase characteristic well and the magnitude characteristic badly; others fall between these classes, approximating both phase and magnitude reasonably well. Before a final choice of approximating function can be made, some measure of quality of the approximation must be formulated. Such a measure of quality is given at a later point; at this point it is necessary only to choose any function of the form of (14) which has roughly the characteristics desired and to proceed from it as a starting point.

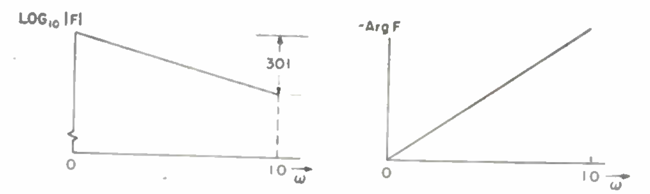


Fig. 5—Partial specification of requirements for illustrative example

[10] In the following, F_r will always represent a factor with a single real zero, σ_r.

Knowledge of the functions used for Butterworth filter characteristics suggests that a function with poles distributed on a semicircle as in Fig. 2(b) will give roughly the kind of characteristics desired. Hence a first approximation is

$$F(\lambda) = \frac{1}{(\lambda+1.00)(\lambda+0.50+j0.866)(\lambda+0.50-j0.866)}. \quad (15)$$

Fig. 6, which is a plot of log $|F|$ and Arg F for (15), indicates that the first approximation shows a too

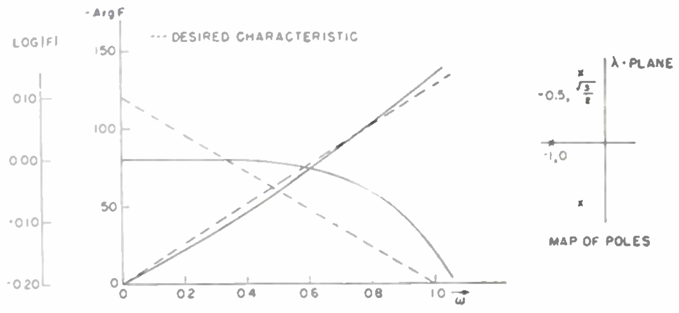


Fig. 6—Log $|F|$ and Arg F for the first approximation to the prescribed characteristics of Fig. 5.

slow decrease in magnitude with frequency. Hence some motion of the poles from their semicircular distribution is desirable. The choice of the first approximation is quite arbitrary, and different first approximations will lead to substantially the same end result. A close first approximation shortens the adjustment process, which is the next step in the procedure.

E. Successive Adjustments

Once the first approximation is chosen, attention is shifted to the problem of determining the changes in pole and zero positions required to effect the desired changes in the approximating function. One observes that the number of variable quantities which may be adjusted is equal to the total number of poles and zeros. In the illustrative example above, these are three adjustable quantities, σ_1, σ_2, and ω_2. In that instance one wishes to choose the combination of values for $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$, which together change the characteristics from those of Fig. 6 to characteristics much nearer those of Fig. 5. A study of the influence of elementary shifts of the critical frequencies, $\Delta\sigma_1$ by itself, $\Delta\sigma_2$ by itself, and $\Delta\omega_2$ by itself, is an enlightening guide to the choice of the combination. This fact suggests the study of

$$\frac{\partial \log|F|}{\partial\sigma_1}, \quad \frac{\partial \operatorname{Arg} F}{\partial\sigma_1}, \quad \frac{\partial \log|F|}{\partial\sigma_2},$$

and so on since

$$\Delta \log|F| = \int_0^{\Delta\sigma_1} \frac{\partial \log|F|}{\partial\sigma_1} d\sigma_1 + \int_0^{\Delta\sigma_2} \frac{\partial \log|F|}{\partial\sigma_2} d\sigma_2 + \int_0^{\Delta\omega_2} \frac{\partial \log|F|}{\partial\omega_2} d\omega_2$$

$$\cong \text{(for small changes)} \frac{\partial \log|F|}{\partial\sigma_1}\Delta\sigma_1 + \frac{\partial \log|F|}{\partial\sigma_2}\Delta\sigma_2 + \frac{\partial \log|F|}{\partial\omega_2}\Delta_{\omega_2}. \quad (16)$$

Similarly,

$$\Delta \operatorname{Arg} F \cong \text{(for small changes)} \frac{\partial \operatorname{Arg} F}{\partial\sigma_1}\Delta\sigma_1 + \frac{\partial \operatorname{Arg} F}{\partial\sigma_2}\Delta\sigma_2 + \frac{\partial \operatorname{Arg} F}{\partial\omega_2}\Delta\omega_2. \quad (17)$$

Fortunately, the use of the logarithm of F has separated the influence of individual critical frequencies as indicated by (3), and one has, for instance,

$$\frac{\partial \log|F|}{\partial\sigma_1} = \frac{-\partial \log|\lambda - \sigma_1|}{\partial\sigma_1}. \quad (18)$$

From (18) it is apparent that the derivative functions can be normalized and presented in curves just as was done for the logarithm and argument curves in section C. Further discussion of the use of the derivative curves will be given subsequent to the statement of expressions for the derivative functions.

F. Normalization of the Derivative Functions

From the foregoing, one appreciates that the functions which it is sufficient for all cases to consider are

$$\frac{\partial \log|F_c|}{\partial\sigma_c}, \quad \frac{\partial \log|F_c|}{\partial\omega_c}, \quad \frac{\partial \log|F_r|}{\partial\sigma_r}; \quad \frac{\partial \operatorname{Arg} F_c}{\partial\sigma_c},$$

$$\frac{\partial \operatorname{Arg} F_c}{\partial\omega_c}, \quad \text{and} \quad \frac{\partial \operatorname{Arg} F_r}{\partial\sigma_r}$$

[see (5), (6), (7), (10), and (11)]. Evaluation of the derivatives is an elementary exercise when one employs the relations

$$\ln|F_c(\lambda)|_{\lambda=j\omega} = \tfrac{1}{2}\ln F_c(\lambda)F_c(-\lambda)|_{\lambda=j\omega} \quad (19)$$

and

$$\operatorname{Arg} F_c(\lambda)|_{\lambda=j\omega} = \frac{1}{2j}\ln\frac{F_c(\lambda)}{F_c(-\lambda)}\Bigg|_{\lambda=j\omega}. \quad (20)$$

Carrying through the steps of differentiating and normalizing gives for $\partial \log|F_c|/\partial\sigma_c$

$$\omega_c \frac{\partial \log|F_c|}{\partial\sigma_c} = \frac{2}{2.305}\,\frac{\frac{\sigma_c}{\omega_c}\left[\left(\frac{\sigma_c}{\omega_c}\right)^2+1+\left(\frac{\omega}{\omega_c}\right)^2\right]}{\left|\frac{F_c}{\omega_c^{\,2}}\right|^2}. \quad (21)$$

A sketch of $\omega_c\partial\log|F_c|/\partial\sigma_c$ as a function of ω/ω_c for various values of the parameter σ_c/ω_c is given in Fig. 7. This figure reveals that shifts in the displacement from the imaginary axis of poles or zeros influence the magnitude of the function most in a frequency range nearest the critical frequency being shifted. Through use of accurate charts corresponding to Fig. 7 one can predict

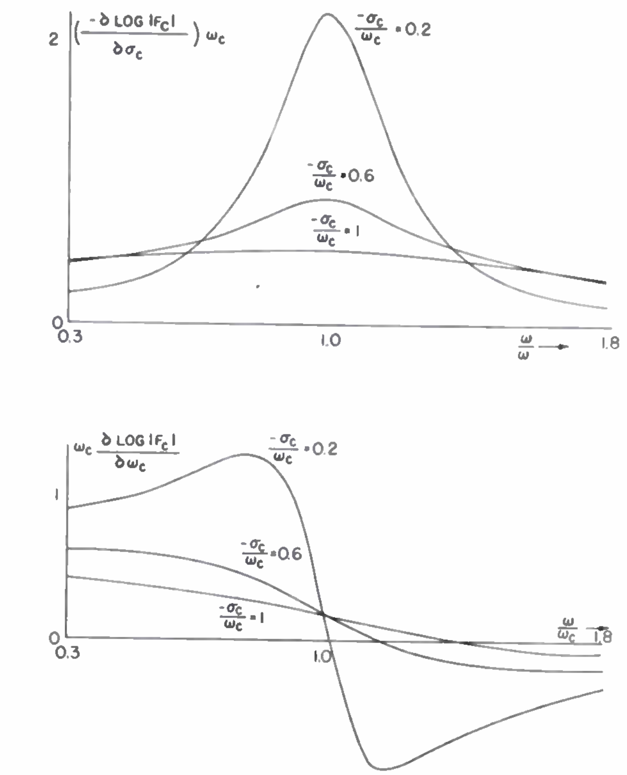


Fig. 7—Sketches of

$$\omega_c \frac{\partial \log |F_c|}{\partial \sigma_c} \quad \text{and} \quad \omega_c \frac{\partial \log |F_c|}{\partial \omega_c}$$

as functions of ω/ω_c for various values of σ_c/ω_c.

accurately the influence of a shift of a critical frequency. Alternately, one has suitable information for prescribing what kind of shift and how large a shift is necessary to obtain a required change in the magnitude characteristic.

Through a similar procedure one can obtain

$$\omega_c \frac{\partial \log |F_c|}{\partial \omega_c} = \frac{2}{2.305} \frac{\left[1 + \left(\frac{\sigma_c}{\omega_c}\right)^2 - \left(\frac{\omega}{\omega_c}\right)^2\right]}{\left|\frac{F_c}{\omega_c^2}\right|^2}. \tag{22}$$

Equation (22) and the corresponding sketch in Fig. 7 indicate the influence on the magnitude characteristic of a shift parallel to the imaginary axis of a conjugate pair of critical frequencies.

After a straightforward manipulation, one has for $\partial \operatorname{Arg} F_c/\partial \sigma_c$

$$\omega_c \frac{\partial \operatorname{Arg} F_c \text{ (degrees)}}{\partial \sigma_c} = 57.3 \times 2 \frac{\frac{\omega}{\omega_c}\left[\left(\frac{\sigma_c}{\omega_c}\right)^2 - 1 + \left(\frac{\omega}{\omega_c}\right)^2\right]}{\left|\frac{F_c}{\omega_c^2}\right|^2}. \tag{23}$$

Through a similar procedure

$$\omega_c \frac{\partial \operatorname{Arg} F_c \text{ (degrees)}}{\partial \omega_c} = \frac{57.3 \times 3 \frac{\omega}{\omega_c} \frac{\sigma_c}{\omega_c}}{\left|\frac{F_c}{\omega_c^2}\right|^2}. \tag{24}$$

Fig. 8 shows sketches of $\omega_c(\partial \operatorname{Arg} F_c/\partial \sigma_c)$ and $\omega_c(\partial \operatorname{Arg} F_c/\partial \omega_c)$ for various values of the parameter σ_c/ω_c. Figs. 7 and 8 give complete information on the influence on magnitude and phase characteristics for any shifts of a conjugate pair of complex critical frequencies.

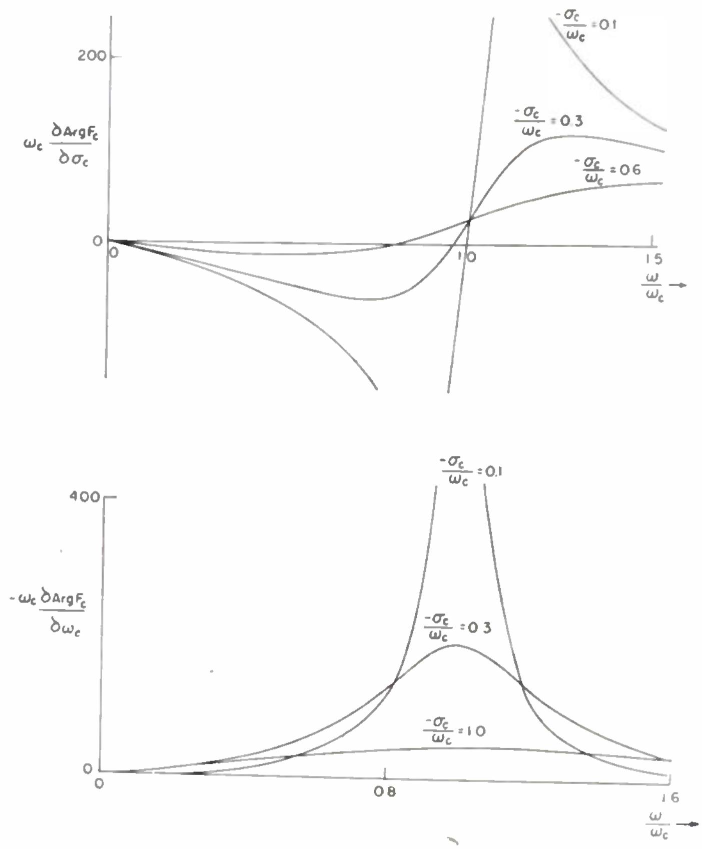


Fig. 8—Sketches of $\omega_c(\partial \operatorname{Arg} F_c/\partial \sigma_c)$ and $\omega_c(\partial \operatorname{Arg} F_c/\partial \omega_c)$ as functions of ω/ω_c for various values of σ_c/ω_c.

In an analogous manner to that just indicated for conjugate critical frequencies, the corresponding derivative functions for real critical frequencies can be determined.

$$-\sigma_r \frac{\partial \log |F_r|}{\partial \sigma_r} = \frac{-1}{2.305\left[1 + \left(\frac{\omega}{\sigma_r}\right)^2\right]} \tag{25}$$

$$-\sigma_r \frac{\partial \operatorname{Arg} F_r}{\partial \sigma_r} = \frac{57.3 \frac{\omega_c}{-\sigma_r}}{1 + \left(\frac{\omega}{\sigma_r}\right)^2}. \tag{26}$$

Fig. 9 presents sketches of $-\sigma_r(\partial \text{Arg } F_r/\partial\sigma_r)$ and $-\sigma_r(\partial \log |F_r|/\partial\sigma_r)$.

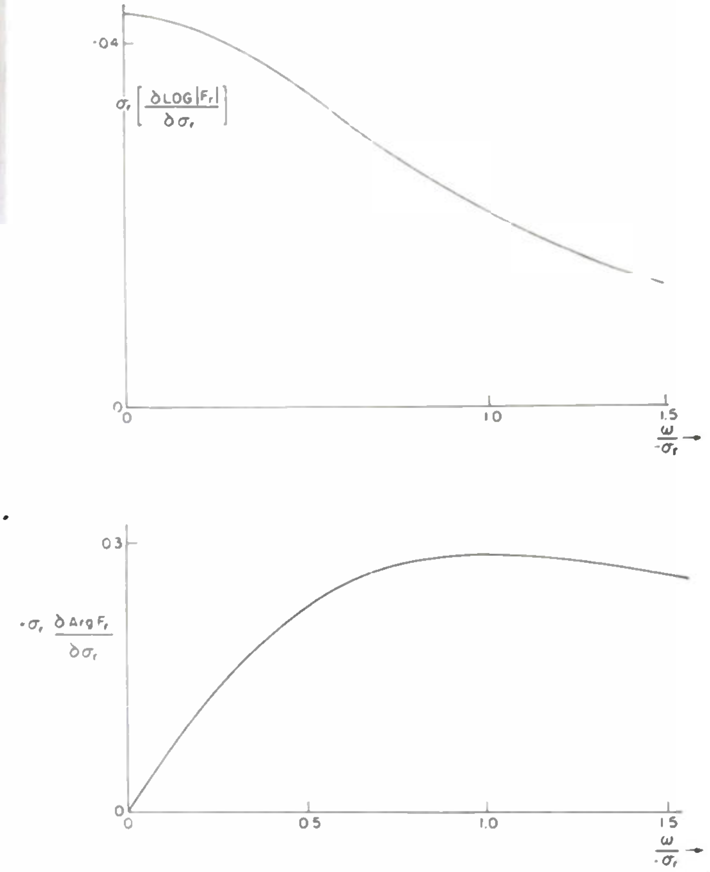


Fig. 9—Sketches of $-\sigma_r(\partial \log |F_r|/\partial\sigma_r)$ and $-\sigma_r(\partial \text{Arg } F_r/\partial\sigma_r)$ as functions of $\omega/-\sigma_r$.

G. Illustration of the First Stage of the Adjustment Procedure

The preliminary step to stating what changes should be made in the pole and zero positions is the sketching of the derivative functions for the case at hand. A convenient medium to illustrate the process is the illustrative example of section D. One observes that the deviation of the magnitude characteristic from the desired is much more serious than that of the phase characteristic. Accordingly, it is appropriate for the first adjustment to base the choice of changes on the deviation of the magnitude characteristic alone. Fig. 10 shows plots of the desired change in the magnitude characteristic and the derivative functions. On the basis of Fig. 10 the size changes to be used to give the desired changes in the magnitude characteristic can be estimated. For instance, it is clear that any positive increment in σ_1 will cause an improvement since the first estimate gives insufficient amplification in the low range. Rough calculations suggest the values $\Delta\sigma_1 = 0.2$, $\Delta\sigma_2 = -0.2$ and $\Delta\omega_2 = -0.1$. Fig. 11 shows the second approximation corresponding to these changes. As long as it is obvious, without a more precise method, how to obtain improvement in the characteristics, one continues to use this simple method of adjustment. Fortunately, for many problems, specifications allow considerable tolerance and sufficiently close approximations can be arrived at without going to a more accurate method.

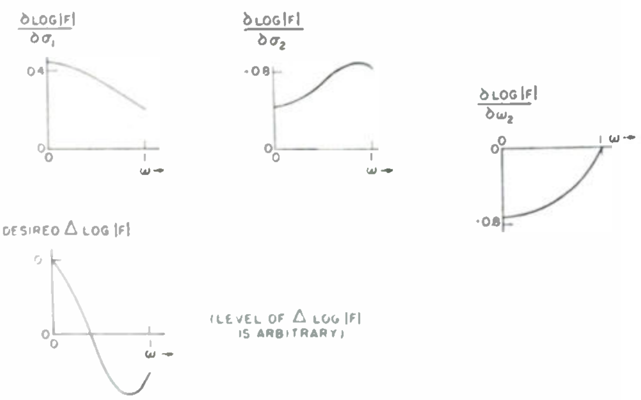


Fig. 10—Derivative curves and desired change in magnitude characteristic for the example of Fig. 6.

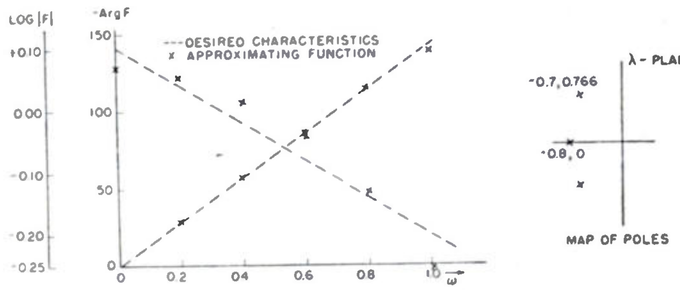


Fig. 11—Second approximation to the characteristics of Fig. 5.

Actually, as the present example illustrates, the simple procedure carefully applied leads to a very nice approximation. After a few more adjustments of the nature of that illustrated above, one arrives at the characteristics of Fig. 12.

The method of adjustment illustrated in the foregoing paragraphs becomes ineffective in complicated situations. For instance, the derivation of the characteristics at one frequency may indicate one shift of a pole, while the deviation at a different frequency may indicate an entirely different shift. In such a situation it is difficult to prescribe appropriate shifts of the critical frequencies. If one tries to prescribe the shifts by the simple procedure illustrated, he finds himself in the predicament of one trying to solve a set of simultaneous equations one at a time. A further defect of the simple meth-

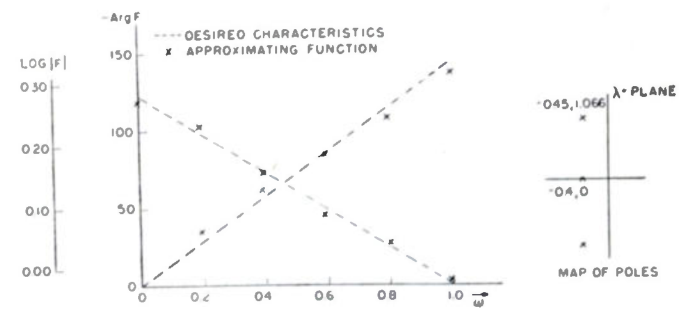


Fig. 12—Approximation to characteristics of Fig. 5 obtained by the first stage of the adjustment procedure.

od of adjustment lies in the fact that one stops the procedure when further improvement cannot be obtained. He may be uncertain as to whether or not someone else can obtain a better approximation by stumbling upon a more appropriate combination of critical-frequency shifts.

At this point a more precise method of adjustment is illustrated which will handle the difficult cases. Moreover, it gives the designer a very strong signal when he has arrived at the "best" approximation (he will have had to define a criterion for "best," however).

II. Final Adjustments of the Positions of Poles and Zeros

The problem which must be solved more adequately in the final adjustment of the position of poles and zeros is the choice of $\Delta\sigma$'s and $\Delta\omega$'s in (16) and (17). The means employed to prescribe the shifts must account for the effect of all shifts simultaneously.

At this point it is helpful to consider the derivative curves and to observe that changes of 10 to 20 per cent in the pole positions cause rather small changes in the derivative. This fact indicates that in (16) and (17) the derivatives may be considered constant for small changes. Equations 16 and 17,

$$\Delta \log |F| \cong \frac{\partial \log |F|}{\partial \sigma_1}\Delta\sigma_1 + \frac{\partial \log |F|}{\partial \sigma_2}\Delta\sigma_2 + \frac{\partial \log |F|}{\partial \omega_2}\Delta\omega_2, \tag{16}$$

$$\Delta \operatorname{Arg} F \cong \frac{\partial \operatorname{Arg} F}{\partial \sigma_1}\Delta\sigma_1 + \frac{\partial \operatorname{Arg} F}{\partial \sigma_2}\Delta\sigma_2 + \frac{\partial \operatorname{Arg} F}{\partial \omega_2}\Delta\omega_2, \tag{17}$$

suggest a mathematically convenient formulation of the problem of final adjustment, which is most simply understood by a reference to the problem of Fig. 12. In Fig. 13 are shown plots of the derivative functions corresponding to the critical-frequency positions of Fig. 12. Fig. 14 shows plots of the desired changes in the magnitude and phase from that illustrated in Fig. 12. (Note that a constant change in level or a linear change in argument may be prescribed in addition to the changes of Fig. 14.) The problem of the choice of $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$ is the choice of these quantities such that, when the functions of (a) of Fig. 13 are multiplied by the proper $\Delta\sigma_1$, those of (b) are multiplied by the proper $\Delta\sigma_2$ and those of (c) are multiplied by the proper $\Delta\omega_2$ and the sums formed in accordance with (16) and (17). These sums approximate the functions of Fig. 14 in the best possible manner. Clearly, "the best possible manner" is, in general, not a perfect fit. One criterion of "the best possible manner" is that the integral of the square of the deviation subsequent to the chosen shifts be a minimum. The method of choosing $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$ for such a criterion is well known. One forms a set of normal orthogonal functions from linear combinations of the derivative functions and evaluates the optimum $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$ by the evaluation of appropriate integrals. However, the formation of the normal orthogonal functions is so laborious because of the evaluation of the integrals that the method is entirely impractical.

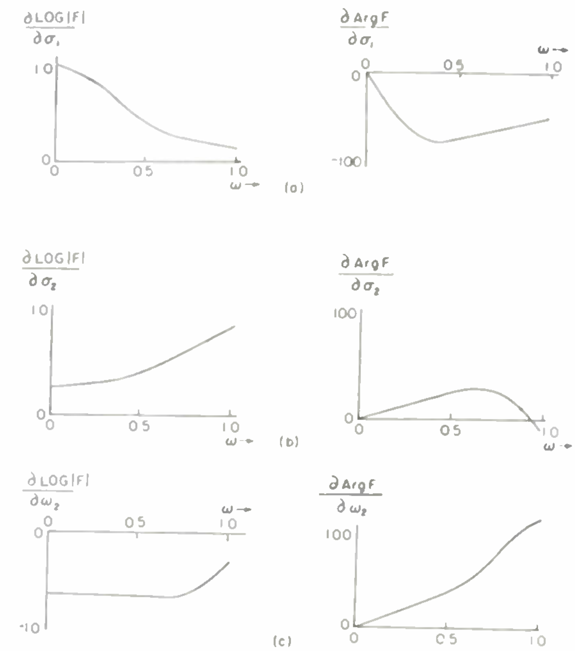


Fig. 13—Sketches of derivative functions for the pole positions of Fig. 12.

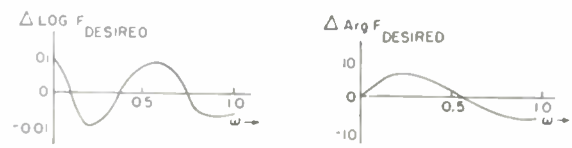


Fig. 14—Desired changes in magnitude and phase from Fig. 12.

An alternate method which is based on the same kind of approach but is far simpler is that of approximating at a set of points rather than at every point from $\omega=0$ to $\omega=1$. The labor involved in forming the normal orthogonal functions arises because of the fact that one uses an infinite set of samples to get an average. It is easy to see by consideration of Figs. 13 and 14 that, if one finds the sums of (16) and (17) matching well with Fig. 14 at five or six points on the range from $\omega=0$ to $\omega=1$, the match will be good elsewhere as the derivative functions and the desired deviations are smooth. The labor involved in matching at eleven points (six on the magnitude curve and five on the phase curve) is very much reduced. To formulate the procedure of solution, suppose a set of points has been chosen from the range $\omega=0$ to 1 and the values of the functions of Figs. 13 and 14 are indicated at those points as shown in Fig. 15. One is under no compulsion to choose the same matching frequencies for the magnitude and the argument curves though he is choosing $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$ which apply to both at the same time. Ordinarily, it is expedient to choose matching frequencies where the worst deviations occur or where the derivative curves are large. In any event one guides the choice of matching frequencies to fit the needs of the particular case.

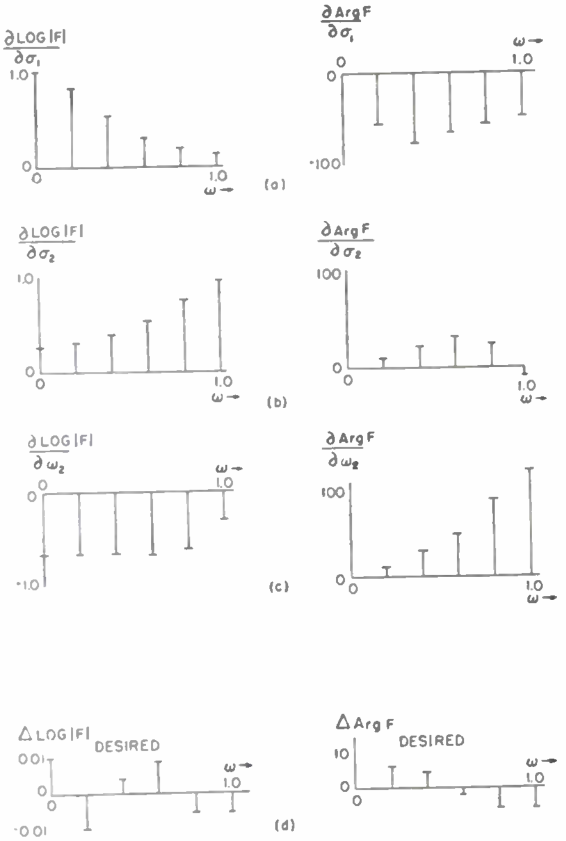


Fig. 15—Values of functions of Figs. 13 and 14 at selected matching points.

The ordinates of Fig. 15 are of two different kinds; some are in terms of logarithms to the base 10 and some are in terms of degrees. Before proceeding one must put all of the ordinates on a common basis. In the problem at hand a deviation of 0.05 in log $|F|$ will be said to be as serious as a deviation of 10 degrees in the argument. Accordingly, all quantities referring to arguments should be divided by 200 to have the quantities on a uniform basis. One appreciates at this point that if the magnitude or phase at a particular frequency is more important than others it may be given a heavier weight. If in Fig. 15 the phase shift at $\omega=1$ is to be more closely controlled than at other frequencies, the ordinates of argument curves there might be divided by 100 instead 200, for instance.

A compact notation for the functions of Fig. 15 (after weighting) is vector notation. The ordinates of the functions define vectors of eleven dimensions. Call F_D a vector in 11 space associated with the weighted ordinates of (d), $F_{\sigma 1}$ a vector in 11 space associated with the ordinates of (a), $F_{\sigma 2}$ a vector associated with (b), and $F_{\omega 2}$ a vector associated with (c). For the situation illustrated in Fig. 15, one has the following vectors:

The problem to be solved, expressed in vector terminology, is to select $\Delta\sigma_1$, $\Delta\sigma_2$ and $\Delta\omega_2$ so that

$$F_D \cong \Delta\sigma_1 F_{\sigma 1} + \Delta\sigma_2 F_{\sigma 2} + \Delta\omega_2 F_{\omega 2}. \tag{28}$$

The Δ's found to solve the problem stated in vector terms are appropriate to be used as shifts of pole positions. Clearly, the approximation can not be made exact in the general case when the number of dimensions is greater than the number of adjustable quantities as is the case here, eleven being larger than three. Accordingly, one must establish a measure of approximation and, on the basis of it, choose the Δ's. A mathematically convenient and physically practical approximation is that in which the sum of squares of deviations at the points considered is minimized. In vector terms, the Δ's are chosen so that

$$[F_D - \Delta\sigma_1 F_{\sigma 1} + \Delta\sigma_2 F_{\sigma 2} + \Delta\omega_2 F_{\omega 2}]\cdot[\text{Same}] \text{ is minimized.} \tag{29}$$

The solution for the Δ's of (28) is most conveniently done by choosing a set of normal orthogonal vectors which are linearly dependent with $F_{\sigma 1}$, $F_{\sigma 2}$, and $F_{\omega 2}$ and first approximately F_D with them. Finally, one is led to the Δ's of (28). A simple set of normal orthogonal vectors is

$$\left.\begin{aligned} F_{n1} &= a_{11}F_{\sigma 1} \\ F_{n2} &= a_{21}F_{\sigma 1} + a_{22}F_{\sigma 2} = b_{21}F_{n1} + a_{22}F_{\sigma 2} \\ F_{n3} &= a_{31}G_{\sigma 1} + a_{32}F_{\sigma 2} + a_{33}F_{\omega 2} \\ &= b_{31}F_{n1} + b_{32}F_{n2} + a_{33}F_{\omega 2} \end{aligned}\right\}. \tag{30}$$

The a's are chosen to fulfill the following relationships:

$$\left.\begin{array}{lll} F_{n1}\cdot F_{n1} = 1 & & \\ F_{n1}\cdot F_{n2} = 0 & F_{n2}\cdot F_{n2} = 1 & \\ F_{n1}\cdot F_{n3} = 0 & F_{n2}\cdot F_{n3} = 0 & F_{n3}\cdot F_{n3} = 1 \end{array}\right\}. \tag{31}$$

The solution of the equations above, or similar equations for a case of any number of poles and zeros, is fortunately simple. The first row of (31) involves only a_{11}, and one solves for a_{11} first. The second row (using F_{n1} as found from the first and the expression for F_{n2} in terms of F_{n1} and $F_{\sigma 2}$) involves only b_{21} and a_{22}. One can solve for b_{21} in terms of a_{22} and then, using the second equation in the second row (a quadratic in a_{22} only), one can evaluate a_{22} and a_{21}. The third row of (31) (using F_{n1} and F_{n2} and the expression for F_{n3} in terms of F_{n1}, F_{n2}, and $F_{\omega 2}$) involves b_{31}, b_{32}, and a_{33}. By solving the equations from left to right and expressing the b's in terms of a_{33}, one finds the equations are simple and can be solved one at a time. Only the final equation is a quadratic but involfes a_{33} only. Hence by considering the equations in the proper order, one completely avoids the solution of general simultaneous equations.

$$\left.\begin{array}{llllllllllll} F_D = & +0.018, & -0.010, & +0.004, & +0.008, & -0.006, & -0.006, & +0.030, & +0.024, & -0.004, & -0.027, & -0.027 \\ F_{\sigma 1} = & +1.08, & +0.87, & +0.54, & +0.33, & +0.22, & +0.15, & -0.29, & -0.37, & -0.33, & -0.29, & -0.25 \\ F_{\sigma 2} = & +0.29, & +0.33, & +0.40, & +0.56, & +0.76, & +0.98, & +0.059, & +0.11, & +0.17, & +0.13, & -0.055 \\ F_{\omega 2} = & -0.69, & -0.69, & -0.71, & -0.73, & -0.64, & -0.33, & +0.065, & +0.15, & +0.26, & +0.44, & +0.60 \end{array}\right\}. \tag{27}$$

The simplicity indicated above applies to the determination of orthogonal vectors regardless of the number. The degeneracy arises from the special way in which the orthogonal vectors are obtained.

For the illustrative example at hand, one obtains the following:

$$\left.\begin{array}{lll} a_{11}=0.5902 & & \\ a_{21}=-0.3608 & a_{22}=0.748 & \\ a_{31}=\ \ 1.063 & a_{32}=0.6356 & a_{33}=1.463 \end{array}\right\}. \tag{32}$$

$$\left.\begin{array}{l} F_{n1}=+0.6374, +0.5135, +0.3187, +0.1948, +0.1298, +0.0885, -0.1712, -0.2184, -0.1948, -0.1712, -0.1476 \\ F_{n2}=-0.1144, -0.0200, +0.1335, +0.3177, +0.5010, +0.6870, +0.1331, +0.1958, +0.2248, +0.1862, +0.1355 \\ F_{n3}=+0.3228, +0.1250, -0.2103, -0.3609, -0.2191, +0.2997, -0.1756, -0.1038, +0.1376, +0.4180, +0.5768 \end{array}\right\}. \tag{33}$$

Once the normal orthogonal vectors are selected, one approximates F_D in terms of them.

$$F_D = c_1F_{n1} + c_2F_{n2} + c_3F_{n3}. \tag{34}$$

This approximation involves choice of the c's so that

$$[F_D - (c_1F_{n1} + c_2F_{n2} + c_3F_{n3})]\cdot[\text{Same}] \tag{35}$$

is a minimum. This problem is particularly easy to solve since the vectors used are an orthogonal set. Once the c's of (34) are selected, the optimum values for $\Delta\sigma_1$, $\Delta\sigma_2$, and $\Delta\omega_2$ can be obtained directly. Considering (35) and observing the orthgonality of the vectors, one has

$$[\quad]\cdot[\quad] = F_D\cdot F_D - 2c_1F_{n1}\cdot F_D + c_1^2 - 2c_2F_{n2}\cdot F_D + c_2^2 - 2c_3F_{n3}\cdot F_D + c_3^2. \tag{36}$$

To minimize (35), one determines the c's by setting the partial derivatives with respect to σ_1, σ_2, and ω_2 of (36) equal to zero. This step gives

$$\left.\begin{array}{l} \dfrac{\partial}{\partial c_1}[\quad]\cdot[\quad] = 2c_1 - 2F_{n1}\cdot F_D = 0 \\ \dfrac{\partial}{\partial c_2}[\quad]\cdot[\quad] = 2c_2 - 2F_{n2}\cdot F_D = 0 \\ \dfrac{\partial}{\partial c_3}[\quad]\cdot[\quad] = 2c_3 - 2F_{n3}\cdot F_D = 0 \end{array}\right\}. \tag{37}$$

One observes that the second derivatives of (37) with respect to the c's give 2 and, accordingly, c's chosen to satisfy (37) give a minimum of (35). Consequently,

$$\left.\begin{array}{l} c_1 = F_{n1}\cdot F_D \\ c_2 = F_{n2}\cdot F_D \\ c_3 = F_{n3}\cdot F_D \end{array}\right\}. \tag{38}$$

Finally, (38) leads to the optimum pole shifts indicated in (39).

$$\left.\begin{array}{l} \Delta\sigma_1 = c_1a_{11} + c_2a_{21} + c_3a_{31} \\ \Delta\sigma_2 = \qquad\quad\ \ c_2a_{22} + c_3a_{32} \\ \Delta\omega_2 = \qquad\qquad\qquad\quad\ c_3a_{33} \end{array}\right\}. \tag{39}$$

In connection with the illustrative example being considered, one has

$$\left.\begin{array}{lll} c_1 = 0.00175, & \Delta\sigma_1 = -0.039, & \sigma_1 = -0.439 \\ c_2 = 0.00086, & \Delta\sigma_2 = -0.024, & \sigma_2 = -0.474 \\ c_3 = 0.0374, & \Delta\omega_2 = -0.055, & \omega_2 = \ \ \ 1.011 \end{array}\right\}. \tag{40}$$

A plot of the characteristics of the approximating function given by (40) is a part of Fig. 16. One can see by comparison of the o's and the x's how the final adjustment has changed the characteristics. Quantity 35 changed from 0.0033 to 0.0021. In this case the improvement has been quite small; this was to be expected since the example is simple and the first stage of

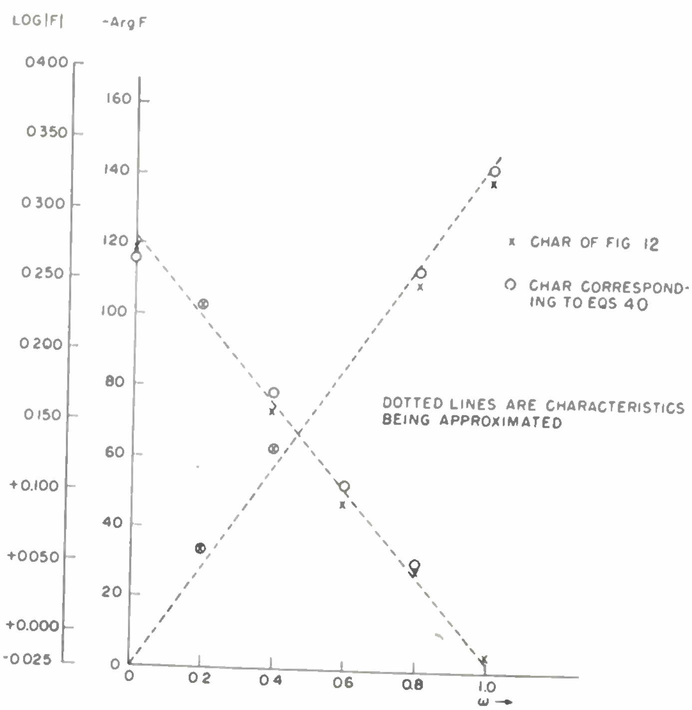


Fig. 16—Characteristics of approximating functions.

adjustment is very effective in simple cases. However, one can now state definitely that it is impossible to obtain a better approximation than the "o" characteristics of Fig. 16 by further shifts of the poles. It is necessary to recall that the measure of quality of the approximations has been chosen, and the approximation arrived at is best in the sense of the chosen measure. Different weightings of magnitude and phase characteristics would have results in somewhat different approximations.

In cases where the first stage of approximation gives a far less satisfactory result than that of the accompanying illustrative example, it may be necessary to use more than one adjustment in the final state. This will be necessary only if the shifts required are so large that the derivative functions change a large amount. A simple method is available (6) to compensate for changes in the derivative functions over the range of large shifts. This technique makes it possible to employ any family of normal orthogonal vectors over a range of critical-frequency changes which is much larger than would otherwise be possible. Naturally, the result is that the optimum adjustment can be obtained in a fewer number of steps, and one avoids the labor of obtaining the normal orthogonal vectors a corresponding number of times.

An Annular Corrugated-Surface Antenna*

E. M. T. JONES†, ASSOCIATE, IRE

Summary—The far-zone pattern of the symmetrically excited annular corrugated-surface antenna is shown to be uniform in the azimuthal direction and polarized in a direction perpendicular to the antenna surface, and to have the major lobe directed slightly above the plane of the antenna. An analysis of the antenna immersed in an infinite ground plane is presented and compared with experimental results on the antenna in a finite ground plane.

Introduction

THE ANNULAR corrugated-surface antenna pictured in Fig. 1 is a cylindrical geometry counterpart of the rectangular corrugated-surface antenna discussed by Reynolds and Lucke.[1]

The annular corrugated-surface antenna supports a surface-guided wave that travels in an outward radial direction at a velocity less than that of light, and has a component of electric field tangential to the surface. In practice it is found that the surface wave is easily excited from the end of a coaxial line, with the center conductor extended a quarter wavelength above the surface.

The method of analysis consists of determining the properties of the surface wave by an approximate matching of the tangential fields above and below the surface of the corrugations. The radiation pattern of the antenna is determined by the Green's function method.

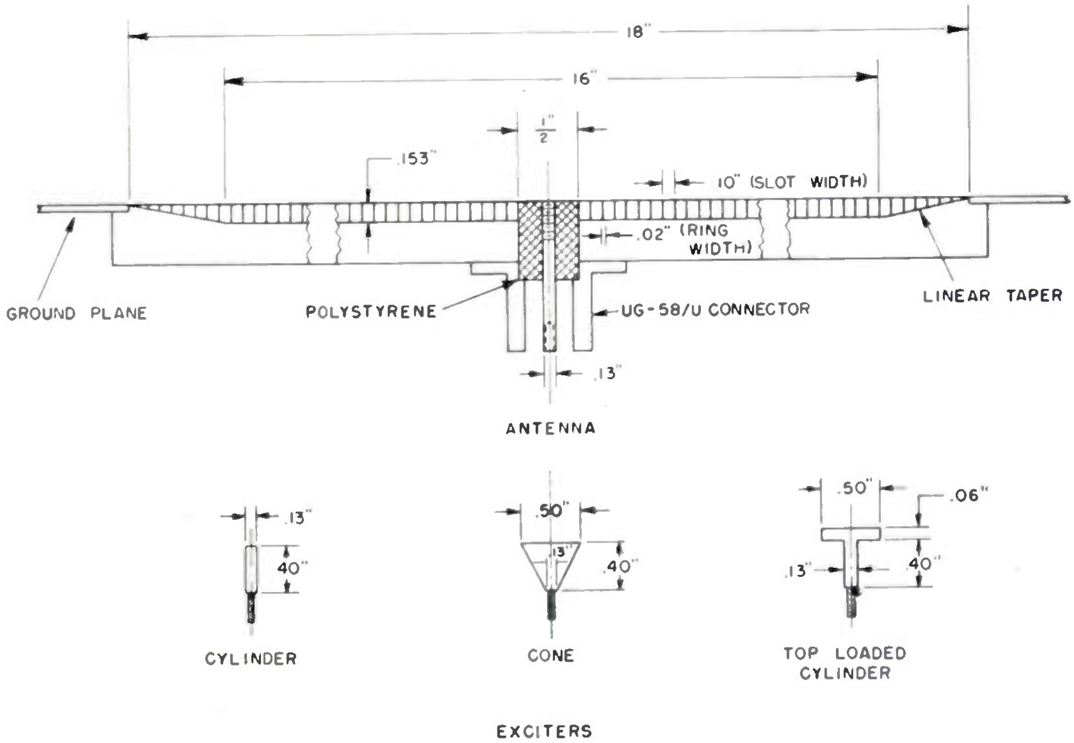


Fig. 1—A principal diameter cross section of the experimental annular corrugated-surface antenna showing dimensions.

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† Stanford Research Institute, Stanford, Calif.

[1] D. K. Reynolds and W. S. Lucke, "Corrugated end-fire antennas," *Proc. NEC* (Chicago), vol. 6, pp. 16–28; 1950.

Analysis of the Antenna

The properties of the annular corrugated-surface antenna can be determined as follows:

Let us choose as cylindrical co-ordinates on the surface ρ', ϕ', and z', and as co-ordinates above the surface

ρ, ϕ, and z, as illustrated in Fig. 2. We will confine the present analysis to fields that have no variation in the ϕ' direction. This restriction allows us to use a corrugated surface with uniform slot depth to obtain a surface-guided wave whose phase velocity in the radial direction is nearly constant for radii greater than a quarter wavelength.

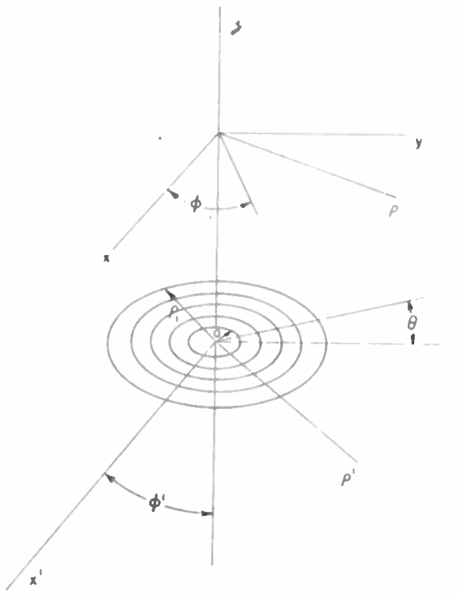

Fig. 2—The co-ordinate system used to describe the annular corrugated-surface antenna.

Borrowing a result from Stratton,[2] we can write down immediately the components of the field existing above the slots as

$$\begin{aligned} E_\rho &= -jh\beta H_1^{(2)}[\beta\rho]e^{j(\omega t-hz)} \\ E_z &= \beta^2 H_0^{(2)}[\beta\rho]e^{j(\omega t-hz)} \\ H_\phi &= j\frac{k^2}{\mu_0\omega}\beta H_1^{(2)}[\beta\rho]e^{j(\omega t-hz)}, \end{aligned} \tag{1}$$

where

$$k^2 - \beta^2 = h^2.$$

The mode existing in each annular slot, in the limit of infinitesimally narrow slots, is just the TEM coaxial transmission-line mode. Even for finite-sized slots the TEM mode is the principal mode if the slot width is much less than a free-space wavelength. The ratio of the transverse electric to magnetic field of the TEM mode in each slot is just

$$\frac{E_\rho}{H_\phi} = j\sqrt{\frac{\mu_0}{\epsilon_0}}\tan kd = j\eta\tan kd, \tag{2}$$

where d is the slot depth.

[2] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 360; 1941.

At the plane $z=z'$, it is necessary that the ratio of the tangential electric to tangential magnetic field be continuous. The ratio of the tangential fields seen looking in the positive z direction is

$$\left.\frac{E_\rho}{H_\phi}\right|_{z=z'} = -\frac{\omega\mu_0 h}{k^2}. \tag{3}$$

The average value of the ratio of the tangential fields at $z=z'$ seen looking in the negative z direction is different from the value given by (2) because $E_\rho=0$ along the barrier rings. The expression is

$$\left.\frac{E_\rho}{H}\right|_{z=z} = j\left(\frac{t}{t+w}\right)\eta\tan kd, \tag{4}$$

where t is the slot width and w the ring width.

Combining (3) and (4) we find

$$h = -jk\left(\frac{t}{t+w}\right)\tan kd. \tag{5}$$

Therefore, the field of the annular corrugated-surface antenna decays in an exponential fashion in the positive z direction for $m\pi \leq kd \leq m\pi+(\pi/2)$ where $m=0, 1, 2 \cdots$. These are the only allowable ranges of slot depths that will permit a surface-guided wave to propagate.

The propagation constant, β, of the surface wave in the ρ' direction is

$$\beta = k\sqrt{1+\left(\frac{t}{t+w}\right)^2\tan^2 kd}. \tag{6}$$

Equation (6) is exact for the case of infinitely many corrugations per wavelength along the surface, but is in error by a small amount for a finite number of corrugations per wavelength.

The far-zone field of the annular corrugated-surface antenna immersed in an infinite ground plane can be determined by the Green's function method. The magnetic vector at the point $r(\rho, \phi, z)$ within a volume V, surrounded by a surface S (with outward directed normal $\bar{n}$), and containing no charges or currents is given by Stratton[3] as

$$\overline{H}(r) = \frac{1}{4\pi}\int_S \{-j\omega_0[\bar{n}\times\overline{E}(r')]G(r,r') - [\bar{n}\times\overline{H}(r')] \times\nabla G(r,r') - [n\cdot\overline{H}(r')]\nabla G(r,r')\}da, \tag{7}$$

where r' is a point on the surface S and $G(r, r')$ is a Green's function satisfying the wave equation

$$(\nabla^2+k^2)G(r,r') = -\delta(r-r'), \tag{8}$$

in which $\delta(r-r')$ is the three-dimensional Dirac delta function. The surface S is chosen as the plane $z'=0$ and the hemisphere at infinity $(r\rightarrow\infty)$. It is also assumed that the fields existing at $z'=0$ are excited by some

[3] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 466; 1941.

ource at $z<0$. It is possible to choose the free-space Green's function so that $\partial G(r, r')/\partial z' = 0$, at $z' = 0$. Equation (7) then simplifies to

$$\overline{\Pi}(r) = \frac{1}{4\pi}\int_S \{-j\omega\epsilon_0[n \times \overline{E}(r')]G(r, r')\}da. \qquad (9)$$

The appropriate Green's function is

$$G(r, r') = \frac{1}{4\pi}\left[\frac{e^{-jkR_1}}{R_1} + \frac{e^{-jkR_2}}{R_2}\right]. \qquad (10)$$

where

$$R_1 = \sqrt{\rho^2 + \rho'^2 - 2\rho\rho' \cos(\phi - \phi') + (z - z')^2}$$

$$R_2 = \sqrt{\rho^2 + \rho'^2 - 2\rho\rho' \cos(\phi - \phi') + (z + z')^2}$$

At $z' = 0$, $R_1 = R_2 = R'$. Also for $\rho \gg \rho_1$, as is true in the far-zone field,

$$R' \cong \sqrt{\rho^2 + z^2}\left[1 - \frac{\rho\rho' \cos(\phi - \phi')}{\rho^2 + z^2}\right]. \qquad (11)$$

Transforming to spherical co-ordinates by the relations $\rho^2 + z^2 = R^2$ and $R \cos \phi = \rho$, we find

$$G[R, (\phi - \phi'), \theta] = \frac{1}{2\pi}\frac{e^{-jkR}}{R} e^{jk\rho'} \cos(\phi - \phi') \cos\theta. \qquad (12)$$

Substituting (12) into (9) we find

$$H_\phi = \frac{+j\omega\epsilon_0 e^{-jkR}}{8\pi^2 R}\int_0^{2\pi}\int_a^{\rho_1} E_\rho(\rho')e^{jk\rho\rho} \cos(\phi - \phi') \cos\theta \cdot \cos(\phi - \phi')\rho' d\rho' d\phi', \qquad (13)$$

where a and ρ^1 are the inner and outer radii of the corrugated surface, respectively.

Integration with respect to ϕ' in (13) and substitution of the explicit expression for $E_\rho(\rho')$ from (1) gives

$$H_\phi = \frac{-j\omega\epsilon_0\beta h e^{j(\omega t - kR)}}{4\pi R} \cdot \int_a^{\rho_1} H_1^{(2)}[\beta\rho']J_1(k\rho' \cos\theta)\rho' d\rho'. \qquad (14)$$

Integration of (14) gives the magnetic component of the far-zone field as

$$\begin{aligned} H_\phi = {} & \frac{-j\omega\epsilon_0\beta h e^{j(\omega t - kR)}}{[\beta^2 - k^2 \cos^2\theta]4\pi R} \\ & \cdot \{k\rho_1 \cos\theta H_1^{(2)}[\beta\rho_1]J_0[k\rho_1 \cos\theta] \\ & - \beta\rho_1 H_0^{(2)}[\beta\rho_1]J_1[k\rho_1 \cos\theta] \\ & - ka \cos\theta H_1^{(2)}[\beta a]J_0[ka \cos\theta] \\ & + \beta a H_0^{(2)}[\beta a]J_1[ka \cos\theta]\}. \end{aligned} \qquad (15)$$

Experimental Results

An annular corrugated-surface antenna has been constructed to operate in the 4-cm wavelength region, and has the dimensions shown in Fig. 1. The various types of exciters shown can easily be inserted in the end of the coaxial feed line. The depth of the slots is tapered at the outer periphery of the antenna to make a gradual transition from the corrugated-surface mode to that existing over the ground plane.

Theoretical and experimental values for the normalized velocity of propagation of the surface wave along the antenna are shown in Fig. 3. It is seen that the approximate theory agrees only moderately well with the experimental results; but agreement is best when the number of corrugations per surface wavelength is largest (small kd), in agreement with the previous explanation.

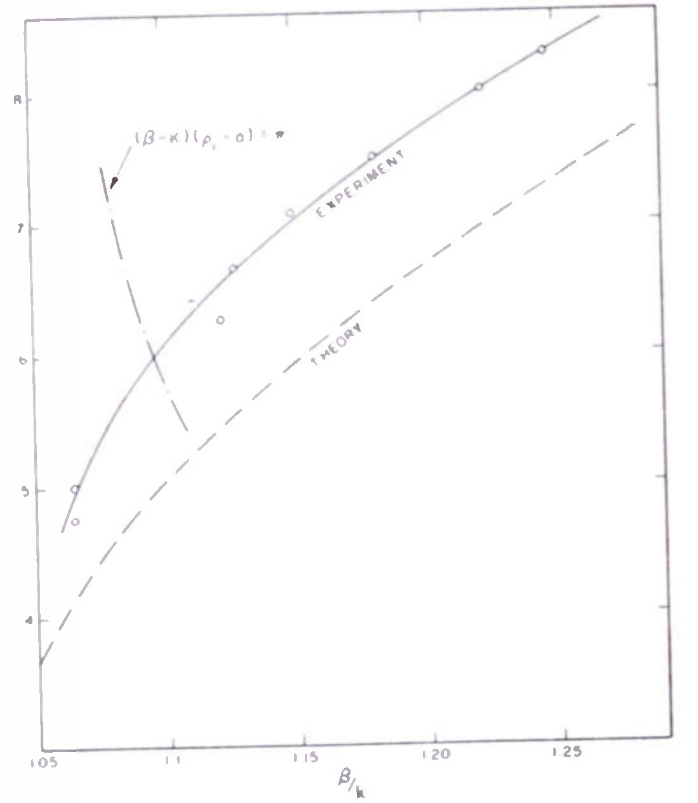

Fig. 3—A comparison of theoretical and experimental surface-wave radial propagation velocity (which is proportional to β/k), versus frequency (which is proportional to kd).

The curve labeled $(\beta - k)(\rho_1 - a) = \pi$ represents the optimum gain condition for a uniformly illuminated end-fire antenna. This condition is also very close to the optimum gain condition for exponentially illuminated surfaces, and, hence, has been used as a design criterion for the annular-corrugated antenna. This curve intersects the experimental phase-velocity curve at $\beta = 1.595$.

The measured amplitude of the tangential electric field of the surface wave is shown as a function of the radial distance in Fig. 4. The theoretical variation of tangential electric field, labeled $|H_1^{(2)}(1.595)|$, is shown for comparison. There is a marked discontinuity in the measured amplitude of the surface wave over the region of tapered slot depth, and a corresponding small standing wave is superimposed on the outward traveling

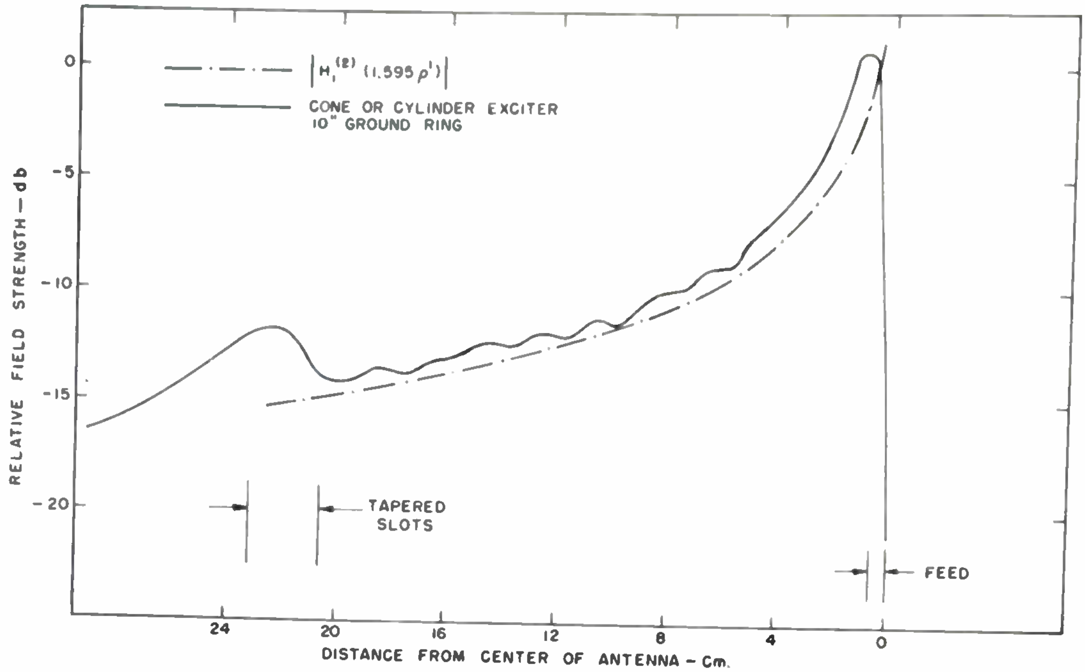

Fig. 4—A comparison of the theoretical and experimental radial tangential field along the annular corrugated surface.

surface wave. The cone and cylinder exciters were found to start the surface wave with equal efficiency, while the top-loaded cylinder excited the surface wave only weakly.

The measured amplitude of tangential electric field over the exciter is very low, so that the tacit assumption of zero field, made in the analysis of the far-zone pattern, is well approximated.

The theoretical far-zone pattern of the antenna immersed in an infinite ground plane for 4.24-cm wavelength operation is shown in Fig. 5. Fig. 6 shows the measured radiation pattern of the antenna, immersed in a 10-inch ground plane and driven by a cone exciter at 4.24-cm wavelength. The agreement between the calculated and measured patterns is quite good, except that the main lobe of the experimental pattern appears about 5 degrees farther above the surface than that of the theoretical pattern. This pattern tilt is characteristic of antennas operating in finite ground planes.

The upper operating frequency limit of this type of antenna is determined by the condition that $(\beta-k)(\rho_1-a)\simeq 2\pi$, at which point the gain is very low. The lower frequency limit is set by the condition that the radius of the antenna is on the order of two free-space wavelengths, when the gain is again low.

Conclusion

The theoretical analysis of the annular corrugated-surface antenna has been shown to predict the behavior of the antenna quite faithfully. This type of antenna should find application in the microwave frequency spectrum as a beacon antenna where an antenna having a low silhouette is important.

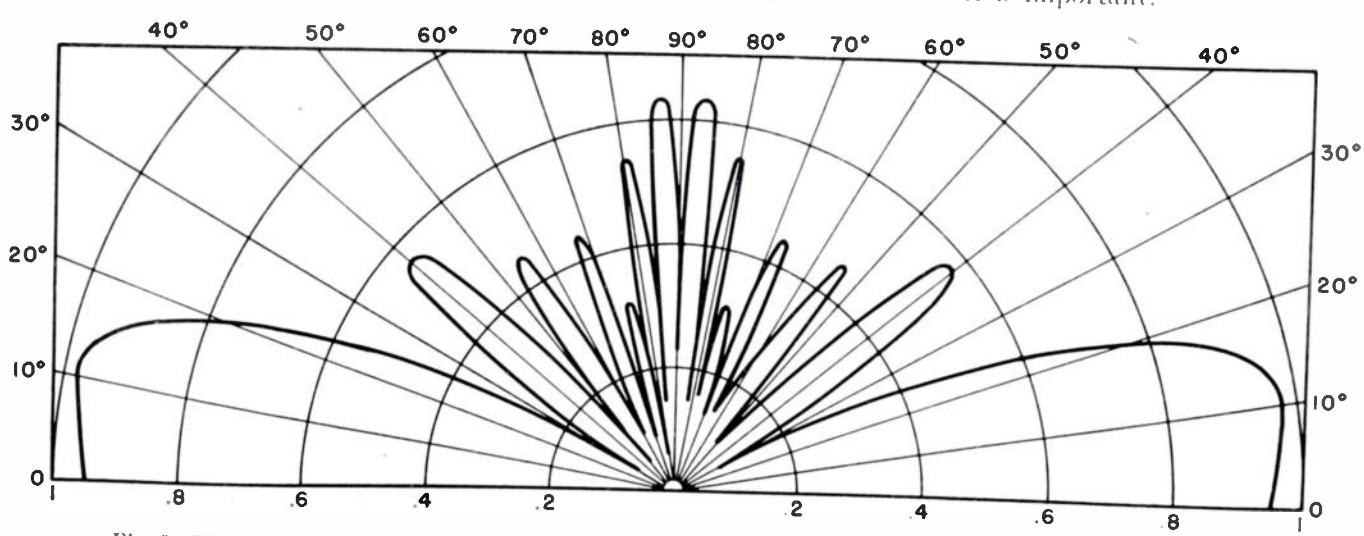

Fig. 5—Theoretical field pattern of the annular corrugated-surface antenna immersed in an infinite ground plane, operating at a wavelength of 4.24 cm.

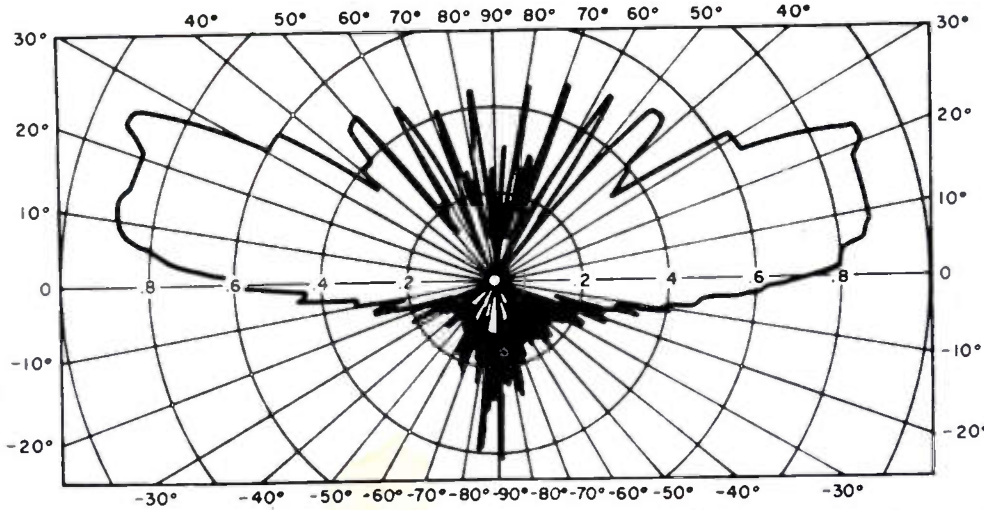

Fig. 6—Experimental field pattern of the annular corrugated-surface antenna immersed in a 10-inch ground ring, operating at a wavelength of 4.24 cm.

Acknowledgment

The research reported in this paper was done under the auspices of the Evans Signal Laboratory, Signal Corps Contract W36-039 sc-44524. The writer is indebted to J. V. N. Granger, J. T. Bolljahn, W. S. Lucke, A. S. Dunbar, and D. K. Reyonolds for many valuable discussions, and to Sarah Hornig, who computed the antenna pattern.

Discussion on

The Synthesis of *RC* Networks to Have Prescribed Transfer Functions*

H. J. ORCHARD

Ernst A. Guillemin:[1] The procedure given by Mr. Orchard is precisely that contained in my M.I.T. Radiation Laboratory Report No. 43 of 10/11/44 to which Messrs. J. L. Bower and P. F. Ordung[2] refer in their paper. This report contains the method given in Mr. Orchard's paper, discussed for both an impedance and an admittance basis, together with methods for developing the resultant lattice into an unbalanced form (similar to those in the Bower-Ordung paper), and a discussion of factors upon which the resultant gain depends.

Since vacuum tubes need to be associated with *RC* networks in order to overcome the effect of loss, it is very desirable that the design be convertible to an unbalanced form such as a ladder, or parallel ladders, or ladders with bridged elements, etc. Such a development of the lattice is further desirable since experience indicates that the element tolerances in many lattice designs are extremely small and that this situation is less severe in the ladder forms. Since the lattice design cannot in all cases be developed in this manner, the procedure given in the R. L. Report 43 was regarded as being of limited interest and its wider publication did not seem advisable, especially in view of the method developed subsequently, which always leads to an unbalanced structure.[3] Its restriction (in general) to minimum-phase networks appears to be relatively minor since one would scarcely use *RC*-networks for this purpose (because of their inherent loss) except in rather unusual circumstances.

H. J. Orchard:[4] I am indebted to Professor Guillemin for drawing my attention to the Radiation Laboratory Report on *RC*-coupling networks. Undoubtedly our methods of deriving basic lattice are identical, but, nevertheless, I feel there remains a certain difference in point of view. Realizing that the *RC*-network will normally require some associated amplification Professor Guillemin is primarily concerned with providing a network to operate between vacuum tubes and so can employ, if necessary, sources and loads with zero or

* Proc. I.R.E., vol. 39, pp. 428–432; April, 1951.

[1] Massachusetts Institute of Technology, Cambridge, Mass.

[2] J. L. Bower and P. F. Ordung, "The synthesis of resistor-capacitor networks," Proc. I.R.E., pp. 263–269; March, 1950. (See footnote[2] on p. 264).

[3] Synthesis of RC-networks, *Jour. of Math. and Physics*, vol. 28, no. 1, p. 22; April 1949.

[4] British Post Office Research Branch, London, England.

infinite impedance. On the other hand, I find it frequently specified that the amplification must be placed at some point in the circuit remote from the network, which is then required to operate between resistive terminations. In these circumstances it is important to know that such terminations can always be "found" in the network, and this was really the main point of my paper. An insertion loss theory for passive transmission networks, in its general form, should surely demand that the prescribed insertion loss function be provided by the network between finite non-zero terminations. Any procedure which relaxes these termination requirements may be an excellent practical expedient but it nevertheless evades the theoretical question. An analogous situation exists in the LC case where the design of a network to work from a resistive generator into an open-circuit or short-circuit load is considerably simpler than a design involving resistive terminations at both ends. No practicing engineer, I think, would deny that though the open-circuit filter may save him design time it is a much more awkward proposition to use.

I must admit that the lattice structure I proposed has many practical disadvantages. However, as an RC ladder network is only possible when the poles of loss occur at negative real values of $p = iw$ (a rather restricted case) one must compare the lattice with some network such as that proposed by Professor Guillemin, i.e. a parallel connection of ladders. The only advantage of the latter is that it is unbalanced. I contend that it is just as difficult to adjust because the poles of loss are, in both cases, produced by a complicated compensating process; either the signals through the *two* lattice arms, or else the sum of *all* the signals through the individual ladders must cancel in the output at the pole frequencies. In the one case we ask stability of impedance for each arm, in the other, stability of transfer function for each ladder. The ideal would appear to be a tandem connection of, perhaps, simple parallel T networks, each producing one conjugate pair of poles of loss, analogous to the canonical tandem sections given by Darlington and others for LC networks. Unfortunately no design method yet exists but it might be a good research field.

Finally, I would like to mention that the extension to non-minimum phase circuits permits the construction of RC all-pass networks with real poles and zeros. Such networks are almost ideal for the pairs of very wide-band phase-splitting networks required for single sideband modulators etc. Examples which have been constructed have been entirely satisfactory and the flat loss was no more than 15 or 20 db.

Ernst A. Guillemin: There is, I believe, even less difference between our viewpoints than Mr. Orchard's reply indicates, inasmuch as my Radiation Laboratory report (article 6) referred to goes into considerable detail regarding possible ladder developments of the lattice and, in connection with the discussion of the zero-frequency response, shows that a shunt resistance is always removable at the terminations and that it is the value of this shunt resistance which significantly controls the zero-frequency loss of the resulting structure. This argument is carried through on both an impedance and an admittance basis with coinciding results.

With regard to the production of zeros of transmission (or poles of loss) our experience (which admittedly is somewhat limited) has shown that it is easier to obtain stability of transfer impedances than that of driving-point impedances. Stated in another way, we have had no difficulty in obtaining zeros of transmission with parallel ladder configurations, but have had great difficulty in realizing the same designs in the lattice form. It is quite possible that the unbalanced character of the parallel ladders was a significant factor in these comparisons. I am quite willing to admit that our experience in this regard is far from sufficient to permit general conclusions to be drawn.

Regarding the complexity of the paralleled ladder configuration, some recent studies indicate that one may in the most unfavorable situation be able to realize a given transfer function with not more than three parallel ladders. I admit that the chief motivation leading to this parallel ladder synthesis scheme lay in the desire to obtain an unbalanced structure. But this specification is so prevalent in practical situations (and was at the time this investigation was done) that little practical value was attached to the lattice procedure, as stated in my original letter.

Mr. Orchard is quite right in pointing out that a method of RC synthesis is needed that yields a cascade connection of component two terminal-pairs which individually place the zeros of transmission in evidence. An investigation of the possibilities of such a procedure is in progress in our network group.

H. J. Orchard: My previous letter, I trust, did not give the impression that, as far as a practical structure is concerned, I would prefer the lattice to a parallel connection of ladders. Assuming one is prepared to accept the limitations on transfer function (minimum-phase is sufficient although not necessary) and one has a design method which permits resistance terminations at both ends then the paralleled ladder is undoubtedly much superior. The suggestion that, in general, no more than three ladders are necessary is interesting; does this imply that an arbitrary Hurwitz polynomial can be expressed as the sum of at most, three polynomials with real non-positive roots?

One method of design which has done Trojan service in connection with LC filters, and which now seems unfashionable, is the image-parameter theory; it is possible that this might be adapted to permit two-terminal pairs comprising, for example, the familiar twin-T null circuit, to be cascaded under matched conditions to give the arrangement we both consider desirable.

Contributors to Proceedings of the I.R.E.

For a photograph and biography of Seymour B. Cohn, see page 1461 of the November, 1951, issue of the Proceedings of the I.R.E.

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E. E. David, Jr. (A'48) was born on January 25, 1925, in Wilmington, N. C. He attended the Georgia Institute of Technology and received the B.A. degree in electrical engineering in 1945.

E. E. David, Jr.

After serving briefly in the U. S. Navy as a fire-control officer, Dr. David joined the staff of the Massachusetts Institute of Technology as a research assistant in June, 1946. There he was concerned with work involving microwave techniques, communication theory, and noise problems. He received the M.S. degree in electrical engineering in 1947 and the Ph.D. degree in 1950, from M.I.T.

In 1950 he joined the Bell Telephone Laboratories, and has been concerned with acoustics research. Dr. David is an associate member of the Acoustical Society of America and a member of Tau Beta Pi, Sigma Xi, Phi Kappa Phi, Eta Kappa Nu, Omicron Delta Kappa, and Phi Eta Sigma.

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Lester M. Field (S'39–M'48–F'52) was born on February 9, 1918, in Chicago, Ill. He received the B.S. degree from Purdue University in 1939 and the Ph.D. degree from Stanford University in 1944. He was acting instructor in 1941 and acting assistant professor from 1942 to 1944 in electrical engineering at Stanford University.

Lester M. Field

In 1944 Dr. Field joined Bell Telephone Laboratories as a member of the magnetron development group and later the electron dynamics group of the physical research department. In 1946, Dr. Field returned to Stanford University as acting associate professor of electrical engineering, advanced to associate professor in 1948, and to professor in 1951.

He is a member of the American Physical Society, the American Society for Engineering Education, Tau Beta Pi, and Sigma Xi.

❖

Lawrence A. Harris (S'45–A'48) was born in Toronto, Canada, on January 25, 1923. He received the B.A.Sc. degree in electrical engineering from the University of Toronto in 1946; the S.M. degree from the Massachusetts Institute of Technology in 1948; and the Sc.D., also from M.I.T., in 1950.

L. A. Harris

From 1946 to 1950 he was research assistant at the Research Laboratory of Electronics at M.I.T., where his work was connected with traveling-wave tube research and electron-gun and -beam design. He was with the electrical engineering faculty at the University of Florida for one year, and is now assistant professor of electrical engineering at the University of Minnesota.

Dr. Harris is a member of the American Physical Society, and member of Sigma Xi.

❖

Edward O. Johnson (S'46–A'49) was born in Hartford, Conn., in 1919. He received the B.S. degree in electrical engineering in 1948 from Pratt Institute of Brooklyn, where he held a Westinghouse Achievement Scholarship. From 1941 to 1945 he served as chief aircraft electronic technician in the U. S. Navy. In 1948 he joined the RCA Laboratories Division. Princeton, New Jersey, where he is now engaged in work on gaseous electronics.

Ed. O. Johnson

Mr. Johnson is an associate member of the I.R.E.

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F. J. Kerr (A'43–SM'49) was born in St. Albans, England, on January 8, 1918. He received the B.Sc. degree from the University of Melbourne in 1937, and the M.Sc. degree in 1940. Since 1940 he has been with the Division of Radiophysics of the Commonwealth Scientific and Industrial Research Organization in Sydney, Australia. During the war he was engaged in the development of radar antennas and other equipment, and in studies of superrefraction. Recently he has been working in radio astronomy. He is at present spending a period at Harvard College Observatory, Cambridge, Mass., studying solar problems.

F. J. Kerr

For a photograph and biography of E. M. T. Jones, see page 490 of the April, 1952, issue of the Proceedings of the I.R.E.

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J. G. Linvill (A'49) was born in Kansas City, Mo., in August, 1919. He was graduated from William Jewell College with an A.B. degree in mathematics and physics in 1941. He entered M.I.T. as a co-operative student in electrical engineering and received an S.B. degree in 1943.

J. G. Linvill

Mr. Linvill joined the electrical engineering staff of M.I.T. in 1943, continuing as a graduate student. In 1949 he received the Sc.D. degree, his field of specialization being network theory. He was successively assistant, instructor, and since 1949 has been assistant professor of electrical engineering. He is currently on a year's leave of absence from M.I.T. as a member of the technical staff of the Bell Telephone Laboratories.

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J. B. McCandless (M'46) was born in Philadelphia, Pa., on June 22, 1917. He attended evening sessions at the Drexel Institute of Technology, and graduated with the degree of M.M. in electrical engineering in 1943.

J. B. McCandless

From 1939 to 1947 Mr. McCandless was employed as a draftsman and designer on electromechanical instruments by the Electro Tachometer Corp. of Philadelphia. He joined the Brown Instrument Division of the Minneapolis-Honeywell Regulator Co. in 1942 as test engineer, and was later transferred to the research and development division of Brown Instrument as project engineer on servomechanisms.

Mr. McCandless worked for the Philco Corp. of Philadelphia as a senior research engineer on radar receivers from 1945 to 1946. During the next two years he was associated with Lavoie Laboratories, Morganville, N. J. and with Electronic Associates, Inc., Long Branch, N. J., where he was project engineer and chief of quality control, respectively, on various research and development items for USN, USA, and USAAF.

In 1948 Mr. McCandless entered government service with the Signal Corps Engineering Laboratories, Fort Monmouth, N. J. He is currently on the staff of the Director of Research.

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M. C. Pease (M'47–SM'51) was born in New York, N. Y., in 1920. He received the B.S. degree in chemistry from Yale University in 1940 and the M.A. degree in physical chemistry from Princeton University in 1943. He worked at the Radio Research Laboratory at Harvard University (a war laboratory for the development of countermeasures for radar), where he was principally concerned with the 1-kw series of cw magnetrons. In 1945 he left this position to accept a commission in the U. S. Navy as a radar officer.

M. C. Pease

After his discharge from the Navy, Mr. Pease joined the Electronics Division of Sylvania Electric Products, Inc. Initially he was concerned with various theoretical problems, largely in the field of microwave networks. He is now manager of Tube-Development Engineering, and has charge of the development of magnetrons, gas switch tubes, thyratrons, strobotrons, and the like.

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R. F. Rinehart was born in Springfield, Ohio, on May 31, 1907. He received the A.B. degree from Wittenberg College in 1930, and the M.A. and Ph.D. degrees in mathematics from Ohio State University in 1932 and 1934. He taught mathematics at Ashland College, Ashland, Ohio, from 1934 to 1937. Since 1937 he has been associated with the mathematics department of Case Institute of Technology where he is now professor of mathematics.

From 1942 to 1945 Dr. Rinehart was a member of the Operations Research Group of the U. S. Navy, engaged in analyses of naval tactics and strategy in subsurface warfare. He was awarded the Medal of Freedom for his work in antisubmarine warfare in the Caribbean Sea Frontier, and the Medal for Merit for achievements in submarine warfare in the Pacific.

R. F. Rinehart

From 1948 to 1950 Dr. Rinehart served with the Research and Development Board of the Department of Defense, first as director of the Planning Division, and subsequently as executive secretary. He was also the acting chairman of that Board for five months following Dr. K. T. Compton's resignation.

Dr. Rinehart is a member of Sigma Xi, the American Mathematical Society, and the Mathematical Association of America.

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For a photograph and biography of S. T. Smith, see page 359 of the March, 1952, issue of the Proceedings of the I.R.E.

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Dr. Tien was an engineer at the Tien-Sun Industrial Company, Shanghai, China, from 1942 to 1947, and was a research assistant at Stanford University from 1949 to 1951. At present he is a research associate in the Electronics Research Laboratory of Stanford University, where he is engaged in microwave-tube research.

Dr. Tien is a member of Sigma Xi.

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W. M. Webster

Mr. Webster is an associate member of the I.R.E.

Correspondence

A Link Between Information and Energy*

With the advent of the transistor and its striking ability to operate on low-power, it is pertinent to inquire what are the ultimate limits upon low power operation of switching systems, computers, and other communication machinery.

Mr. C. E. Shannon has derived the relationship for the capacity of an information channel

$$C = W \log_2 \left(1 + \frac{P}{N}\right) \tag{1}$$

where

C = maximum number of binary bits of information that can be transmitted per second

W = bandwidth of channel in cps

P = average signal power

N = average noise power assumed to be thermal or "white."

* Received by the Institute, February 15, 1952.

Regarding (1), we see that the higher the signal power P the greater the information rate C. One may ask, "At what signal level should we transmit in order to maximize C/P (the information rate per unit of signal power)?"

From (1)

$$\frac{C}{P} = \frac{W}{P} \log_2 (1 + P/N). \tag{2}$$

Taking the derivative of (2) with respect to P and setting it to zero, it is found that an upper bound to C/P is approached as

$$\frac{P}{N} \to 0,$$

and therefore the upper bound is given by

$$\left(\frac{C}{P}\right)_{\text{ub}} = \frac{W}{N} \log_2 e. \tag{3}$$

One may inquire as to the upper bound to W/N. In a circuit of bandwidth W, the minimum available noise power will be the thermal noise power

$$N = \frac{E^2}{4R} = kTW \tag{4}$$

where

k = Boltzmann's constant $= 1.374 \times 10^{-23}$ joule per °K

T = absolute temperature.

From this we have the relationship below as the upper bound to W/N

$$\left(\frac{W}{N}\right)_{\text{ub}} = \frac{1}{kT} \tag{5}$$

and (5) in (3) gives

$$\left(\frac{C}{P}\right)_{\text{ub}} = \frac{1}{kT} \log_2 e. \tag{6}$$

from which we can conclude that, with an available signal power P, the upper bound to the number of binary bits per second that can be transmitted is given by

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Correspondence

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$$P_{lb} = CkT \log_e 2. \tag{8}$$

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Equation (8) says that the signal power an never be less 2.85×10^{-15} watts. Simiarly, a repeater that is to pass information t a rate of 10^6 bits per second must certainly have a battery drain of more than $.85 \times 10^{-15}$ watts.

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C times time gives bits and *P* times time gives energy. Multiplying (8) by one second, we find that at 300°K, one binary bit of nformation will always cost more than 2.85×10^{-21} joules.

Mr. J. R. Pierce has informed the writer that it is possible to derive the same relationship between energy and information by arguments based on the energy available from the exploitation of information as to which half of an otherwise empty box contains a single gas molecule.

J. H. FELKER
Bell Telephone Laboratories, Inc.
Whippany, N. J.

A Note on Booker's Extension of Babinet's Principle*

Some differences exist in the literature today as to the proper interpretation of Babinet's Principle extended to apply to coherent waves. A few words of clarification would seem in order.

Consider an antenna source P to the left of a perfectly conducting, indefinitely thin, infinite plane screen S (Fig. 1(a)). S contains an aggregate of holes of arbitrary shape and position. The total field to the right of S is $(\vec{E_s} = \vec{E_0} + \vec{E_{1s}}, \vec{H_s} = \vec{H_0} + \vec{H_{1s}})$, where $(\vec{E_0}, \vec{H_0})$ is the field due to the source P, and $(\vec{E_{1s}}, \vec{H_{1s}})$ is the field due to the sources induced in the conducting portions of S.

Next, replace P by another source P^* and replace S by its complementary screen C (Fig. 1(b)). On C the conducting and open regions of S have been interchanged. The total field to the right of C is $(\vec{E_c} = \vec{E_0}^* + \vec{E_{1c}}, \vec{H_c} = \vec{H_0}^* + \vec{H_{1c}})$, where $(\vec{E_0}^*, \vec{H_0}^*)$ is the field caused by the source P^* and $(\vec{E_{1c}}, \vec{H_{1c}})$ is the field resulting from the sources induced in the conducting portions of C.

* Received by the Institute, June 25, 1951.

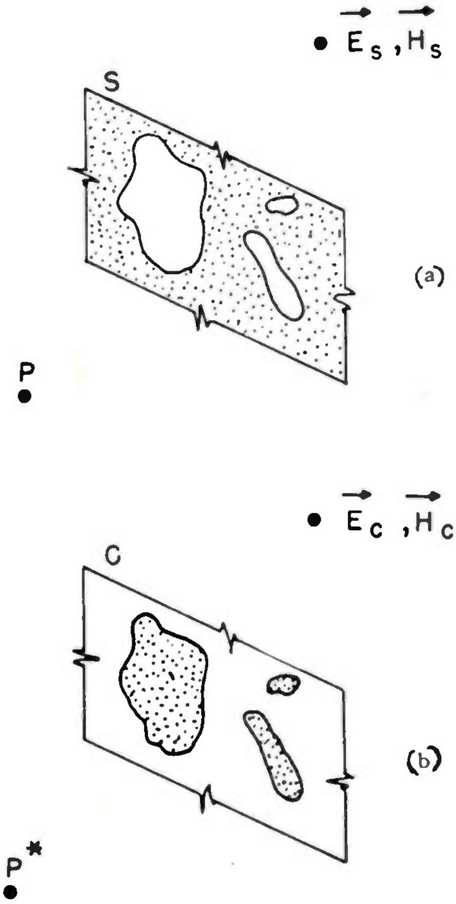

Fig. 1(a) and (b)—Both screens are infinite in extent. The dotted regions represent conductor.

If P^* is chosen so that $\vec{H_0}^* = \vec{E_0}$[1], it can be shown[2,3] that

$$\vec{H_c} = -\vec{E_{1s}}, \quad \vec{E_c} = \frac{\mu}{\epsilon} \vec{H_{1s}} \tag{1}$$

so that

$$\vec{E_s} + \vec{H_c} = \vec{E_0}. \tag{2}$$

Equation (2) is an accurate statement of Babinet's Principle.

If $\vec{E_s}$ *and* $\vec{H_c}$ *are collinear at every point,* (2) can be written as a scalar equation and divided by E_0 to give

$$\frac{E_s}{E_0} + \frac{H_c}{E_0} = 1; \tag{3}$$

and since P^* has been chosen so that $\vec{H_0}^* = \vec{E_0}$, (3) can alternatively be expressed as

$$\frac{E_s}{E_0} + \frac{H_c}{H_0^*} = 1. \tag{4}$$

The prorated diffracted fields may be defined by

$$U_1 = \frac{E_s}{E_0}, \qquad U_2 = \frac{H_c}{H_0^*}$$

to obtain Booker's extension[4] of Babinet's Principle,

$$U_1 + U_2 = 1. \tag{5}$$

This is a scalar relationship which is *only* valid if $\vec{E_s}$ and $\vec{H_c}$ are everywhere collinear. But this condition is met only by a restricted class of apertures. A simple case will serve to illustrate this point. In Fig. 2 a horizontal dipole is shown exciting an infinite conducting screen which has an inclined rectangular slot cut in it. To the right of the screen, $\vec{E_0}$ is horizontally polarized but $\vec{E_{1s}}$ is not. Since $\vec{E_s} = \vec{E_0} + \vec{E_{1s}}$, it follows that $\vec{E_s}$ and $\vec{E_0}$ are not collinear for this case; hence, from (2) $\vec{E_s}$ and $\vec{H_c}$ are not collinear.

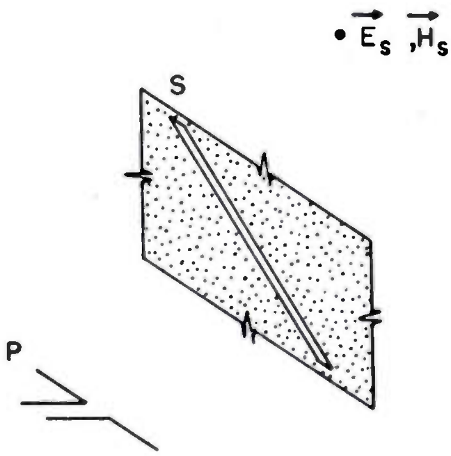

Fig. 2

Thus the scalar relation (5) is not generally applicable. Indeed, it only applies for slots which have symmetry about the polarization axis of the primary source P.

The vector formulation is the one most generally suitable, and the mathematical statement of Babinet's Principle should be

$$\vec{E_s} + \vec{H_c} = \vec{E_0} \tag{2}$$

valid for thin, perfectly conducting, infinite, plane complementary screens excited by conjugate sources.

R. S. ELLIOTT
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[1] Mathematically, this may be achieved as follows: Replace all of the sources in P by their equivalent magnetic currents and charges, thus creating a fictitious antenna P' which has *exactly* the same electromagnetic field as P. Then, at every point in P', replace the magnetic current by an electric current equal to it in magnitude and direction but reversed in time phase. This will be the system P^* which will insure the equality $\vec{E_0} = k\vec{H_0}^*$, where k equals 1 ohm and, for simplicity, has been deleted from the expressions above.

[2] L. G. Huxley, "A Survey of the Principles and Practice of Wave Guides," Cambridge University Press, Cambridge, England; 1947.

[3] E. T. Copson, "An integral equation method of solving plane diffraction problems," *Proc. Roy. Soc.* 186A., p. 100; 1946.

[4] H. G. Booker, "Slot aerials and their relation to complementary wire aerials," *Jour. IEE* (London), vol. 93, pt. IIIA; April, 1946.

Institute News and Radio Notes

IRE Western Convention Plans Progress

The 1952 IRE Western Convention and Electronics Show will be held at the Municipal Auditorium, in Long Beach, Calif., August 27–29.

The Technical Program Committee is relying to an unprecedented extent upon participation by the IRE Professional Groups in procuring papers and arranging sessions. Eleven Professional Groups have committed themselves to arranging a minimum of 16 technical sessions in their special fields. These Groups and their local representatives are as follows: Airborne Electronics, G. M. Greene; Antennas and Propagation, L. C. Van Atta; Audio, J. K. Hilliard; Broadcast and Television Receivers, H. E. Rice; Broadcast Transmission Systems, P. G. Caldwell; Circuit Theory, W. R. Abbott; Electron Devices, H. Q. North, Electronic Computers, H. E. Huskey; Information Theory, A. S. Fulton; Instrumentation, W. D. Hershberger; Radio Telemetry and Remote Control, F. W. Lehan. Members of these Professional Groups are urged to submit papers and to attend the Convention.

The Convention will be of interest to all IRE Members both for the extensive technical program and for the larger West Coast Electronic Manufacturers Association Show.

Technical Committee Notes

The Standards Committee met under the Chairmanship of A. G. Jensen on March 20. The Chairman proposed that H. E. Roys be appointed IRE Representative on ASA Z57, with A. W. Friend as alternate, and the Committee approved the appointment. The Committee discussed a proposal of the **Industrial Electronics** Committee to establish a joint Committee on Magnetic Amplifiers. John Dalke said that the AIEE Committee on Magnetic Amplifiers has agreed to the formation of such a joint committee providing it could be set up within the IRE structure. The Standards Committee voted to request the Executive Committee for approval of the formation of a joint AIEE-IRE Committee on Magnetic Amplifiers, the IRE members of which would constitute a subcommittee of the Standards Committee. Amplifying his previous comments on the relationship between CCIR and the IRE which were made at the last meeting of the Standards Committee, Chairman Jensen told the Committee of four new programs which have been initiated by CCIR Study Group 14 (vocabulary), of which he is chairman. These are: (a) the study of feasibility of adopting the international decimal classification system, (b) to provide a suitable vocabulary for use in IRE and CCIR documents, (c) the consideration of the extension of international frequency designations, and (d) the question of adopting a rationalized mks system in connection with all CCIR work. The Committee then turned its attention to a further consideration of the "Proposed Standards on Receivers: Definitions of Terms," as revised by the Task Group on Receiver Definitions. The Committee will continue discussion of these definitions.

Under the Chairmanship of C. H. Page, the **Circuits** Committee met on March 21. A number of active circuit definitions, submitted by Subcommittee 4.9, were agreed upon.

The **Electronic Computers** Committee met on February 25. Robert Serrell took the chair in the absence of Nathaniel Rochester. It was proposed that a second subcommittee on definitions be established in the East to co-ordinate the work of the Eastern groups with that of W. H. Ware's Committee, now active on the West Coast. It was moved that a subcommittee be established by the Chairman to carry out such work in computer definitions, and a Chairman will be named for this new subcommittee in the near future. The Committee discussed a proposal to establish a group to handle static-storage elements and such technical problems as arise in the development of these devices. Mr. Sands (guest) explained how an informal group had grown out of discussions which took place at the recent AIEE winter meeting. Detailed information was given on the plans of this informal group concerning the standardization of terminology, test methods, the establishment of adequate electronic circuitry, and means of interchanging information. Acting on a suggestion of the Chairman, it was felt that a Subcommittee of the Electronic Computers Committee should be set up to sponsor this informal group and facilitate its task within the framework provided by the main Committee. It was then moved that the proposed subcommittee be named "Subcommittee on Static Storage Elements." Mr. Serrell informed those present that Mr. Rochester had named J. A. Rajchman as Chairman of the new subcommittee.

The **Navigation Aids** Committee convened on March 7, under the Chairmanship of P. C. Sandretto. Consideration was given to a letter from M. W. Baldwin, Vice Chairman of the Standards Committee and Definitions Co-ordinator, regarding the publishing of the "terms" which the Committee now has ready. General Sandretto asked to defer decision on this matter. Harry Davis, who is compiling a list of radar terms, was asked to check his list against the Master Index of IRE Definitions. If the resulting radar list does not exceed 100 terms, the Navigation Aids Committee will postpone publication of the list of terms already reviewed until the new list has been considered. However, if the list contains more than 100 new terms, the committee will recommend to the Standards Committee that the completed list be published.

On March 7, the **Receivers** Committee convened under the Chairmanship of Jack Avins. The supplement to IRE Standard 51 IRE 17.S1 "Open Field Method of Measurement of Spurious Radiation from FM and TV Receivers," was presented for approval. The Committee approved the supplement which will be submitted to the Standards Committee. The Chairman reported on the extended scope of the Committee. K. Jarvis, Chairman of the Single Sideband Receivers Subcommittee reported that the problem of co-ordination between receiver and transmitter groups was yet to be solved. The Subcommittee is being kept intact pending co-ordination. R. F. Shea, Chairman of the Spurious Radiation Subcommittee, reported that progress is being made on the Standard for sweep radiation measurement which is now in preliminary form. Consideration is being given to measurements in the uhf range. There was unanimous approval on revising the 1948 Standard on Methods of Testing TV Receivers to bring the procedures into agreement with current practice and to produce an integrated Standard to replace the current Standard which includes a supplement for intercarrier sound receivers. The Chairman will activate a subcommittee to work on this revision. A vote of commendation was given to I. E. Lempert for his review of the current Standard and the suggestion of much needed changes.

Under the Chairmanship of W. M. Pease, the **Servo-Systems** Committee met on March 25. Plans were discussed for a technical meeting to promote interest in the formation of a Professional Group on Feedback Control Systems. Arrangements for a sectional meeting will be discussed with the Boston Section Chairman. The draft of report, "Present Status of ASA Committee Y10.14 Activities," was reviewed and modified.

The **Symbols** Committee convened on March 28, under the Chairmanship of A. G. Clavier. The Chairman called on C. D. Mitchell, Chairman of the Task Group on Graphical Symbols for Semiconductors, for a report on the work in his group. W. B. Callaway, Chairman of the Task Group on Symbols for Functional Operations of Control, Computing, and Switching Equipment, gave a report on the present status of work in his group. A report by A. F. Pomeroy on the status of ASA standards for Graphical Symbols was given.

Calendar of COMING EVENTS

Conference on Semiconductor Device Research, Univ. of Ill., Urbana, Ill., June 19–20.

AIEE-IRE Telemetering Conference, Los Angeles, Calif., August 26–27

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29

Cedar Rapids IRE Technical Conference, Roosevelt Hotel, Cedar Rapids, Iowa, September 20

National Electronics Conference, Sherman Hotel, Chicago, Ill., September 29–October 1

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

1953 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 23–26

Professional Group News

The paper "Measuring Techniques for Broad-Band Long-Distance Radio Relay Systems," by W. J. Albersheim, which appeared on pages 548–551, of the May, 1952 issue of the PROCEEDINGS, was published with the approval of the IRE Professional Group on Instrumentation.

Airborne Electronics

The Group's Airborne Electronics Conference held in Dayton, Ohio, May 12–14, was a huge success. The first issue of the Airborne Electronics Group's NEWSLETTER, containing 50 pages and issued to the members, was distributed to those attending the Dayton Conference. The TRANSACTIONS (PGAE-3) has been mailed to the Group members.

A Los Angeles Chapter of the Airborne Electronics Group has been formed, and plans are being organized for a Chicago Chapter.

Antennas and Propagation

The Group on Antennas and Propagation sponsored the Spring Meeting of URSI, at the National Bureau of Standards, Washington, D. C., April 21–24. The Group plans to sponsor the 1953 Conference on Radio-Meteorology in Austin, Tex. The publication of four TRANSACTIONS by the Group is planned for each year.

L. C. Van Atta has been appointed representative of this Group to the Sections in the Southern California area.

Audio

The TRANSACTIONS (PGA-7) was mailed to members of the Audio Group in May. It is planned to publish PGA TRANSACTIONS every second month.

The Boston and Cincinnati Chapters of the Audio Group are holding well-attended monthly meetings; excellent papers are being presented. At the Group meeting in the Cincinnati Section, W. E. Stewart, of RCA, Camden, N. J., presented a paper entitled, "High Fidelity Design Factors."

Broadcast and Television Receivers

The contents of the round table discussion held during the IRE 1952 National Convention will be published as the first TRANSACTIONS of the Broadcast and Television Receivers Group.

Sessions were sponsored by this Group at the IRE Cincinnati Section Spring Technical Conference, and the 4th Southwestern IRE Conference and Radio Engineering Show held in Houston, Tex.

H. E. Rice, session organizer, reports that five papers have been obtained for the Group's session at the IRE Western Convention. Papers for the IRE/RTMA Fall Meeting in Syracuse, October 20–22, are also being solicited.

Broadcast Transmission Systems

The Administrative Committee of the Broadcast Transmission Systems Group has selected a list of nominees to be submitted to the membership for election to a three-year term. Lewis Winner has been reelected Chairman of the Group.

A capacity audience registered for the Group's morning and afternoon sessions at the IRE 1952 National Convention, held in the Blue Room of Grand Central Palace. During the Administrative Committee meeting at the Convention, held in conjunction with a dinner and cocktail party, the 1952 program was discussed, and the Chairman announced that a two-day exhibit and technical conference would probably be held in the early Fall.

The Group Chairman attended the final meeting for the season of the Boston Chapter of Broadcast Transmission Systems. Problems were discussed as well as future programs for TRANSACTIONS, NEWSLETTERS, and meetings. It was announced during the meeting that the following members would serve as 1952 officers of the Boston chapter: P. K. Baldwin, Station WHDH, Chairman; S. V. Stadig, Station WBZ-TV, Vice Chairman; and Hollis Gray, Station WHDH, Secretary-Treasurer. The meeting featured a talk by Ralph Harmon, Engineering Manager of Westinghouse Radio Stations, on "Variables Affection Coverage in Vhf and Uhf Television." The sessions were conducted in the auditorium studio of station WBZ-TV.

Communications

The IRE Executive Committee has approved the new Professional Group on Communications. Members of the Group's Administrative Committee are: G. T. Royden (Chairman), J. J. Callahan, Andre Clavier, I. S. Coggeshall, J. N. Craver, M. G. Crosby, D. D. Donald, E. E. Eldredge, W. G. H. Finch, John Hessel, A. A. Kunze, G. M. Neely, A. C. Peterson, Jr., D. S. Rau, W. B. Sullinger, and D. C. Trafton.

This Group will sponsor a Symposium at the Mackay Radio Transmitting Station, Brentwood, L. I., N. Y., on June 21, 1952. Those interested should write to G. T. Royden, Chairman, Mackay Radio and Telegraph Co., 67 Broad Street, New York, N. Y., for details.

Electronic Computers

A successful Electronic Computer Symposium entitled, "Engineering Tomorrow's Computers," was held at the University of California in Los Angeles, April 30–May 1, 1952. The Group is now making plans for a joint IRE/AIEE conference for December 1–3, 1952, in New York, N. Y.

The Group on Electronic Computers has appointed the following committees: Membership and Finance Committee, J. R. Weiner, Chairman; Meetings and Conventions Committee, C. V. L. Smith, Chairman; and Papers Study and Procurement Committee, J. H. Felker, Chairman.

Electron Devices

The Group on Electron Devices plans to sponsor a technical session at the IRE/RTMA Fall Meeting in Syracuse, October 22, 1952.

Consideration is being given to the forming of chapters of the Electron Devices Group in Philadelphia, Princeton, and other interested areas.

Engineering Management

An interested group of IRE members who have held sessions on engineering management at Section meetings is considering an Engineering Management Chapter in Los Angeles, Calif.

Industrial Electronics

The Industrial Electronics Group held a successful symposium on "Electronics and Machines" in Chicago, May 22–23, 1952.

Information Theory

A. S. Hughes of the Los Angeles Chapter on Information Theory has been appointed to organize the session at the IRE West Coast Convention.

(*Continued on page 732*)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

The following issues of TRANSACTIONS were recently published by IRE Professional Groups and additional copies are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N.Y., at the prices listed below.

A list of previously published issues appeared in the March and May, 1952 issues of PROCEEDINGS.

Sponsoring Group	Publication	Group Members	IRE Members	Non-members*
Audio	PGA-7; Editorials, technical papers, and news (48 pages)	$ .90	$1.35	$2.70
Airborne Electronics	PGAE-3; Symposium on the Integration of Electronic Equipment with Air-Frame Design (28 pages)	$ .85	$1.25	$2.50
Instrumentation	PGI-1; Symposium on Subaudio Instrumentation (104 pages)	$1.10	$1.65	$3.30

* Public libraries and colleges can purchase copies at IRE Member rates.

Contributors to Proceedings of the I.R.E.

M. C. Pease (M'47–SM'51) was born in New York, N. Y., in 1920. He received the B.S. degree in chemistry from Yale University in 1940 and the M.A. degree in physical chemistry from Princeton University in 1943. He worked at the Radio Research Laboratory at Harvard University (a war laboratory for the development of countermeasures for radar), where he was principally concerned with the 1-kw series of cw magnetrons. In 1945 he left this position to accept a commission in the U. S. Navy as a radar officer.

M. C. PEASE

After his discharge from the Navy, Mr. Pease joined the Electronics Division of Sylvania Electric Products, Inc. Initially he was concerned with various theoretical problems, largely in the field of microwave networks. He is now manager of Tube-Development Engineering, and has charge of the development of magnetrons, gas switch tubes, thyratrons, strobotrons, and the like.

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R. F. Rinehart was born in Springfield, Ohio, on May 31, 1907. He received the A.B. degree from Wittenberg College in 1930, and the M.A. and Ph.D. degrees in mathematics from Ohio State University in 1932 and 1934. He taught mathematics at Ashland College, Ashland, Ohio, from 1934 to 1937. Since 1937 he has been associated with the mathematics department of Case Institute of Technology where he is now professor of mathematics.

From 1942 to 1945 Dr. Rinehart was a member of the Operations Research Group of the U. S. Navy, engaged in analyses of naval tactics and strategy in subsurface warfare. He was awarded the Medal of Freedom for his work in antisubmarine warfare in the Caribbean Sea Frontier, and the Medal for Merit for achievements in submarine warfare in the Pacific.

R. F. RINEHART

From 1948 to 1950 Dr. Rinehart served with the Research and Development Board of the Department of Defense, first as director of the Planning Division, and subsequently as executive secretary. He was also the acting chairman of that Board for five months following Dr. K. T. Compton's resignation.

Dr. Rinehart is a member of Sigma Xi, the American Mathematical Society, and the Mathematical Association of America.

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For a photograph and biography of S. T. SMITH, see page 359 of the March, 1952, issue of the PROCEEDINGS OF THE I.R.E.

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W = bandwidth of channel in cps
P = average signal power
N = average noise power assumed to be thermal or "white."

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Regarding (1), we see that the higher the signal power P the greater the information rate C. One may ask, "At what signal level should we transmit in order to maximize C/P (the information rate per unit of signal power)?"

From (1)

$$\frac{C}{P} = \frac{W}{P} \log_2 (1 + P/N). \tag{2}$$

Taking the derivative of (2) with respect to P and setting it to zero, it is found that an upper bound to C/P is approached as

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and therefore the upper bound is given by

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T = absolute temperature.

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and (5) in (3) gives

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Correspondence

and that the lower bound to the signal power required to transmit C is

$$P_{lb} = CkT \log_e 2. \quad (8)$$

As an example of the use of these equations, assume that we wish to transmit 10^6 bits of information per second (at 300°K).

Equation (8) says that the signal power can never be less 2.85×10^{-15} watts. Similarly, a repeater that is to pass information at a rate of 10^6 bits per second must certainly have a battery drain of more than 2.85×10^{-15} watts.

A useful figure of merit (having a maximum value of one) can be computed for communication devices. Suppose an *n-p-n* transistor regenerative amplifier could be designed to respond to information at a rate of 10^6 bits per second and required only 1 microwatt of battery power. Its figure of merit would be 2.85×10^{-9}. This shows that in the design of switching systems and computers, we have a long way to go before we reach the fundamental floor under which we cannot operate.

C times time gives bits and P times time gives energy. Multiplying (8) by one second, we find that at 300°K, one binary bit of information will always cost more than 2.85×10^{-21} joules.

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J. H. Felker
Bell Telephone Laboratories, Inc.
Whippany, N. J.

A Note on Booker's Extension of Babinet's Principle*

Some differences exist in the literature today as to the proper interpretation of Babinet's Principle extended to apply to coherent waves. A few words of clarification would seem in order.

Consider an antenna source P to the left of a perfectly conducting, indefinitely thin, infinite plane screen S (Fig. 1(a)). S contains an aggregate of holes of arbitrary shape and position. The total field to the right of S is $(\vec{E}_s = \vec{E}_0 + \vec{E}_{1s},\ \vec{H}_s = \vec{H}_0 + \vec{H}_{1s})$, where $(\vec{E}_0, \vec{H}_0)$ is the field due to the source P, and $(\vec{E}_{1s}, \vec{H}_{1s})$ is the field due to the sources induced in the conducting portions of S.

Next, replace P by another source P^* and replace S by its complementary screen C (Fig. 1(b)). On C the conducting and open regions of S have been interchanged. The total field to the right of C is $(\vec{E}_c = \vec{E}_0^* + \vec{E}_{1c},\ \vec{H}_c = \vec{H}_0^* + \vec{H}_{1c})$, where $(\vec{E}_0^*, \vec{H}_0^*)$ is the field caused by the source P^* and $(\vec{E}_{1c}, \vec{H}_{1c})$ is the field resulting from the sources induced in the conducting portions of C.

* Received by the Institute, June 25, 1951

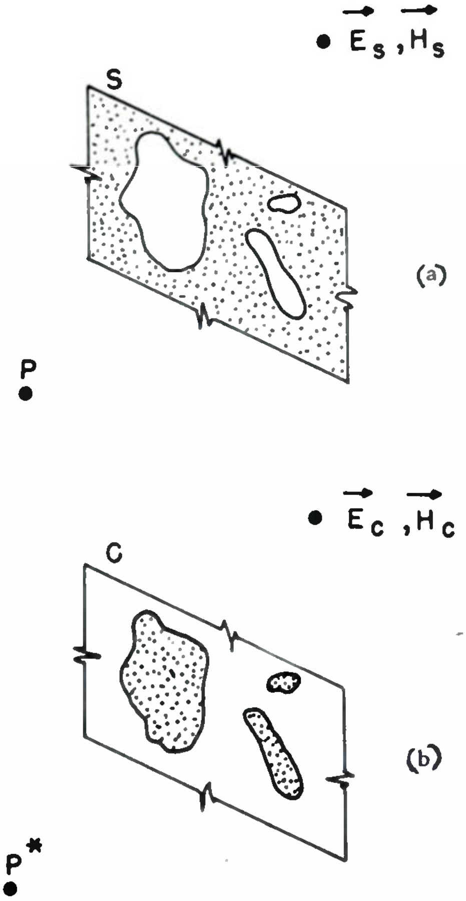

Fig. 1(a) and (b)—Both screens are infinite in extent. The dotted regions represent conductor.

If P^* is chosen so that $\vec{H}_0^* = \vec{E}_0$[1], it can be shown[2,3] that

$$\vec{H}_c = -\vec{E}_{1s}, \quad \vec{E}_c = \frac{\mu}{\epsilon}\vec{H}_1 \quad (1)$$

so that

$$\vec{E}_s + \vec{H}_c = \vec{E}_0. \quad (2)$$

Equation (2) is an accurate statement of Babinet's Principle.

If $\vec{E}_s$ *and* $\vec{H}_c$ *are collinear at every point*, (2) can be written as a scalar equation and divided by E_0 to give

$$\frac{E_s}{E_0} + \frac{H_c}{E_0} = 1; \quad (3)$$

and since P^* has been chosen so that $\vec{H}_0^* = \vec{E}_0$, (3) can alternatively be expressed as

$$\frac{E_s}{E_0} + \frac{H_c}{H_0^*} = 1. \quad (1)$$

The prorated diffracted fields may be defined by

$$U_1 = \frac{E_s}{E_0}, \qquad U_2 = \frac{H_c}{H_0^*}$$

to obtain Booker's extension[4] of Babinet's Principle,

$$U_1 + U_2 = 1. \quad (5)$$

This is a scalar relationship which is *only* valid if $\vec{E}_s$ and $\vec{H}_c$ are everywhere collinear. But this condition is met only by a restricted class of apertures. A simple case will serve to illustrate this point. In Fig. 2 a horizontal dipole is shown exciting an infinite conducting screen which has an inclined rectangular slot cut in it. To the right of the screen, $\vec{E}_0$ is horizontally polarized but $\vec{E}_{1s}$ is not. Since $\vec{E}_s = \vec{E}_0 + \vec{E}_{1s}$, it follows that $\vec{E}_s$ and $\vec{E}_0$ are not collinear for this case; hence, from (2) $\vec{E}_s$ and $\vec{H}_c$ are not collinear.

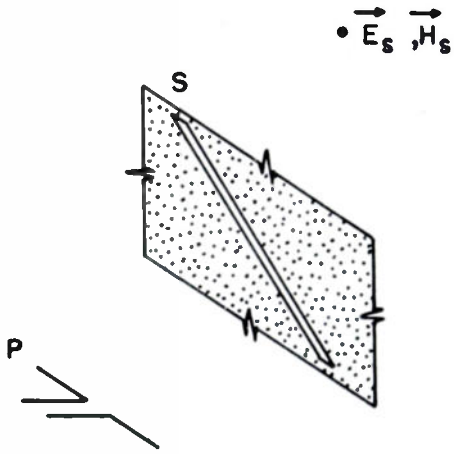

Fig. 2

Thus the scalar relation (5) is not generally applicable. Indeed, it only applies for slots which have symmetry about the polarization axis of the primary source P.

The vector formulation is the one most generally suitable, and the mathematical statement of Babinet's Principle should be

$$\vec{E}_s + \vec{H}_c = \vec{E}_0 \quad (2)$$

valid for thin, perfectly conducting, infinite, plane complementary screens excited by conjugate sources.

R. S. Elliott
Electrical Engineering Department
University of Illinois
Urbana, Ill.

[1] Mathematically, this may be achieved as follows: Replace all of the sources in P by their equivalent magnetic currents and charges, thus creating a fictitious antenna P' which has *exactly* the same electromagnetic field as P. Then, at every point in P', replace the magnetic current by an electric current equal to it in magnitude and direction but reversed in time phase. This will be the system P^* which will insure the equality $\vec{E}_0 = k\vec{H}_0^*$, where k equals 1 ohm and, for simplicity, has been deleted from the expressions above.

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Contributors to Proceedings of the I.R.E.

M. C. Pease (M'47–SM'51) was born in New York, N. Y., in 1920. He received the B.S. degree in chemistry from Yale University in 1940 and the M.A. degree in physical chemistry from Princeton University in 1943. He worked at the Radio Research Laboratory at Harvard University (a war laboratory for the development of countermeasures for radar), where he was principally concerned with the 1-kw series of cw magnetrons. In 1945 he left this position to accept a commission in the U. S. Navy as a radar officer.

M. C. Pease

After his discharge from the Navy, Mr. Pease joined the Electronics Division of Sylvania Electric Products, Inc. Initially he was concerned with various theoretical problems, largely in the field of microwave networks. He is now manager of Tube-Development Engineering, and has charge of the development of magnetrons, gas switch tubes, thyratrons, strobotrons, and the like.

❖

R. F. Rinehart was born in Springfield, Ohio, on May 31, 1907. He received the A.B. degree from Wittenberg College in 1930, and the M.A. and Ph.D. degrees in mathematics from Ohio State University in 1932 and 1934. He taught mathematics at Ashland College, Ashland, Ohio, from 1934 to 1937. Since 1937 he has been associated with the mathematics department of Case Institute of Technology where he is now professor of mathematics.

From 1942 to 1945 Dr. Rinehart was a member of the Operations Research Group of the U. S. Navy, engaged in analyses of naval tactics and strategy in subsurface warfare. He was awarded the Medal of Freedom for his work in antisubmarine warfare in the Caribbean Sea Frontier, and the Medal for Merit for achievements in submarine warfare in the Pacific.

R. F. Rinehart

From 1948 to 1950 Dr. Rinehart served with the Research and Development Board of the Department of Defense, first as director of the Planning Division, and subsequently as executive secretary. He was also the acting chairman of that Board for five months following Dr. K. T. Compton's resignation.

Dr. Rinehart is a member of Sigma Xi, the American Mathematical Society, and the Mathematical Association of America.

❖

For a photograph and biography of S. T. Smith, see page 359 of the March, 1952, issue of the Proceedings of the I.R.E.

❖

Ping King Tien was born in Checkiang, China, on August 2, 1919. He received the B.S. degree in electrical engineering from the National Central University of China in 1942, and from Stanford University the M.S. degree in 1948 and the Ph.D. degree in 1951.

Ping King Tien

Dr. Tien was an engineer at the Tien-Sun Industrial Company, Shanghai, China, from 1942 to 1947, and was a research assistant at Stanford University from 1949 to 1951. At present he is a research associate in the Electronics Research Laboratory of Stanford University, where he is engaged in microwave-tube research.

Dr. Tien is a member of Sigma Xi.

❖

William M. Webster (A'48) was born in Warsaw, N. Y., in 1925. He studied physics at Rensselaer Polytechnic Institute, and at Union College as a Navy V-12 student. He received a B.S. degree in physics and joined the RCA Laboratories Division, Princeton, N. J. He has worked in the fields of solid state devices and vacuum and gaseous electronics.

W. M. Webster

Mr. Webster is an associate member of the I.R.E.

Correspondence

A Link Between Information and Energy*

With the advent of the transistor and its striking ability to operate on low-power, it is pertinent to inquire what are the ultimate limits upon low power operation of switching systems, computers, and other communication machinery.

Mr. C. E. Shannon has derived the relationship for the capacity of an information channel

$$C = W \log_2 \left(1 + \frac{P}{N}\right) \tag{1}$$

where

C = maximum number of binary bits of information that can be transmitted per second
W = bandwidth of channel in cps
P = average signal power
N = average noise power assumed to be thermal or "white."

* Received by the Institute, February 15, 1952.

Regarding (1), we see that the higher the signal power P the greater the information rate C. One may ask, "At what signal level should we transmit in order to maximize C/P (the information rate per unit of signal power)?"

From (1)

$$\frac{C}{P} = \frac{W}{P} \log_2 (1 + P/N). \tag{2}$$

Taking the derivative of (2) with respect to P and setting it to zero, it is found that an upper bound to C/P is approached as

$$\frac{P}{N} \to 0,$$

and therefore the upper bound is given by

$$\left(\frac{C}{P}\right)_{\text{ub}} = \frac{W}{N} \log_2 e. \tag{3}$$

One may inquire as to the upper bound to W/N. In a circuit of bandwidth W, the minimum available noise power will be the thermal noise power

$$N = \frac{E^2}{4R} = kTW \tag{4}$$

where

k = Boltzmann's constant = 1.374×10^{-23} joule per °K
T = absolute temperature.

From this we have the relationship below as the upper bound to W/N

$$\left(\frac{W}{N}\right)_{\text{ub}} = \frac{1}{kT} \tag{5}$$

and (5) in (3) gives

$$\left(\frac{C}{P}\right)_{\text{ub}} = \frac{1}{kT} \log_2 e, \tag{6}$$

from which we can conclude that, with an available signal power P, the upper bound to the number of binary bits per second that can be transmitted is given by

$$C_{\text{ub}} = \frac{P}{kT} \log_2 e \tag{7}$$

Correspondence

and that the lower bound to the signal power required to transmit C is

$$P_{lb} = CkT \log_e 2. \qquad (8)$$

As an example of the use of these equations, assume that we wish to transmit 10^6 bits of information per second (at 300°K).

Equation (8) says that the signal power can never be less 2.85×10^{-15} watts. Similarly, a repeater that is to pass information at a rate of 10^6 bits per second must certainly have a battery drain of more than 2.85×10^{-15} watts.

A useful figure of merit (having a maximum value of one) can be computed for communication devices. Suppose an n-p-n transistor regenerative amplifier could be designed to respond to information at a rate of 10^6 bits per second and required only 1 microwatt of battery power. Its figure of merit would be 2.85×10^{-9}. This shows that in the design of switching systems and computers, we have a long way to go before we reach the fundamental floor under which we cannot operate.

C times time gives bits and P times time gives energy. Multiplying (8) by one second, we find that at 300°K, one binary bit of information will always cost more than 2.85×10^{-21} joules.

Mr. J. R. Pierce has informed the writer that it is possible to derive the same relationship between energy and information by arguments based on the energy available from the exploitation of information as to which half of an otherwise empty box contains a single gas molecule.

J. H. FELKER
Bell Telephone Laboratories, Inc.
Whippany, N. J.

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Some differences exist in the literature today as to the proper interpretation of Babinet's Principle extended to apply to coherent waves. A few words of clarification would seem in order.

Consider an antenna source P to the left of a perfectly conducting, indefinitely thin, infinite plane screen S (Fig. 1(a)). S contains an aggregate of holes of arbitrary shape and position. The total field to the right of S is $(\vec{E}_s = \vec{E}_0 + \vec{E}_{1s},\ \vec{H}_s = \vec{H}_0 + \vec{H}_{1s})$, where $(\vec{E}_0, \vec{H}_0)$ is the field due to the source P, and $(\vec{E}_{1s}, \vec{H}_{1s})$ is the field due to the sources induced in the conducting portions of S.

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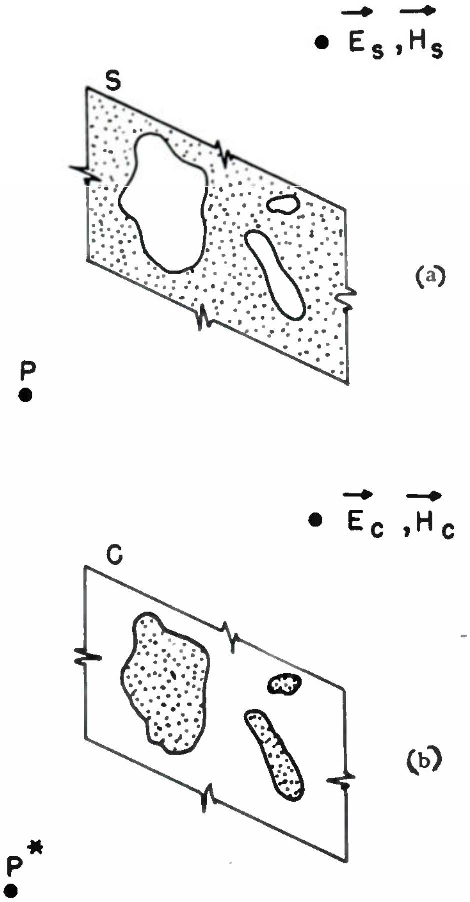

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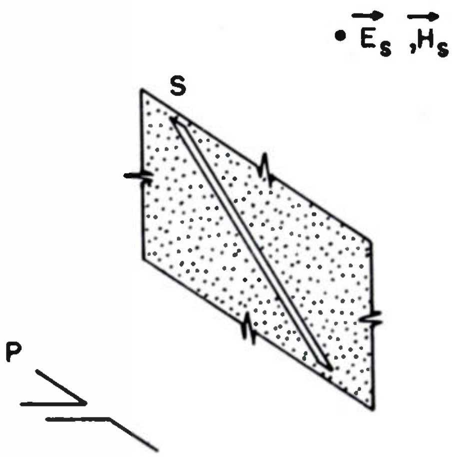

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Institute News and Radio Notes

IRE Western Convention Plans Progress

The 1952 IRE Western Convention and Electronics Show will be held at the Municipal Auditorium, in Long Beach, Calif., August 27–29.

The Technical Program Committee is relying to an unprecedented extent upon participation by the IRE Professional Groups in procuring papers and arranging sessions. Eleven Professional Groups have committed themselves to arranging a minimum of 16 technical sessions in their special fields. These Groups and their local representatives are as follows: Airborne Electronics, G. M. Greene; Antennas and Propagation, L. C. Van Atta; Audio, J. K. Hilliard; Broadcast and Television Receivers, H. E. Rice; Broadcast Transmission Systems, P. G. Caldwell; Circuit Theory, W. R. Abbott; Electron Devices, H. Q. North; Electronic Computers, H. E. Huskey; Information Theory, A. S. Fulton; Instrumentation, W. D. Hershberger; Radio Telemetry and Remote Control, F. W. Lehan. Members of these Professional Groups are urged to submit papers and to attend the Convention.

The Convention will be of interest to all IRE Members both for the extensive technical program and for the larger West Coast Electronic Manufacturers Association Show.

Technical Committee Notes

The Standards Committee met under the Chairmanship of A. G. Jensen on March 20. The Chairman proposed that H. E. Roys be appointed IRE Representative on ASA Z57, with A. W. Friend as alternate, and the Committee approved the appointment. The Committee discussed a proposal of the **Industrial Electronics** Committee to establish a joint Committee on Magnetic Amplifiers. John Dalke said that the AIEE Committee on Magnetic Amplifiers has agreed to the formation of such a joint committee providing it could be set up within the IRE structure. The Standards Committee voted to request the Executive Committee for approval of the formation of a joint AIEE-IRE Committee on Magnetic Amplifiers, the IRE members of which would constitute a subcommittee of the Standards Committee. Amplifying his previous comments on the relationship between CCIR and the IRE which were made at the last meeting of the Standards Committee, Chairman Jensen told the Committee of four new programs which have been initiated by CCIR Study Group 14 (vocabulary), of which he is chairman. These are: (a) the study of feasibility of adopting the international decimal classification system, (b) to provide a suitable vocabulary for use in IRE and CCIR documents, (c) the consideration of the extension of international frequency designations, and (d) the question of adopting a rationalized mks system in connection with all CCIR work. The Committee then turned its attention to a further consideration of the "Proposed Standards on Receivers: Definitions of Terms," as revised by the Task Group on Receiver Definitions. The Committee will continue discussion of these definitions.

Under the Chairmanship of C. H. Page, **the Circuits** Committee met on March 21. A number of active circuit definitions, submitted by Subcommittee 4.9, were agreed upon.

The **Electronic Computers** Committee met on February 25. Robert Serrell took the chair in the absence of Nathaniel Rochester. It was proposed that a second subcommittee on definitions be established in the East to co-ordinate the work of the Eastern groups with that of W. H. Ware's Committee, now active on the West Coast. It was moved that a subcommittee be established by the Chairman to carry out such work in computer definitions, and a Chairman will be named for this new subcommittee in the near future. The Committee discussed a proposal to establish a group to handle static-storage elements and such technical problems as arise in the development of these devices. Mr. Sands (guest) explained how an informal group had grown out of discussions which took place at the recent AIEE winter meeting. Detailed information was given on the plans of this informal group concerning the standardization of terminology, test methods, the establishment of adequate electronic circuitry, and means of interchanging information. Acting on a suggestion of the Chairman, it was felt that a Subcommittee of the Electronic Computers Committee should be set up to sponsor this informal group and facilitate its task within the framework provided by the main Committee. It was then moved that the proposed subcommittee be named "Subcommittee on Static Storage Elements." Mr. Serrell informed those present that Mr. Rochester had named J. A. Rajchman as Chairman of the new subcommittee.

The **Navigation Aids** Committee convened on March 7, under the Chairmanship of P. C. Sandretto. Consideration was given to a letter from M. W. Baldwin, Vice Chairman of the Standards Committee and Definitions Co-ordinator, regarding the publishing of the "terms" which the Committee now has ready. General Sandretto asked to defer decision on this matter. Harry Davis, who is compiling a list of radar terms, was asked to check his list against the Master Index of IRE Definitions. If the resulting radar list does not exceed 100 terms, the Navigation Aids Committee will postpone publication of the list of terms already reviewed until the new list has been considered. However, if the list contains more than 100 new terms, the committee will recommend to the Standards Committee that the completed list be published.

On March 7, the **Receivers** Committee convened under the Chairmanship of Jack Avins. The supplement to IRE Standard 51 IRE 17.S1 "Open Field Method of Measurement of Spurious Radiation from FM and TV Receivers," was presented for approval. The Committee approved the supplement which will be submitted to the Standards Committee. The Chairman reported on the extended scope of the Committee. K. Jarvis, Chairman of the Single Sideband Receivers Subcommittee reported that the problem of co-ordination between receiver and transmitter groups was yet to be solved. The Subcommittee is being kept intact pending co-ordination. R. F. Shea, Chairman of the Spurious Radiation Subcommittee, reported that progress is being made on the Standard for sweep radiation measurement which is now in preliminary form. Consideration is being given to measurements in the uhf range. There was unanimous approval on revising the 1948 Standard on Methods of Testing TV Receivers to bring the procedures into agreement with current practice and to produce an integrated Standard to replace the current Standard which includes a supplement for intercarrier sound receivers. The Chairman will activate a subcommittee to work on this revision. A vote of commendation was given to I. E. Lempert for his review of the current Standard and the suggestion of much needed changes.

Under the Chairmanship of W. M. Pease, the **Servo-Systems** Committee met on March 25. Plans were discussed for a technical meeting to promote interest in the formation of a Professional Group on Feedback Control Systems. Arrangements for a sectional meeting will be discussed with the Boston Section Chairman. The draft of report, "Present Status of ASA Committee Y10.14 Activities," was reviewed and modified.

The **Symbols** Committee convened on March 28, under the Chairmanship of A. G. Clavier. The Chairman called on C. D. Mitchell, Chairman of the Task Group on Graphical Symbols for Semiconductors, for a report on the work in his group. W. B Callaway, Chairman of the Task Group on Symbols for Functional Operations of Control, Computing, and Switching Equipment, gave a report on the present status of work in his group. A report by A. F. Pomeroy on the status of ASA standards for Graphical Symbols was given.

Calendar of COMING EVENTS

Conference on Semiconductor Device Research, Univ. of Ill., Urbana, Ill., June 19–20.

AIEE-IRE Telemetering Conference, Los Angeles, Calif., August 26–27

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29

Cedar Rapids IRE Technical Conference, Roosevelt Hotel, Cedar Rapids, Iowa, September 20

National Electronics Conference, Sherman Hotel, Chicago, Ill., September 29–October 1

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

1953 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 23–26

Professional Group News

The paper "Measuring Techniques for Broad-Band Long-Distance Radio Relay Systems," by W. J. Albersheim, which appeared on pages 548–551, of the May, 1952 issue of the PROCEEDINGS, was published with the approval of the IRE Professional Group on Instrumentation.

Airborne Electronics

The Group's Airborne Electronics Conference held in Dayton, Ohio, May 12–14, was a huge success. The first issue of the Airborne Electronics Group's NEWSLETTER, containing 50 pages and issued to the members, was distributed to those attending the Dayton Conference. The TRANSACTIONS (PGAE-3) has been mailed to the Group members.

A Los Angeles Chapter of the Airborne Electronics Group has been formed, and plans are being organized for a Chicago Chapter.

Antennas and Propagation

The Group on Antennas and Propagation sponsored the Spring Meeting of URSI, at the National Bureau of Standards, Washington, D. C., April 21–24. The Group plans to sponsor the 1953 Conference on Radio-Meteorology in Austin, Tex. The publication of four TRANSACTIONS by the Group is planned for each year.

L. C. Van Atta has been appointed representative of this Group to the Sections in the Southern California area.

Audio

The TRANSACTIONS (PGA-7) was mailed to members of the Audio Group in May. It is planned to publish PGA TRANSACTIONS every second month.

The Boston and Cincinnati Chapters of the Audio Group are holding well-attended monthly meetings; excellent papers are being presented. At the Group meeting in the Cincinnati Section, W. E. Stewart, of RCA, Camden, N. J., presented a paper entitled, "High Fidelity Design Factors."

Broadcast and Television Receivers

The contents of the round table discussion held during the IRE 1952 National Convention will be published as the first TRANSACTIONS of the Broadcast and Television Receivers Group.

Sessions were sponsored by this Group at the IRE Cincinnati Section Spring Technical Conference, and the 4th Southwestern IRE Conference and Radio Engineering Show held in Houston, Tex.

H. E. Rice, session organizer, reports that five papers have been obtained for the Group's session at the IRE Western Convention. Papers for the IRE/RTMA Fall Meeting in Syracuse, October 20–22, are also being solicited.

Broadcast Transmission Systems

The Administrative Committee of the Broadcast Transmission Systems Group has selected a list of nominees to be submitted to the membership for election to a three-year term. Lewis Winner has been reelected Chairman of the Group.

A capacity audience registered for the Group's morning and afternoon sessions at the IRE 1952 National Convention, held in the Blue Room of Grand Central Palace. During the Administrative Committee meeting at the Convention, held in conjunction with a dinner and cocktail party, the 1952 program was discussed, and the Chairman announced that a two-day exhibit and technical conference would probably be held in the early Fall.

The Group Chairman attended the final meeting for the season of the Boston Chapter of Broadcast Transmission Systems. Problems were discussed as well as future programs for TRANSACTIONS, NEWSLETTERS, and meetings. It was announced during the meeting that the following members would serve as 1952 officers of the Boston chapter: P. K. Baldwin, Station WHDH, Chairman; S. V. Stadig, Station WBZ-TV, Vice Chairman; and Hollis Gray, Station WHDH, Secretary-Treasurer. The meeting featured a talk by Ralph Harmon, Engineering Manager of Westinghouse Radio Stations, on "Variables Affection Coverage in Vhf and Uhf Television." The sessions were conducted in the auditorium studio of station WBZ-TV.

Communications

The IRE Executive Committee has approved the new Professional Group on Communications. Members of the Group's Administrative Committee are: G. T. Royden (Chairman), J. J. Callahan, Andre Clavier, I. S. Coggeshall, J. N. Craver, M. G. Crosby, D. D. Donald, E. E. Eldredge, W. G. H. Finch, John Hessel, A. A. Kunze, G. M. Neely, A. C. Peterson, Jr., D. S. Rau, W. B Sullinger, and D. C. Trafton.

This Group will sponsor a Symposium at the Mackay Radio Transmitting Station, Brentwood, L. I., N. Y., on June 21, 1952. Those interested should write to G. T. Royden, Chairman, Mackay Radio and Telegraph Co., 67 Broad Street, New York, N. Y., for details.

Electronic Computers

A successful Electronic Computer Symposium entitled, "Engineering Tomorrow's Computers," was held at the University of California in Los Angeles, April 30–May 1, 1952. The Group is now making plans for a joint IRE/AIEE conference for December 1–3, 1952, in New York, N. Y.

The Group on Electronic Computers has appointed the following committees: Membership and Finance Committee, J. R. Weiner, Chairman; Meetings and Conventions Committee, C. V. L. Smith, Chairman; and Papers Study and Procurement Committee, J. H. Felker, Chairman.

Electron Devices

The Group on Electron Devices plans to sponsor a technical session at the IRE/RTMA Fall Meeting in Syracuse, October 22, 1952.

Consideration is being given to the forming of chapters of the Electron Devices Group in Philadelphia, Princeton, and other interested areas

Engineering Management

An interested group of IRE members who have held sessions on engineering management at Section meetings is considering an Engineering Management Chapter in Los Angeles, Calif.

Industrial Electronics

The Industrial Electronics Group held a successful symposium on "Electronics and Machines" in Chicago, May 22–23, 1952.

Information Theory

A. S. Hughes of the Los Angeles Chapter on Information Theory has been appointed to organize the session at the IRE West Coast Convention.

(*Continued on page 732*)

Transactions of IRE Professional Groups

The following issues of TRANSACTIONS were recently published by IRE Professional Groups and additional copies are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N.Y., at the prices listed below.

A list of previously published issues appeared in the March and May, 1952 issues of PROCEEDINGS.

Sponsoring Group	Publication	Group Members	IRE Members	Non-members*
Audio	PGA-7; Editorials, technical papers, and news (48 pages)	$ .90	$1.35	$2.70
Airborne Electronics	PGAE-3; Symposium on the Integration of Electronic Equipment with Air-Frame Design (28 pages)	$ .85	$1.25	$2.50
Instrumentation	PGI-1; Symposium on Subaudio Instrumentation (104 pages)	$1.10	$1.65	$3.30

* Public libraries and colleges can purchase copies at IRE Member rates.

IRE Inventor Honored

Lee De Forest (*right*), inventor of the audion tube and a founder of the modern electronics industry, presents one of his original tubes to Reginald Hawkins, chief of the science and technology division, New York Public Library. The tube was included in an exhibition of Dr. De Forest's life and inventions on display in the library.

Student Branches Hear IRE Speakers

The IRE Syracuse Section, in addition to its regular activities during the past year, has obtained speakers from among its membership and from the engineers of the General Electric Company for several of the local IRE Student Branches. This important activity helps broaden the perspective of engineering students by bringing to them practicing engineers from the industry.

The arrangements for the speakers has been co-ordinated by A. D. Haddecke, under Chairman of the IRE Syracuse Section, D. C. Pinkerton. Speakers from the General Electric Company Electronics Park, Syracuse, N. Y., have been presented at Syracuse University, Cornell University, and Clarkson University IRE Student Branches. Among the speakers have been: I. C. Abrahams, A. R. Arsem, R. E. Hansen, J. L. Jones, P. W. Howells, L. O. Krause, C. G. Lob, and G. A. Schupp. Their subjects included: uhf antenna design, vacuum tubes, color television, missile guidance, dual frequency-all weather radar, and television studio equipment.

Professional Group News (*Cont.*)

Instrumentation

The first Transactions containing the technical papers presented by the Group at the IRE National Convention has been sent to the Instrumentation Group.

The IRE/AIEE/RTMA Symposium on Progress in Quality Electronics Components, held in Washington, D. C., May 5–7, was sponsored by the Instrumentation Group. The Group is planning to sponsor a joint IRE/AIEE Conference on High-Frequency Measurements in January, 1953, and to co-sponsor with AIEE the West Coast Symposium on Electronic Components in May, 1953. The session to be sponsored by the Group at the Western Convention is entitled, "Data Handling Systems."

Director R. L. Sink and L. E. Packard have been elected and E. P. Felch and S. C. Lawrence reelected to the Administrative Committee of the Instrumentation Group for three-year terms.

Nuclear Science

The Newsletter of the Nuclear Science Group will be continued in the near future. Plans are being made for the Group's 1952 Fall meeting to be held either at Argonne or Oak Ridge, Tenn.

Microwave Electronics

The recently formed Group on Microwave Electronics has announced the members of the Group's Administrative Committee. They are: Ben Warriner (Chairman), Andre Clavier, P. D. Coleman, D. D. King, H. B. Marvin, J. G. McCann, W. W. Mumford, G. A. Rosselot, Harald Schutz, and G. C. Southworth.

This Group will hold a Symposium this Fall, and plans have been laid to solicit members throughout the industry. Those interested in joining the Group should contact the Technical Secretary at IRE Headquarters. Attention is called to the fact that IRE Student Members are now eligible for Professional Group membership.

IRE Western Convention Sessions

The IRE Professional Groups which will sponsor sessions at the IRE West Coast Convention are as follows: Audio, Antennas and Propagation, Airborne Electronics, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Electronic Computers, Electron Devices, Information Theory, and Instrumentation.

Medical Electronics

The Administrative Committee of the new Group on Medical Electronics consists of the following members: L. H. Montgomery (Chairman), W. C. Ballard, Jr., Britton Chance, Henry Grundfest, Julia F. Herrick, and V. K. Zworykin.

Vehicular Communications

Papers are being solicited by A. A. MacDonald of Westinghouse Electric Corporation, 2519 Wilkens Ave., Baltimore, Md., for the Fall meeting of the Vehicular Communications Group.

Radio Telemetry and Remote Control

The Group will co-sponsor the Joint IRE/AIEE Telemetering Conference to be held in Los Angeles, Calif., August, 26–27, 1952. The conference will not be a part of the IRE West Coast Convention.

Quality Control

Arrangements are being made by the Group on Quality Control to list prospective speakers and to notify Section chairmen of their availability

The first Transactions of the Quality Control Group will contain the papers presented at the Toronto meeting.

AIEE/IRE Conference Proceedings Available

A "Review of Electronic Digital Computers," the proceedings of the joint AIEE-IRE Computer Conference held in Philadelphia last December, is now available in a bound volume.

The book contains the full text and discussion of papers presented at the meeting, and is expanded to include technical details which were excluded at the conference due to time limitations.

Orders for the book may be placed through R. S. Gardner, American Institute of Electrical Engineers Headquarters, 33 West 39 St., New York, N. Y. The price is $3.50.

A Tribute from Former Vice-President Jorgen Rybner

Dear Sirs:

Upon returning to normal life after having completed certain special duties, my first task is to express towards the President and the Board of Directors of The Institute of Radio Engineers my thanks for your kind letter at the completion of my term of office as Vice President of the Institute for 1951.

I beg to thank you once more for the honor bestowed on my country and on me by my election to this post, and to express the admiration of radio engineers all over the world for brilliant leadership assumed by our colleagues in the United States and manifested through The Institute of Radio Engineers.

We appreciate the generous way in which the achievements of American radio engineers are being disseminated through the Proceedings, and thus giving stimulus to workers in this field everywhere and, in the best possible way, promoting international co-operation.

With the best wishes for the future of The Institute of Radio Engineers and its officers, I am—

Very sincerely yours,
Jorgen Rybner

New IRE Sections Formed in Little Rock and Binghamton

The Board of Directors, at its March meeting, approved the establishment of IRE Sections in Little Rock, Ark., and Binghamton, N. Y. The territory of the Little Rock Section will include the state of Arkansas. The Binghamton Section was formerly a Subsection of the Syracuse Section.

These recent additions bring the total number of IRE Sections to 62.

Report of the Secretary—1951

To the Board of Directors,
The Institute of Radio Engineers

Gentlemen:

The Secretary again submits the usual Annual Report covering the operations of the Institute for the year 1951, which presents information concerning membership, fiscal status, editorial activities, technical activities, Section activities, and Student Branches.

Growth continues in every department. While the gain in new members was about 4,500, the net gain was small due to an abnormal shrinkage of Student Members. Pages of the Proceedings, both editorial and advertising, increased appreciably. Three new Sections bring the total to 60 not counting subsections (subsections total 14), 4 new student branches bring the total to 109, and Professional Groups have increased by 6 to a total of 16 embracing an aggregate membership of 13,000.

Editorial activities, of which special note should be taken, include the innovation of occasionally producing one issue devoted to one subject exclusively, the inauguration of Professional Group Transactions and the publication of tutorial papers.

General activity may be visualized by noting the number of meetings held, stated as follows, approximately:

1. Administrative	18
2. Editorial	14
3. Technical committees	135
4. Joint meetings (conferences, symposia, etc.)	37
5. Joint Technical Advisory Committee	6
6. Meetings (National Convention, etc.)	12
7. National Television System Committees of the Radio and Television Manufacturers Association	55
Total, excluding Section meetings	277
8. Sections	602
9. Subsections	69
Grand Total	948

Special mention is made of the substantial growth of the Professional Group system, the active participation of Groups in sponsoring symposia, issuing Transactions, and providing specialized papers for the Proceedings, with the likelihood of several new Groups becoming established soon.

The 1951 National Convention constitutes another example of the Institute's expanding services to the field which it serves. 22,919 persons attended to participate in 43 technical sessions with 210 papers and to see 277 exhibits, all in excess of those for the previous National Convention.

The Secretary notes, therefore, that the year just closed leaves with it a very satisfactory record. It also leaves with the management the task of continuing the search for solutions to problems inevitably associated with growth, among which is that of space for headquarters' activities, now becoming inadequate.

Respectfully submitted,

Haraden Pratt

Haraden Pratt,
Secretary

February 7, 1952.

Membership

At the end of the year 1951, the membership of the Institute, including all grades, was 29,408, an increase of 406, or 1.4 per cent over the previous year. The 406-member increase in 1951 was less than 3,332 and 2,233, the increases for 1949 and 1950, respectively. The percentage increase was 14 per cent in 1949, and 8 per cent in 1950. The net increase in membership during 1951 was greatly reduced by the 42 per cent decrease in members in the Student grade. The membership trend from 1912 to date is shown graphically in Fig. 1.

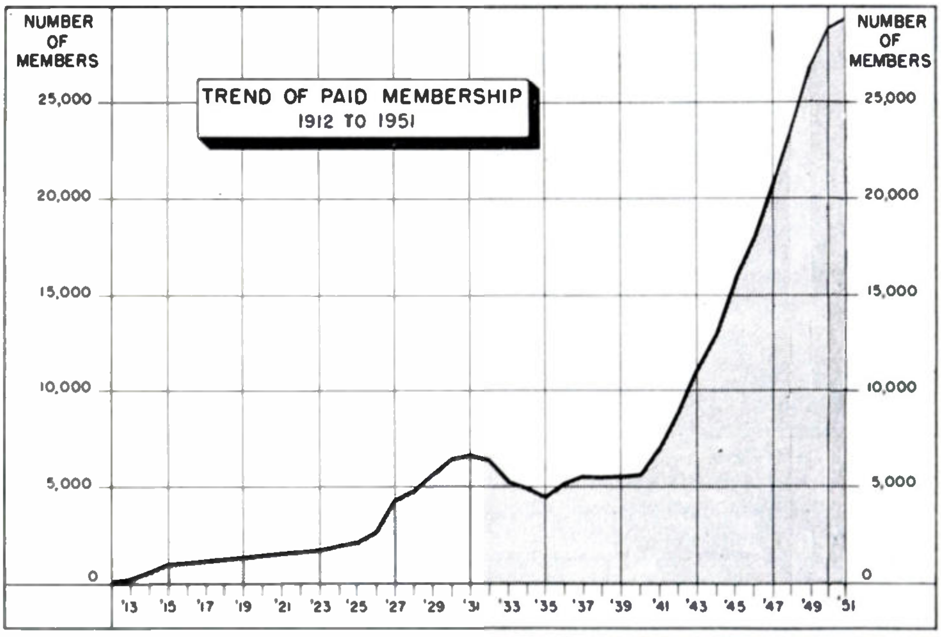

Fig. 1

Actual membership figures for 1949, 1950, and 1951 are shown in Table I. Of the 16,549 nonvoting Associates, 4,154 have been in that grade for more than five years. 3,059 Students transferred to the Associate grade in 1951, compared to 1,786 in 1950, which reflects a satisfactory trend. The membership ratio of Associates to higher grades is approximately 2 to 1.

It is with deep regret that this office records the death of the following members of the Institute during the year 1951:

Fellows

John Adrian Balch (M'17–F'40)
Francis J. Behr (A'13–M'13–F'20–L'49)
Edward Bennett (M'17–F'18)
Robert H. Marriott (A'12–M'12–F'15–L'49)
Chester W. Rice (A'16–M'26–F'28)

Senior Members

Cyrus D. Backus (A'19–M'26–SM'43)
Quincy A. Brackett (M'41–SM'43)
Vivian M. Brooker (A'27–M'35–SM'43)
Louis G. Caldwell (M'29–SM'43)
Hartman B. Canon (A'41–M'42–SM'43)
P. K. Chatterjea (M'41–SM'43)
Wayne G. Eaton (M'30–SM'43)
Joseph C. Ferguson (A'41–SM'46)
Herman T. Kohlhaas (SM'46)
Garrett W. Lewis (A'44–SM'50)
Frederick P. Morf (SM'47)
Keron C. Morrical (SM'44)
Edmund R. Morton (SM'50)
Edwin L. Powell (A'14–M'29–SM'43)
N. S. Subba Rao (SM'45)

TABLE I.—Total Membership Distribution by Grades

Grade	As of Dec. 31, 1951		As of Dec. 31, 1950		As of Dec. 31, 1949	
	Number	Per Cent of Total	Number	Per Cent of Total	Number	Per Cent of Total
Fellow	345	1.2	323	1.1	284	1.1
Senior Member	3,070	10.4	2,726	9.4	2,421	9.0
Member	4,043	13.7	3,803	13.1	3,559	13.3
Associate	17,523***	59.6	14,590**	50.3	12,309*	46.0
Student	4,427	15.1	7,560	26.1	8,196	30.6
Totals	29,408		29,002		26,769	

* Includes 1,260 Voting Associates. ** Includes 1,159 Voting Associates. *** Includes 974 Voting Associates.

Members

Arthur A. Bertram (A'44–M'49)
Alfred J. Bremmer (M'46)
Harold M. Brett (M'44)
Theodore A. Brindle (M'50)
J. W. Fleming (A'39, M'49)
Henry J. Oosterling (M'45)
Philip M. Raycer (A'48–M'48)
John E. Shaw (M'48)
Emerson W. Thomas (M'48)
Albert J. Woodcock (A'41–M'48)

Voting Associates

Ralph E. Adams (A'32–VA'39)
Jerome Burbank (A'30–VA'39)
Arvid P. Sunnergren (A'28–VA'39)

Associates

Andrew Aguele (A'46)
John C. Bell (A'41)
Richard N. Chambers (A'49)
Jessie Raymond Collins (A'40)
Arthur P. Copson (S'48–A'50)
James C. Drumm (A'51)
Waldo D. Elmer (A'44)
Oscar Hachoogan (A'45)
Peter L. Harbury (S'43–A'45)
Donald M. Harring (A'45)
Henry S. Kaminski (A'50)
Thomas R. Knudsen (A'50)
Henry J. Lawton (A'51)
Frederic F. Lee (A'50)
Alfred R. Leek (S'45–A'47)
Ronald W. S. Marsano (A'44)
Eli Ossofsky (S'43–A'46)
A. V. Platter (A'49)
Charles H. Rehnke (A'49)
Lloyd D. Rodgers, Jr. (A'50)
Colin G. Ross, Jr. (A'41)
Alvin Sachs (A'45)
Roger M. Sanderford, Jr. (S'42–A'44)
Ralph W. Smith, Jr. (A'49)
Ernest Uribe (A'50)
Oliver D. Westerberg (A'45)
John L. Wiley (A'46)

Students

Allan Briggs (S'48)
Harvey B. Glasser (S'48)
Thomas J. Ricotta (S'50)

Fiscal

A condensed summary of income and expenses for 1951 is shown in Table II, and a balance sheet for 1951 is shown in Table III.

TABLE II.—SUMMARY OF INCOME AND EXPENSES, 1951

Income		
Advertising	$345,418.45	
Member Dues and Convention	495,850.95	
Subscriptions	51,695.11	
Sales Items, Binders, Emblems, etc.	20,463.13	
Inv stments Income	13,260.88	
Miscellaneous Income	683.36	
TOTAL INCOME		$927,371.88
Expenses		
PROCEEDINGS Editorial Pages	$190,257.31	
Advertising Pages	188,025.57	
Directory	67,240.06	
Section and Student Branch Rebates	31,643.94	
Sales Items	11,006.97	
Miscellaneous Printing	2,185.37	
General Operations	204,784.16	
Convention Cost	127,080.62	
TOTAL EXPENSES		$822,224.00
Surplus		$105,147.88
Reserve for Depreciation		6,925.89
NET SURPLUS		$ 98,221 99

TABLE III.—BALANCE SHEET—DECEMBER 31, 1951

Assets		
Cash and Accounts Receivable	$276,312.09	
Inventory	13,352.53	
TOTAL CURRENT ASSETS		$ 289,664 62
Investments at Cost	601,013.42	
Building and Land at Cost	356,687.30	
Furniture and Fixtures at Cost	85,803.60	
Other Assets	19,139.20	
TOTAL		1,062,643.52
TOTAL ASSETS		$1,352,308.14
Liabilities and Surplus		
Accounts Payable	$ 18,718.46	
Federal Taxes on Emblems, etc.	91.76	
Total Current Liabilities		$ 18,810.22
*Deferred Liabilities		315,896.42
Total Liabilities		334,706 64
Reserve for Depreciation		31,348.54
Surplus—Donated	595,406.43	
Surplus—Earned	390,846.53	
TOTAL SURPLUS		986,252.96
TOTAL LIABILITIES AND SURPLUS		$1,352,308.14

* 1952 Items, PROCEEDINGS for members and subsc ibers, Advertising, and Convention Service.

TABLE IV.—VOLUME OF PROCEEDINGS PAGES.

	1951	*1950*	*1949*	*1948*
Editorial	1,628	1,516	1,532	1,592
Advertising	1,424	1,016	872	860
Total	3,052	2,532	2,404	2,452

Editorial Department

During the year 1951 there were published in the PROCEEDINGS OF THE I.R.E. a total of 3,052 pages, including covers; a marked increase over previous years. Of these, 1,628 were editorial pages and 1,424 were advertising pages. The number of PROCEEDINGS pages published in each of the last four years is given in Table IV. The number of editorial pages published each year since 1913 is shown in Fig. 2.

Technical papers totaling 173 were published in 1951, as against 186 in 1950. In addition, 8 IRE Standards were published in 1951 as against 6 in 1950. Authorship of these papers was by 239 individuals of whom 175, or 73 per cent, were members of the Institute. In 1950, 182 of the 252 authors, or 73 per cent, were members.

The volume of papers submitted for publication continues at a high rate. During 1951, 241 papers totaling an estimated 1,680 PROCEEDINGS pages were submitted, or an average of 20 papers and 140 pages per month. During 1950, 271 papers of 1,547 pages were received, or 23 papers and 129 pages per month.

The backlog of papers on hand in the Editorial Department at the end of 1951 consisted of 153 papers totaling 704 PROCEEDINGS pages, of which 67 papers or 303 pages had been accepted for publication, the remainder being under review. This represented no significant change over the backlog total at the end of 1950 of 141 papers or 777 pages.

The year 1951 was marked by four important innovations in publication policies which will substantially increase the effectiveness with which the publications of the Institute fulfill the technical information needs of the membership.

(1) The IRE Professional Groups, with the assistance of the Editorial Department, inaugurated a quick, inexpensive method of providing group members with technical papers in their particular fields of interest by issuing their own technical publications, called TRANSACTIONS, to their respective memberships. The Editorial Department published 6 TRANSACTIONS for the Audio Group and 2 for the Airborne Electronics Group.

(2) The publication in October of the special color-television issue of PROCEEDINGS marked the first issue of PROCEEDINGS devoted to one subject. The 400-page issue was the largest issue of PROCEEDINGS ever published.

(3) Manuscripts of lengthy descriptive or mathematical papers, which cannot be published in the PROCEEDINGS in full-length form due to space considerations, will be deposited (with the author's consent) with the American Documentation Institute in Washington, D. C., where microfilm or photocopies may be purchased by interested readers. Abstracts of these papers will be published in the PROCEEDINGS accompanied in each case by a footnote advising readers of the availability and costs of copies of the full-length papers.

(4) Publication of a series of valuable and outstanding tutorial papers on a wide variety of topics of both present and historical interest was begun with the appearance of seven such papers during 1951, prepared by authorities in the field. The tutorial papers series is sponsored by the Tutorial Papers Subcommittee of the IRE Committee on Education.

The Editorial Department continued to co-operate closely with IRE Professional Groups on the publication of technical papers, abstracts of papers delivered at Group-sponsored conferences, and news items of interest to Group members in order to give this important Institute activity full expression in the pages of the PROCEEDINGS. During the year, nine papers were published in the PROCEEDINGS which had been referred to and recommended by Professional Groups for publication.

The 1951 IRE Directory was the largest Directory published to date, containing 764 pages including covers, of which 418 were membership listings and information, 346 were advertisements and listings of manufacturers and products. The Directory was issued in October, 1951.

The Editorial Department is directed by Editor Alfred N. Goldsmith in matters of editorial policy, content, and format, and by Executive Secretary George W. Bailey in matters of finance and administration. Both function through Technical Editor E. K. Gannett and an effective staff. It has been greatly assisted by the counsel and co-operation unstintingly given by the members of the Board of Editors, Papers Review Comnittee, and the Administrative Committee of the Board of Editors.

Standards on Transducers: Definitions of Terms, 1951.

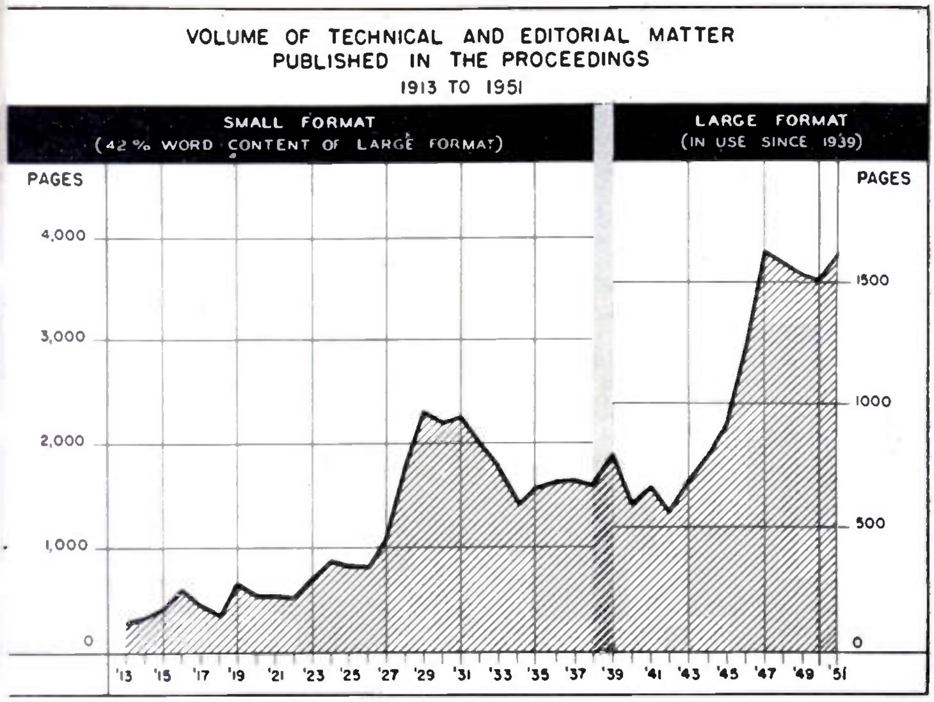

Fig. 2

Technical Activities

[*Technical Committees.*] During 1951, 23 IRE Technical Committees, with their subcommittees, task groups, and planning committees for conferences or symposia, held 114 meetings at the Institute Headquarters, one meeting at the Waldorf-Astoria Hotel during the National Convention, and twenty meetings at other points, a grand total of 135 meetings.

A report on the activities of the IRE Technical Committees and their subcommittees is compiled by the office of the Technical Secretary and published each month in the PROCEEDINGS.

The policy of publishing Standards in the PROCEEDINGS was adopted in late 1949. Issues of the 1951 PROCEEDINGS containing Standards are identified by a red band on the cover and spine. This method of publication is still being carried on and reprints of Standards appearing in the PROCEEDINGS may be purchased from headquarters for a nominal sum. The following eight Standards were published during 1951:

Standards on Circuits: Definitions of Terms in Network Topology, 1950.
Standards on Television: Methods of Measurement of Electronically Regulated Power Supplies, 1950.
Standards on Electronic Computers: Definitions of Terms, 1950.
Standards on Abbreviations of Radio-Electronic Terms, 1951.
Standards on Electroacoustics: Definitions of Terms, 1951.
Standards on Pulses: Definitions of Terms, Part I, 1951.
Standards on Radio Receivers: Open Field Method of Measurement of Spurious Radiation from Frequency Modulation and Television Broadcast Receivers, 1951.

The following Standards, prepared by IRE Technical Committees, were approved by the Standards Committee in 1951 and will be published in forthcoming issues of the PROCEEDINGS:

Standards on Gas-Filled Counter Tubes: Definitions of Terms, 1952.
Standards on Electron Tubes: Methods of Testing Gas-Filled Radiation Counter Tubes, 1952.

The Annual Review Committee has prepared its survey, the Radio Progress Report During 1951, which appeared in the April, 1952 issue of the PROCEEDINGS.

The Master Index of Terms which was prepared in 1949 was revised in February, 1951 by the Office of the Technical Secretary and made available to all IRE committees and, upon request, to other societies.

Approval was secured in April, 1951 for the formation of a new Technical Committee on Servo-Systems under the Chairmanship of Professor W. M. Pease of the Massachusetts Institute of Technology.

The Executive Committee authorized the change in the name of the Modulation Systems Committee to the Committee on Information Theory and Modulation Systems. The name of the Electron Tubes and Solid State Devices Committee was changed to the Committee on Electron Devices. The Committee on Mobile Communications was renamed the Committee on Mobile Communications Systems.

Joint Committees. Twelve meetings of committees jointly sponsored by IRE and other outside organizations were held at IRE headquarters. Secretarial service was provided for several of these meetings. During the year 37 conferences, symposia, technical sessions, and national meetings were sponsored by IRE Professional Groups, by technical committees, or co-sponsored by IRE and other organizations. The major portion of the work involved was the responsibility of the Office of the Technical Secretary.

The Second Joint IRE/AIEE/NBS High Frequency Measurements Symposium, sponsored by the Joint IRE/AIEE Committee on High Frequency Measurements was held in Washington, D. C. on January 10–12, 1951. The conference attendance of 590 was most enthusiastic. It was held as a part of the celebration of the semicentennial of the National Bureau of Standards. The IRE Professional Group on Instrumentation also co-operated in the organization of the conference.

The IRE participated for the first time in the program of the National Convention of the Institute of Aeronautical Sciences in January, which was held at the Hotel Astor. The IRE co-operated with the Institute of Aeronautical Sciences in sponsoring a full day's technical session devoted to Electronics in Aviation.

At the request of the Department of State, the appointment of A. G. Jensen was approved as Chairman of the new United States committee to handle questions involved in the work assigned to Study Group XIV of the International Radio Consultative Committee. Mr. Jensen's nomination as a representative for the Executive Committee of the United States delegation of the International Radio Consultative Committee was also approved by IRE.

The National Association of Radio and Television Broadcasters during 1951 accepted the invitation to join the IRE, RTMA, and SMPTE on the Steering Committee which will be known as the Joint Committee for Inter-Society Co-ordination (JCIC). This committee was formerly known as the IRS Steering Committee, established to avoid duplication of effort among the technical committees of IRE, RTMA, and SMPTE working in fields associated with television.

Professional Group System. There are sixteen Professional Groups presently active in the following fields: Airborne Electronics, Antennas and Propagation, Audio, Broadcast and Television Receivers, Broadcast and Transmission Systems, Circuit Theory, Electronic Computers, Electron Devices, Engineering Management, Information Theory, Industrial Electronics, Instrumentation, Nuclear Science, Quality Control, Radio Telemetry and Remote Control, and Vehicular Communications. The total membership is in excess of 13,000 and is increasing steadily. It is anticipated that pending petitions to form five new IRE Professional Groups in the fields of Medical Electronics, Biological Electronics, Radio Communications, Microwave Electronics, and Basic Sciences will be approved by the Executive Committee. The Growth of the Professional Group system is evidenced by the fact that at the 1951 National Convention seven groups sponsored eleven symposia. Each Group nominated a representative to serve on the Technical Program Committee for the Convention.

IRE Sections are constantly informed of the activities of the Groups through the media of newsletters, conference notices, and minutes from the office of the Technical Secretary. Steps are being taken in the various Sections towards the stimulation of

Group activities. As a result of several Group membership drives, many new members have been enrolled in IRE. During the year, a number of Groups sponsored joint symposia, national meetings, and technical sessions with other professional societies. All Groups are charged with the procurement of papers for publication in the PROCEEDINGS. The Professional Group Manual was again revised in 1951.

The Professional Group on Nuclear Science sponsored a very successful Annual Symposium at the Brookhaven National Laboratory on December 3–4. It was well attended and enthusiastically received.

The Professional Group on Electronic Computers co-operated with the IRE Technical Committee on Computers to co-sponsor with AIEE an outstandingly successful Conference on Electronic Computers, in Philadelphia, December 10–12. A total of 872 persons attended the conference.

Joint Technical Advisory Committee. Six meetings of JTAC were held during 1951. Volume VI, *Proceedings of the JTAC*, was published. This volume included: Section I, Official Correspondence Between the Federal Communications Commission and the Joint Technical Advisory Committee (IRE-RTMA) with Other Items Pertinent to the Action of the JTAC; and Section II, Approved Minutes of Meetings of the Joint Technical Advisory Committee (IRE-RTMA), both covering the period July 1, 1949–June 30, 1950. Volume VII of the *Proceedings* for the subsequent year ending June 30, 1951 was also authorized and prepared. It will be distributed in January, 1952. The preparation of a paper on the "Conservation of the Radio Spectrum," was undertaken and will appear as JTAC Volume VIII. The membership for the year July 1, 1951 to June 30, 1952 was appointed. I. J. Kaar replaced John V. L. Hogan as Chairman and Ralph Bown was elected unanimously as Vice Chairman. On July 31, 1951, Haraden Pratt resigned from membership on JTAC to become Telecommunications Advisor to the President of the United States. A. V. Loughren was elected unanimously to replace Mr. Pratt.

A Subcommittee on Land-Mobile Channel Allocations was appointed on October 9, 1951, to act as consultant to the JTAC in the preparation of a reply to the FCC regarding its request for information on the subject of channel allocations for communication services. One meeting of this subcommittee was held in 1951, at which time the problem was outlined and the assignment made.

NTSC. In addition to IRE committee meetings, the National Television System Committee held 55 meetings at headquarters. The Technical Secretary acts as Secretary for Panel 18, which is the co-ordinating panel for the National Television System Committee. By request, the office of the Technical Secretary provided secretarial assistance several times for other of the NTSC panels.

Section Activities

We were glad to welcome three new Sections into the Institute during the past year. They are as follows:

Oklahoma City (July), 1951
Phoenix (January), 1951
Tulsa (May), 1951

The total number of Sections is now 60. There has been an increase in the membership of 25 Sections.

The Subsections of Sections now total 14, the following being formed in 1951:

Palo Alto (San Francisco Section), January, 1951
Rome (Syracuse Section), May, 1951

Student Branches

The number of Student Branches formed during 1951 was 4, all of which operate as Joint IRE-AIEE Branches. The total number of Student Branches is now 109, 66 of which operate as Joint IRE-AIEE Branches.

Following is a list of the Student Branches formed during 1951: Johns Hopkins University, University of Kansas, Montana State College, and Villanova College.

IRE People

Herbert S. Bennett (A'42–M'43–SM'44) has been appointed as director of research and engineering of the engineering and production division of Dynamic Electronics-New York, Inc., it was recently announced.

H. S. BENNETT

Dr. Bennett was born on June 13, 1917, in New York, N. Y. He received the B.E.E. degree and the M.S.E.E. degree from the College of the City of New York, in 1938, and 1939, respectively. He received the M.S. degree in physics in 1947, and the D.Sc. degree in electrical engineering in 1951, both from the Polytechnic Institute of Brooklyn.

Dr. Bennett was a radio engineer for the United States Signal Corps from 1939–1942, and was then commissioned to serve as a technical officer with the Signal Corps, and later, with the United States Air Force until 1946. At present, he holds a reserve rank of lieutenant colonel. After returning to civilian employment in 1946 as a research and development engineer with the United States Air Force at Watson Laboratories, he then became chief of the planning and analysis branch of these laboratories. From 1950–1952, he was chief of the engineering branch at the Electronic Warfare Center, Fort Monmouth, N. J.

Dr. Bennett is a member of the IRE Admissions Committee and was active on the Technical Program Committee for the 1947 IRE National Convention. He is a member of Sigma Xi, the Scientific Research Association of America, the National Society of Professional Engineers, the American Institute of Electrical Engineering, Tau Beta Pi, and Eta Kappa Nu. He is the author of papers and reports on researches in electromagnetic diffraction problems.

❖

Joseph Racker (A'51) has been named president of the recently expanded Joseph Racker Company, a firm of radar consultants and editors.

JOSEPH RACKER

Mr. Racker was born on October 30, 1921, in New York, N. Y. He received the B.S. degree in physics from Brooklyn College in 1942, and the M.S.E.E. degree from Brooklyn Polytechnic Institute in 1948. During World War II, Mr. Racker served with the United States Army as a radar officer, and has had more than six years of experience with International Telephone and Telegraph Corporation system companies. He is the co-author of the text "Pulse Techniques," and author of "Fundamentals of Microwave Communications."

Herbert A. Elion (A'52) has been appointed to the research staff of Paul Rosenberg Associates, consulting physicists. Formerly, he was a projects engineer at Freed Radio Corporation and the M. W. Kellogg Company.

HERBERT A. ELION

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WASHINGTON, D. C.—APRIL 21-24, 1952

SUMMARIES OF TECHNICAL PAPERS

1. MODERN CONCEPTS IN AMPLIFIER THEORY.—J. M. Pettit, *Stanford University*.—This paper will present a resume of several concepts in the theory of vacuum-tube amplifiers which have led not only to better understanding of this class of electronic devices but to substantial advances in their capabilities. The discussion will be limited to the so-called "wide-band" amplifier, in both its low-pass ("video") and band-pass forms, using conventional tubes and lumped-element networks.

For the low-pass case, the principal applications require excellence of transient response, rather than any specific variation of amplification with respect to frequency. Direct evaluation of the transient response becomes practicable through the Laplace transformation; from it can be derived the basic laws of multistage amplifiers, together with a moderately satisfactory understanding of the circuitous relationships between transient and steady-state response.

For the band-pass case, the applications usually involve the rejection of unwanted signals in adjacent channels, but may in addition call for excellent transient response within the desired pass band. Achieving a compromise on these requirements, with the greatest efficiency of tubes and circuits, calls for a clear understanding of the problem and good analysis techniques. Recent progress in network theory has provided tools like the electrostatic potential analogy, which yields, among other benefits, a beautiful visualization of the physical behavior of wide-band amplifiers.

2. UPPER ATMOSPHERE PHYSICAL CHARACTERISTICS.[1]—M. Nicolet, *The Pennsylvania State College* (On leave from Royal Meteorological Institute of Belgium, Brussels, Belgium).—Recent physical data concerning the constitution of the upper atmosphere and solar emissions are used in a discussion regarding some properties of the ionospheric regions.

Numerical values are given for the electron collision frequency as a function of height. The transition region of dissociation of molecular oxygen is considered and the dissociation of molecular nitrogen is studied. Diffusive separation is shown to contribute to the vertical distribution of the atmospheric constituents. The penetration of relevant solar radiation is discussed with respect to the formation of ionospheric layers.

3. ANTIFRICTION BEARINGS AS RADIO NOISE GENERATORS.—Harold E. Dinger and J. E. Raudenbush, *Naval Research Laboratory*.—During World War II, an unfamiliar kind of radio noise was noticed that proved to be a source of irritation to those engaged in interference reduction. It was found that certain individual machines employing anti-friction bearings, would, at times, produce a wide-band radio noise that failed to respond to the usual treatments. At first, it was thought to be a static phenomenon and was often referred to as "bearing static" and some cases proved to be such. However, many cases were found in which a different mechanism of noise generation was apparently involved. For example, in one motor-generator unit employing three bearings and which created a strong source of noise, even when all commutation brushes were removed, a grounding brush was applied to the rotating shaft without effect. When a separate brush was applied near each bearing, the noise level dropped; the reduction in level being roughly proportional to the contact area and the brush pressure. A limited investigation of this phenomenon has recently been conducted by employing a special test assembly wherein shaft-to-ground voltages and currents can be studied under controlled conditions. Investigation has shown that bearings interrupt a circuit across which stray potentials usually exist and that the effect varies greatly with different lubricants. In general, it is proportional to the resistance across the bearing which may vary from a fraction of an ohm to several megohms. Several results have been observed that may have application in other fields of investigation.

4. REGULARITIES IN THE BEHAVIOR OF REGIONS E AND F OF THE IONOSPHERE.—J. W. Findlay, *Carnegie Institution of Washington* (On leave from Cambridge University, Cambridge, England).—The paper describes recent work by Ratcliffe on the regularities in the $F2$ region of the ionosphere, the results of which are in process of publication in the *Journal of Geophysical Research*, and also a study of the phase path of echoes from Region E.

The regularities in the behavior of the $F2$ layer are shown by using a simple method for analyzing routine (h', f) records which is similar to the graphical method of Booker and Seaton. Records from Watheroo, Huancayo, and College have been analyzed to determine the total number, (n), of electrons below the level of maximum electron density in a column of unit cross section in the $F2$ region. This analysis shows that n is closely related to the sun's zenith angle z whereas the maximum electron density N_m is not simply related to z. The well-known anomalies when N_m is studied as a function of time of day, time of year and geographical position all seem to disappear when n is studied instead of N_m. An anomaly observed at Huancayo is discussed and a relation between thickness and height of the $F2$ layer which may be of use in ionospheric forecasting is described.

The results of three years observations of the phase of echoes of about 2 mc in frequency from region E are described. These show that the phase height of the region varies qualitatively, but not quantitatively, as would be expected for a simple Chapman region. The effects of fadeouts in increasing the absorption and also the phase heights for echoes from the region are described and measurements are given of the velocities of movement of clouds of ionization in the lower part of the E layer.

5. NORMAL TROPOSPHERIC PROPAGATION DEEP INTO THE EARTH'S SHADOW: THE PRESENT STATUS OF SUGGESTED EXPLANATIONS.—Thomas J. Carroll, *Massachusetts Institute of Technology*.—During the last seven years, experimental observations of field strength measured well beyond the horizon from high-power transmitters through the vhf and microwave frequency range have shown much higher fields than conventional theory had predicted in the absence of ducts. The three types of suggested explanations have been (a) scattering by omnipresent turbulence in the atmosphere; (b) wave scattering connected with earth roughness; (c) partial internal reflections from the normal stratification of the troposphere itself. No definitive judgment can yet be made concerning the relative validity of these explanations, but the physical ideas underlying each hypothesis and the claims of their protagonists as of the end of 1951 will be summarized and checked against supporting evidence in the hope of setting the stage for the conflicts of views expected to develop later in the meeting.

Independently of the explanation which ultimately wins acceptance, it seems generally agreed that these fields will have practical importance for reliable point-to-point communication at ranges larger than formerly envisaged at vhf and microwave frequencies.

6. A MICROCALORIMETER FOR THE MEASUREMENT OF ABSOLUTE MICROWAVE POWER.—Alan C. Macpherson, *National Bureau of Standards*.—A matched termination is heated with the rf power to be measured. The temperature rise of the termination is measured and the calorimeter calibrated with known dc power. The present model operates at X-band at power levels from approximately 1 to 15 mw, but the method is applicable to other frequencies in waveguide and coaxial line. The sensitivity is approximately 10 μm. To reduce the effect of ambient temperature changes the Joule twin calorimeter system is employed. The two identical reflectionless terminations each consist of shorted sections of waveguide in which a conventional barretter

[1] The research reported in this paper has been supported by the Geophysical Research Division of the Air Force Cambridge Research Center under Contract AF19(122)-44.

* Co-sponsored by the IRE Professional Group on Antennas and Propagation and the USA National Committee of URSI.

is mounted. One termination is used only as a reference point from which the temperature rise of the other is measured. A 20-junction, copper-to-constantan thermopile is connected between the two terminations. Both the rf power and the dc calibrating power are dissipated in the barretter. The fundamental question of the accuracy of a dc-substitution calorimeter concerns the equivalence of the dc power and the unknown power. This was investigated both experimentally and theoretically and the estimated accuracy of the present calorimeter is 1 per cent.

7. NBS MAGNETIC ATTENUATOR.—Frank Reggia, *National Bureau of Standards.*—The need for an electrically-controlled variable attenuator for microwave frequencies has brought about the development of an inexpensive type of magnetic attenuator for coaxial transmission lines. Its operation depends upon the interaction between the electromagnetic field within the coaxial line, which contains a magnetic ferrite as its rf energy-dissipating material, and an external magnetic field applied perpendicularly to the axis of the line. As a result of this interaction, instantaneous changes in attenuation are obtained.

The new device is simple in construction. It requires no movable parts, mechanical controls, or slotted sections which heretofore have characterized conventional variable attenuators used at microwave frequencies. Because the external magnetic field is produced by an electromagnet, it can be operated either manually or automatically from a proximate or remote position. Small control currents in the field coils are changed simply and precisely with a multiturn Helipot potentiometer.

The magnetic attenuator has been used successfully to amplitude modulate the rf output of microwave oscillators with sine-wave voltages of frequencies from dc to above 10,000 cps. It has also been used as a control device in a degenerative feedback circuit to automatically stabilize the output of microwave generators and as a microwave switch. Further developments are now in progress.

8. THE MEASUREMENT OF *Q* OF RESONANT CAVITIES IN THE NORMAL AND SUPERCONDUCTING STATE.—C. J. Grebenkemper and J. P. Hagen, *Naval Research Laboratory.*—In the investigation of the phenomena of superconductivity at microwave frequencies the usual procedure followed is to construct a resonant cavity of the metal under study and measure its electric Q factor. Since the loss in the cavity itself is extremely small the Q factor will be very large. It is desirable to measure the loss which is proportional to the resistance with as much precision as possible. Conventional techniques such as the resonance-curve method or standing-wave measurements become less reliable for very high values of Q due to oscillator instability.

In our measurements a decrement method is used. This type measurement is very rapid and versatile. It can be used to cover a large range of Q values ranging from 10^4 to 10^7 with a good deal of precision. A recurrent pulse of RF energy of about 1 microsecond duration is used to excite the resonator. The decay of energy in the resonator is observed on an oscilloscope after the oscillator is shut off. Time is measured either by a precision delay circuit or by timing markers generated by a crystal-controlled oscillator. The receiver is a conventional superheterodyne. The signal source is a reflex klystron oscillator.

Measurements of this type have been made in the 10 kmc and 24 kmc regions. Some of the general results will be discussed.

9. AN ELECTRONIC RATIO METER FOR REFLECTION COEFFICIENT MEASUREMENT.—L. A. Rosenthal, J. L. Potter, and G. M. Badoyannis, *Rutgers University.*—A vacuum-tube ratio meter has been developed which, when integrated with two directional couplers, will allow for the direct measurement of reflection coefficients. The ratio meter consists of two electronic voltmeters working into a conventional ratio meter movement. The meter deflection depends only on the ratio of the average full rectified values of the input signals. The frequency response of this instrument is limited to below 200 kc, and it is therefore made to respond to the modulation of the RF signal. In the absence of modulation a converter accessory is provided. In operation two signals are taken from a reflectometer, that has been inserted in the transmission system, and applied to the ratio meter. The reflection coefficient can be measured continuously as adjustments are made in the line or at the termination.

The principles of design of the ratio meter, and the theory of application are presented together with some experimental observations. This work has been sponsored by the Rome Air Development Center under contract AF28(099)-33.

10. A MICROWAVE POWER COMPARATOR.—K. C. C. Gunn and K. O. Holmes, *Air Force Cambridge Research Center.*—In some cases it is necessary to equalize RF powers in two or more circuits to a high degree of accuracy. In order to determine when equality has been reached, an instrument has been developed which is capable of detecting a change of 0.02 db in power at *S* band.

This instrument consists of a hybrid junction, rotating eccentric attenuators phased 180 degrees apart in the two input arms, a matched load on one output arm, and a crystal detector on the other output arm. The maximum attenuation of the attenuators is equalized by conventional means and on the other half of the cycle, they are removed from the waveguide. Effectively they shut off the power in the input arms alternately. The amplitude of the varying voltage at the detector is determined by the input powers, the position of the attenuator wheels, and the phase relation between the two input powers. Successive voltage peaks are observed on an oscilloscope.

In the original model a variation of input power of 0.02 db produced an observable deflection on the oscilloscope. The power comparator has high sensitivity at low-power levels, compares powers almost instantaneously, and compares powers in the same transmission line using the same detector at the same power level.

11. SHORT-PERIOD SKY-WAVE FADING OF CW EMISSIONS.—H. P. Hutchinson, *Department of the Army.*—A study of the short-period sky-wave fading of cw emissions shows that observed characteristics of these fading distributions are functions of the relative amplitudes of the received "*o*" and "*x*" magneto-ionic components. When the received magneto-ionic components are essentially equal at the receiving point, severe and rapid fading occurs and variations in short-period fading are maximum. However, when a single mode of propagation is suppressed, the variations in short-period fading distributions of the received field are reduced and become log-normal in character.

A new parameter is suggested to replace the Rayleigh distribution in cases where a more accurate formulation of short-period fading characteristics is desired.

Finally, a description is given of a practical method of emitting cw radio transmissions, which yields reflections from the ionosphere of a single ionospheric mode, thereby obtaining some man-made positive control over short-period sky-wave fading.

12. THE LIMITING POLARIZATION OF MAGNETO-IONIC WAVES.—J. Feinstein, *National Bureau of Standards.*—The failure of ray theory in the transition region between the ionosphere and free space is reviewed. A first-order wave theory which takes account of the mode interaction which must exist in the transition region between isotropic and anisotropic media is developed. and compared with Booker's criterion for the position at which limiting polarization is assumed. As is well known, for an exponential decay of electron density, the usual situation encountered in practice, this criterion becomes indeterminate; the theory is applied to determine the limiting polarizations for this case.

The wave treatment is also applied to ascertain the effect of a lower layer in altering the characteristic polarization of a downcoming wave reflected from an upper layer, as a function of the ray polarizations and electron-density distribution of the lower layer. The results of this study are applied to low-frequency waves reflected from the *E*-layer and passing through a *D*-region, with a view toward indicating the magnitude of any possible effect.

13. CHARACTERISTIC WAVES.—A. J. Mallinckrodt, W. Snyder, R. A. Helliwell, *Stanford University.*—The familiar quasi-homogeneous (ray) treatment of the ionosphere shows that in a slowly-varying ionosphere there are two and only two possible types of independent propagating waves, the magneto-ionic components. In the case of normal incidence these components have the well-recognized property that the polarization of a reflected wave of one type is identical to that of an incident wave of the same type. However, it does not seem to have been appreciated that generally for linear anisotropic reflecting media of any degree of inhomogeneity there exist two and only two characteristic polarizations having the latter property.

It is shown that the characteristic polarizations can be determined from three experiments in which three different known polarizations are transmitted and the resulting reflected polarizations are measured. The information thus obtained is also sufficient to determine the ratio of the reflection coefficients of the two characteristic waves. In the special case of the homogeneous ionosphere it is shown that the number of experiments of the above type can be reduced to two.

14. POLARIZATION CONTROL AND MEASUREMENT IN IONOSPHERE VERTICAL-INCIDENCE ECHO RANGING.—M. G. Morgan, *Dartmouth College.*—Two vertical half-rhombic, or delta, antennas are placed at right angles to each other on a single supporting structure. These are excited from a dual-channel, pulsed transmitter. The relative phase and amplitude of the synchronous signals applied to the two antennas are independently adjustable. In this way, any possible elliptic polarization can be imparted to the transmitted wave. Ionospheric echoes are received on the same antennas. The voltages developed at the antenna terminals are amplified with identical gain and phase shift in a dual-channel receiver. The polarization loci of the echoes received are displayed on a cathode-ray tube.

Measurements have been made at Hanover, New Hampshire during the winter months of 1951–1952, in the hours between sunrise and noon, and at frequencies near F_2-layer criticals. When the magneto-ionically split components of the echoes can be resolved on a time basis, the polarization is found to be independent of the transmitted polarization and to have sense defined as: ordinary, north-to-west; extraordinary, north-to-east. Axial ratios of the elliptic polarization loci are about 0.5 and decrease with increasing frequency. The major axes usually lie along the northwest and northeast direction, either direction occurring at different times for either magneto-ionic component. It is found that large attenuation of either component can be obtained by adjustment of the transmitted polarization. Particular sensitivity to relative phase of the transmitter channels is noted and insensitivity to relative amplitude. Enhancement of the unattenuated echo occurs, with maximum enhancement of the one corresponding to maximum attenuation of the other.

15. PLANE WAVES IN THE IONOSPHERE.—H. B. Keller, *New York University.*—For plane-waves incident on a plane ionosphere with electron density and collisional frequency arbitrary functions of height, Maxwell's equations are reduced to a system of four linear first-order differential equations. This system of "generalized transmission-line" equations is then solved by an extension of the Peano-Baker matrizant method. The solution is a uniformly convergent series which is interpreted as four "waves" which undergo continuous multiple reflections of all orders, each wave contributing to all of the others. This representation fails at points (or intervals) called "reflection levels," and a different representation of the solution is obtained for such intervals. This exhibits the existence of only three waves at a reflection level, two of the previous waves having become strongly coupled and identical.

The ionospheric properties are considered explicitly with a view to determining when the previous analysis is most simplified. Various cases are examined and the quantities necessary to compute the solution are obtained for vertical incidence.

16. CONTROL OF ANNULAR SLOT EXCITATION BY SELECTIVE DIELECTRIC FILLING.*—D. J. Angelakos and R. W. Bickmore, *University of California.*—In general, the principal H-plane radiation pattern of an arbitrarily excited annular slot is directly related to the form of the excitation function in the slot.† This paper describes a technique of modifying a coaxially excited annular slot (which when uniformly excited has an omnidirectional H-plane pattern) by means of selective dielectric filling so as to produce a cardioid H-plane pattern. Equations are derived which relate the angular distribution, length and dielectric constant of the filling to the slot conductance and characteristic impedance. The assumption is made that any net slot reactance has been compensated directly at the slot, this being the first requisite of a low vswr.

An antenna built on these principles is described and experimentally determined E- and H-plane radiation patterns and impedances are presented.

Although the paper is primarily concerned with the production of a cardioid pattern, the method described is sufficiently general to be applicable to other pattern configurations which may be desired from coaxially fed annular slot radiators.

* This work sponsored by the Office of Naval Research and the Bureau of Ships under contract N7-onr29529.

† R. W. Bickmore, Report No. 168 Antenna Laboratory, University of California; June 15, 1950.

17. ANTENNA PATTERN CALCULATION FOR ASYMMETRICAL APERTURE DISTRIBUTIONS.—C. C. Allen, *General Engineering Laboratory.*—A method is presented for calculating the radiation patterns and gain of antennas having asymmetrical phase and amplitude distributions. This method, which determines the radiation pattern from the actual distribution across the aperture, makes use of automatic punch card machines to perform a numerical integration. The patterns obtained for several distributions are discussed for amplitude and phase asymmetry considered both separately and in combination. Results are given for typical distributions obtained with asymmetrical parabolic cylinders. Applications to scanning antenna analysis and the determination of the phase of the radiated pattern are presented.

18. THEORY OF WAVEGUIDE-FED SLOTS RADIATING INTO PARALLEL-PLATE REGIONS.—H. Gruenberg, *National Research Council, Canada.*—Slotted waveguide arrays feeding into parallel-plate regions have been used in some high-speed scanners. Parallel-plate regions also have been used for the suppression of second-order beams of high-gain arrays.

A theoretical expression is derived for the conductance of a longitudinal shunt-slot in a rectangular guide when the slot is radiating into a parallel-plate region of arbitrary plate spacing. Some peculiarities of the theoretical results are discussed. There is good agreement between theory and experiment.

19. FACTOR OF MERIT FOR AIRCRAFT ANTENNA SYSTEMS FOR THE FREQUENCY RANGE FROM 3–30 MC.*—Ernest J. Moore, *Stanford Research Institute.*—Lucke has proposed a measure of the effectiveness of different antenna systems when these are used as part of the same, specified communication circuit. This measure rates the antenna system according to the average information capacity of the link. To a first approximation this can be shown to be simply related to the average signal-to-noise ratio at one of the terminals of the link. The average signal-to-noise ratio will be used here as a factor of merit for the antenna system.

Over the frequency range considered transmission takes place by way of the ionosphere whose properties need to be accounted for by the rating system. In evaluating aircraft antennas, the relative location of transmitter and receiver, time and place of transmission, and the frequency actually used are all expressed as probability distributions.

The factor of merit is obtained as a weighted integral over the gain function of the aircraft antenna. The weighting function contains operational data of the communication system and statistical information of ionosphere behavior under the transmission conditions encountered.

* This work was supported by the U. S. Air Force under Contract No. AF 33 (038)-7850.

20. THE MEASUREMENT OF VARIATIONS IN ATMOSPHERIC REFRACTIVE INDEX.—George Birnbaum, H. E. Bussey, and R. R. Larson,* *National Bureau of Standards.*—A recording microwave refractometer has been adapted for measuring variations in the refractive index (ri) of the atmosphere and for determining scale of ri inhomogeneities. The following features are discussed: cavity design and arrangement, response time of instrument, and calibration.

With the aid of a small wind tunnel, the thermal errors attendant on sampling the atmosphere with cavity resonators have been investigated. These errors arise because of a change in temperature of the air sample on coming in contact with the cavity, and because of thermal expansion effects on the cavity frequency. The former effect may amount to 20 per cent; methods for reducing the latter are suggested. Other possible errors in the measurement of ri, those due to condensation and adsorption of water vapor, are apparently not significant.

Observations during August, 1951 were made with refractometers and meteorological equipment installed on two levels of a 420-foot tower at the Brookhaven National Laboratory, L. I., N. Y. It has been found from a preliminary data analysis that ri variations were caused mostly by variations in water-vapor density. The ri data at 410 feet show a wide range of Fourier components, corresponding to wavelengths of roughly 10 to 1,000 meters.

* Now with the Jacobs Instrument Co., Bethesda, Maryland.

21. DIRECTLY RECORDED TROPOSPHERIC REFRACTIVE INDEX FLUCTUATIONS AND PROFILES.*—C. M. Crain, *The University of Texas.*—This paper presents typical tropospheric measurements obtained with a direct-reading microwave refractometer. Included are the following: (a) Measurements to an altitude of 5,000 feet of refractive-index fluctuations over the Atlantic Ocean and coastal areas in the vicinity of Lakehurst, N. J., in April, 1951. The refractometer was carried aloft in a lighter-than-air airship. (b) Measurements to an altitude of 10,000 feet of refractive index fluctuation and profiles over the vicinity of Dayton, Ohio in June, 1951. For these measurements the refractometer was carried in a C-46 aircraft. (c) Airplane-carried refractometer measurements up to an altitude of 3,000 feet off the coast of South-Central California in October, 1951.

There is also included a brief discussion on the principle of operation of the refractometer and possible sources of errors in the measurements described. Slides of the equipment installation in an aircraft are shown.

* This work was sponsored by Office of Naval Research under Contract N5ori 136 P.O.I.

22. TROPOSPHERIC PROPAGATION WELL BEYOND THE HORIZON.—Thomas J. Carroll, *Massachusetts Institute of Technology.*—Observed vhf and microwave fields well beyond the horizon from high-powered transmitters are attributed to partial internal reflection from the normal troposphere with its index of refraction decreasing smoothly and continuously with height. A simple estimate of such reflection from the normal troposphere gives value well beyond the horizon of the order of magnitude of the observed average fields at frequencies throughout the vhf and microwave range. The conventional effective earth's radius correction for refraction fails to consider the tropospheric reflection effect on the field beyond the horizon. This failure is traced to the misbehavior of the linear index model at great heights. Either a bilinear model for the normal troposphere with a unit index above 30,000 feet, or an exponential model which falls asymptotically to unity at great heights, gives modes whose eigenvalues depend both on the index gradient and on the total index change through the atmosphere. Observations by Pickard and Stetson are believed to give strong qualitative experimental support for the importance of the total index change in a correct wave theory of weak nonoptical fields. Comparison is possible with the hypothesis of scattering from atmospheric turbulence with respect to the following: height gain, pulse-shape preservation, antenna beam width and gain observations, short- and long-term fading characteristics, divergence factor considerations, and number of arbitrary parameters involved.

23. PARTIAL REFLECTIONS IN TROPOSPHERIC PROPAGATION.—Joseph Feinstein, *National Bureau of Standards.*—The work on partial reflections produced by the gradient of refractive index associated with the standard atmosphere has proceeded along two lines: the revision of the standard theory required to theoretically justify the existence of these reflections, and development of more exact methods of calculating this type of field contribution.

The phase integral method of obtaining the eigenvalues employed in the mode residue expansion has been found to correspond solely to the earth-diffracted field; it is analogous to the geometric optic treatment which results from the use of the method of stationary phase in the interference region. The existence of additional eigenvalues, and in some cases of branch cuts, corresponding to the partial reflections present in an inhomogeneous atmosphere is indicated.

Quantitatively, the partial reflections originating from the field in the vicinity of the intersection of the horizon rays of transmitter and receiver has been found to be of most importance in contributing to the fields far beyond the horizon. This is brought about by the relatively slow variations in phase with height below the horizon level. The work of Pekeris has been utilized in this region. The results indicate a smaller than first-power wavelength dependence, and a saturating antenna height-gain factor. The antenna lobe structure generally plays a minor role because of the rapid phase variations of the contributions above the horizon.

24. TROPOSPHERIC PROPAGATION BEYOND THE HORIZON.—Martin Katzin, *Naval Research Laboratory.*—It has been proposed by Feinstein and Carroll that partial reflections at high levels in the standard atmosphere, due to the gradient of refractive index, can propagate far beyond the horizon by a free-space mechanism, and so would predominate over the diffracted field deep in the diffraction region. It is shown that these partial reflections are, in fact, taken into account in the rigorous normal mode theory, although they are neglected in the WKB approximation thereto. The inclusion of these partial reflections results in a modification of the characteristic values of the normal modes, but nothing more. Hence, for the standard atmosphere, in which all modes are leaky, the field sufficiently far beyond the horizon decays exponentially with distance. Thus, the observations far beyond the horizon of consistently higher fields than given by the normal mode theory cannot be attributed to the standard atmosphere over a spherical earth.

25. CONCERNING THE RADIO FIELD DUE TO INTERNAL REFLECTIONS IN THE STRATIFIED ATMOSPHERE.—L. J. Anderson and J. F. Colwell, *Navy Electronics Laboratory.*—The phenomenon of reflection of plane waves from a continuously varying medium has been investigated from a theoretical standpoint for some time.[1,2] Recently, it has been suggested that, depending on the position and type of receiving and transmitting antennas, such reflections may account for the large field strength observed beyond the optical horizon.[3] In this paper Feinstein's method is used to calculate the field strength for a standard linear refractive index and for a duct-type refractive index profile. Experimental results obtained at several different frequencies are compared with the calculated values and with other theoretical values.

[1] R. Gans, "Fortflanzung des Lichtes durch ein inhomogenes Medium," *Ann. Phys.*, vol. 47, pp. 709–736; 1915.

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

[3] J. Feinstein, "Tropospheric propagation beyond the horizon," *Jour. Appl. Phys.*, vol. 22, p. 1292; October, 1951.

26. ATMOSPHERIC NOISE IN THE VERY LOW-FREQUENCY RANGE*.—J. S. Barlow, G. W. Frey, and J. B. Newman, *The Johns Hopkins University.*—The vlf contributions to atmospheric noise of lightning flashes, having the "typical waveforms" of the current surges on overhead transmission lines, and of a typical daytime week, as received with a wide-band amplifier, are evaluated. The power spectrum of the flash waveform is flat from 180 to 20,000 cycles, falling off at lower and higher frequencies. The power spectrum of the daytime tweek is flat in the range 100 to 3,000 cps with a (+20 db) peak at the power-weighted mean frequency of the waveform (5,850 cps), followed by a sharp drop with increasing frequency.

The bases of the flash analysis following the method of Thomas and Burgess for the hf domain are empirical, being satisfactory only for a very nontypical flash. Our current work is toward an improved flash analysis consistent with the phenomonological data and the streamer theory of discharge for long gaps at atmospheric pressure. However, agreement of our analysis with more recent results of other investigators indicates that the outcome is not sensitive to the details of the mechanism.

The conclusions from the available information suggest that the maximum of atmospheric noise power usually lies at 6 to 8 kc Kessler and Goyder give the figure of 10 kc.

* Work done under auspices of the Office of Naval Research

27. AN APPROACH TO THE APPLICATION OF SUNSPOT-CYCLE CORRECTION TO ATMOSPHERIC RADIO NOISE PREDICTION.—Edna Shultz, *National Bureau of Standards.*—Studies show that there is a lag of atmospheric radio noise cycle behind sunspot cycle of the order of 19 months, both for the present cycle and for the one corresponding to a 125 months' series of atmospheric noise measurements beginning January, 1923. No statistically significant difference in lag was found between cycles, times of day, frequencies, or latitude. Knowledge of this phase shift may result in 10-to-15 db improvement in predictions.

28. DETERMINATION OF EFFECTIVE BANDWIDTHS OF RADIO NOISE METERS FOR IMPULSE AND RANDOM TYPE NOISE.—Francis T. Nicholson, *The University of Pennsylvania.*—When using radio noise measuring equipment, it is usually considered desirable to be able to express meter indications in such a way that they will be independent of the selectivity characteristics of the instrument. In order to accomplish this, in the case of impulse-type noise, knowledge of the effective impulse bandwidth of the noise meter is necessary. "Effective impulse bandwidth" is defined here as the ratio between the peak response of the meter to an impulse and the spectral intensity of the impulse.

A method is outlined whereby present knowledge concerning the response of ideal selective networks to impulses is used to determine the relationship between the effective impulse bandwidth of a practical noise meter and the shape of its selectivity curve. This method is applicable to high Q selective circuits of any number of stages utilizing either single-tuned circuits or double-tuned circuits with less than critical coupling. Results of experiments are given, which demonstrate that for the circuits presently being used in commercial noise meters the effective impulse bandwidth is very close to the bandwidth at the "6-db down" points on the selectivity curve. The distinction between the effective bandwidth for impulse-type noise and the effective bandwidth for random-type noise is emphasized, and reference is made to several methods for determining the latter.

29. THE OGIVER—A RADIO NOISE METER.—A. W. Sullivan, *The University of Florida*.—The Ogiver is a radio noise meter designed to measure the probability distribution of the envelope of atmospheric noise in terms of noise voltages exceeded with any degree of probability between 0.01 and 0.99. Measurement is made of the per cent of time that the noise wave exceeds certain arbitrary levels of intensity, and this is shown to be equivalent to direct probability measurement.

A major advantage of this type of noise meter is that measurements are made of the noise envelope without the use of special detector time constants as are found in the quasi-peak and peak types of noise meters. This equipment is also readily adaptable to noise measurements at the output of several types of detection systems.

Data obtained with the Ogiver are basically those necessary to obtain the probability distribution of the noise. These distributions can be measured over limited time intervals, or instantaneously. In some cases the entire distribution may not be desired, and measurements can be made of the time variations of a single percentile level of noise intensity.

This measure of noise yields a better description of noise conditions than the conventional types, and shows promise of correlating closely with the subjective effects of noise interference on some communication systems.

30. SOLAR FLARES AND ASSOCIATED 200-MC/SEC RADIATION.—Helen W. Dodson and E. Ruth Hedeman, *McMath-Hulbert Observatory of the University of Michigan*, and Lief Owren,* *Cornell University*.—Solar radio-frequency radiations on 200 mc have been compared with solar activity observed photographically and visually. The most clear-cut relationships have been found for radio emissions associated with solar flares.

* On leave from the Institute of Theoretical Astrophysics, University of Oslo.

The present study covers 141 flares which occurred during the period August, 1948 to December, 1950. A solar flare is a complex phenomenon, and it is not certain which aspect of it is related to the associated radio event.

The 200-mc events may be divided into seven categories: major bursts including outbursts, minor bursts, micro bursts, series of bursts, small rise in base level, short-duration noise storms, nulls.

Eighty per cent of the flares produced some form of enhancement in the 200-mc radiation. With but four significant exceptions, all of the flares for which no distinctive radio event occurred were relatively unimportant solar phenomena. For 20 flares for which the energy excess of the radio event has been measured there appears to be a direct relationship between this and the importance of the flare. Comparisons of the times of onset of the flare and the radio event indicate that major, minor, and micro bursts occur at the very commencement of the optical rise. Series of bursts begin in a number of cases before the start of the optical flare; noise storms and base-level increases start at or near flare maximum. Such delayed events increase in intensity as the flare fades and often attain greatest intensity when the flare has faded completely.

31. THE EFFECT OF VELOCITY DISTRIBUTION IN A MODULATED ELECTRON STREAM.—D. A. Watkins, *Hughes Aircraft Company*.—A method of solving electron-beam problems is described which takes into account the thermal velocity spread. The method is based on the Liouville theorem. Two such problems are solved by means of a power series. In the case of small-signal velocity modulation of a drifting stream, the velocity spread distorts the shape of the standing wave of current which is produced in addition to lengthening the plasma wavelength and the electronic wavelength. For a drifting stream with full shot noise in each velocity class at the input, a standing wave of noise current is produced with deep minima. The size of the first minimum is found to increase as the square of frequency and with increasing velocity spread. That full uncorrelated shot noise can produce a standing wave of convection current is somewhat surprising. The present analysis demonstrates that space charge is the important factor in determining the behavior of noise currents in electron streams even at high frequencies, and that velocity spread plays a less important role.

32. NOISE IN ELECTRON BEAMS WITH A VELOCITY DISTRIBUTION.—L. R. Walker, *Bell Telephone Laboratories*.—A one-dimensional theory indicates that, for a stream of electrons, homogeneous in dc velocity, moving in a drift space, the noise power in a narrow frequency range should exhibit a periodic variation with distance, going to zero at the minima. Experimentally, Cutler and Quate find a residual noise at the minima. An estimate is made here of the contribution to this excess noise arising from the spread of dc velocities in the beam. In the one-dimensional theory the noise current satisfies a linear integral equation. This may be solved exactly by Laplace transforms, but, the result is, in general, difficult to use. For the practical case where the dc velocity spread is small compared to the average dc velocity, the integral equation may be replaced approximately by a differential equation which may be solved. The approximation is valid over several periods of the noise variation and the error is calculable. An expression for the noise at the minima may now be found. The method may be generalized to calculate the noise current in some accelerating regions.

33. SPACE-CHARGE WAVES IN MAGNETICALLY FOCUSED BEAMS.—M. Chodorow and L. Zitelli, *Stanford University*.—Any electron beam, if disturbed in any way, as by velocity modulation or density modulation, will exhibit so-called plasma oscillations, which depend on the beam density. Detailed behavior of these oscillations will be determined by the nature of the adjacent boundaries and by any focusing magnetic or electric fields. The simplest geometrical configuration is an electron beam in the drift tube of a klystron. Previous analyses have been for the case of either an infinite magnetic field or zero magnetic field, which are not the commonest forms of focusing used. The analysis has been extended to include the case of a finite magnetic field and in particular the Brillouin field in which the magnetic forces are just sufficient to counteract the divergent space-charge forces. The equations have been formulated in such a way that it is possible to examine the behavior of the space-charge waves as the magnetic field is varied from zero to infinity and to trace the change of the characteristic propagation constants.

There are two sets of such waves, only one of which is really significant for describing the space-charge behavior in the gridless klystron. These waves change their character as the magnetic field increases so that at zero field all the current is in the surface, due to the displacement of the boundary, with no volume current. This changes so that at the Brillouin field about a third of the current is in the volume, and at infinite fields where there is no transverse motion, all the current is in the volume. When initial conditions are satisfied, the second set of waves always has amplitude coefficients of the second order and can be neglected.

34. SPACE-CHARGE WAVE—GENERAL THEORY.—Philippe A. Clavier, *Sylvania Electric Products, Inc*.—A true space-charge wave propagation is defined as compared to a modified cold-mode propagation.

In the true space-charge wave medium, problems occur in which gradients in the cross section are of primary importance; in other problems, a unidimensional analysis is sufficient.

Results will be given for a beam of constant velocity and variable current densities in the cross section. Amplification occurs in the forward direction for given configurations of current densities.

A closed-form analysis in the unidimensional case when the beam voltage varies arbitrarily along the path of the beam is carried out, and results for different important cases are given.

35. THE EFFECT OF THERMAL VELOCITIES ON THE DC BEHAVIOR OF DIODES.—Philip Parzen, *Federal Telecommunication Laboratories.*—Hahn[1] has suggested that it is possible to account for the effect of the thermal velocities by including a pressure term in the equation of motion, similar to that in the flow of gases. The pressure is proportional to the electron density and an equivalent electron temperature. If one assumes further that the electron temperature does not vary between the cathode and anode, then the resultant system of equations may be solved in closed form for the variation of velocity and potential in the diode. This extends Hahn's numerical solution for the region between the virtual cathode and the anode. The region between the cathode and virtual cathode was not considered by Hahn. This solution compares favorably with that obtained by Langmuir[2] with a large amount of numerical labor. There are, however, numerical differences which may be attributed to lack of constancy of the electron temperature.

[1] W. C. Hahn, Proc. I.R.E., vol. 36, pp. 1115–1121; September, 1948.
[2] I. Langmuir, *Phys. Rev.*, vol. 17, p. 441; 1921.

36. A NEW METHOD OF CALIBRATING FIELD-STRENGTH MEASURING EQUIPMENT.—Harold E. Dinger and William E. Garner, *Naval Research Laboratory.*—Two methods have been in general use for the calibration of field-strength measuring equipment employing loop antennas. In one method the standard field is created by a known current flowing through a loop, in the other by a known current flowing through a transmission line erected in a shielded room. Both of these methods require elaborate setups to obtain a high degree of accuracy. Of these two methods the loop-to-loop method is considered the most accurate. Both methods are rather inconvenient for production calibration and are subject to errors which may be cumulative.

A new method of calibrating field-strength measuring equipment using shielded loop antennas is now being investigated. The basic concept is that of using the shielded loop as a transformer. The voltage that is induced in the shield of a loop due to the presence of a field can be calculated from a well-known equation. This value of voltage can then be applied across the shield gap by a signal generator with the loop turns open circuited. The meter to be calibrated is then connected to the loop and the meter should indicate a field strength equal to that which was used to calculate the induced voltage in the shield. This method, referred to as the "shield-injection" method, is easily performed in practice and requires only a standard signal generator.

37. CHARACTERISTICS OF MICROWAVE PRINTED LINES WITH APPLICATIONS TO BAND-PASS MICROWAVE FILTER DESIGN.—M. Arditi and J. Elefant, *Federal Telecommunication Laboratories.*—Microwave printed lines have been developed as substitutes for waveguides and coaxial lines. This development is a part of a program of circuit miniaturization extended to the uhf field. The physical requirements of such printed lines make it difficult to apply the usual techniques of microwave measurements in waveguides or coaxial lines. The application of the principles of the method developed by G. Deschamps in a companion paper provides a unique method of measuring various parameters of the microwave printed line in a simple but accurate manner. Examples of such measurements as applied to various types of microwave printed lines are described. The characteristics of transducers between a coaxial line feeder and a microwave printed line are analyzed experimentally. Matching conditions and broadbanding of the transducer are studied. This method gives also very easily the insertion losses of the transducer and the losses of the line. The scattering matrix coefficients of various types of susceptances produced by obstacles in the line are studied. Equivalent circuits are deduced. From the values of the susceptances, the design characteristics of band-pass microwave filters are obtained. Experimental results are given.

38. APPLICATION OF NONEUCLIDEAN GEOMETRY TO THE ANALYSIS OF WAVEGUIDE JUNCTIONS.—George Deschamps, *Federal Telecommunication Laboratories, Inc.*—Various practical problems involving bilinear transformations can be solved by non-Euclidean geometry. Two models of Lobachewsky geometry illustrate this point. They form a natural domain for representing reflection coefficients. One is the Smith chart, where non-Euclidean distances to the center are standing-wave ratios in decibels. The other model may be simpler sometimes as its geodesics are straight lines.

Transformation from one model to the other is easy and can prove an important property; chords joining images by a bilinear transformation of opposite points on a circle intersect. From their intersection, the image of the center or "iconocenter" of the circle can be constructed.

A waveguide junction is characterized by a bilinear transformation between input and output reflection coefficients. Some relations are given between scattering parameters (reflection and transmission coefficients) of a junction and the corresponding geometric transformation. This highly accurate method of measurement applies to lossy junctions also.

Measurements through a junction, lossy or not, sometimes cannot be avoided or are even desirable. An equivalent circuit need not be computed. A direct-reading "hyperbolic protractor" is described that gives corrected standing-wave ratios as if the junction were not there.

39. GENERATION OF STANDARD FREQUENCIES USING A SELECTIVE SPECTRUM GENERATOR.—R. Guenther and A. Hahnel, *Signal Corps Engineering Laboratory.*—To generate highly accurate and stable standard RF frequencies over a wide range, from a single low-frequency standard, it is common practice to use a harmonic generator, for example, a nonlinear device producing harmonics of the primary standard. The result is a spectrum which spreads over the entire frequency range from the fundamental to the highest required frequency. Hence, the energy within a narrow band or for the individual harmonic is rather low. In many applications, however, only a small fraction, or even a single frequency of the entire spectrum is used. In such a case it would be desirable to generate a restricted band of frequencies and concentrate all the energy in this narrow band. The selective spectrum generator described in this paper provides such a restricted harmonic spectrum in a very simple manner. It consists of a tuned RF oscillator, which determines the center of the narrow spectrum, and a pulse generator or shaper with the repetition rate derived from the standard frequency. If the oscillator is pulsed in such a way that the oscillations will start with the same RF phase at the start of each pulse, the resultant wave shape is periodical in the period of the pulse frequency. Hence, the spectrum contains harmonics of the standard frequency only. An analytical treatment shows that the envelope of the spectrum can be varied between extreme narrow and moderately wide bands. The result of these calculations is used to show how the different design parameters influence the envelope of the frequency spectrum. Experiments conducted on such a circuit show good agreement with the theoretical study as evidenced by samples of oscillograms of wave shape and corresponding frequency spectrum for typical examples.

The selective spectrum generator proved to be very useful as a simple frequency standard for calibration purposes and frequency controls and in precision frequency meters.

40. LOW-FREQUENCY PROPAGATION IN AN EXPONENTIAL IONOSPHERIC LAYER.—J. Shmoys, *New York University.*—The paper is concerned with the study of propagation of low-frequency plane waves normally incident on a horizontally stratified ionosphere with exponentially increasing charge density and oblique magnetic field. The treatment parallels that of Whipple of the similar problem of oblique incidence and vertical magnetic field. It is shown that in the case of low collision frequency (the assumption used by Whipple) coupling is small unless the ionization gradient is very high. Furthermore, the two rays are both reflected: one partially at the level where $\omega_0 \sim \omega$, the other totally at the level where $\omega_0 \sim (\omega\omega_H)^{1/2}$. Because of the special form of the differential equations in the case of exponential electron density variation, the reflection coefficients can be calculated explicitly.

A similar study of both oblique incidence with vertical magnetic field and vertical incidence with oblique magnetic field has been made in the case of large collision frequency.

41. THEORETICAL AND EXPERIMENTAL INVESTIGATION OF THE GROUP HEIGHTS OF REFLECTION OF 150-KC/S RADIO WAVES VERTICALLY INCIDENT ON THE IONOSPHERE.*—Norman Davids and Rune Lindquist,† *The Pennsylvania State College.*—This paper is

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† On leave from the Chalmers Institute of Technology, Gothenberg, Sweden.

concerned with the theoretical and experimental determination of the group heights of reflection of long radio waves vertically incident on the ionosphere. The theoretically expected group heights for 250 microsecond Gaussian-shaped pulses are determined utilizing: (1) a wave-theory treatment including coupling, (2) a Chapman-like ionospheric model including all variables and their height variations, and (3) a fundamental radio frequency or carrier component of 150 kc.

From the theoretical viewpoint the continuous-wave (cw) results of Gibbons and Nertney (1951) (1952) are extended by including the dispersive characteristics of the above model. This is accomplished by determining the received pulse characteristics by means of a response function developed from a suitable Fourier-Hermite series, which includes the frequency dependent effects of the model. It is found that the rapid changes in polarization near the lower edge of E region, called the "coupling region," are very effective in generating reflected waves or pulses under suitable conditions, for example, late night hours associated with low E-layer critical frequencies. However, the dispersive characteristics are small so that almost negligible group retardation is to be expected. At a higher level in the layer, rapid variations in the index of refraction for one of the characteristic wave components occur in what is called the "reflection region." Here the time delays are large for low critical-frequency models.

Some geometrical discussion is included concerning polarization and coupling factors. If not considered, then the ordinary and extraordinary modes can be given a representation (in a certain complex space) as orthogonal principal directions.

The above theory is compared with the experimentally observed group heights on a diurnal and seasonal basis. Good agreement is obtained and is further strengthened by comparison with some recent measurements of R. E. Jones on the diurnal change in phase (cw) paths.

The occurrence of a third echo in addition to the regular first hop E echoes is reported. This apparently occurs from levels above the maximum ionization density of the E layer, height range 100 to 200 km, and appears to be associated with low nighttime E-layer critical frequencies and magnetically disturbed conditions.

Correlations of experimental results with sunrise, solar flares, and so forth are also considered.

42. IONOSPHERIC WIND MEASUREMENTS AT 150 KC/S.*—G. H. Millman, *The Pennsylvania State College.*—The results of ionospheric wind measurements utilizing the three spaced-receiver technique at long-wave lengths are given. Evidence is presented which indicates the presence of diurnal and seasonal variations of upper-atmosphere wind movements. The measured wind magnitudes range from 20 to 300 meters per second, with most probable values between 40 and 100 meters per second.

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By the method of Briggs-Phillips (1950), the "angular spreading" of the downcoming reflected wave is found to be approximately 8.5 degrees as compared to the theoretical prediction of 7.5 degrees. This appears to verify the fact that the ionosphere acts as a diffracting screen at long wavelengths.

The calculated rms line-of-sight speeds of ionospheric scattering centers are found to be about 6–8 meters per second. Agreement is found between the speeds obtained by the Booker-Ratcliffe-Shinn (1950) autocorrelation method and Ratcliffe's (1948) speed of fading method.

A study of the facing rate of the first reflected echo indicates the presence of fades of $\frac{1}{4}$ to 15 minutes duration with maximum between 1 and 2 minutes. The facing rate is found to increase with greater indices of geomagnetic activity (kw).

Several other factors and correlations are also discussed.

43. TURBULENCE IN THE LOWER IONOSPHERE AS DEDUCED FROM INCREMENTS IN ABSORPTION AND PHASE PATH AT 150 KC/S.*—R. E. Jones, G. H. Millman, and R. J. Nertney.—*The Pennsylvania State College.*—It is important in the study of ionospheric winds to determine the height of the "diffraction screen" which produces the variations in radio-wave field strength, or other characteristics, measured at ground level. A method which may be used to determine this height is developed.

It is shown that, under proper conditions, the mean electronic collisional frequency associated with the electron clouds which are presumed to form the "diffraction screen" may be determined by studying change in phase-path records obtained simultaneously with absorption records. This value of collisional frequency may then be related to height through an ionospheric model.

Single recording stations are used in this comparison of the phase pattern and the amplitude pattern at ground level. The statistical information necessary to the theoretical development is obtained, in the usual way, from a triangular arrangement of receiving stations which record the amplitude of the signal.

Preliminary calculations show that the screen which produces the daytime ionospheric winds measured at 150 kc lies in a collisional frequency range from 2×10^6 to 8×10^6 sec^{-1}. This would correspond to a height range from 70 to 80 km for our ionospheric model. These results appear to be in good agreement with recent work by Kellogg (1951) from the meteorological viewpoint.

* The research reported in this paper has been supported by the Geophysical Research Division of the Air Force Cambridge Research Center under Contract AF 19 (122)-44.

44. SPORADIC-*E* STRATIFICATION AND CORRELATION WITH LOW-FREQUENCY SOUNDINGS.—R. A. Helliwell, *Stanford University.*—The virtual heights of nighttime sporadic-E reflections are examined for evidence of stratification and correlation with virtual heights measured at low frequencies. For the month of August, 1951, the heights of sporadic-E appearing at 2.4 mc on the routine h'f records from a C-3 recorder range from 97 to 137 km and show a pronounced maximum of occurrence at 100 km with less well-defined maxima at 106 km and 112 km. At 310 kc there is evidence of maxima at the same three heights with an additional maximum of occurrence at 93–94. When two or more traces appear on the same C-3 record (but usually not in the same frequency range), the height separations of these traces show a tendency to be independent of the height of the lowest trace. The frequency of occurrence of these height separations shows peaks at 7 km and 14 km, which are better defined than the peaks in the frequencies of occurrence of the absolute heights.

To test for correlation between the heights at low and high frequencies the lowest height measured on 2.4 mc was plotted against the heights of all echoes recorded on 310 kc at corresponding times. The resulting scatter plot shows a pronounced grouping of points along a line representing one-to-one correlation.

It is concluded from the study that: (1) sporadic-E tends to occur at certain preferred heights which for nighttime during the month of August are 100 km, 106 km, and 112 km; (2) the differences of different reflection heights may be less variable than the heights themselves; (3) low-frequency echoes and sporadic-E echoes are correlated in virtual height and are probably reflected from the same layers.

45. CORRELATION VERSUS LINEAR TRANSFORMS.—Marcel J. E. Golay, *Signal Corps Engineering Laboratories.*—A criticism is made of the emphasis recently placed by communication engineers on correlative concepts and devices. Specific suggestions, illustrated by an example borrowed from radar, are offered about other possible attacks on the problems presented by the submergence of wanted data in noise.

46. A METHOD FOR THE CONSTRUCTION OF OPTIMUM CODES.—David A. Huffman, *Massachusetts Institute of Technology.*—An optimum method of assigning codes to members of an ensemble of messages is developed. For an ensemble consisting of a finite number of messages, such a method yields a minimum average number of digits. The coding procedure is applicable whatever the number of coding symbols may be.

47. A NOTE ON MOVING POLES IN NONLINEAR OSCILLATING SYSTEMS.—William B. Wrigley, *Georgia Institute of Technology* and *Air Force Cambridge Research Center.*—Since poles of the complex immittance of a linear system represent the decrements and frequencies of rotating phasors in the linear time domain, it is suggested that a nonlinear system might be represented by moving poles whose instantaneous decrements and frequencies are associated with phasors rotating in the nonlinear or time-distorted phase space. This idea is applied to the analysis of a class of nonlinear oscillation generators of the second order which is described by the differential equation, $\ddot{X}-N_1(\dot{X}, X)=0$.

48. RISE-TIME MODULATION.—Maxime G. Kaufman, *Naval Research Laboratory.*—During investigations of the loss of pulse amplitude across the RC coupling between stages of a video-amplifier, it became apparent that the *amplitude* of the pulse across the resistance was a function of the *steepness* of the applied pulse. Analysis of this condition and application of the principle to pulse modulation offered a new system of modulation to which the term "rise-time modulation" has been assigned.

Rise-time modulation may be defined as a method of pulse modulation wherein the intelligence conveyed by the modulating voltage is imparted to the pulses in such manner as to control the rate of rise of the pulses. Likewise, the rate of fall may be modulated ("fall-time modulation"). Such a method of pulse modulation has the following features: (1) It eliminates the need for synchronization; (2) it permits the pulse repetition rate to vary); (3) it carries several channels simultaneously.

49. TRANSFER EFFICIENCY.—Donald K. Weaver, Jr., *Stanford Research Institute.*—Transfer efficiency is defined as the ratio of output power to input power for a network. This paper describes a relationship between efficiency and changes in the input impedance caused by changes in the load impedance. This relation is useful in measuring the efficiency of antenna couplers.

50. RANDOM PROCESSES IN WAVE PROPAGATION.—W. S. Ament, *Naval Research Laboratory.*—Part I. Transmission. For wave propagation through a randomly scattering medium, theory predicts statistical averages of certain measurements. Familiar theories are discussed in terms of the measurements they relate to. For instance, the "extinction cross section" of a large opaque object is simply shown to be twice its geometric cross section; the relevant measurement is not one of transmitted power flux, but an interferometric measurement of the average phase and amplitude of a coherent wave passing through a cloud of such objects. A similar interferometric measurement relates to the attenuation currently calculated from the Booker-Gordon scatter-ing formula. This is shown by two new derivations of the propagation constant of a "blobby" medium. The simpler measurement of average transmitted power-flux requires a complicated theory of photon transport. Theory and measurements relating to angle-of-arrival are intermediate between those relating to power-flux and propagation constant.

Part II. Reflection. Similar considerations hold for reflection from a rough surface. Specular reflection is an interferometric concept. Lambert's law and back-scatter are power-flux concepts. A theory of specular reflection is logically prior to a theory of back scatter. The requirements for such theories are discussed through simple examples.

51. A METHOD FOR EVALUATING TRENDS IN TIME SERIES OF TROPOSPHERIC RADIO FIELD-STRENGTH DATA.—Philip L. Rice, *National Bureau of Standards.*—This paper describes an evaluation of the seasonal trend in a series of hourly median field strengths corresponding to one hour of the day. Of three methods investigated in order to evaluate time trends, a type of harmonic analysis was found to be the most successful. The other two methods were an adaptation of the smoothing theory of Weiner and Levinson and a method using orthogonal polynomials.

Two powerful tools of the statistician, the autocorrelation function and the variate difference method, enable us to refine our analysis of time trends to a point where we hope to obtain a maximum amount of correlation with similar trends in relevant meteorological information.

52. A FORMULA FOR THE TRANSMISSION LOSS OF SPACE WAVES PROPAGATED OVER IRREGULAR TERRAIN.—Kenneth A. Norton, *National Bureau of Standards.*—It is shown that the effects of irregularities in the height of the terrain on space-wave intensity calculations may be adequately described in terms of the coefficients of a second-degree polynomial fitted by the method of least squares to an appropriate portion of the terrain between the transmitting and receiving sites. The mean-square residual height variation from such a polynomial provides a measure, through Rayleigh's criterion of roughness, as to the usefulness of the resulting field-strength formula. If the frequency under consideration is sufficiently low to satisfy this criterion, then the three coefficients in the second-degree polynomial fitting the terrain are found to play important roles in determining the effective heights of the transmitting and receiving antennas, and the effective curvature of the transmission path. A constant lapse of refractive index with height may be taken into account by a modification in the effective path curvature.

The utility of the formula on frequencies between 92 and 1,047 mc is established by comparing the calculated and measured fields for a variety of terrain in the vicinity of the Cheyenne Mountain transmitting station of the National Bureau of Standards.

53. MEASUREMENT OF THE EFFECT OF IRREGULAR TERRAIN ON DIRECTIVE ANTENNA PATTERNS.—R. S. Kirby, J. M. Taff, and H. S. Moore, *National Bureau of Standards.*—In co-operation with the United States Army Signal Corps, the National Bureau of Standards has undertaken the measurement of directive antenna patterns over irregular terrain in the Fort Dix, N. J., area. Measurements of the patterns of directive antennas were made while driving around the transmitters at relatively constant distances of 0.4 mile, 10 miles, and 30 miles with a mobile field, strength recording vehicle. Measurements were made at 49 mc, 141.75 mc, 239 mc, and 460 mc with vertically polarized waves. At the recording unit omnidirectional ground plane antennas were used on a telescopic mast which could be operated at heights of from 10 feet to 30 feet above the ground. Because of the great number of wires, trees, and other obstacles over the road, most of the pattern measurements were made with receiving antenna heights of 12 feet and 15 feet.

In addition to the measurements made around the transmitters, receiving antenna pattern measurements were made using a 460-mc corner reflector antenna with a gain of approximately 10 db. These measurements were made at spot locations around the 460-mc transmitter.

54. THE CONSTANTS OF THE EQUATION FOR THE REFRACTIVE INDEX OF AIR.—Ernest K. Smith, Jr., *National Bureau of Standards.*—Recent measurements at the National Bureau of Standards, the University of Texas, the National Physical Laboratory, and elsewhere have indicated that the conventional constants ($A = 79°K/mb$, $B = 4800°K$) in the expression for the refractive index of moist air $(n-1)\cdot 10^6 = \frac{A}{T}(p + \frac{B}{T})$ should be revised. Various laboratories appear to have arrived at this conclusion independently with the result that there are at least four different sets of constants in current use. In much of propagation work the absolute value of the refractive index of the atmosphere is of small moment. However in some work it is important and it seems highly desirable to decide upon a particular set of constants. It is also advisable to agree on the amount of future improvement in the constants which will make it worthwhile to respecify them. This paper attempts to give "best" values of the constants in the least squares sense, and also suggests a criterion for deciding whether a future change would be worthwhile.

55. EFFECT OF PARTICLE SHAPE AND COMPOSITION ON MICROWAVE ATTENUATION AND SCATTERING BY PRECIPITATION.—Walter Hitschfeld, Kenrich Gunn, T. W. R. East, and J. S. Marshall, *McGill University.*—Calculations indicate that when a small spherical ice particle melts to a water drop, its scattering increases by a factor 5. Most of this increase is effected by the melting of only a thin coating.

The scattering by a small ice particle is almost independent of its shape, but if the corresponding water drop were lengthened to an axial ratio of 10 to 1, its scattering is calculated (for $\lambda = 3$ cm) to increase 20 to 30 times. A water-coated ice crystal with 10:1 axes presumably scatters like a distorted water drop, and so scatters 20 to 30 times as much as a water sphere, or up to 150 times as much as any ice particle of the same mass.

The total absorption by a small water-coated ice sphere is as much as, or more than that by the all-melted drop. The effect of distortion from sphericity on total absorption is roughly as much as, probably more than, the effect on scattering. Thus while an ice particle absorbs negligibly regardless of shape, a thin water coating may run the absorption up to 30 or more times that of a water drop of the same mass.

Intense scattering by melting snow has already been observed as a "bright bank" in precipitation echo just below the freezing level. Strong attenuation may be anticipated in the same region, the attenuation being greater than that by rain in roughly the same proportion as the echo intensity.

56. A SWEEP-FREQUENCY IONOSPHERE RECORDER FOR THE LOW FREQUENCIES.—J. C. Blair, J. N. Brown, and J. M. Watts, *National Bureau of Standards.*—Application of the techniques heretofore used in high-frequency ionosphere recorders is described. The beat-frequency method of generating wide-frequency sweeps is used, covering frequencies from 50 kc to 1,000 kc in a short time without band switching. Advantage is gained by the use of transformers containing ferromagnetic cores in the wide-band transmitter amplifiers, but the antenna system, for practical reasons, is very inefficient. Sample records are shown.

57. SCATTER-SOUNDING: A TECHNIQUE FOR STUDY OF THE IONOSPHERE AT A DISTANCE.—O. G. Villard, Jr. and A. M. Peterson, *Stanford University.*—It is the purpose of this paper to summarize the current state of knowledge in regard to oblique-incidence long-distance backscattering, and to suggest a number of practical applications. It appears that "scattersounding"—studying the reflecting properties of the ionosphere by means of energy backscattered from the earth—offers a valuable new technique for studying the characteristics of the ionosphere at a distance.

Recent research has shown (1) that the majority of oblique-incidence backscattering occurs from the earth, and not from layers or clouds in the ionosphere, (2) that the backscattering coefficient of the open sea is not noticeably different from that of land, (3) that the employment of relatively long pulse lengths enormously increases the strength of the returned echoes, and (4) that time delay to the leading edge of the scatter echo is a reasonably accurate measure of ground distance to the edge of the skip zone, except when this distance is less than 1,000 km in the case of F-layer propagation. A result of item (3) is that a peak transmitter power of less than a kilowatt is ample for many applications.

Within the limitation of (4) above, scattersounding may be used to determine the muf for either F-layer or sporadic-E transmission, and is especially useful as a means for plotting at a distance the growth and development of sporadic-E clouds of ionization. For this work antenna directivity is essential; that of a three-element Yagi has been found to be satisfactory. A simple lumped-constant tr circuit is described. Antenna rotation and ppi display of scatter echoes shows at a glance the areas to which communication is possible at a given time and frequency. A statistical test of this technique for communication prediction has been devised, which takes advantage of the wide geographical distribution of radio amateurs. The results have confirmed expectations, both for F-layer and sporadic-E layer transmission.

Curves are presented by which oblique-incidence time delay may be converted into the desired skip distance. Errors to be expected when diurnal and seasonal average layer heights are used in the conversion process are shown to be small when transmission paths are long. An overlay is developed which, when used with vertical incidence p'f records, determines the equivalent height of reflection for use in scatter-echo interpretation. Theoretical reasons for the observed increase in echo amplitude with increased pulse widths are discussed.

58. F-REGION EFFECTS OF SOLAR ECLIPSE AT SUNRISE, SEPTEMBER 1, 1951.—H. W. Wells, *Carnegie Institution of Washington.*—The annular eclipse of September 1, 1951 started before ground sunrise along the East Coast of the United States, but did not reach maximum phase until later. Three high-speed ionospheric stations were operated by the DTM, CIW for the eclipse observations. Locations at Charlottesville, Virginia; Derwood, Maryland; and Chincoteague, Virginia established a west-to-east chain with local time difference of approximately twelve minutes. The maximum phase of eclipse occurred at a time (twenty to thirty minutes after ground sunrise) when normal rate of production of ionization (established by control observations) was very high.

The results show absence of any ion production at *any station* for a period of approximately one-half hour centered on time of eclipse maximum. From the moment when two-thirds of the sun was covered, through the maximum phase (+91 per cent), and until one-third of the sun was uncovered, no ionization was generated.

Several possible explanations are discussed: (1) emitting sources near center of sun's disc, (2) uniform solar emission but an effective limb darkening, and (3) an atmosphere on the moon.

59. IONOSPHERE REFLECTION COEFFICIENTS BY VARIATIONAL TECHNIQUE.—J. Lurye, *New York University.*—One of the important problems of ionosphere work is to predict the nature of the electromagnetic wave reflected by the ionosphere when a wave of known form is incident upon it. In the present paper, this problem is treated by considering a plane ionosphere model for which the electron density and earth's magnetic field are continuous but otherwise arbitrary functions of the height. It is shown that if a plane wave impinges on such a model at an arbitrary angle of incidence, then four complex reflex coefficients suffice to characterize the reflected wave as to its intensity, phase, and polarization. Thus the problem becomes one of calculating these four numbers. The attack on this problem is based on an adaptation of a method due to the physicist, J. Schwinger, wherein, with the aid of an exact integral equation for the electric field in the ionosphere, one derives variational formulas for the four reflection coefficients. These formulas may then be used either to calculate the coefficients numerically or to obtain approximate expressions for them in terms of the various physical parameters of the problem.

Finally, as a consequence of a certain symmetry property of the variational formulas, a reciprocity theorem for ionospheric propagation is deduced.

60. DISTANT RADIO-COMMUNICATION THEORY.*—M. J. DiToro, *Federal Telecommunication Laboratories.*—Distant radio communication at high frequencies is difficult because the transmission medium—the earth ionosphere duct—is time variable, noisy, and shows dispersive or multipath transmission with consequent fading of the received signal. Because most of these factors are random and not under design control, one can treat the ionospheric communication problem only on a statistical basis in terms, for example, of such things as the probability that a transmitted pulse or bit of information will be received correctly. On this basis it is shown that, by a purely numerical experiment wherein random number tables are used to simulate fading and noise, it is possible to appraise various telegraph transmission systems without the costly process of building them and then testing their performance.

An approximate, but simple, analysis is given of receiver signal detection, by which is meant generation of dc for operation of an output printing device. It is shown that the important parameter here is the average or expectation of the difference in dc between a received space and mark signal, divided by the square root of the variance of this difference. Incoherent square law detection is compared with coherent detection using the matched filter or, what is the same thing, correlation.

The use of diversity transmission to overcome signal fading at the receiver is considered. The important statistical data regarding time-varying ionospheric transmission are obtained from the auto and cross-correlation functions of the received signal envelopes for the various diversities. For the correlated fading of signals in two transmission channels, a simple design formula is shown which predicts the improvement in the use of diversity.

A brief description will also be given of a simple acoustic ionosphere analog simulating time varying multipath transmission.

* This paper gives some of the theoretical results obtained in a project for the Wright Air Development Center, Dayton, Ohio.

61. SYMPOSIUM ON THE MEASUREMENT OF ATMOSPHERIC NOISE.—An informal presentation and discussion of material by A. W. Sullivan and J. M. Barney, *University of Florida*, W. Q. Crichlow, *National Bureau of Standards*, Ralph Showers, *University of Pennsylvania*, and others.—In measuring atmospheric noise phenomena it is desirable to obtain a complete understanding of the noise process. This is accomplished through fine structure studies (waveforms) and statistical analyses of noise as function of time and frequency. These studies are primarily concerned with the study of noise characteristics over relatively short time intervals, for example, the order of integrating times to be used in the noise meters.

Before the measured characteristics of noise can be generally useful, it is necessary to evaluate the response of noise meters in such a manner as to determine the effect of the noise meter on the characteristics of noise at the output. Particularly, the predetection and postdetection bandwidths, the phase response, and automatic-gain-control time-constant of the noise meter will influence the measured noise characteristics. The effects of these must be determined if noise measurements made with different noise meters are to be referred to a common basis.

The various parameters of atmospheric noise should be related to the interference effects of noise on communication systems in order to determine which parameters of noise are to be measured in the field. This amounts to correlating the various objective measures of noise intensity with the subjective annoyance value of the noise on certain communication systems.

After the characteristics of noise have been evaluated, a noise meter has been designed, and a useful parameter of noise has been chosen, it is then necessary to undertake a long-term measurement program to study noise conditions as a function of geographical position, time, and frequency. Data will then be obtained to facilitate prediction of required signal intensities for reliable communication over given paths, at given times and frequencies.

62. **THE DIFFERENCES IN THE RELATIONSHIP BETWEEN IONOSPHERIC CRITICAL FREQUENCIES AND SUNSPOT NUMBER FOR DIFFERENT SUNSPOT CYCLES.**—S. M. Ostrow and M. PoKempner, *National Bureau of Standards.*—The approximately linear relationship between ionospheric critical frequencies and sunspot numbers is well known. Examination of ionospheric data from the few locations at which ionosphere observations have been made for substantial parts of two sunspot cycles shows appreciable differences in the slope of the line of correlation between critical frequencies and sunspot numbers for different cycles. This is an additional indication that the Zurich sunspot number is only an approximate index of the solar activity responsible for ionospheric ionization. The differences between cycles should be taken into account in the preparation of ionospheric radio-frequency propagation predictions.

63. **CONTINENTAL MAPS FOR FOUR IONOSPHERE DISTURBANCES.**—R. S. Lawrence, *National Bureau of Standards.*—Observations of *F*-2 critical frequencies from thirteen stations in North America were used to study characteristics of subauroral ionospheric disturbances on a continental scale. Contour maps of percentage deviation from the monthly median were drawn at two-hour intervals for four storms. Positive and negative deviations as great as 60 per cent were noted. The maps reveal that two of these disturbances had rather well-defined centers of 2,000- to 3,000-km diameter which moved across the continent. The other two storms had no centers, but exhibited a marked dependence upon geomagnetic latitude. During these latter disturbances which occurred in winter, the critical frequency was enhanced in regions below and reduced in regions above approximately 50-degrees geomagnetic latitude. Some additional evidence indicates that the reversal takes place at a lower latitude in the summer.

64. **RELATIONSHIPS BETWEEN AURORAS AND SPORADIC-*E* ECHOES.**—R. W. Knecht, *National Bureau of Standards.*—During March, 1951 a series of visual auroral observations were made simultaneously with ionospheric soundings at Point Barrow, Alaska. Observations were made every 15 minutes during the dark hours of ten successive clear nights. Some 400 simultaneous observations were made. Auroras were present during about 90 per cent of these observations. Analysis indicates that quiet auroras, 45 degrees or higher above the horizon, were correlated with certain sporadic-E region echoes. The more intense the aurora, the higher the maximum frequency returned. The closer the auroral form was to the zenith, the shorter the range of the "E's" echoes. Active auroral displays were usually accompanied by a sharp increase in radio-wave absorption. Local geomagnetic variations coincident with auroral displays were also examined.

65. **THEORY OF RADIO SCATTERING FROM THE AURORA.**—R. K. Moore, *Sandia Corporation.*—It is postulated that radio signals returned from the aurora may be scattered by columns of ionization created by incoming auroral protons. The scattering from each column may be treated in the same manner as the "whistles" from meteor ionization, but because of the large number of columns created in a short time the "whistles" blend into a fading spectrum. Experimental determination of the fading spectra of such signals leads to curves which correlate well with those calculated by the theory. The velocities indicated agree in order of magnitude with that found by Gartlein's spectrographic observations. The calculated distances at which radar echoes should be observed agree with experiment and bear out the observation that signals are not heard or seen from overhead auroras. Density calculations cannot be complete because columnar size is not known, but indications are that this size may possibly be of the order of $1\frac{1}{2}$ meters in diameter.

66. **THE LENGTH OF IONIZED METEOR TRAILS.**—L. A. Manning, O. G. Villard, Jr., and A. M. Peterson, *Stanford University.*—By studying the rate of coincident detection of meteors at two widely separated locations, as a function of the differential time of first detection, it is possible to deduce the distribution in length of the meteoric ionization columns. A precise determination would require, in addition, that the velocity of the individual meteors used in the test be known. However, in the absence of such data, an approximate distribution of trail lengths can still be deduced, whose mean value is probably representative of actual conditions.

For equipment generating about a kilowatt of continuous wave output, and capable of detecting meteors at a total rate of 450 meteors per hour, the mean trail length was found to be 25 or 30 km. Calculations based upon this value of trail length, and upon the use of a corrected rate of meteoric arrival, show that meteors up to the 6th magnitude were detected during the test.

The method to be described provides a measure of trail length dependent upon the ability of the trail to reflect a signal at normal incidence. The accuracy of the determination of frequency of occurrence of trails of various lengths is independent of the rate of detection, but is proportional to the square root of the period of observation.

67. **GUIDED-WAVE CONCEPT IN ELECTROMAGNETIC THEORY.**—N. Marcuvitz, *Polytechnic Institute of Brooklyn.*—As is known, field representations in terms of guided waves, or modes, are characteristic both of the field operators and the geometrical symmetry of an electromagnetic region. The simplicity of such representations is manifest by the reformulation of a vector-field problem as a scalar transmission-line problem descriptive of the individual modal behavior. In relatively simple problems, involving say an arbitrarily oriented dipole antenna in a stratified region, the transmission-on-line problem is solved readily since there is no coupling between modes. In general diffraction problems intermode coupling exists, and the transmission line problem is solvable rigorously only in restricted cases. When the mode representation used is not rapidly convergent, to obtain the over-all field solution it is necessary to synthesize, for example, sum or integrate, the individual modal solutions. In free-space problems this may be effected by integration in the complex wave-number plane, a procedure which, under certain circumstances, may be avoided if an alternative and more convergent mode representation exists. To illustrate the above modal analysis and synthesis there will be treated (a) the field of a dipole antenna both in a planar and spherically stratified region (the general plane and spherical earth problem) and (b) the diffraction of a plane wave by two obliquely displaced semi-infinite planes in free space. The related questions of surface waves and alternative representations will be discussed in this connection.

68. **A FURTHER STUDY OF THE PATTERNS OF SINGLE SLOTS ON CIRCULAR CONDUCTING CYLINDERS.***—S. Sensiper, W. G. Sterns, and T. T. Taylor, *Hughes Aircraft Company.*—The azimuthal patterns of both axial and circumferential slots on circular conducting cylinders have been carefully calculated in both magnitude and phase, and some experimental checks have been obtained. The calculated patterns show that in the semicircle over which the slot is optically visible the magnitude and particularly the phase of the patterns are very similar to those of a similarly situated slot in an infinite ground plane. This conclusion has significant implications in the design of an antenna involving several slots on a cylinder. On the semicircle over which the slot is optically invisible and particularly near the midpoint of this range the pattern is very well represented by $E_\pi \cos v_1 (\pi-\phi)$ where E_π is the value of the pattern at $\phi=\pi$ (the point opposite the slot) and v_1 is complex. Thus the field in either one of the rear quadrants resembles the voltage on an open-circuited lossy transmission line.

The implications of the above-noted form of the field pattern behind the slot led to the consideration of an expression for the field which is quite different from the usual one originally employed. By an exact transformation of the usual expression it is possible to show that the far field is given by the expansion $\sum_{m=1}^{\infty} A_m \cos v_m (\pi-\phi)$, where v_m is complex. Near $\phi=\pi$, the first term of this series is dominant, and the results of this approach agree with those noted above. The procedure and its significance are quite closely related to the problem of electromagnetic wave propagation over a sphere which has been of considerable interest for some time. The various aspects of the cylindrical problem are discussed in some detail.

* Work described in this paper was carried out through the sponsorship of the Air Force Cambridge Research Laboratory under contract AF 19 (122)-454.

69. A SYNTHESIS METHOD FOR CIRCULAR AND CYLINDRICAL ANTENNAS COMPOSED OF DISCRETE ELEMENTS.*—T. T. Taylor, *Hughes Aircraft Company.*—Antennas composed of discrete elements equally spaced in angle around a circle or circular cylinder are studied with the objective of designing such antennas to produce required azimuthal radiation patterns.

Much has already been written upon this subject under the assumption that a continuous distribution of elementary sources will be an acceptable solution to the design problem or at least will form a step in the attainment of an acceptable solution. In the present writing, however, it is felt that something may be gained by analyzing the problem from the beginning on the basis of discrete elements. The question of how many elements are needed is discussed in detail, and it is shown that the envelope of the excitation coefficients is not necessarily equivalent to the continuous solution available by other methods. Practical procedures for finding the envelope of the excitation coefficients—and hence the coefficients themselves—are outlined.

* Work described in this paper was carried out through the sponsorship of the Air Force Cambridge Research Laboratory under contract AF 19 (122)-454.

70. THE GEOMETRICAL OPTICS FIELD AT A CAUSTIC.*—Irwin Kay, *New York University.*—The asymptotic expansion of a wave field in powers of $1/k$, where k is the wave number, for large k has as its lowest order term what is commonly known as the geometrical optics field. The caustics of geometrical optics are those point sets on which the zero order term becomes infinite. It is well known that caustics may exist even where the exact wave field is perfectly regular. An investigation of reflection from cylindrical walls of arbitrary cross section shows that the occurrence of caustic points means a change in character of the asymptotic expansion of the true field such that the lowest order term is no longer independent of k, but actually contains a factor k raised to a positive power. There also occurs a jump in phase along a ray passing through a caustic which, as is well known, equals $\pi/2$ in the case of a focal point, but which may differ from $\pi/2$ in the case of more general types of caustics. In addition the geometrical optics field is worked out in detail for the case of a plane wave incident on a parabolic cylinder, and the field is obtained in its lowest order at the focus and in the neighborhood of the focus.

* This work was supported by the Geophysics Research Division, Air Force Cambridge Research Center, under Contract No. AF-19 (122)-42.

71. INVESTIGATION OF A SURFACE-WAVE LINE FOR LONG-DISTANCE COMMUNICATION.—G. Goubau, C. Sharp, and S. W. Attwood, *Signal Corps Engineering Laboratories.*—The paper discusses results of an experimental study on a two-mile, single-conductor, surface-wave transmission line, for the frequency range of 100 to 300 mc. The objective of these measurements was to determine whether the theoretically expected low attenuation over a frequency band of more than 100 mc could be realized and to what extent field distortions, caused by supports and bends, were detrimental. Also the effect of weather conditions was investigated.

The results of the experiments clearly demonstrate that the surface-wave line has practical aspects for long-distance transmission.

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P. S. Veneklasen

7. ELECTRON DEVICES

G. D. O'Neill, *Chairman*
R. M. Ryder, *Vice-Chairman*

E. M. Boone — R. B. Janes
R. S. Burnap — I. E. Mouromtseff
J. W. Clark — L. S. Nergaard
W. J. Dodds — P. A. Redhead
George Espersen — H. J. Reich
C. E. Fay — A. C. Rockwood
W. R. Ferris — A. L. Samuel
M. S. Glass — W. G. Shepherd
T. J. Henry — R. W. Slinkman
J. W. Kearney — H. L. Thorson
G. R. Kilgore — C. M. Wheeler

SUBCOMMITTEES

7.2 Cathode-Ray and Television Tubes

R. B. Janes, *Chairman*
R. G. Stoudenheimer

7.3 Gas Tubes

H. L. Thorson, *Chairman*

J. H. Burnett — D. E. Marshall
W. A. Depp — Wm. Schmitt
E. J. Handley — W. W. Watrous
R. A. Herring — W. M. Webster
H. H. Wittenberg

7.4 Power-Output High-Vacuum Tubes

C. E. Fay, *Chairman*

J. W. Clark — I. E. Mouromtseff
H. D. Doolittle — E. E. Spitzer
T. A. Elder — Carol Veronda
G. A. Espersen — W. G. Wagener
J. W. Greer — C. M. Wheeler
A. K. Wing

7.4.1 Magnetron Working Group Under POHVT

G. A. Espersen, *Chairman*
R. R. Moats, *Secretary*

Norman Alstad — M. S. Glass
R. S. Briggs — G. R. Kilgore
H. W. Cole — E. C. Okress
T. P. Curtis — H. Ricker
S. Weston

7.6 Small High-Vacuum Tubes

T. J. Henry, *Chairman*

E. M. Boone — E. H. Hurlburt
R. C. Hergenrother — I. Meth
L. N. Heynick — W. T. Millis
M. E. Hines — R. W. Slinkman
W. G. Shepherd

7.6.1 Klystron Task Group

W. G. Shepherd, *Chairman*

E. M. Boone — A. E. Harrison
B. C. Gardner — T. J. Henry
C. T. Goddard — V. R. Learned
E. G. Linder

7.6.2 Task Group on Traveling-Wave Tubes

M. E. Hines, *Chairman*

H. D. Arnett — R. G. E. Hutter
W. J. Dodds — J. R. Pierce
L. M. Field — S. E. Webber

7.6.3 Task Group on Noise Under SSHVT

R. M. Ryder, *Chairman*

J. J. Freeman — D. O. North
W. A. Harris — E. K. Stodola

7.7 Solid-State Devices

R. M. Ryder, *Chairman*

J. R. Ambrose — R. W. Grantham
S. J. Angello — J. Kurshan
C. B. Brown — H. L. Owens
W. C. Dunlap — W. J. Pietenpol
J. R. Flegal — B. J. Rothlein

7.9 Conference on Semiconductor Device Research

R. M. Ryder, *Chairman*
J. P. Jordan, *Vice-Chairman, Secretary*

A. E. Anderson — Jerome Kurshan
W. L. Emery — Harry Owens

7.10 Conference on Electron Tube Research

W. J. Dodds, *Chairman*
P. A. Redhead, *Vice-Chairman*

C. C. Cutler — V. R. Learned
B. A. Kazan — S. D. Robertson

8. ELECTRONIC COMPUTERS

Robert Serrell, *Chairman*
D. R. Brown, *Vice-Chairman*

R. D. Elbourn — C. V. L. Smith
E. L. Harder — R. L. Snyder, Jr.
E. Lakatos — W. H. Ware
G. W. Patterson — C. R. Wayne
J. A. Rajchman — J. R. Weiner
Nathaniel Rochester — C. F. West
Way Dong Woo

SUBCOMMITTEE

8.3 Static Storage Elements

J. Rajchman, *Chairman*

A. O. Black — E. Gelband
H. Bonn — M. K. Haynes
H. R. Brownell — M. Littmann
T. C. Chen — W. N. Papian
E. A. Sands

8.4 Definitions (Eastern Division)

N. Rochester, *Chairman*

R. R. Everett — Linder Hobbs
G. W. Patterson

8.5 Definitions (Western Division)

W. H. Ware, *Chairman*

H. D. Husky — W. E. Smith
H. F. Larson — W. S. Speer
L. D. Stevens

9. FACSIMILE

R. J. Wise, *Chairman*
Henry Burkhard, *Vice-Chairman*

C. K. Clauer — F. A. Hester
A. G. Cooley — John V. Hogan
I. H. Franzel — Pierre Mertz
J. H. Hackenberg — C. J. Young

10. INDUSTRIAL ELECTRONICS

John Dalke, *Chairman*
Eugene Mittelmann, *Vice-Chairman*

G. P. Bosomworth — H. W. Parker
J. M. Cage — S. I. Rambo
E. W. Chapin — Walther Richter
J. E. Eiselein — W. C. Rudd
C. W. Frick — E. H. Schulz
G. W. Klingaman — C. F. Spitzer
H. R. Meahl — W. R. Thurston
J. N. Mennie — M. P. Vore
P. E. Ohmart — Julius Weinberger

SUBCOMMITTEES

10.1 Definitions for Induction and Dielectric Heating

C. F. Spitzer, *Chairman*

C. P. Bosomworth — J. L. Dalke
J. M. Cage — Eugene Mittelmann
Walther Richter

10.2 Dielectric Measurements

John Dalke, *Acting Chairman*

G. W. Klingaman — J. H. Mennie
H. R. Meahl — C. F. Spitzer

10.4 Methods of Measurement

Eugene Mittelmann, *Chairman*

J. M. Cage — S. I. Rambo
J. L. Dalke — E. H. Schulz
W. R. Thurston

11. INFORMATION THEORY AND MODULATION SYSTEMS

W. G. Tuller, *Chairman*
J. G. Kreer, Jr., *Vice-Chairman*

P. L. Bargellini — D. D. Grieg
W. R. Bennett — I. V. Harrington
L. A. DeRosa — H. Kohler

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E. R. Kretzmer
V. D. Landon
Nathan Marchand
L. A. Meacham
B. M. Oliver
Winslow Palmer
Dale Pollack
J. R. Ragazzini

12. NAVIGATION AIDS

P. C. Sandretto, *Chairman*
C. J. Hirsch, *Vice-Chairman*
H. R. Mimno, *Vice-Chairman*

W. B. Burgess
Harry Davis
R. E. Gray
C. J. Hirsch
H. I. Metz
K. A. Norton
Winslow Palmer
L. M. Sherer

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12.2 Standard DF Measurements

E. D. Blodgett, *Chairman*
J. Kaplan, *Vice-Chairman*
J. L. Allison, *Secretary*
R. Silberstein, *Alternate Secretary*

A. D. Bailey
H. I. Butler
L. M. Harding
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A. A. Kunze
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H. R. Mimno
W. M. Richardson
J. A. Solga
J. O. Spriggs
C. A. Strom, Jr.
S. R. Thrift
J. H. Trexler

12.3 Measurement Standards for Navigation Systems

Fred Moskowitz, *Chairman*
S. B. Fishdein, *Secretary*

Paul Adams
Paul Hansel
M. Leifer
George Litchford
J. T. MacLemore
Charles Marshburn
Irving Stokes
Vernon Weihe

14. PIEZOELECTRIC CRYSTALS

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W. P. Mason, *Vice-Chairman*

W. L. Bond
W. G. Cady
J. K. Clapp
W. A. Edson
Clifford Frondel
W. D. George
Edward Gerber
Hans Jaffe
S. Roberts
P. L. Smith
K. S. Van Dyke

15. RADIO TRANSMITTERS

M. R. Briggs, *Chairman*
A. E. Kerwien, *Vice-Chairman*

E. L. Adams
J. H. Battison
T. J. Boerner
H. R. Butler
L. K. Findley
Harold Goldberg
J. B. Heffelfinger
P. J. Herbst
J. B. Knox
L. A. Looney
J. F. McDonald
John Ruston
Berthold Sheffield
I. R. Weir

SUBCOMMITTEES

15.1 FM Transmitters

John Ruston, *Chairman*

John Bose
J. R. Boykin
Nathan Marchand
Perry Osborne
H. P. Thomas

15.2 Radio Telegraph Transmitters up to 50 MC

H. R. Butler, *Chairman*

J. L. Finch
J. F. McDonald
Berthold Sheffield
F. D. Webster
I. R. Weir

15.3 Double Sideband AM Transmitters

J. B. Heffelfinger, *Chairman*

E. L. Adams
R. E. Beetham
L. T. Findley
L. A. Looney

15.4 Pulse-Modulated Transmitters

Harold Goldberg, *Chairman*

Ross Bateman
L. L. Bonham
W. F. Cook
H. Kohler
G. F. Montgomery
W. E. Newlon
W. K. Roberts
B. D. Smith
W. G. Tuller
R. M. Whitehorn

15.5 Single-Sideband Radio Communication Transmitters

A. E. Kerwien, *Chairman*

M. R. Briggs
A. Brown
John Costas
H. E. Goldstine
Leonard Kahn
E. A. Laport
J. B. Singel

15.6 Television Broadcast Transmitters

P. J. Herbst, *Chairman*

F. J. Bias
Ervin Bradburd
T. M. Gluyas, Jr.
L. A. Looney
John Ruston

16. MOBILE COMMUNICATION SYSTEMS

F. T. Budelman, *Chairman*
Alexander Whitney, *Vice-Chairman*

G. M. Brown
A. V. Dasburg
D. B. Harris
C. M. Heiden
C. N. Kimball, Jr.
D. E. Noble
J. C. O'Brien
David Talley
George Teomney
R. W. Tuttle

17. RECEIVERS

Jack Avins, *Chairman*
J. D. Reid, *Vice-Chairman*

K. A. Chittick
R. De Cola
L. M. Harris
K. W. Jarvis
J. K. Johnson
W. R. Koch
I. J. Melman
C. R. Miner
Garrard Mountjoy
F. R. Norton
Leon Riebman
L. M. Rodgers
W. F. Sands
S. W. Seeley
R. F. Shea
S. C. Spielman
W. O. Swinyard
F. B. Uphoff
R. S. Yoder

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17.3 Single Sideband Receivers

K. Jarvis, *Chairman*
D. E. Norgaard

17.4 Spurious Radiation

R. F. Shea, *Chairman*

E. W. Chapin
K. A. Chittick
H. E. Dinger
W. R. Koch
B. J. Lawrence
T. S. Loeser
R. A. Maher
F. L. Pampel
D. P. Post
Frank Pounsett
John Rennick
L. M. Rodgers
E. H. Schrot
H. Schwarz
S. W. Seeley
C. G. Seright
W. O. Swinyard
H. J. Tyzzer
R. S. Yoder

17.5 Annual Review Material

L. M. Harris, *Chairman*

17.8 Task Group to Review I.E.C. Proposal on Standardization of Measurements of Receivers for AM Broadcast Transmissions

R. DeCola, *Chairman*

H. Adler
W. F. Cotter
J. K. Johnson
E. R. Pfaff
S. C. Spielman
W. O. Swinyard
R. Yoder

17.9 Automatic Frequency Control

F. B. Uphoff, *Chairman*

19. SOUND RECORDING AND REPRODUCING

A. W. Friend, *Chairman*
Lincoln Thompson, *Vice-Chairman*

S. J. Begun
M. S. Corrington
R. M. Fraser
E. W. Kellog
C. J. Lebel
A. P. G. Peterson
H. E. Roys
Harry Schecter
C. F. West
R. E. Zenner

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19.1 Magnetic Recording

A. P. G. Peterson, *Chairman*

W. H. Erikson
J. H. McGuigan
Ward Shepard, Jr.
R. E. Zenner

19.2 Mechanical Recording

Lincoln Thompson, *Chairman*

W. S. Bachman
Theodore Lindenberg
A. R. Morgan
R. C. Moyer
F. W. Roberts
M. S. Royston
Arthur S. R. Toby

19.3 Optical Recording

R. M. Fraser, *Chairman*

Price Fish
Everett Miller
Charles Townsend

21. SYMBOLS

A. G. Clavier, *Chairman*
K. E. Anspach, *Vice-Chairman*

W. J. Everts
W. A. Ford
R. T. Haviland
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C. D. Mitchell
C. Neitzert
A. F. Pomeroy
M. B. Reed
M. P. Robinson
M. S. Smith
H. R. Terhune
H. P. Westman

SUBCOMMITTEES

21.2 Graphical Symbols for Semiconductors

C. D. Mitchell, *Chairman*

L. F. Hescock
F. J. Lingel
R. M. Ryder
A. D. Shaw
E. H. Ulm

21.3 Task Group on Graphical Symbols for Functional Operations of Control, Computing and Switching Equipment

W. B. Callaway, *Chairman*

J. C. Bayles
C. W. Frank
J. D. Hood
A. Lesti
B. McMillan
F. J. Roehm
J. L. Wagner
R. M. Walker

22. TELEVISION SYSTEMS

R. E. Shelby, *Chairman*
R. M. Bowie, *Vice-Chairman*

W. F. Bailey
M. W. Baldwin, Jr.
A. H. Brolly
J. E. Brown
K. A. Chittick
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D. G. Fink
P. C. Goldmark
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J. L. Hollis
A. G. Jensen
I. J. Kaar
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Leonard Mautner
J. Minter
J. H. Mulligan, Jr.
A. F. Murray
J. A. Oiumet
D. W. Pugsley
Martin Silver
David Smith
M. E. Strieby
A. Talamini

SUBCOMMITTEES

22.1 Definitions

M. W. Baldwin, Jr., *Chairman*

A. V. Bedford
F. J. Bingley
J. W. Christensen
C. J. Hirsch

(*Subcommittee 22.1, cont.*)

D. L. McAdam
E. Seminski
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W. T. Wintringham

23. VIDEO TECHNIQUES

W. J. Poch, *Chairman*
A. J. Baracket, *Vice-Chairman*

K. B. Benson
J. L. Jones
Stephen Doba, Jr.
J. E. Keister
V. J. Duke
L. L. Lewis
G. L. Fredendall
C. G. Pierce
R. L. Garman
B. F. Tyson
R. E. Graham
J. F. Wiggin

SUBCOMMITTEES

23.2 Utilization, Including Video Recording: Methods of Measurement

R. L. Garman, *Chairman*

K. B. Benson
L. D. Grignon
J. M. Brumbaugh
Sidney Kraemer
V. J. Duke
H. Milholland
George Gordon
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23.3 Video Systems and Components: Methods of Measurement

A. J. Baracket, *Chairman*

I. C. Abrahams
G. L. Fredendall
R. Betts
J. R. Hefele
R. DeBaun
A. Lind
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23.4 Video Signal Transmission: Methods of Measurement

J. L. Jones, *Chairman*

K. B. Benson
R. M. Morris
S. Brand
Keith Mullenger
R. D. Chipp
R. S. O'Brien
Stephen Doba, Jr.
C. A. Rosencrans
L. R. Moffett
E. H. Schreiber
B. F. Tyson

24. WAVE PROPAGATION

Newbern Smith, *Chairman*
H. W. Wells, *Vice-Chairman*

E. W. Allen, Jr.
D. E. Kerr
S. L. Bailey
J. E. Keto
H. G. Booker
Morris Kline
C. R. Burrows
M. G. Morgan
T. J. Carroll
R. K. Moore
A. B. Crawford
K. A. Norton
A. E. Cullum, Jr.
H. O. Peterson
W. S. Duttera
George Sinclair
R. E. Fontana
R. L. Smith-Rose
E. G. Fubini
A. W. Straiton
I. H. Gerks
A. H. Waynick
M. C. Gray
J. W. Wright
T. W. Wright

SUBCOMMITTEES

24.1 Standards and Practices

H. O. Peterson, *Chairman*

W. S. Duttera
George Sinclair
F. M. Greene
E. F. Vandivere
J. W. Wright

24.2 Theory and Application of Tropospheric Propagation

Thomas Carroll, *Chairman*

H. G. Booker
Marion C. Gray
A. B. Crawford
A. W. Straiton
R. P. Wakeman

24.3 Theory and Application of Ionospheric Propagation

M. G. Morgan, *Chairman*

R. Bateman
R. A. Helliwell
K. W. Bowles
C. W. McLeish
J. H. Meek

24.4 Definitions and Publications

George Sinclair, *Chairman*

A. B. Crawford
I. H. Gerks
J. E. Eaton
Karl Jansky
E. G. Fubini
Martin Katzin
D. E. Kerr

24.5 Annual Review

H. W. Wells, *Chairman*

24.6 Radio Astronomy

C. R. Burrows, *Chairman*

A. E. Covington
F. T. Haddock, Jr.
J. Hagen

25. MEASUREMENTS AND INSTRUMENTATION

F. J. Gaffney, *Chairman*
P. S. Christaldi, *Vice-Chairman*

J. L. Dalke
G. A. Morton
Harold Dinger
C. D. Owens
G. L. Fredendall
A. P. G. Peterson
W. D. George
J. G. Reid, Jr.
W. J. Mayo-Wells
J. R. Steen
Ernst Weber

SUBCOMMITTEES

25.1 Basic Standards and Calibration Methods

W. D. George, *Chairman*

S. L. Bailey
G. L. Davies
F. J. Gaffney

25.2 Dielectric Measurements

John Dalke, *Chairman*

C. A. Bieling
Fred A. Muller

25.3 Magnetic Measurements

C. D. Owens, *Chairman*

E. J. Smith

25.4 Audio-Frequency Measurements

A. P. G. Peterson, *Chairman*

25.5 Video-Frequency Measurements

G. L. Fedendall, *Chairman*

Joseph Fisher
H. A. Samulon
Richard Palmer
W. R. Thurston

25.6 High-Frequency Measurements

E. I. Green, *Chairman*

R. A. Braden
G. B. Hoadley
I. G. Easton
E. W. Houghton
F. J. Gaffney
H. E. Webber

25.9 Measurements of Radio Activity

G. A. Morton, *Chairman*
P. W. Davison, *Acting Secretary*

R. L. Butenhoff
F. D. Lewis
D. L. Collins
A. Lovoff
L. F. Curtiss
M. A. Schultz
W. W. Schultz

25.10 Oscillography

P. S. Christaldi, *Chairman*

M. J. Ackerman
E. R. Haberland
R. R. Batcher
W. J. Jones
Edward Durbin
H. M. Joseph
Rudolf Feldt
A. L. Stillwell
Howard Vollum

25.11 Statistical Quality Control

J. R. Steen, *Chairman*

25.13 Telemetering

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R. E. Colander
F. Lehan
J. F. Brinster
E. E. Lynch
C. K. Duff
M. G. Pawley
A. P. Gruer
G. M. Thynell
R. L. Harding
F. L. Verwiebe
C. H. Hoeppner
G. F. C. Weedon
M. V. Kievert
G. E. White
W. A. Wildhack

25.14 Electronic Components

J. G. Reid, Jr., *Chairman*

G. B. Devey
A. W. Rodgers
J. C. P. Long
N. H. Taylor
J. H. Muncy
A. E. Zdobysz

26. SERVO-SYSTEMS

W. M. Pease, *Chairman*

W. K. Linvill
Paul Travers
E. A. Sabins
R. B. Wilcox
S. B. Williams

SUBCOMMITTEE

26.1 Terminology for Feedback Control Systems

J. C. Lozier, *Chairman*

Martin Cooperstein
L. H. O'Neill
J. O. Edson
Ralph Redemski
Thomas Flynn
C. F. Rehberg
L. Goldman
Felix Sweig

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*Akron, University of: P. C. Smith
*Alabama Polytechnic Institute: R. M. Steere
*Alberta, University of: J. W. Porteous
*Arizona, University of: H. E. Stewart
*Arkansas, University of: R. E. Lanier
*British Columbia, University of: L. R. Kersey
*Brooklyn, Polytechnic Institute of, (Day Division): J. A. Foecke
*Brooklyn, Polytechnic Institute of, (Evening Division): A. B. Giordano
*Brown University: Appointment later
*Bucknell, University of: B. H. Bueffel, Jr.
*California Institute of Technology: W. H. Pickering
*California State Polytechnic College: Clarence Radius
*California, University of: L. J. Black
California, University of at Los Angeles: E. F. King
Carleton College: G. R. Love
*Carnegie Institute of Technology: G. H. Royer
*Case Institute of Technology: J. D. Johannesen
Cincinnati, University of: A. B. Bereskin
*Clarkson College of Technology: Joseph Salerno
*Colorado, University of: P. W. Carlin
*Columbia University: J. R. Ragazzini
*Connecticut, University of: H. L. Heydt
*Cooper Union: J. B. Sherman
*Cornell University: True McLean
Dartmouth College: M. G. Morgan
*Dayton, University of: Appointment later
*Delaware, University of: H. S. Bueche
*Denver, University of: Herbert Reno

* Colleges with approved Student Branches.

*Detroit, University of: Thomas Yamauchi
*Drexel Institute of Technology: R. T. Zern
Duke University: H. A. Owen
Evansville College: J. F. Sears
*Fenn College: K. S. Sherman
*Florida, University of: S. P. Sashoff
*George Washington University: W. S. Carley
*Georgia Institute of Technology: M. A. Honnell
Harvard University: E. L. Chaffee
*Illinois Institute of Technology: G. F. Levy
*Illinois, University of: P. F. Schwarzlose
*Iowa, State University of: L. A. Ware
*Iowa State College: G. A. Richardson
*John Carroll University: J. L. Hunter
*Johns Hopkins University: Ferdinand Hamburger, Jr.
*Kansas State College: J. E. Wolfe
*Kansas, University of: D. G. Wilson
*Kentucky, University of: N. B. Allison
*Lafayette College: F. W. Smith
*Lehigh University: D. E. Mode
*Louisiana State University: L. V. McLean
*Louisville, University of: S. T. Fife
*Maine, University of: W. J. Creamer, Jr.
*Manhattan College: T. P. Canavan
Manitoba, University of: R. G. Anthes
*Marquette University: Appointment later
*Maryland, University of: G. F. Corcoran
*Massachusetts Institute of Technology: E. A. Guillemin, W. H. Radford
McGill University: F. S. Howes
*Miami, University of: F. B. Lucas
*Michigan College of Mining and Technology: R. J. Jones
*Michigan State College: I. O. Ebert
*Michigan, University of: L. N. Holland
*Minnesota, University of: Leroy Anderson
*Mississippi State College: P. T. Hutchison
*Missouri School of Mines and Metallurgy: R. E. Nolte
*Missouri, University of: G. V. Lago
*Montana State College: R. C. Seibel
*Nebraska, University of: Charles Rook
Nevada, University of: I. J. Sandorf
*Newark College of Engineering: D. W. Dickey
New Hampshire, University of: A. L. Winn
*New Mexico College of Agriculture and Mechanic Arts: H. A. Brown
*New Mexico, University of: A. D. Wooten
*New York, College of the City of: Harold Wolf
*New York University: Philip Greenstein
*North Carolina State College: G. B. Hoadley
*North Dakota, University of: Clifford Thomforde
*Northeastern University: G. E. Pihl
*Northwestern University: A. H. Wing, Jr.
*Notre Dame, University of: H. E. Ellithorn
*Ohio State University: G. E. Mueller
*Oklahoma Agricultural and Mechanical College: A. L. Betts
Oklahoma, University of: C. L. Farrar
*Oregon State College: A. L. Albert
*Pennsylvania State College: C. R. Ammerman
*Pennsylvania, University of: D. F. Hunt
*Pittsburgh, University of: John Brinda, Jr.
*Pratt Institute: David Vitrogan
*Princeton University: N. W. Mather
*Purdue University: R. P. Siskind
Queens University: H. H. Stewart
*Rensselaer Polytechnic Institute: H. D. Harris
*Rhode Island, University of: R. S. Haas
Rice Institute: C. R. Wischmeyer
Rose Polytechnic Institute: H. A. Moench
*Rutgers University: J. L. Potter
*St. Louis University: G. E. Dreifke
*San Diego State College: D. C. Kalbfell
*San Jose State College: Harry Engwicht
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*Seattle University: Appointment later
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Union College (Nebr.): M. D. Hare
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*Utah State Agricultural College: Clayton Clark
*Utah, University of: M. E. Van Valkenburg
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*Virginia Polytechnic Institute: R. R. Wright
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The Annual Index to these Abstracts and References, covering those published in the Proc. I.R.E. from February, 1951, through January, 1952, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

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ACOUSTICS AND AUDIO FREQUENCIES

534.231+621.396.671 **1179**

The Representation of the Radiation Field of Two Radiators by means of Constant-Phase and Constant-Amplitude Curves—Stenzel. (See 1179.)

534.231:534.121.1 **1180**

Effects of a Finite Circular Baffle Board on Acoustic Radiation—T. Nimura and Y. Watanabe. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 79–93; 1950.) Analysis of the field of a circular disk vibrating in a finite concentric circular baffle indicates that a baffle radius of about half the wavelength of the lowest frequency to be used gives a power increase almost equal to that for an infinite baffle. Experimental and theoretical results are in good agreement.

534.26 **1181**

Diffraction of Sound Waves at an Uneven Surface—L. M. Brekhovskikh. (*Compt. Rend. Acad. Sci.* (URSS), vol. 79, pp. 585–588; August 1, 1951. In Russian).

534.321.9:534.232 **1182**

Ultrasonic Generators and their Applications—M. Alixant. (*Radio Tech. Dig.* (France), vol. 5, pp. 271–278 and 299–325; 1951.) A survey paper with 165 references.

534.321.9:538.652 **1183**

Factors in the Generation of Ultrasonic Oscillations by means of Ferromagnetic Sediments—H. H. Rust and P. Pilz. (*Z. angew. Phys.*, vol. 3, pp. 379–382; October, 1951.) Volume-magnetostriction measurements on a series of powdered materials are shown graphically and discussed. A hollow sphere of material with length-magnetostriction properties gives large volume variations for a relatively small applied magnetic field.

534.7 **1184**

Neuropsychological Relations between Phonation and Hearing—R. Husson. (*Ann. Télécommun.*, vol. 6, pp. 273–277; October, 1951.)

534.7 **1185**

[Audible] Objective and Subjective Resonance Effects with Short Trains of Pulses—O. Sala. (*Frequenz*, vol. 5, pp. 250–258; September, 1951.)

534.76 **1186**

Auditory Perspective—A Study of the Biological Factors related to Directional Hearing—H. G. Kobrak. (*Jour. Soc. Mot. Pic. Telev. Eng.*, vol. 57, pp. 328–335; October, 1951.) The physiological aspects of stereophonic hearing are discussed. Properties of the sound signal significant for location are examined, and the role of the central nervous system in integrating the binaural stimuli is described. The influence of experience and training is stressed.

534.78 **1187**

Audio Problems in Aircraft Communication—I. H. Bowker. (*Tele-Tech*, vol. 10, pp. 41–43, 90; October, 1951.) High external noise levels and large variations of air pressure are the two major factors in reducing speech intelligibility in aircraft. The effect of altitude on speech production, on hearing, and on microphone and earphone performance is discussed and the design of special types of equipment to minimize noise effects and to increase speech intelligibility is considered.

534.84 **1188**

Notes on some Problems of Room Acoustics—S. Sawade. (*Elektrotech. Z.*, vol. 71, pp. 245–246; May 5, 1950.) Introduction of the concept of "reverberation radius," i.e., the distance from an omnidirectional source at which the pressures due to direct and reflected sound are equal, enables simple rules to be applied to the arrangement of loudspeakers and microphones, in particular for avoiding acoustic feedback; it also facilitates understanding of the significance of the experimentally found optimum reverberation time.

534.844.1:621.396.615.11 **1189**

Equipment for Acoustic Measurements: Part 2—A Portable Tone Source Developed for Use in Room Acoustics—Mayo and Beadle. (See 1382.)

534.846 **1190**

Influence of Chandeliers on the Acoustics of Theatres and Concert Halls—L. Villard. (*Bull. Tech. suisse romande*, vol. 77, pp. 243–245; September 8, 1951.) The subject is discussed in connection with the rebuilding of the Grand Theatre at Geneva. Many halls with good acoustics have in the past been illuminated by chandeliers, which provide a substantial amount of nonselective sound absorption and also act as sound diffusers; their replacement by modern lighting fittings may have an adverse effect.

621.317:621.395.813 **1191**

A New Evaluation of the Transmission Quality of a Telephone System—G. Fontanellaz. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 29, pp. 384–390; October 1, 1951. In German and French.) The C.C.I.F. has introduced a new criterion, the "equivalent attenuation for intelligibility." This is explained, and subjective tests are described for determining its value. The old criterion, "reference equivalent" which took account only of loudness, is shown to be inadequate.

621.395.625.3 **1192**

A New Method for Measurement of the Speed of Magnetic Tapes—P. H. Werner. (*Tech. Mitt. scheiz. Telegr.-Teleph. Verw.*, vol. 29, pp. 390–392; October 1, 1951. In French.)

621.395.92.001.4 **1193**

Testing of Components of Hearing Aids—F. Müller. (*Funk u. Ton*, vol. 5, pp. 466–473; September, 1951.) Short description of methods of testing microphones, amplifiers, earpieces and bone-conduction devices. From these measurements the over-all characteristics of a hearing aid can be determined. See also 904 of May.

ANTENNAS AND TRANSMISSION LINES

621.3.018.78†:621.315.212 **1194**

Distortion of a Signal Transmitted by a Perfectly Homogeneous Coaxial Line—R. Cazenave. (*Câbles & Trans.* (Paris), vol. 5, pp. 279–314; October, 1951.) For frequencies up to about 10 kmcps the telegraphy equation represents coaxial-line propagation sufficiently closely, provided the coefficients are treated as functions of frequency. By considering separately the front and tail end of the received signal, corresponding respectively to the highest and lowest frequencies transmitted, an approximate solution is obtained, the total signal duration and the front-end curves for unit-pulse, unit-step and rectangular-wave signals being determined in the absence of and for small values of the leakage conductance. A numerical example illustrates the practical application of the results; a comparison of the performance of 2.6/9.4 mm and 5/18 mm coaxial cables operating over long distances clearly shows the superiority of the latter.

621.392 1195
Survey of Radio-Frequency Transmission Lines and Wave Guides—E. S. Winlund. (*Proc. Radio Club Amer.*, vol. 28, pp. 1–64; 1951. Editor's remarks, p. 65.) Material published during the period 1919–1936 is surveyed; the majority of the 684 references given deal with later publications, up to 1951.

621.392.26 1196
The Relative Power-Carrying Capacity of High-Frequency Waveguides—H. M. Barlow. (*Proc. IEE*, Part III, vol. 99, pp. 21–27; January, 1952.) An assessment is made of this capacity, within the limits of electrical breakdown, for the wave modes commonly used inside waveguides of rectangular or circular cross section, or outside a solid wire. This shows that for hollow guides the *H* modes are inferior to the *E* modes and both compare unfavorably with the TEM wave of a parallel-strip transmission line and with the waves supported by the solid-wire surface guide. The latter is particularly suitable for millimeter wavelengths.

621.392.26 1197
Diaphragms in Waveguides—H. H. Meinke. (*Fernmeldetech. Z.* vol. 4, pp. 431–435; October, 1951.) The application and exact quantitative treatment of basic types of diaphragm are demonstrated by referring their measured characteristics to their equivalent electrical circuits.

621.392.26:621.317.329 1198
Determination of Aperture Parameters by Electrolytic-Tank Measurements—S. B. Cohn. (PROC. I.R.E. vol. 40, p. 33, January, 1952.) Correction to paper noted in 725 of April.

621.392.26:621.396.67 1199
Radiation from an Aerial located inside a Rectangular Waveguide—M. Jessel. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 783–785; October 8, 1951.) The field radiated in this case is found from that radiated by the same antenna in free space by considering the radiation from the electrical images of the antenna in the walls of the waveguide. Radiation from slots can be calculated similarly.

621.392.5:681.142 1200
Magnetostriction Storage Systems for a High-Speed Digital Computer—R. Millership, R. C. Robbins and A. E. De Barr. (*Brit. Jour. Appl. Phys.*, vol. 2, p. 304; October, 1951.) A 60-μs delay line of the type described in 2107 of 1951 (Bradburd) consists of a thin Ni or Ni-Fe tape about 30 cm long, terminated to suppress reflections and threaded through a transmitting coil near one end and a receiving coil near the other. The advantages of this type of delay line over the mercury type when used in a storage system are: (*a*) greater simplicity and ruggedness; (*b*) smaller insertion loss; (*c*) availability of output at any point along the delay line.

621.396.67+534.231 1201
The Representation of the Radiation Field of Two Radiators by means of Constant-Phase and Constant-Amplitude Curves—H. Stenzel. (*Arch. elekt. Übertragung*, vol. 5, pp. 447–454 and 517–526; October and November, 1951.) The radiation process is represented by two vectors whose locus diagrams are hyperbolic spirals, position on which corresponds to distance from the respective radiator. By displacing the resultant vector and taking the readings on the graduated spirals the required curves can be drawn directly, without interpolation. The points of special significance on the curves are (*a*) the zero-amplitude points, (*b*) the phase crossover points and (*c*) the amplitude crossover points. The corresponding coordinates are given respectively by (*a*) the points of intersection of the spirals, (*b*) the points of contact of the common tangents and (*c*) the feet of the common normals. A set of equations is given for finding these points, and values are tabulated.

621.396.67 1202
A Dipole with a Tuned Parasitic Radiator—R. King. (*Proc. IEE*, Part III, vol. 99, pp. 6–14; January, 1952.) A more accurate determination is made of the electric field, the front/back ratio and the input impedance of an antenna with a single parasitic radiator, using recently derived approximate second-order self and mutual impedances. Theoretical curves for the field and front/back ratio are compared with the experimental data of other investigators, the apparent disagreement being explained by failure to coordinate correctly the measured and computed quantities. A method is outlined in which the theoretical requirements are closely approached in the experimental arrangement; good agreement is obtained between theory and experiment for the electric field and front/back ratio.

621.396.67 1203
On the Theory of Antennae with Discontinuous Thickness—S. Uda and Y. Mushiake. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 105–116; 1950.) Hallén's theory is extended to antennas with thickness discontinuities, such as a telescopic dipole, and approximate formulas for input impedance are derived. Experimental results showed qualitative agreement with theory.

621.396.67 1204
An Experimental Investigation of the Dielectric-Rod Antenna of Circular Cross Section excited in the Dominant Mode—C. W. Horton and C. M. McKinney, Jr. (*Jour. Appl. Phys.*, vol. 22, pp. 1246–1249; October, 1951.) The radiation pattern and gain were measured at 9.275 kmcps for four series of dielectric antennas of circular cross section, constructed of polystyrene or lucite. Only one parameter in the rod geometry was varied in each series of measurements. The optimum working conditions for dielectric-rod antennas are deduced.

621.396.67 1205
A Broadside Dielectric Antenna—G. E. Mueller. (PROC. I.R.E., vol. 40, pp. 71–75; January, 1952.) An antenna is described which uses the properties of a nonuniform dielectric transmission line to produce a broadside directive pattern. Simple array theory is developed for prediction of the main features of the radiation patterns.

621.396.67:621.392.43 1206
Automatic Impedance Matcher—True. (See 1224.)

621.396.67:621.396.619.13 1207
High-Power F.M. Antenna Design—M. B. Sleeper. (*FM-TV*, vol. 11, pp. 11–12; October, 1951.) Description of the modified doughnut-type antenna and feeder system at station WMIT, on Clingman's Peak, N.C. Adjustable stubs replaced the flat plates previously used for tuning. Heating elements were added to prevent icing.

621.396.67:621.397.6 1208
Television Broadcast Antennas—J. E. Keister. (*Gen. Elec. Rev.*, vol. 54, pp. 19–22; October, 1951.) A review of the transmitting antenna systems at present in use for television broadcasting. Omnidirectional horizontal radiation and impedance matching over a wide frequency range are the general requirements. The sidefire helical antenna is a convenient radiator at uhf.

621.396.67:621.397.6 1209
Television in Buenos Aires—(*Rev. telegr. Electronica* (Buenos Aires), no. 468, pp. 623, 636; September, 1951.) An account of the erection of the antenna system, which consists of eight symmetrical elements mounted one above the other at the top of a 50-m lattice mast, each element comprising three horizontal folded dipoles.

621.396.67.012.71:517.512.2:621.39.001.11 1210
Fourier Analysis and Negative Frequencies—Shaw. (See 1422.)

621.396.677 1211
A Method for Calculating the Current Distribution of Tschebyscheff Arrays—D. Barbiere. (PROC. I.R.E., vol. 40, pp. 78–82; January, 1952.) The computation of the expressions for the current elements is reduced by a simplification, with no loss of exactness, of Dolph's equations for the optimum current distribution for a linear equispaced broadside array (2487 of 1946 and 2685 of 1947). A computation table for a 24-element array is given.

621.396.677 1212
Directional Antenna Arrays of Elements Circularly disposed about a Cylinder Reflector—R. F. Harrington and W. R. Le Page. (PROC. I.R.E., vol. 40, pp. 83–86; January, 1952.) 1951 I.R.E. National Convention paper. "The general solution for the field pattern of a circular array is adapted to include the effect of a concentric reflecting cylinder. Two solutions are presented, one giving the field as a Fourier series, and the other as an infinite series of Bessel functions. The results are general, being applicable to any array dimensions and for arbitrary distribution of excitation. The solution is idealized to the extent of assuming a continuous current sheet, rather than discrete elements, and an infinitely long cylindrical reflector."

621.396.677.5 1213
Calculation of the Radiation Distribution for a Rhombic Aerial with Arbitrary Termination Impedance—E. G. Hoffmann. (*Funk u. Ton*, vol. 5, pp. 518–525; October, 1951.) Calculation of the current distribution is based on that for a lossless two-wire line. The vector potentials in each arm of the antenna are calculated, and the resultant distant-field strength deduced as a function of reflection coefficient. This coefficient may be equal to or less than unity and can be chosen to give two equal side lobes of minimum amplitude. This is illustrated by a numerical example.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.7:621.387.4 1214
[Pulse] Amplitude Selectors—R. Wahl. (*Jour. Phys. Radium*, vol. 12, pp. 67A–74A; October, 1951.) Selectors for one-channel, four-channel or ten-channel operation are obtained by using two, five or eleven identical selector units. Operation is independent of the shape, duration ($>5\times10^{-8}$ sec) and time separation (>12 μs) of pulses. Channel widths and threshold levels are adjustable.

621.3.018.78†:621.396.619.16:621.392.52 1215
Pulse Distortion arising in F.M. Pulse Transmissions—P. A. Mann. (*Telefunken Ztg.* vol. 24, pp. 140–142; October, 1951.) The Dirac build-up function for an *n*-stage band-pass filter with critical coupling is expressed in terms of a Bessel function of order $(n-1/2)$. From this the distortion of a unit

pulse in passage through the filter can be evaluated. A 15-stage filter is taken as a numerical example, and graphs show that the time to reach the first maximum is practically independent of the frequency swing or of the phase linearity of the filter, but the tendency to overshoot is greater the greater the swing or filter nonlinearity.

621.314.2.012.3 1216
Transformer Iron Losses—N. H. Crowhurst. (*Electronic Eng.*, vol. 23, pp. 396–403; October, 1951.) Four charts are presented for application in af transformer design. They are based on analysis of samples of ordinary-grade transformer iron and mumetal, and simplify the determination of all necessary design parameters.

621.314.6.011.1 1217
The Theory of the Linear R.M.S.-Value Rectifier—O. Schmid. (*Arch. elekt. Übertragung*, vol. 5, pp. 459–463; October, 1951.) The operation of the circuit described by Boucke (687 of 1951) is analyzed for inputs of any wave form. For alternating voltages that can be represented by algebraic functions a closed expression can be derived for the ratio of rectified voltage to peak voltage which is a function of the ratio of the charge and discharge time-constants. For inputs corresponding to transcendental functions the ratio rectified-voltage/peak-voltage is found by a graphical method. Curves are plotted from which the best time-constants ratio and the wave form errors of first and second type can be determined. See also 339 of March.

621.314.7:546.289 1218
Transistor Circuit Design—G. Raisbeck. (*Electronics*, vol. 24, pp. 128–132, 134; December, 1951.) An explanation of the application of duality technique to the design of transistor amplifier, oscillator, modulator and multivibrator circuits, starting from the corresponding tube circuits.

621.316.86 1219
Nonlinear Resistors with Sintered-Semiconductor Base—N'Guyen Thien-Chi and J. Suchet. (*Ann. Radioélect.*, vol. 6, pp. 291–298; October, 1951.) The resistors are made of a material with a sintered-carborundum base, have a low temperature coefficient of resistance and a voltage/current relation of the form $V = kI^{0.2}$. Two types are available, type D (0.8-w rating) for operation at a few tens of volts and 1 ma current, and type H (15-w rating) at similar voltages and 100 ma current. They may be applied in the protection against breakdown of circuits containing mainly inductance, as shunts or multipliers in electrical instruments, and for voltage regulation.

621.318.435:621.3.015.3 1220
Transient Eddy-Current Phenomena of Saturable-Reactor Core—T. Kikuchi. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 94–104; 1950.)

621.318.572:621.385.8].001.8 1221
Electron-Beam Switches—F. Schröter. (*Telefunken Ztg.*, vol. 24, pp. 171–186; October, 1951.) A well illustrated review of various types and their applications.

621.319.45 1222
Capacitors with Ceramic Dielectric: Performance and Operating Characteristics—A. Danzin. (*Onde élect.*, vol. 31, pp. 342–356 and 406–412; August/September and October, 1951.) Detailed review based on official tests of capacitors for commercial and industrial use. Two groups are distinguished: (*a*) stable, with normal permittivity; (*b*) ferroelectric. The forms of construction adopted for various applications, particularly in receivers and transmitters, are described with illustrations.

621.319.4 1223
Tantalytic Capacitors—L. W. Foster. (*Gen. Elec. Rev.*, vol. 54, pp. 30–38; October, 1951.) Description of electrolytic capacitors using tantalum-foil electrodes. A capacitance of approximately 25 μF/in.^{3} at 150-v dc rating is obtained. The Ta_2O_5 insulating film can operate at an electrical stress of over 19,000 v/mil. Tantalytic capacitors are considerably smaller than corresponding ones of other types.

621.392.43:621.396.67 1224
Automatic Impedance Matcher—V. True. (*Electronics*, vol. 24, pp. 98–102; December, 1951.) Description of equipment for matching a 35-foot whip antenna to a 50-Ω coaxial feeder over the frequency range 2–18 mcps. Matching is performed by a cantilever network in which the capacitance of one branch of the network is adjusted by a servomechanism controlled by a circuit which determines the phase angle between feeder current and voltage; the capacitance of the other branch of the network is similarly controlled by a circuit which measures the total load impedance of the feeder. The swr in no case exceeds 1.25. With minor circuit changes, the equipment can be adapted for different frequency ranges, power levels, types of load, and feeder characteristic impedances.

621.392.5 1225
General Input-Output Relations for Linear Networks—L. A. Zadeh. (PROC. I.R.E., vol. 40, p. 103; January, 1952.) Outline of a method of analysis based on resolution of signals into a set of elementary components by means of a certain relation, which replaces that used for the resolution of signals into exponential components when using the Laplace or Fourier transform technique.

621.392.5 1226
Some General Theorems for Non-Linear Systems possessing Resistance—W. Millar. (*Phil. Mag.*, vol. 42, pp. 1150–1160; October, 1951.) "In the case of a resistive network, the dissipation is divided into two parts—the 'content' and 'co-content'—which are duals of each other. The dissipation itself has stationary properties in linear but not (in general) in nonlinear networks, but it can be shown that the 'content' and 'co-content' have stationary and additive properties in the nonlinear case. The idea of 'content' is extended to reactive systems, and it is shown that the total content of any system in motion is an invariant." See also 1227 below.

621.392.5 1227
Some General Theorems for Non-Linear Systems possessing Reactance—C. Cherry. (*Phil. Mag.*, vol. 42, pp. 1161–1177; October, 1951.) "The quantity called the co-energy (the dual of energy) is shown to possess stationary properties (maximum or minimum) and superposition properties; this is sufficient to establish the concept of an 'equivalent element' for any 2-terminal system of like elements (all-inductor, all-capacitor, all springs, etc.). The unfamiliar 'rectangle representation' of a circuit of linear resistors is explained and extended to the nonlinear case, including reactive elements. It is shown that the equations of motion of a nonlinear system, possessing reactance, may be expressed in Lagrangian form, thus emphasizing the importance of co-energy and also showing that the Principle of Duality is applicable. Finally, systems are considered possessing mutual inductance and moving magnetic circuits (as in rotating machines)." See also 1226 above.

621.392.5 1228
Conditions of Validity of Matrix Analysis for Quadripole Assemblies: Applications to Feed-back Networks—A. Kaufmann. (*Onde élect.*, vol. 31, pp. 396–405 and 446–452; October and November, 1951.) The basic method of interconnecting two quadripoles so that they may be represented by a single matrix, and the representation of any 4-terminal network so that two matrices of order 2 are substituted for a matrix of order 3, are outlined. Impedance relations are derived which must be satisfied for the matrix calculation to be valid in the case of (*a*) parallel-parallel, (*b*) series-series, (*c*) series-parallel, and (*d*) parallel-series connection of input and output of two quadripoles. Particular examples of each case are noted. Different networks and their corresponding matrix relations are shown in a series of tables. The method is particularly useful for studying feedback networks; this is illustrated by examples.

621.392.5:519.241.1 1229
Note on "Correlation Functions and Power Spectra in Variable Networks"—B. D. Steinberg. (PROC. I.R.E., vol. 40, p. 103; January, 1952.) Extension of Zadeh's work (586 of 1951) to derive the correlation function of a system function in which the variables are separable. In this case the correlation function is the product of two others, relating respectively to the time-dependent and frequency-dependent parts of the system function.

621.392.5:517.755 1230
Hilbert Transforms and Bayard-Bode Relations—F. H. Raymond. (*Ann. Télécommun.*, vol. 6, pp. 262–272; October, 1951.)

621.392.5:546.431.824–31 1231
Barium Titanate Delay Lines—L. M Orman and L. G. Callahan, (*Electronics*, vol 24, pp. 224, 248; December, 1951.) Mechanical vibration waves may be excited in BaTiO ceramic by the application of an electric field, provided that the material between the electrodes has been previously polarized. The preparation of thin sheets of the substance for use in delay lines is outlined. Undesired reflections from the ends of the sheet may be damped out by coating the sheet with clay or paraffin wax.

621.392.5 1232
Electronic Delay System for Flash-Bulb Release—J. P. Œhmichen. (*Toute la Radio*, no. 156, pp. 151–153; June, 1951.) The flash bulb is operated by a thyratron energized from a cathode-coupled double-triode "flip-flop" circuit, the time constant of which is adjusted by a potentiometer with scale calibrated in seconds or microseconds.

621.392.5.029.64 1233
The Ferromagnetic Faraday Effect at Microwave Frequencies and its Applications—The Microwave Gyrator—C. L. Hogan. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 1–31; January, 1952.) A new type of gyrator [see 301 of 1951 (Tellegen and Klaus)] dependent on the Faraday rotation of the plane of polarization of an em wave has been developed. Analysis of wave propagation through a ferromagnetic material with dielectric and magnetic loss gives a formula for the Faraday rotation which indicates that in materials such as ferrites large rotations are to be expected and should be independent of frequency. Results of measurements on a MnZn ferrite, which gave a rotation of about 120°/cm, were in very good agreement with theory. Gyrators constitute low-loss wide-band devices with many possible applications, including one-way transmission systems, microwave circulators, microwave switches, electrically controlled variable attenuators, and modulators.

621.392.52:621.314.2 1234
The Transformer Properties of Choke Filters—W. Taeger. (*Funk u. Ton*, vol. 5, pp. 526–535; October, 1951.) Certain low-pass

arrangements, in particular choke filters, function near the resonance point very much like transformers. The ratio M/R where M is the ratio of output voltage to input current on open circuit and R the actual load resistance) represents the transformation ratio u of the general quadripole and nearly corresponds to the mutual inductance of a transformer on open circuit. The ratio of output to input voltage of the choke filter, assuming losses are small, is u, the ratio of output to input current being $1/u$, but the phase difference between output and input voltages is only 90°. Doherty's modulation circuit makes use of the transformer properties of choke filters.

621.392.52:621.396.611.21 1235
The Maximum Bandwidth of Narrow-Band Quartz-Lattice Filters—W. Rave. (*Arch. elekt. Übertragung*, vol. 5, pp. 455–458; October, 1951.) In filters comprising capacitors and crystals only, the highest attainable bandwidth depends almost entirely on the properties of the crystals. Numerical calculations are made for X-cut (frequency range 50–300 kcps). AT-cut (frequency range 300 kcps–6 mcps) and BT-cut crystals (frequencies over 6 mcps). The highest value of relative bandwidth, obtained at 50 kcps, is 0.4 per cent. Between about 800 kcps and 6 mcps the bandwidth is <0.25 per cent; above 6 mcps and at 300 kcps the value is <0.125 per cent. By including inductances in all the lattice arms the bandwidth can be increased to about 10 per cent.

621.392.52.012.3 1236
Calculation Aids and Simple Formulae for the Approximate Determination of the Parameters of Two-Stage Band-Pass Filters—E. William. (*Funk u. Ton*, vol. 5, pp. 545–554; October, 1951.) Tables and charts, developed from approximation equations, are presented from which the required data can be read off to within about 1 per cent.

621.392.54† 1237
U.H.F. Oscillator Attenuator—F. Reggia. (*FM-TV*, vol. 11, pp. 16–17, 23; October, 1951.) The construction, performance and applications are described of a linear attenuator in which an external magnetic field is used to alter the loss characteristics of a microwave-energy-dissipating material in a transmission line. An attenuator using similar principles has been described by Miller (337 of 1950).

621.395.645:621.395.97 1238
The Broadcasting-Network Amplifier Rack, Type 48, of the German Post Office—E. A. Pavel, H. v. Schau and W. Schwenn. (*Fernmeldetech. Z.*, vol. 4, pp. 452–457; October, 1951.) Description of the high-fidelity line equipment for program transmission on four incoming or outgoing lines.

621.396.611.21 1239
Quartz Crystal Vibrators as Circuit Elements—H. E. Pearson. (*P.O. Elec. Eng. Jour.*, vol. 44, part 3, pp. 124–126; October, 1951.) Factors to be considered in the manufacture of quartz vibrators are discussed and values are given for the equivalent-circuit and other important parameters of types made by the British Post Office for operation in the range 1 kcps–40 mcps.

621.396.611.3 1240
An Analysis of Triple-Tuned Coupled Circuits—N. W. Mather. (PROC. I.R.E., vol. 40, p. 82; January, 1952.) Correction to paper abstracted in 2745 of 1950.

621.396.611.3.011.21 1241
Input-Admittance Characteristics of a Tuned Coupled Circuit—R. A. Martin and R. D. Teasdale. (PROC. I.R.E., vol. 40, pp. 57–61; January, 1952.) A steady-state analysis for a high-Q primary and low-Q secondary, with emphasis on the phase variation of admittance with frequency. Dimensionless curves of magnitude and phase illustrate the effect of varying the coupling and Q of the secondary.

621.396.611.4 1242
Representation of the Complete System of Natural Oscillations of Cylindrical Cavity Resonators with Horizontally Stratified Dielectric—E. Ledinegg and P. Urban. (*Acta Phys. austriaca*, vol. 3, pp. 320–341; March, 1950.) Results previously obtained by Ledinegg (4 of 1943) for cavities with homogeneous dielectric are shown to be valid also when the dielectric is stratified parallel to the base of the cylinder. Explicit expressions are derived for the possible field distributions. The calculation is made first for a "smooth" layer structure and is extended to incompletely smooth structures. Resonators of this type are of interest in measurement technique.

621.396.611.4:538.566 1243
Effect of Deformation of a Cylindrical Cavity Resonator on the Wave Numbers of the E_{010} and E_{011} Modes—R. Müller. (*Z. Naturf.*, vol. 5a, pp. 332–334; June, 1950.) The magnitude of the effect is calculated and its significance in relation to Essen and Gordon-Smith's measurements of the propagation velocity of em waves (3488 of 1948) is discussed.

621.396.615.17 1244
Cathode-Coupled Pulse Generator—F. A. Benson and G. V. G. Lusher. (*Wireless Eng.*, vol. 29, pp. 12–14; January, 1952.) Positive pulses of amplitude about 30 v, duration about 1 μs and rise time slightly less than 0.5 μs are derived from a square-wave input by means of a flip-flop arrangement incorporating a highly damped oscillatory circuit.

621.396.645 1245
Review of British Amplifiers: Part 1—General Constructional and Design Practices in the Better Amplifiers. Discussion of the Acoustical Amplifier—J. Moir. (*FM-TV*, vol. 11, pp. 30–32, 40; October, 1951.)

621.396.645 1246
New Miniature Intermediate-Frequency Amplifier—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 143–145; October, 1951.) Description of the National Bureau of Standards Model VI, a 7-tube amplifier for the frequency range 20–100 mcps; construction is simplified by providing separate subassemblies for (*a*) all the inductors, (*b*) all the capacitors and (*c*) all the tube shields.

621.396.645 1247
The Series Amplifier—E. L. Crosby, Jr. (*Radio and Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 12–13, 30; October, 1951.) Various amplifier units can be much reduced in size by the connection of tubes in series. The anode of the first tube is connected through its load to the cathode of the second, and so on. Coupling capacitors and also the decoupling elements can then be omitted.

621.396.645.018.424:621.317.792 1248
Wideband Pre-amplifier—F. Horner. (*Wireless Eng.*, vol. 29, pp. 19–26; January, 1952.) The amplifier was developed to increase the sensitivity of the aural-comparison method described by Thomas (169 of 1951). It is located at the antenna, away from the rest of the equipment, and operates unattended. Intermodulation between received signals is reduced by restricting the response to the required frequency band of 2.5–20 mcps by careful design of the final cathode-follower stage, and by making the voltage gain no greater than is necessary to achieve the desired sensitivity. The sensitivity is such that with a receiver of 10 kcps bandwidth, and in the absence of atmospherics, a cw signal with a field strength of about 0.05 μv/m is intelligible in the presence of set noise only.

621.396.645.029.3 1249
A Tunable Shunt Selector-Rejector for Audio Amplifiers—O. G. Villard, Jr. (*Rev. Sci. Instr.*, vol. 22, pp. 726–729; October, 1951.) Arrangements are described for adjusting the frequency response of an af amplifier for experimental or other temporary purposes. An auxiliary tube with an appropriate RC feedback loop is shunted externally across the last voltage-amplifying tube of the amplifier. Design theory is discussed and illustrated by an example giving component values and measured performance.

621.396.645.029.3 1250
A Single-Ended Push-Pull Audio Amplifier—A. Peterson and D. B. Sinclair. (PROC. I.R.E., vol. 40, pp. 7–11; January, 1952.) 1951 I.R.E. National Convention paper. Describes a circuit which provides a direct output to a grounded load, and avoids the need for close magnetic coupling between the halves of the primary of an output transformer. Practical circuits and the application of negative feedback are discussed.

621.396.645.211:621.385.3 1251
High-Frequency Characteristics of Resistance-Coupled Triode Amplifiers—J. W. Sauber. (PROC. I.R.E., vol. 40, pp. 48–49; January, 1952.) A semigraphical method of calculating the steady-state hf response, taking account of interelectrode capacitances.

621.396.645.35 1252
The Automatic Compensation of Zero-Drift Errors in Direct-Coupled Feedback Systems—R. G. Lex: F. A. Summerlin. (*Proc. IEE*, Part II, vol. 98, pp. 641–642; October, 1951.) Discussion on 1867 of 1951 and author's reply.

621.396.822 1253
Note on Resistance Fluctuations and the Flicker Effect—M. Surdin. (*Physica, 's Grav.*, vol. 17, pp. 548–550; May, 1951. In French.) Published experimental results on resistance fluctuations and flicker effect are reviewed. Macfarlane's theory (4087 of 1947) that flicker effect is due to fluctuations of potential barrier is shown to be compatible with van der Ziel's theory (3035 of 1950) that in oxide cathodes the effect is due to fluctuations of conductivity of the oxide layer, the mechanism involved being identical with that responsible for resistance fluctuations.

621.318.4 1254
Bauelemente der Nachrichtentechnik. Teil 3: Spulen [Book Review]—H. Nottebrock. Publishers: Schiele & Schön, Berlin, 1950, 264 pp., 12 DM. (*Arch. elekt. Übertragung*, vol. 5, p. 485; October, 1951.) Gives practical information on coils (including transformers) for telecommunication applications.

621.396.615 1255
Theory and Design of Valve Oscillators. [Book Review]—H. A. Thomas. Publishers: Chapman & Hall, London, 2nd ed., 317 pp., 36s. (*Electrician*, vol. 147, p. 1207; October 19, 1951.) Five additional chapters have been included on uhf, vm, RC, crystal, and magnetron oscillators, and a certain amount of rearrangement of the original important matter on frequency stabilization has taken place.

GENERAL PHYSICS

519.24 1256
The Best Method of Correcting for the Uncertainty involved in the Discrete, Discontinuous Nature of the Data in the Analysis of an Experiment—P. Vernotte. (*Compt. Rend.*

Acad. Sci. (Paris), vol. 233, pp. 735–736; October 1, 1951.)

534.111 1257
The Alternating-Current-Maintained Pendulum—N. Minorsky. (*Compt Rend. Acad. Sci.* (Paris), vol. 233, pp. 728–729; October 1, 1951.) The differential equations are developed and solved for the system comprising a pendulum carrying a piece of iron in the field of a coil carrying ac. The excitation of the pendulum is of nonlinear parametric type (1338 of 1951), no rational relation existing between the respective frequencies of the ac and the pendulum oscillations.

534.213.4 1258
Some Remarks on the Coupling of Two Ducts—A. E. Heins. (*Jour. Math. Phys.*, vol. 30, pp. 164–169; October, 1951.) A mathematical treatment of the propagation of waves in two channels presenting different boundary conditions and coupled end to end.

535.361.2 1259
Scattering of Electromagnetic Waves from Two Concentric Spheres—A. L. Aden and M. Kerker. (*Jour. Appl. Phys.*, vol. 22, pp. 1242–1246; October, 1951.) "A solution is given for the problem of the scattering of plane electromagnetic waves from a sphere with a concentric spherical shell. The solution is general, and under appropriate conditions is reduced to the well-known solution for scattering from a single sphere."

535.37+537.311.33 1260
Radiationless Transitions of Electrons in Crystals—F. Stöckmann. (*Z. Phys.*, vol. 130, pp. 477–479; October 9, 1951.) A theory of radiationless transitions is advanced which depends on the fact that the effective radius of the Coulomb field of an impurity center is equal to or greater than the free electron paths in valence crystals. Observations of luminescence and inhibition in phosphors and of conduction in semi-conductors are adduced in support of the theory.

535.43 1261
Scattering of Plane Waves by Soft Obstacles: Part 2—Scattering by Cylinders, Spheroids, and Disks—E. W. Montroll and R. W. Hart. (*Jour. Appl. Phys.*, vol. 22, pp. 1278–1289; October, 1951.) "Approximate closed-form analytical expressions are derived for the total and differential scattering cross sections of cylindrical, prolate-spheroidal, and disk-shaped scatterers which subject scalar plane waves to only a small phase shift." Part 1: 2139 of 1951.

537/538 1262
General Solutions of the Equations of Electrostatics and Magnetostatics—É. Durand (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1008–1010; October 29, 1951.) Solutions are derived directly from an identity which is the vector analogue of Green's scalar identity.

537.221 1263
Contact Potential Differences—I. F. Patai and M. A. Pomerantz. (*Jour. Frank. Inst.*, vol. 252, pp. 239–260; September, 1951.) A historical survey of the subject, an outline of fundamental theoretical aspects of contact between metals, and a description of methods of measuring contact potentials. An extensive bibliography is given.

537.311.1 1264
On Electrostatic Plasma Oscillations in Metals—J. A. Kok. (*Physica*, *'s Grav.*, vol. 17, pp. 543–547; May, 1951.) Equations derived for plasma oscillations in a gas are applicable to conditions in metals, provided Fermi-Dirac rather than Maxwell statistics are used.

537.311.33:621.385.032.216 1265
P-N Transition of an Oxide-Coated Cathode—Ishikawa, Sato, Okumura and Saski. (See 1472.)

537.311.37 1266
A General Formula for the Conductivity of a Gas containing Free Electrons—L. G. H. Huxley. (*Proc. Phys. Soc.*, vol. 64, pp. 844–861; October 1, 1951.) Electron drift in gases is discussed in terms of the method of free paths, for constant and alternating electric fields with and without applied magnetic field; free-path distortion due to the electric field is taken into account. Formulas are derived for the drift velocity of the centroid of the electron group; the current density and gas conductivity are found directly from the drift velocity. The formulas are relevant to the theory of conduction in metals and semiconductors, the Hall effect and wave propagation in the ionosphere.

537.56 1267
Energy Losses of Charged Particles in a Very Strongly Ionized Medium (Ionic Plasma)—B. Kwal. (*Jour. Phys. Radium*, vol. 12, pp. 805–810; October, 1951.) Theoretical discussion of processes in electronic discharges in gases; conditions are probably similar in a large part of interstellar space.

537.58 1268
Thermal Ion Source for Negative Ions—H. Hintenberger. (*Helv. Phys. Acta*, vol. 24, pp. 307–309; September 20, 1951. In German.)

537.581-13:537.566 1269
The Ionization in the Incandescent Gases of Jet-Propulsion Mechanisms—C. Klein. (*Ann. Telecommun.*, vol. 6, pp. 287–289; October, 1951.) Anomalies observed in the guiding by radio of rockets with their engine working are attributed to thermal ionization of the incandescent gases ejected. The mechanism of this ionization and its influence on the propagation of radio waves is investigated. As in the case of the ionosphere, the electron concentration N and the mean collision frequency determine the course of the phenomenon. The method used by Eggert Saha in 1920 for calculating N for stellar atmospheres is applied; values obtained are consistent with u.s.w. reflection but are very different from Goercke's experimental results (657 of April). Physico-chemical study of the combustion process shows that in most cases Saha's formula is inapplicable because thermal equilibrium is not attained; calculation based on Saenger's theory that ionization is due to molecular collisions prior to the establishment of equilibrium leads to results in better agreement.

538.12+538.65 1270
The Force and Torque acting on Magnetized Bodies in a Magnetic Field—W. Döring. (*Ann. Phys.* (Lpz.), vol. 9, pp. 363–372; October 15, 1951.)

538.12+538.65 1271
Magnetic Field and Torque for a Magnetic Ellipsoid in a Permeable Medium and an External Field—H. Diesselhorst. (*Ann. Phys.* (Lpz.), vol. 9, pp. 316–324; October 15, 1951.)

538.221 1272
The Mechanism of Discontinuities in Magnetization—T. Hofbauer and K. M. Koch. (*Z. Phys.*, vol. 130, pp. 409–414; October 9, 1951.) The effect of superposing on a main ac magnetizing field an auxiliary field of higher frequency is investigated experimentally. For a given value of the main magnetizing field the effect of the auxiliary field is to increase the value of saturation magnetization and remanence. The effect is related to that produced by hf biasing in magnetic sound recording.

538.221 1273
The Study of Ferromagnetic Resonance using the Complex-Premeability Diagram—B. Pistoulet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1015–1017; October 29, 1951.)

538.566 1274
Wave Packets, the Poynting Vector, and Energy Flow: Part 4—Poynting and Macdonald Velocities in Dissipative Anisotropic Media—C. O. Hines. (*Jour. Geophys. Res.*, vol. 56, pp. 535–544; December, 1951.) It is found preferable to replace the Poynting vector by Macdonald's to obtain physically suitable results, but even then the direction obtained for the energy flow differs from that found by wave-packet methods, which probably give the better results. Part 3: 2697 of 1951.

538.613 1275
Correlation of the Faraday and Kerr Magneto-optical Effects in Transmission-Line Terms—C. H. Luhrs. (PROC. I.R.E., vol. 40, pp. 76–78; January, 1952.) The Faraday effect is derived from phase considerations and the Kerr effect from impedance considerations.

530.145 1276
Quantum Mechanics of Particles and Wave Fields. [Book Review]—A. March. Publishers: J. Wiley & Sons, New York, 1951, 292, pp., $5.50. (*Jour. Frank. Inst.*, vol. 252, p. 270; September, 1951.) "The book provides a text which may help clarify the latest conceptions and mathematical formalism of the theory of fields and particles."

538 1277
Modern Magnetism. [Book Review]—L. F. Bates. Publishers: Cambridge University Press, London, 3rd ed., 506 pp., 30s. (*Electrician*, vol. 147, pp. 1207–1208; October 19, 1951.) "Much fresh material is incorporated, including new hf techniques developed during the last war and the very striking elaborations of the domain concept resulting from important work in America, England and France, and their bearing on the interpretation of the hysteresis cycle."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.621.396.822 1278
Radio Waves and Astronomy—H. H. Klinger. (*Funk u. Ton*, vol. 5, pp. 474–485; September, 1951.) An outline of the methods of radio astronomy and discussion of the rf radiation from the sun, the galaxy and extragalactic sources.

523.72+523.854]:621.396.822 1279
New Data on the Radiation of Electromagnetic Waves from the Sun and the Galaxy—G. G. Getmantsev. (*Uspekhi Fiz. Nauk*, vol. 44, pp. 527–557; August, 1951.) A review of recent publications on the subject, including many Russian papers.

523.746"1951.07/.09" 1280
Provisional Sunspot-Numbers for July to September, 1951—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 56, p. 604; December, 1951. *Z. Met.*, vol. 5, p. 349; November, 1951.)

523.755 1281
A New Radio Method for Measuring the Electron Density in the Solar Corona—K. E. Machin and F. G. Smith. (*Nature* (London), vol. 168, pp. 599–600; October 6, 1951.) The suggested method uses the "occultation" of the radiation from radio stars situated in directions near that of the sun. The solar radiation is prevented from masking the weaker radiation from the star by using an antenna system insensitive to sources subtending a comparatively large angle. The determination of the effective radius of the sun for this effect at a number of radio frequencies would allow the

distribution of electron density with height to be deduced, for densities as low as $10^4/cm^3$. Tests at frequencies of 210, 81.5 and 38 mcps, carried out during a period of abnormal sunspot radiation, gave no useful results. Antennas of greater resolving power will be used in future experiments.

523.78 1282

Observation of the Annular Eclipse of the Sun, 1st September 1951—F. Bosson, É. J. Blum, J. F. Denisse, E. Leroux and J. L. Steinberg. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 917–919; October 22, 1951.) Measurements are reported of the intensity of radiation on wavelengths of 3.12 and 178 cm received during the eclipse at a location close to Markala in the French Sudan. The observations on 3.12 cm confirm that the sun's radiation is more intense at the limb. The apparent diameter corresponding to radiation on 178 cm is 1.4 times the apparent optical diameter, and the coronal radiation on that wavelength constitutes half the total radiation.

523.8:621.396.822]:522.92 1283

A Diffraction Theory of the Scintillation of Stars on Optical and Radio Wavelengths—C. G. Little. (*Mon. Not. R. Astr. Soc.*, vol. 111, pp. 289–302; 1951.) The refraction theory of scintillation at optical wavelengths requires excessive atmospheric density gradients, and fails to explain the observed scintillation in color; also the physiological explanations are insufficient to explain all the observations. The effects may be explained by Fresnel diffraction at a nonhomogeneous atmospheric layer; this theory requires much smaller density gradients. A similar theory explains the observed intensity flucutations of radio waves from discrete extraterrestrial sources.

523.8:621.396.822]:535.42 1284

The Diffraction of Radio Waves in Passing through a Phase-changing Ionosphere—A. Hewish. (*Proc. Roy. Soc. A*, vol. 209, pp. 81–96; October 8, 1951.) Discussion of the diffraction by the ionosphere of waves from radio stars, and the resultant field at the earth's surface. It is assumed that the wave emerges from the ionosphere with constant amplitude but with lateral variations of phase. The cases of simple sinusoidal variation and random variation of phase are considered. Knowledge of the amplitude and phase variations at the ground enables the average magnitude of the phase deviations produced by the ionosphere, and their lateral extent, to be calculated. Comparison of results on different wavelengths enables an estimate to be made of the distance of the effective diffracting screen from the plane of observation. Experimental results indicate that the ionospheric irregularities have a lateral extent of the order of 5 km and cause phase deviations of 1 to 2 radians for radiation of wavelength 6.7 m.

523.852.32:621.396.822 1285

Radio Emission from the Andromeda Nebula—R. H. Brown and C. Hazard. (*Mon. Not. R. Astr. Soc.*, vol. 111, pp. 357–367; 1951.) Detailed account of work noted in 629 of 1951. The results indicate that extragalactic rf radiation accounts for about 1 per cent of the total rf radiation reaching the earth.

550.38"1951.04/.06" 1286

International Data on Magnetic Disturbances, Second Quarter, 1951—J. Bartels and J. Veldkamp. (*Jour. Geophys. Res.*, vol. 56, pp. 601–603; December, 1951.)

550.38"1951.07/.09" 1287

Cheltenham [Maryland] Three-Hour-Range Indices *K* for July to September, 1951—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 56, p. 604; December, 1951.)

550.385"1951.04/.09" 1288

Principal Magnetic Storms [April–Sept. 1951]—(*Jour. Geophys. Res.*, vol. 56, pp. 605–608; December, 1951.)

551.510.52:551.594.13 1289

Electrical Conductivity of Air in the Troposphere—R. C. Callahan, S. C. Coroniti, A. J. Parziale and R. Patten. (*Jour. Geophys. Res.*, vol. 56, pp. 545–551; December, 1951.) Fair-weather measurements for widely separated areas in the U.S.A. indicate equality of positive-ion and negative-ion conductivity throughout the 35,000-foot altitude range investigated. The results are in agreement with those calculated from a formula based on Thomson's theory of volume recombination.

551.510.535 1290

The Absorption of Ultraviolet Solar Radiation and the Temperature of the Upper Atmosphere at a Height of about 100 Km—J. Gauzit. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1048–1049; October 29, 1951.) The mean kinetic energy of oxygen atoms in the upper atmosphere at the instant of their creation by photodissociation corresponds to a high temperature which rises from 800° to about 3,800° K between 90 and 130 km height.

551.510.535 1291

Some Phenomena of the Upper Atmosphere—S. Chapman. (*Proc. Phys. Soc.*, vol. 64, pp. 833–844; October 1, 1951.) A review of recent progress in knowledge of the height distribution of the temperature and constitution of the atmosphere, and of winds and tides in the upper atmosphere. Ionospheric phenomena observed near the magnetic equator are discussed, especially the abnormal intensification of electric current flow.

551.510.535 1292

A Quick Method for Analysing Ionospheric Records—J. A. Ratcliffe. (*Jour. Geophys. Res.*, vol. 56, pp. 463–485; December, 1951.) A transparent scale carries families of calculated h'/f curves (h' is virtual height and f the frequency) for a series of values of layer thickness and critical frequency. Experimental h'/f records can then be matched with the calculated ones and the thickness parameters read on the scale. The method is applied to parabolic, linear, and square-root electron-distribution laws and provides a simple way of finding the total electron content in the F_2 region. Allowances for the effect of the F_1 layer are simply made.

551.510.535 1293

An Unusual Ionospheric Disturbance observed in Adélie Land—M. Barre and K. Rawer. (*Ann. Géophys.*, vol. 6, pp. 309–317; October/December, 1950.) See 2997 of 1951.

551.510.535 1294

Some Regularities in the F_2 Region of the Ionosphere—J. A. Ratcliffe. (*Jour. Geophys. Res.*, vol. 56, pp. 487–507; December, 1951.) Many of the anomalies of f_0F_2 values vanish when the total electron content per unit column (n) is considered instead. Records from Watheroo, Huancayo, and Alaska show that the value of n at noon is closely related to the sun's zenith angle over the whole range of values. The diurnal change in n does not show the "bite-out" seen on f_0F_2 curves, and the change in n with magnetic dip shows no equatorial minimum.

551.510.535 1295

The Effective Midday Recombination Coefficient of the F_2 layer calculated from the Values of Critical Frequency observed during the Mogel-Dellinger Effect on 19th Nov. 1949—W. Becker and W. Dieminger. (*Z. Naturf.*, vol. 5a, pp. 308–311; June, 1950.) The value 1.9×10^{-10} is found as a lower limit; this is of the same order as previously published nighttime values. The rate of production of charge carriers in the F_2 layer at the start of the disturbance was twice as great as normal.

551.510.535:537.56 1296

'Isanomalies' of F_2 Ionization—O. Burkard. (*Jour. Geophys. Res.*, vol. 56, pp. 595–600; December, 1951. In German.) Discussion of the information which can be derived from sets of curves indicating places with corresponding F_2-layer variations.

551.510.535:551.557 1297

Winds in the Ionosphere—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 178–179; December, 1951.) A short explanation is given of the method of observation in which pulse signals from a central transmitter are received, after reflection from the ionosphere, at three stations located at the corners of a right-angled triangle. From records of the times at which fading of the signals occurs at the three stations, an estimate is made of the direction and speed of the ionospheric regions responsible for the fading pattern of the received signals. The N.B.S. transmitter at Sterling, Va., operates at 2.3 mcps and emits pulses of duration 0.2 ms, repetition rate 60 per second and peak power of 10 kw. Typical records of ionospheric wind velocities obtained at Sterling, Va., and at Cambridge, England, show close correlation, indicating that the ionospheric winds belong to a world-wide circulation system.

551.510.535:551.557 1298

Systematic Ionospheric Winds—C. D. Salzberg and R. Greenstone. (*Jour. Geophys. Res.*, vol. 56, pp. 521–533; December, 1951.) Detailed account of the work referred to in 1297 above.

551.510.535:621.3.087.5 1299

Spot-Frequency Ionospheric Recording—A Combination of Sweep- and Fixed-Frequency Techniques—H. W. Wells. (*Jour. Geophys. Res.*, vol. 56, pp. 613–615; December, 1951.) A panoramic recorder is used but instead of photographing the complete $h'f$ trace, only three selected frequencies are recorded. The cycle is repeated at intervals of 25 seconds and the photographic paper is moved on slightly after each frequency in the cycle. The resulting record is a combination of three $h't$ records at different frequencies. Changes in structure with time are much more easily observed than on a series of full $h'f$ records.

551.510.535:621.396.11 1300

Contribution to the Study of the Electron Distribution in the Ionosphere and of the Absorption of Short Waves—É. Argence, M. Mayot and K. Rawer. (*Ann. Géophys.*, vol. 6, pp. 242–285; October/December, 1950.) Based on a comprehensive survey of recently obtained data on the temperature distribution and the dissociation of O_2 in the atmosphere, a model for the *D* and *E* layers is proposed with parabolic distribution of temperature below 100 km and a rapid decrease of ionizable molecules above that height. The frequency variation of absorption due to collisions is investigated; approximate formulas are obtained assuming a stratified structure with parabolic distribution of μ^2 (where μ is the refractive index) for each layer and an exponential distribution of collision number. Special functions introduced in the theory are discussed and numerical values are tabulated.

551.510.535:621.396.11 1301

Group Velocities and Group Heights from the Magneto-ionic Theory—Shinn and Whale. (see 1405.)

551.510.535:621.396.11 1302

Radio-Wave Propagation at Oblique Incidence including the Lorentz Polarization

Term—J. M. Kelso. (PROC. I.R.E., vol. 40, pp. 87–97; January, 1952.) The true height of reflection, ray path, reflection coefficient, range and group path in the ionosphere, all of which depend only on the ionosphere, are first studied. Then, for both plane- and curved-earth geometry, the ground range, group path, reflection coefficient as a function of ground range, and muf are examined. The ionosphere is considered as plane throughout the treatment, which has the following restrictions: all limitations on the use of the Chapman distribution hold; the earth's magnetic field is neglected; the angular operating frequency is assumed to be greater than the collision frequency; the absorption per vacuum wavelength is assumed to be small; ray theory is used.

551.594.21 **1303**

The Distribution of Electricity in Thunderclouds—D. J. Malan and B. F. J. Schonland. (*Proc. Roy. Soc. A*, vol. 209 pp. 158–177; October 23, 1951.) Various independent methods of measuring the heights of origination of successive lightning strokes are discussed; the values obtained support the hypothesis of a columnar distribution of negative charge.

551.594.222 **1304**

Point-Discharge Currents and the Earth's Electric Field—W. C. A. Hutchinson. (*Quart. Jour. R. Met. Soc.*, vol. 77, pp. 627–632; October, 1951.) Measurements are reported of the current flowing to earth through a point set up at a height of 12 m, and of the electric field near the ground. Point-discharge current of either sign increases with the square of the field; the relation tends to direct proportionality for high values of field. The influence of local space charge on the field measurements is discussed.

551.594.6 **1305**

Recording of Atmospherics on Board the *Commandant Charcot*, 1950–1951 Cruise—R. Bureau and J. J. Vaury. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1049–1051; October 29, 1951.) The third antarctic cruise of the *Commandant Charcot* was followed by one across the Pacific, during which further centers of atmospherics were located. The apparatus was the same as that used on the second cruise [see 637 of 1951 (Bureau and Barré)]. The whole of the Pacific west of 120°W seems to be a diffuse source of atmospherics. A typical recording for a 24-hour period on 18th–19th April 1951 shows the approximately linear rise of the level of atmospherics starting at sunset. Maximum activity for the Pacific centers occurs at a later hour than for known continental centers, at 1700–1800 local time.

LOCATION AND AIDS TO NAVIGATION

621.396.9 **1306**

Radar Technique—F. Michelssen. (*Frequenz*, vol. 5, pp. 258–259; September, 1951.) A brief historical note of developments in Germany during the prewar and early war periods.

621.396.9:621.396.822 **1307**

The Detection of Pulse Signals near the Noise Threshold—R. E. Spencer. (*Jour. Brit.*, IRE, vol. 11, pp. 435–454; October, 1951.) Existing literature is surveyed, with emphasis on simpler and more intuitive physical considerations and their application to radar. The discussion is first in terms of a single pulse and then of repeated pulses Matching the IF response to the frequency spectrum of the pulse is considered. The probability that the presence of a signal can be distinguished is shown to be independent of the detector law provided that it gives a monotonic increase of output with increased input. When signal and noise are of comparable magnitude, there is loss of performance if filtration is incorrectly distributed between the IF and video-frequency stages. Methods of presentation are discussed and the main practical conclusions are summarized.

621.396.93:621.396.812 **1308**

Disturbances caused by the Atmosphere in Aids to Navigation—E. Vassy. (*Onde Élect.*, vol. 31, pp. 379–383; October, 1951.) Sources of error in direction-finding systems are discussed generally, including the effects of tilts and asymmetry in the higher ionized layers, and of lateral refraction and diffuse reflection in the troposphere.

621.396.932.1/.2 **1309**

The Requirements for Radio Aids at Sea—F. J. Wylie. (*Jour. Inst. Nav.*, vol. 4, pp. 327–344; October, 1951.) Radio aids to navigation are required as alternatives to astronomical and terrestrial fixing when these cannot be used, and the accuracy required is that obtained by the visual methods in good conditions. The extent to which shipborne and shore-based direction-finding equipment and loran, consol, Decca and radar systems satisfy the requirements is discussed. An account is given of the coverage obtained by existing systems, and possible lines of development are indicated.

621.396.932.1 **1310**

A New Shore-Based Radar Equipment—E. Fennessy. (*Jour. Inst. Nav.*, vol. 4, pp. 345–350; October, 1951.) A general description of the Decca harbor radar equipment, its performance and applications. A method which has been developed for the radio transmission of information from the antenna and rf assembly to a remote display unit might be adapted to provide displays in vessels in the service area of the radar equipment.

621.396.933 **1311**

Hyberbolic Navigation Systems in Germany—E. Roessler. (*Elektrotech. Z.*, vol. 72, pp. 567–572; October 1, 1951.) Discussion of various systems, particularly Decca and its advantages over the single-frequency system. By a suitable linking of three hyperbola systems, both coarse and fine navigation facilities can be provided for an area greater than Germany, using only four frequencies.

621.396.933 **1312**

Long-Range Radar for Controlling Aircraft: Part 1—The Operation of London Radar—D. W. Watkins. (*Jour. Inst. Nav.*, vol. 4, pp. 402–409; October, 1951. Discussion, pp. 413–414.) An account of early experiments, operational trials, the services at present available, with an outline of future developments. Part 2: 1313 below.

621.396.933 **1313**

Long-Range Radar for Controlling Aircraft: Part 2—Traffic Control at the Royal Aircraft Establishment—G. G. Harris. (*Jour. Inst. Nav.*, vol. 4, pp. 409–413; October, 1951. Discussion, pp. 413–414.) Description of the radar control system at Farnborough, which is designed primarily for speed and flexibility. The approach control is the dominating control and extends in stages over the local or visual control as weather conditions deteriorate, providing talk-down facilities in the extreme case. Part 1: 1312 above.

621.396.933 **1314**

Radio Aids to Airways Navigation. The Australian Visual Aural Radio Range System—H. White and F. B. Partridge. (*Jour. Inst. Eng.* (Aust.), vol. 23, pp. 167–181; September, 1951.) The basic aid adopted for use in Australia consists of a visual aural radio range (V. A. R.) in conjunction with distance-measuring equipment. The V. A. R., operating at about 120 mcps, is an adaptation of the American ρ-θ system (1158 of 1951) and provides four ranges, two with revival and two with aural presentation. A description is given of the general principles and ground equipment, and the radiation patterns are considered in detail. The airborne equipment is also described.

621.396.933 **1315**

Modern Radio Aids to Civil Aviation—The D.M.E. Project of the Department of Civil Aviation—J. H. Gerrand. (PROC. I.R.E. (Aust.), vol. 12, pp. 275–282; September 1951.) "The history of the Australian D.M.E. project is reviewed and the main technical features of the equipment are briefly described. The choice of beacon sites is discussed together with the proposed program of installation testing."

621.396.933(98) **1316**

Radar as an Aid to Air Navigation in the Arctic—K. R. Greenaway. (*Jour. Inst. Nav.*, vol. 4, pp. 399–401; October, 1951.) Summary of paper read at meeting of Canadian Institute of Navigation, Montreal, September 1950. The use of radar for the determination of drift and groundspeeds, and for pin-pointing, is discussed. The special difficulties encountered in the Arctic are considered. It is concluded that search radar ranks next in importance to astronomical navigation for flights in the Arctic.

MATERIALS AND SUBSIDIARY TECHNIQUES

534.321.9.001.8:621.315.615/.617:539.5 **1317**

Mechanical Properties of Polymers at Ultrasonic Frequencies—W. P. Mason and H. J. McSkimin. (*Bell Sys. Tech. Jour.*, vol. 31, pp 122–171; January, 1952.) Different types of measurement methods are described for determining the reaction of polymer materials, in solid or liquid form or in solution, to longitudinal and shear waves over a wide range of frequencies. The relaxation frequencies are determined by a dispersion in the velocity, attenuation constants or characteristic impedance of the material. The various types of relaxation observed are explained by certain motions of the polymer chain or molecule which determine the toughness, impact strength and elasticity of the material.

535.37 **1318**

A Striking Difference in the Activation Susceptibility of Potassium- and Sodium-Halide Phosphors—P. Pringsheim. (*Acta Phys. austriaca*, vol. 3, pp. 396–404; March, 1950.)

535.37 **1319**

Size Effects in the Luminescence of ZnS Phosphors—G. C. Wallick. (*Phys. Rev.*, vol. 84, p. 375; October 15, 1951.) Thermoluminescence measurements for crystals of 9ZnS.CdS-Cu and Zns-Ag were made by saturation excitation at −160°C followed by a uniform increase of temperature to 100°C at a rate of 2° per minute. Evidence is given of many more shallow traps in small crystals (average diameter 5.5μ) than in large ones (average diameter 14μ). The decay of phosphorescence with time was, however, found to be essentially independent of crystal size.

535.371.07 **1320**

A Fluorescent Cadmium-Iodide Screen under the Influence of Visible and Ultraviolet Radiation, Cathode Rays, μ Rays and X Rays—S. Schlivitch. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1023–1024; October 29, 1951.) A strongly luminescent phosphor is prepared by mixing lead nitrate and cadmium iodide. When used in a cr tube, the screen is luminescent for voltages over 600 v, the spectrum being in the region where the eye is highly sensitive.

537.311.33:538.214 **1321**

Magnetism of Free Charge-Carriers in Semiconductors—W. Kanzig and N. Maikoff. (*Helv. Phys. Acta*, vol. 24, pp. 329–331; September 20, 1951. In German.) Measurements of the magnetic susceptibility χ of grey semiconducting tin indicate an increase of χ with rising temperature proportional to the number of free charge-carriers.

537.311.33:538.221 1322
Resistivity of Ferrites of Zinc, Nickel, Cobalt, Magnesium and Copper as a Function of Temperature—L. Bochirol. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 736–738; October, 1951.) Measurements on sintered ferrite disks over the temperature range 100°–700°C are reported and examined in relation to the theory of electronic semiconductors.

537.311.33:546.28 1323
A. C. Characteristics of Si *p-n* Junction—S. Komagata, M. Hatoyama, M. Shibuya, W. Sasaki, T. Yamamoto and M. Kikuchi. (*Jour. Appl. Phys.*, vol. 22, p. 1290; October, 1951.) Graphs are given for the conductance and susceptance of *p-n* junctions in the frequency range 200 cps to 100 kcps, and for the resistance and capacitance, at various frequencies, as functions of bias voltage.

537.311.33:546.289 1324
The Effects of Pressure and Temperature on the Resistance of *p-n* Junctions in Germanium—H. H. Hall, J. Bardeen and G. L. Pearson. (*Phys. Rev.*, vol. 84, pp. 129–132; October 1, 1951.) According to Shockley's theory (379 of 1950), the low-voltage resistance of a *p-n* junction is proportional to exp (E_G/kT), where E_G is the energy gap. Measurements of the change with pressure of the characteristics of a junction in a Ge single crystal indicate a resistance change of 12.5 per cent, corresponding to a change of E_G of about 3.1×10^{-4} ev, for a pressure change of 10,000 lb/in.² Analysis of measurements made at temperatures between 16.5° and 20.5°C gives values of E_G averaging about 0.72 ev; these values are in agreement with those obtained from the change in intrinsic resistivity with temperature and pressure.

537.311.33:621.396.822 1325
A Theory of Electrical Fluctuations in Semiconductors—M. Surdin. (*Jour. Phys. Radium*, vol. 12, pp. 777–783; October, 1951.) When a constant current is passed through a semiconductor filament or through a metal/semiconductor contact, the magnitude of the observed fluctuations across the terminals is greater by some powers of 10 than that to be expected from the thermal or shot effect. An explanatory theory is advanced based on the fluctuations of the number of conduction electrons in the case of filaments, and on the number of donor centers in the barrier layer in the case of metal/semiconductor contacts.

538.221 1326
The Frequency Dependence of Magnetic After-Effect in Powder Cores—T. Einsele and F. Baur. (*Z. angew. Phys.*, vol. 3, pp. 373–376; October, 1951.) The complex permeability of a powder core is determined from the results of a series of bridge measurements at frequencies between 30 cps and 100 kcps on a coil surrounding the core. After subtracting the effect of eddy currents (determined by extrapolation of Q measurements), the extent of the magnetic after-effect alone on the complex permeability is determined. The after-effect has different characteristics at low and high frequencies. It is not connected with the increase of permeability with temperature. Core losses at small field strengths and at frequencies up to 100 kcps are largely dependent on the after-effect.

538.221 1327
Ferromagnetic Resonance in Cobalt Ferrite at High Temperature—T. Okamura, Y. Torizuka and Y. Kojima. (*Phys. Rev.*, vol. 84, p. 372; October 15, 1951.)

538.221 1328
Low-Frequency Dispersion in Ni- and Co-Ferrites—K. Kamiyoshi. (*Phys. Rev.*, vol. 84, pp. 374–375; October 15, 1951.)

538.221 1329
The Magnetic Properties of α-Ferric Oxide—W. P. Osmond. (*Proc. Phys. Soc.*, vol. 64, pp. 931–932; October 1, 1951.) Discussion and reconciliation of contending views advanced by Chevallier and Snoek (2731 of 1951) and Néel (3161 of 1949).

538.221:538.24 1330
The Value of the Spontaneous Magnetization of Binary Nickel Alloys as a Function of Temperature—J. J. Went. (*Physica*, 's Grav., vol. 17, pp. 596–602; June, 1951.)

538.221:538.652 1331
Linear Magnetostriction of Homogeneous Nickel Alloys—J. J. Went. (*Physica*, 's Grav., vol. 17, pp. 98–116; February, 1951.) The magnetostriction was measured as a function of alloy composition, induction, and temperature; a comprehensive table of results is given and discussed.

538.24 1332
The Magnetization of Steels in Weak Magnetic Fields: Effects of Time, Stresses, Shocks, Transverse Magnetic Fields—L. Lliboutry. (*Ann. Phys.* (Paris), vol. 6, pp. 731–829; September/October, 1951.) Full treatment. See also 2230 of 1950.

538.249 1333
The Temperature Dependence of the Magnetic Viscosity of Ferromagnetic Metals—E. F. Kuritsyna. (*Compt. Rend. Acad. Sci.* (URSS), vol. 79, pp. 233–236; July 11, 1951. In Russian.)

539.234 1334
Notes on the Metallization of Surfaces by Evaporation in Vacuum—L. Dunoyer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 919–921; October 22, 1951.) Details are given of (*a*) pretreatment of the surface, (*b*) choice of metal for heater, and (*c*) a method of increasing the hardness of the deposit, in relation to the deposition of Al on glass.

546.23:548.55:537.311.33 1335
Electrical Properties of Selenium: Part 2—Microcrystalline Selenium—H. W. Henkels. (*Jour. Appl. Phys.*, vol. 22, pp. 1265–1278; October, 1951.) The electrical properties of selenium have been studied in their dependence on history of preparation, i.e., as functions of initial temperature of the liquid selenium, quench procedure, nucleation procedure, temperature of crystallization, and time of crystallization. Single crystals and also a microcrystalline matrix of these crystals were investigated. The effect of the addition of small quantities of sulphur, oxygen and iodine was determined. The frequency dependence of resistivity was also studied. Carrier densities and effective mobilities are estimated from data on thermoelectric power and resistivity. The results obtained, and their relation to other observations and to theories of semiconductors, are discussed. Part 1: 698 of April.

546.23.03 1336
Studies on Selenium and its Alloys: Report 1—Physical Properties of Selenium—T. Sato and H. Kaneko. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 45–54; 1950.) Includes measurements of expansion coefficient, density, specific heat, melting and boiling points, dependence of electrical resistance on heat treatment, and resistance at high temperature.

546.28:537.534.9:621.314.6 1337
Properties of Ionic Bombarded Silicon—R. S. Ohl. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 104–121; January, 1952.) The change of the rectifying properties of Si was studied as a function of ion velocity, intensity of bombarding current, length of time of bombardment, kind of gas (H, He, N, Ar), and the temperature of the specimen during bombardment. It was found that Si contaminated with B to the point where it shows little rectification can be modified by bombardment to make its rectifying properties better than those of most unbombarded materials.

546.289:537.312.67 1338
The Electrical Conductivity of Liquid Germanium—R. W. Keyes. (*Phys. Rev.*, vol. 84, pp. 367–368; October 15, 1951.) The resistivity of liquid Ge near the melting point was found to be 60μ Ω cm (about 1/15 that of solid Ge at the melting point), with a positive temperature coefficient of about 1 or 2 parts in 10^4 per 1°C. The purity of the Ge was stated to be over 99.9 per cent.

546.321.85:621.3.011.5 1339
Difference between the Dielectric Constants of Free and Clamped KH_2PO_4 Crystals—H. Baumgartner. (*Helv. Phys. Acta*, vol. 24, pp. 326–329; September 20, 1951. In German.) Measurements in the temperature range −155° to −120°C confirm the Curie-Weiss linear relation between the reciprocal of the dielectric constant and the temperature, both for a crystal effectively clamped by exciting it at its 201st harmonic (10 mcps) and for a free crystal excited at its fundamental frequency of 1 kcps. The graph for the free crystal is displaced about 4°, toward higher temperatures, relative to that for the clamped crystal. The results are in complete agreement with Müller's theory.

546.431.824–31 1340
Ferroelectricity *versus* Antiferroelectricity in Barium Titanate—M. H. Cohen. (*Phys. Rev.*, vol. 85, p. 369; October 15, 1951.) Discussion of an antiferroelectric state leads to the conclusion that the ferroelectric properties of $BaTiO_3$ cannot be explained by the simple dipole-dipole interaction model suggested by Slater (2188 of 1950).

546.431.824–31 1341
Theory of Barium Titanate: Part 2—A. F. Devonshire. (*Phil. Mag.*, vol. 42, pp. 1065–1079; October, 1951.) The theory given in part 1 (663 of 1950) is extended, and expressions are obtained for the piezoelectric constants, the elastic coefficients for constant field and the dielectric constants for constant strain in terms of other physical constants of the material. These quantities are plotted as functions of temperature, and comparison is made with tubes found experimentally. The relations between the constants of the ceramic and those of the single crystal are discussed briefly.

546.431.824–31:621.3.011.5 1342
Is the Ferroelectric Transition of Barium Titanate at 120°C of the First or the Second Kind?—W. Känzig and N. Maikoff. (*Helv. Phys. Acta*, vol. 24, pp. 343–356; September 20, 1951. In German.) Investigation of the effect of a biasing field on the permittivity of single-domain $BaTiO_3$ crystals in the range 20°–180°C confirms Devonshire's theory (663 of 1950). The transition at 120°C appears to be an adiabatic transition of the first kind.

546.817.221:621.314.63 1343
The Rectifying Properties of Lead Sulphide—I. I. Adrianiva and I. L. Sokol'skaya. (*Zh. Tekh. Fiz.*, vol. 21, pp. 713–714; June, 1951.) According to modern theory, contact rectification takes place when the metal is positive with respect to the semiconductor. Experiments with PbS and a tungsten point indicated that if the applied voltage is increased up to 1–1.5 v the rectified current changes its direction. Also, if the rectifying contact is placed in vacuum,

the current decreases and then changes its sign. An explanation of the phenomena is advanced.

546.824–31:621.315.612.4 1344
Electric and Thermoelectric Properties of Partially Reduced (Blue) Titanium Dioxide—B. I. Boltaks, F. I. Vasenin and A. E. Salunina. (*Zh. Tekh. Fiz.*, vol. 21, pp. 532–546; May, 1951.) Experiments were conducted for determining, within a wide range of temperatures, the specific conductivity and thermoelectric emf of samples of TiO_2 at various degrees of reduction. The results obtained are explained from the standpoint of modern theories on the conductivity of semiconductors.

546.–31:621.315.612.4 1345
Dielectric Properties of Various Preparations of Titanium Dioxide—I. N. Belyaev, N. S. Novosil'tsev, A. L. Khodakov and M. S. Shul'man. (*Zh. Tekh. Fiz.*, vol. 21, pp. 547–551; May, 1951.) The effect of impurities on the dielectric properties of TiO_2 has been studied on six different samples. The results obtained are shown in curves and tables.

548.0:537.228.1 1346
Growing Piezoelectric Crystals—H. Mosaner and M. Wurl. (*Arch. elekt. Übertragung*, vol. 5, pp. 463–467; October, 1951.) An examination is made of factors influencing the development of single crystals of commonly used types, grown from an aqueous solution whose temperature is reduced under controlled conditions. A constant rate of growth of only a few millimeters per day is desirable. The temperature reduction must be determined in relation to the rate of growth to give the required size.

548.0:538.114 1347
Interatomic Distances and Ferromagnetism in Spinels—R. S. Weisz. (*Phys. Rev.*, vol. 84, p. 379; October 15, 1951.) Discussion with particular reference to the hypothesis that spin interaction is inversely proportional to the distance from a metal ion to a nearest neighbor (i.e., an oxygen ion) and thence to another metal ion.

621.315.612.6:537.311 1348
Breakdown and Electric Conductivity of Glass—K. J. Keller. (*Physica, 's Grav.*, vol. 17, pp. 511–530; May, 1951.)

621.315.616 1349
Synthetic Materials—P. Domin. (*Fernmeldetech. Z.*, vol. 4, pp. 461–467; October, 1951.) Review of various plastics and their derivation. Mechanical and electrical properties are tabulated.

621.396.611.21:549.514.51 1350
Frequency Shift of Piezoelectric Oscillations of Quartz under High Pressure—A. Michels and J. P. Pérez. (*Physica, 's Grav.*, vol. 17, pp. 563–564; May, 1951.) Measurements are reported on two quartz resonators, an AT cut with frequency 942 kcps and a BT cut with frequency 6.04 mcps at atmospheric pressure. The slope of the frequency/pressure curve is positive for the AT and negative for the BT cut.

666.1 1351
Technical Control in Glass Manufacture—J. H. Partridge and E. Preston. (*GEC Jour.*, vol. 18, pp. 212–220; October, 1951.) Descriptions are given of an automatic method of controlling the temperature of a large glass-melting tank furnace, a photoelectric pyrometer for measuring accurately the temperature of glass as it is fed to machines, and an instrument for measuring the diameter of glass tubing during the drawing process. The results of applying these controls are demonstrated.

MATHEMATICS

517.6:621.396.611.1 1352
The Impulse Integral, a Counterpart of the Duhamel Step Integral—R. Lueg, M. Päsler and W. Reichardt. (*Ann. Phys.* (Lpz.), vol. 9, pp. 307–315; October 15, 1951.) If a time-variable disturbing force acts on a linear system and G is one of the system parameters, the function $G(t)$, representing the time dependence of G under the action of the disturbing force, may be represented by the Duhamel integral. It may also be represented in a much simpler way by an expression whose integrand differs from the Duhamel integral in that the effect of the step function (unit step) is replaced by the effect of a delta function (needle impulse). Hence the term impulse integral. It is the simplest form of many possible expressions, and leads, under stated conditions, to a multiplication law for finding Fourier components.

517.9 1353
On a Relaxation Method for Eigenvalue Problems—S. H. Crandall. (*Jour. Math. Phys.*, vol. 30, pp. 140–145; October, 1951.)

517.9 1354
The Application of Relaxation Methods to the Solution of Differential Equations in Three Dimensions: Part 1—Boundary Value Potential Problems—D. N. de G. Allen and S. C. R. Dennis. (*Quart. Jour. Mech. Appl. Math.*, vol. 4, part 2, pp. 199–208; June, 1951.)

518.5 1355
The Logarithmic-Complex-Number Plane and the Complex-Number Calculator—W. de Beauclair. (*Z. Ver. Dtsch. Ing.*, vol. 93, pp. 955–957; October 21, 1951.) The complex-number plane with rectangular- and polar-coordinate network can be so transformed that multiplication and division of complex numbers reduce to simple addition and subtraction. This forms the basis of a simple calculator.

681.142 1356
The Binac—A. A. Auerbach, J. P. Eckert, Jr., R. F. Shaw, J. R. Weiner and L. D. Wilson. (Proc. I.R.E., vol. 40, pp. 12–29; January, 1952.) A comprehensive description of a high-speed electronic digital computer. The Binac consists of a main computing section, input-output equipment, and a mercury delay line of 512-word capacity. It is a combination of a few basic circuits, classified as diode matrices, switching gates, and reshaping gates; the latter allow a "pulse-envelope" representation of information. Crystal diodes are used in switching and gating. Crystal gates are used with electric delay lines to form a serial binary adder. Input data are supplied to the memory from a keyboard or magnetic tape through a synchronizer, which is also used to transfer data from the memory to magnetic tape or an electric typewriter.

681.142 1357
Logical Description of some Digital-Computer Adders and Counters—H. J. Gray, Jr. (Proc. I.R.E., vol. 40, pp. 29–33; January, 1952.) Description of some of the circuits incorporated in the EDVAC computer.

681.142:517.942.9 1358
A Three-Dimensional Electrical Potential Analyser—S. C. Redshaw. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 291–295; October, 1951.) A resistance network is used in an electrical-analogue method of solving problems involving the three-dimensional form of Laplace's equation.

681.142:621.392.5 1359
Magnetostriction Storage Systems for a High-Speed Digital Computer—Millership, Robbins and De Barr. (See 1200.)

517.63 1360
Die zweidimensionale Laplace-Transformation, eine Einführung in ihre Anwendung zur Lösung von Randwertproblemen nebst Tabellen von Korrespondenzen (The Two-Dimensional Laplace Transformation, an Introduction to its Application in the Solution of Boundary Problems, with Tables of Transforms). [Book Review]—D. Voelker and G. Doetsch. Publishers: Birkhäuser, Basle, 1950, 259 pp., bound, 43 Swiss francs; paper-covered, 39 Swiss francs. (*Z. angew. Phys.*, vol. 3, p. 398; October, 1951.) "The book is written just as the scientific practicing engineer requires it."

519.2 1361
Introduction to the Theory of Probability and Statistics. [Book Review]—N. Arley and K. R. Buch. Publishers: Chapman & Hall, London, 1950, 236 pp., 32s. (*Beama Jour.*, vol. 58, p. 340; October, 1951.) "The treatment given to the subject matter is clear and concise, wisely illustrated with actual fully worked examples, many of a numerical nature."

MEASUREMENTS AND TEST GEAR

389.6 1362
Recent Developments and Techniques in the Maintenance of Standards—M. M. Postgate. (*Nature* (Lond.), vol. 168, pp. 594–597; October 6, 1951.) Report of a symposium at the National Physical Laboratory, 21st–22nd May 1951. Most of the standardizing laboratories of Europe, the Commonwealth, and North America were represented and also many research and industrial organizations in Britain. An outline is given of the various subjects discussed. The full proceedings are to be published shortly by H.M. Stationery Office.

529.78 1363
A Precision Radio Time-Signal System—D. W. R. McKinley and B. E. Bourne. (*Canad. Jour. Tech.*, vol. 29, pp. 428–434; October, 1951.) Accurate "seconds" pulses are produced from a 100-kcps quartz-crystal frequency standard, coded, and radiated from a 220-mcps transmitter with line-of-sight coverage. The average error in the interval between two pulses is 10^{-7} sec at the source, or 10^{-6} sec. as observed at distant stations. The system has been used for continuous automatic time recording at the various stations of a meteor-observation network.

621.317.31:621.396.645 1364
Reduction of Measuring Time of Low-Current Amplifiers without Reduction of Sensitivity—J. Taieb. (*Compt. Rend. Acad. Sci.*, Paris), vol. 233, pp. 785–787; October 8, 1951.) The time constant of the amplifier input circuit is reduced by eliminating stray capacitance associated with a high input resistance; this is done by immersing the resistor in an appropriate electrostatic field. In one experimental arrangement the resistor was inserted through a pair of parallel plates of which one was earthed while the other was connected to a negative feedback circuit. Measuring time is so short that drift effects are eliminated. Some numerical results obtained with different values of input resistance are tabulated.

621.317.329:621.392.26 1365
Determination of Aperture Parameters by Electrolytic-Tank Measurements—S. B. Cohn. (Proc. I.R.E., vol. 40, p. 33; January, 1952.) Correction to paper noted in 725 of April.

621.317.333.4.015.7:621.315.2 1366
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621.317.355.3+538.569.4 1367
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621.317.335.3 + 538.56[.4]:533.1 1368

U.H.F. Measurements of Dielectric Constants and Absorption Coefficients of Gases and Vapours—A. Gennaoui. (*Helv. Phys. Acta*, vol. 24, pp. 333–334; September 20, 1951. In French.) Short outline of a cavity-resonator method of measurement, details of which are to be published later.

621.317.335.3.029.63 1369

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621.317.335.3.029.64 1370

Measurement of the Electrical Properties of Highly Absorbing Dielectrics on Centimetre Waves by the 'Infinite Layer' Method—N. V. Aleksandrov. (*Zh. tekh. Fis.*, vol. 21, pp. 647–651; June, 1951.) The dielectric under test is introduced into a cylindrical waveguide in which H_1 waves are propagated. Assuming no reflection from the far end, the dielectric constant and loss angle can be calculated from the distribution of the field in front of the dielectric. The theory of the method is given, with experimental results.

621.317.335.3.029.64:546.217 1371

The Refractive Indices and Dielectric Constants of Air and its Principal Constituents at 24000 Mc/s—L. Essen and K. D. Froome. (*Proc. Phys. Soc.*, vol. 64, pp. 862–875; October 1, 1951.) Full account of work noted in 1707 of 1951.

621.317.336.029.64 1372

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621.317.34.092:621.397.24 1373

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621.317.725 1376

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621.317.76 1377

The Secondary-Standard U.H.F. Generator of the Compagnie générale de T.S.F.—M. Denis and H. Fovet. (*Ann. Radioélect.*, vol. 6, pp. 336–350, October, 1951. *Onde élect.*, vol. 31, pp. 384–395, October, 1951.) A 3.750-mcps quartz crystal controlled frequency is multiplied to 270 mcps (F_0) and applied to a Type-XM 14 klystron multiplier. No dulation is applied to the klystron multiplier at a frequency $F_0/3$, so that integral multiples $pF0/3$ are available where $24 \leq p \leq 48$. Further subdivision is attained by interpolation, using a standard variable frequency oscillator. The normal range of 2,160–4,320 kmcps can be extended to 0.6–30 kmcps by frequency multiplication methods. Accuracy is within about 5 parts in 10^4, but can be improved to 1 part in 10^4. Auxiliary equipment available enables the meter to be used for calibrating wavemeters, for frequency measurement, and for estimation of the depth of parasitic modulation.

621.317.761 1378

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621.317.79:621.396.611.21.012.8 1379

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621.317.79:621.396.67.029.64 1380

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621.317.792:621.396.645.018.424 1381

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621.396.615.11:534.844.1 1382

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621–52:621.389 1386

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621.384.6 1388

The Linear Electron Accelerator as a Pulsed Neutron Source—B. T. Feld. (*Nucleonics*, vol. 9, pp. 51–57; October, 1951.)

621.385.833 1389

Study of Low-Power Electrostatic [electron] Lenses: the Second Approximation—F. Bertein. (*Jour. Phys. Radium*, vol. 12, pp. 25A–31A; October, 1951.) Continuation of work noted in 3082 of 1951.

621.385.833 1390

An Independent Electrostatic [electron] Lens with Minimum Elliptical Astigmatism—E. Regenstreif. (*Jour. Phys. Radium*, vol. 12, pp. 760–761; July/September, 1951.)

621.385.833 1391

Theory of the Elliptical Electrostatic

the current decreases and then changes its sign. An explanation of the phenomena is advanced.

546.824–31:621.315.612.4 1344

Electric and Thermoelectric Properties of Partially Reduced (Blue) Titanium Dioxide—B. I. Boltaks, F. I. Vasenin and A. E. Salunina. (*Zh. Tekh. Fiz.*, vol. 21, pp. 532–546; May, 1951.) Experiments were conducted for determining, within a wide range of temperatures, the specific conductivity and thermoelectric emf of samples of TiO_2 at various degrees of reduction. The results obtained are explained from the standpoint of modern theories on the conductivity of semiconductors.

546.–31:621.315.612.4 1345

Dielectric Properties of Various Preparations of Titanium Dioxide—I. N. Belyaev, N. S. Novosil'tsev, A. L. Khodakov and M. S. Shul'man. (*Zh. Tekh. Fiz.*, vol. 21, pp. 547–551; May, 1951.) The effect of impurities on the dielectric properties of TiO_2 has been studied on six different samples. The results obtained are shown in curves and tables.

548.0:537.228.1 1346

Growing Piezoelectric Crystals—H. Mosaner and M. Wurl. (*Arch. elekt. Übertragung*, vol. 5, pp. 463–467; October, 1951.) An examination is made of factors influencing the development of single crystals of commonly used types, grown from an aqueous solution whose temperature is reduced under controlled conditions. A constant rate of growth of only a few millimeters per day is desirable. The temperature reduction must be determined in relation to the rate of growth to give the required size.

548.0:538.114 1347

Interatomic Distances and Ferromagnetism in Spinels—R. S. Weisz. (*Phys. Rev.*, vol. 84, p. 379; October 15, 1951.) Discussion with particular reference to the hypothesis that spin interaction is inversely proportional to the distance from a metal ion to a nearest neighbor (i.e., an oxygen ion) and thence to another metal ion.

621.315.612.6:537.311 1348

Breakdown and Electric Conductivity of Glass—K. J. Keller. (*Physica, 's Grav.*, vol. 17, pp. 511–530; May, 1951.)

621.315.616 1349

Synthetic Materials—P. Domin. (*Fernmeldetech. Z.*, vol. 4, pp. 461–467; October, 1951.) Review of various plastics and their derivation. Mechanical and electrical properties are tabulated.

621.396.611.21:549.514.51 1350

Frequency Shift of Piezoelectric Oscillations of Quartz under High Pressure—A. Michels and J. P. Pérez. (*Physica, 's Grav.*, vol. 17, pp. 563–564; May, 1951.) Measurements are reported on two quartz resonators, an AT cut with frequency 942 kcps and a BT cut with frequency 6.04 mcps at atmospheric pressure. The slope of the frequency/pressure curve is positive for the AT and negative for the BT cut.

666.1 1351

Technical Control in Glass Manufacture—J. H. Partridge and E. Preston. (*GEC Jour.*, vol. 18, pp. 212–220; October, 1951.) Descriptions are given of an automatic method of controlling the temperature of a large glass-melting tank furnace, a photoelectric pyrometer for measuring accurately the temperature of glass as it is fed to machines, and an instrument for measuring the diameter of glass tubing during the drawing process. The results of applying these controls are demonstrated.

MATHEMATICS

517.6:621.396.611.1 1352

The Impulse Integral, a Counterpart of the Duhamel Step Integral—R. Lueg, M. Päsler and W. Reichardt. (*Ann. Phys.* (*Lpz*), vol. 9, pp. 307–315; October 15, 1951.) If a time-variable disturbing force acts on a linear system and G is one of the system parameters, the function $G(t)$, representing the time dependence of G under the action of the disturbing force, may be represented by the Duhamel integral. It may also be represented in a much simpler way by an expression whose integrand differs from the Duhamel integral in that the effect of the step function (unit step) is replaced by the effect of a delta function (needle impulse). Hence the term impulse integral. It is the simplest form of many possible expressions, and leads, under stated conditions, to a multiplication law for finding Fourier components.

517.9 1353

On a Relaxation Method for Eigenvalue Problems—S. H. Crandall. (*Jour. Math. Phys.*, vol. 30, pp. 140–145; October, 1951.)

517.9 1354

The Application of Relaxation Methods to the Solution of Differential Equations in Three Dimensions: Part 1—Boundary Value Potential Problems—D. N. de G. Allen and S. C. R. Dennis. (*Quart. Jour. Mech. Appl. Math.*, vol. 4, part 2, pp. 199–208; June, 1951.)

518.5 1355

The Logarithmic-Complex-Number Plane and the Complex-Number Calculator—W. de Beauclair. (*Z. Ver. Dtsch. Ing.*, vol. 93, pp. 955–957; October 21, 1951.) The complex-number plane with rectangular- and polar-coordinate network can be so transformed that multiplication and division of complex numbers reduce to simple addition and subtraction. This forms the basis of a simple calculator.

681.142 1356

The Binac—A. A. Auerbach, J. P. Eckert, Jr., R. F. Shaw, J. R. Weiner and L. D. Wilson. (Proc. I.R.E., vol. 40, pp. 12–29; January, 1952.) A comprehensive description of a high-speed electronic digital computer. The Binac consists of a main computing section, input-output equipment, and a mercury delay line of 512-word capacity. It is a combination of a few basic circuits, classified as diode matrices, switching gates, and reshaping gates; the latter allow a "pulse-envelope" representation of information. Crystal diodes are used in switching and gating. Crystal gates are used with electric delay lines to form a serial binary adder. Input data are supplied to the memory from a keyboard or magnetic tape through a synchronizer, which is also used to transfer data from the memory to magnetic tape or an electric typewriter.

681.142 1357

Logical Description of some Digital-Computer Adders and Counters—H. J. Gray, Jr. (Proc. I.R.E., vol. 40, pp. 29–33; January, 1952.) Description of some of the circuits incorporated in the EDVAC computer.

681.142:517.942.9 1358

A Three-Dimensional Electrical Potential Analyser—S. C. Redshaw. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 291–295; October, 1951.) A resistance network is used in an electrical-analogue method of solving problems involving the three-dimensional form of Laplace's equation.

681.142:621.392.5 1359

Magnetostriction Storage Systems for a High-Speed Digital Computer—Millership, Robbins and De Barr. (See 1200.)

517.63 1360

Die zweidimensionale Laplace-Transformation, eine Einführung in ihre Anwendung zur Lösung von Randwertproblemen nebst Tabellen von Korrespondenzen (The Two-Dimensional Laplace Transformation, an Introduction to its Application in the Solution of Boundary Problems, with Tables of Transforms). [Book Review]—D. Voelker and G. Doetsch. Publishers: Birkhäuser, Basle, 1950, 259 pp., bound, 43 Swiss francs; paper-covered, 39 Swiss francs. (*Z. angew. Phys.*, vol. 3, p. 398; October, 1951.) "The book is written just as the scientific practicing engineer requires it."

519.2 1361

Introduction to the Theory of Probability and Statistics. [Book Review]—N. Arley and K. R. Buch. Publishers: Chapman & Hall, London, 1950, 236 pp., 32s. (*Beama Jour.*, vol. 58, p. 340; October, 1951.) "The treatment given to the subject matter is clear and concise, wisely illustrated with actual fully worked examples, many of a numerical nature."

MEASUREMENTS AND TEST GEAR

389.6 1362

Recent Developments and Techniques in the Maintenance of Standards—M. M. Postgate. (*Nature* (Lond.), vol. 168, pp. 594–597; October 6, 1951.) Report of a symposium at the National Physical Laboratory, 21st-22nd May 1951. Most of the standardizing laboratories of Europe, the Commonwealth, and North America were represented and also many research and industrial organizations in Britain. An outline is given of the various subjects discussed. The full proceedings are to be published shortly by H.M. Stationery Office

529.78 1363

A Precision Radio Time-Signal System—D. W. R. McKinley and B. E. Bourne. (*Canad. Jour. Tech.*, vol. 29, pp. 428–434; October, 1951.) Accurate "seconds" pulses are produced from a 100-kcps quartz-crystal frequency standard, coded, and radiated from a 229-mcps transmitter with line-of-sight coverage. The average error in the interval between two pulses is 10^{-7} sec. at the source, or 10^{-6} sec. as observed at distant stations. The system has been used for continuous automatic time recording at the various stations of a meteor-observation network.

621.317.31:621.396.645 1364

Reduction of Measuring Time of Low-Current Amplifiers without Reduction of Sensitivity—J. Taieb. (*Compt. Rend. Acad. Sci.*, Paris), vol. 233, pp. 785–787; October 8, 1951.) The time constant of the amplifier input circuit is reduced by eliminating stray capacitance associated with a high input resistance; this is done by immersing the resistor in an appropriate electrostatic field. In one experimental arrangement the resistor was inserted through a pair of parallel plates of which one was earthed while the other was connected to a negative feedback circuit. Measuring time is so short that drift effects are eliminated. Some numerical results obtained with different values of input resistance are tabulated.

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A Review of Progress in Electric Furnaces—D. M. Dovey and I. Jenkins. (*GEC Jour.*, vol. 18, pp. 194–211; October, 1951.) Descriptions are given of both well-known and novel types of furnace operated by rf, arc and resistor heating.

621.384.6 1388

The Linear Electron Accelerator as a Pulsed Neutron Source—B. T. Feld. (*Nucleonics*, vol. 9, pp. 51–57; October, 1951.)

621.385.833 1389

Study of Low-Power Electrostatic [electron] Lenses: the Second Approximation—F. Bertein. (*Jour. Phys. Radium*, vol. 12, pp. 25A–31A; October, 1951.) Continuation of work noted in 3082 of 1951.

621.385.833 1390

An Independent Electrostatic [electron] Lens with Minimum Elliptical Astigmatism—E. Regenstreif. (*Jour. Phys. Radium*, vol. 12, pp. 760–761; July/September, 1951.)

621.385.833 1391

Theory of the Elliptical Electrostatic

[electron] Lens for Transgaussian Conditions—É. Regenstreif. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 854–856; October 15, 1951.) The theory previously given (2973 of 1951) for gaussian conditions is extended to cover transgaussian ray paths (2314 of 1950). See also *Ann. Radioélect.*, vol. 6, pp. 299–317; October, 1951.

621.385.833 1392

Some Formulae of Electron Optics—M. Bernard. (*Jour. Phys. Radium*, vol. 12, pp. 761–762; July/September, 1951.) Little used formulas of geometrical optics, due to Cotes and others, are found useful for calculations on electron lenses.

621.385.833 1393

Gaussian Components in the Electrostatic Lens formed by Two Coaxial Cylinders of Equal Diameter—P. Grivet and M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 788–790; October 8, 1951.)

621.385.833 1394

A New Mathematical Model of an Electron Lens—P. Grivet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 921–923; October 22, 1951.) A formula is developed which expresses the axial variation of the field more correctly than Glaser's formula (2205 of 1941) for cases where the decrease of the field is due to screening.

621.385.833 1395

On a Magnetic Electron-optical System Analogous to a Planoconcave Lens—I. L. Tsukkerman. (*Zh. Tekh. Fiz.*, vol. 21, pp. 599–600; May, 1951.)

621.385.833 1396

Investigation of Field Distribution in Symmetrical Electron Lens—J. R. Shah and L. Jacob. (*Jour. Appl. Phys.*, vol. 22, pp. 1236–1241; October, 1951.)

621.387.4 1397

Relative Sensitivities of Windowless and End-Window Counters—W. L. Graf, C. L. Comar and I. B. Whitney. (*Nucleonics*, vol. 9, pp. 22–27; October, 1951.)

621.387.4:549.211 1398

Some Aspects of the Counting Properties of Diamond—G. P. Freeman and H. A. Van der Velden. (*Physica, 's Grav.*, vol. 17, pp. 565–572; June, 1951.)

621.387.424 1399

Geiger Counters with Xenon-Oxygen Filling.—O. Riedel. (*Z. Naturf.*, vol. 5a, pp. 331–332; June, 1950.)

778.37:621.383.001.8:535.61–15 1400

The Performance of Image Convertors as High-Speed Shutters—R. C. Turnock. (*Proc. IEE*, Part II, vol. 98, pp. 635–641; October, 1951.) Characteristics of two types of image converter are discussed and an explanation of their operation as shutters is suggested. See also 2250 of 1951 (Hogan).

621.38.001.8 1401

Theory and Application of Industrial Electronics [Book Review]—J. M. Cage. Publishers: McGraw-Hill, New York, 1951, 290 pp., 40s. 6d. (*Electronic Eng.*, vol. 23, pp. 409–410; October, 1951.) "One of the well-known ... Electrical and Electronic Engineering series edited by Terman."

PROPAGATION OF WAVES

538.566.2 1402

The Propagation of Electromagnetic Waves in an Atmospheric Duct—T. Kahan and G. Eckart. (*Z. Naturf.*, vol. 5a, pp. 334–342; June, 1950.) See 1465 of 1951.

621.396.11 1403

A Geometric Interpretation of the II-Wave and Coupling Factor in Ionospheric Long-Wave Theory—N. Davids. (*Jour. Geophys. Res.*, vol. 56, pp. 611–612; December, 1951.) Analysis shows that the II waves are simply the components of the normal mode along the principal axes of a system of rectangular coordinates in the complex plane. The coupling factor for the two normal modes of propagation corresponding to the cw solution of the Appleton-Hartree equations is found to represent the space rate of twist of the polarization axes.

621.396.11 1404

Spanish Method for the Prediction of Optimum Working Frequencies at Any Distance—R. Gea Sacasa. (*Rev. Telecomunicación* (Madrid), vol. 7, pp. 3–13; September, 1951.) Paper presented before the C.C.I.R., June 1951. See 2614, 2615, 2877 and 2878 of 1950.

621.396.11:551.510.535 1405

Group Velocities and Group Heights from the Magneto-ionic Theory—D. H. Shinn and H. A. Whale. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 85–105; 1952.) The group velocity in the ionosphere above southeastern England is calculated for the ordinary and for the extraordinary wave over a wide range of frequencies; results are presented in the form of curves. The number of values calculated is sufficient for group delays at vertical incidence to be computed. Corresponding values for various angles of dip are used to obtain the group-height/frequency curves for various magnetic latitudes, for the ordinary ray incident vertically on a layer with parabolic distribution of ionization. It is shown that estimates of layer thickness neglecting the magnetic field are correct for the magnetic equator but are too high for other latitudes by amounts rising to 53 per cent at magnetic latitude 62°; estimates of height of maximum ionization neglecting the magnetic field are approximately correct. A new method of estimating these parameters is proposed. The effect of the field on oblique-incidence propagation is also discussed.

621.396.11:551.510.535 1406

Contribution to the Study of the Electron Distribution in the Ionosphere and of the Absorption of Short Waves—Argence, Mayot and Rawer. (See 1300.)

621.396.11:551.510.535 1407

Radio-Wave Propagation at Oblique Incidence including the Lorentz Polarization Term—Kelso. (See 1302.)

621.396.11.029.6 1408

Tropospheric Propagation beyond the Horizon—J. Feinstein. (*Jour. Appl. Phys.*, vol. 22, pp. 1292–1293; October, 1951.) Recent calculations of the field strength of vhf waves, based on a smooth-earth diffraction theory, give results far below the observed values in the shadow region, even when refraction produced by the standard atmosphere is allowed for. But in these calculations the contribution made by partial reflections, caused by the gradient of refractive index in the atmosphere, has been neglected. A mathematical treatment which accounts for these reflections is outlined, and results deduced from it are shown graphically.

621.396.11.029.62 1409

The Anomalous Propagation of Radio Waves in the 1–10-Metre Band—F. H. Northover. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 106–129; 1952.) Theory is developed according to which anomalous propagation in the 1–10-m band is satisfactorily accounted for by high-level subsidence inversions.

621.396.11.029.62:621.396.65 1410

Metre-Wave Links beyond the Optical Range—L. Sacco. (*Poste e Telecomunicazioni*, vol. 19, pp. 465–471; October, 1951.) A comparison of various methods of estimating the performance of such links. Two main types of formulas are distinguished, (*a*) those based on diffraction at a sphere, and (*b*) those based on optical diffraction at an edge. These formulas are assembled in an appendix. Using data obtained from the Monte Cavo-Monte Serpeddi link, pairs of values of field strength and modified radius (R_{10}) of the earth are calculated. The distributions of R_{10} for all methods of type (*a*) are such that very favorable atmospheric conditions must be assumed, while those for methods of type (*b*) give much more probable results. Further evidence supporting this conclusion is adduced from results on the Grasse-Calenzana link, the attenuation calculated by type (*b*) methods being in much closer agreement with measurements.

621.396.11.029.64 1411

Measurements of the Parameters Involved in the Theory of Radio Scattering in the Troposphere—C. M. Crain and J. R. Gerhardt. (PROC. I.R.E., vol. 40, pp. 50–54; January, 1952.) A brief discussion of the recording apparatus used for measurements of the fluctuations of refractive index and temperature in the lower atmosphere, particularly over the height range of 10 inches to 50 feet above the ground, and discussion of the problem of atmospheric turbulence at such levels. Records of refractive index and temperature fluctuations for various atmospheric conditions are given.

621.396.812.029.64 1412

Volume Integration of Scattered Radio Waves—A. H. Lagrone. (PROC. I.R.E., vol. 40, p. 54; January, 1952.) The general equation for radio scattering developed by Booker and Gordon (1757 of 1950) is extended to a volume-integral equation, which gives the total scattered-power density per unit cped at a receiver point relative to the power radiated per unit solid angle by an isotropic source. The propagation conditions assumed are stated. See also 2525 of 1951 (Lagrone et al.).

621.396.812.3.029.64 1413

Selective Fading of Microwaves—A. B. Crawford and W. C. Jakes, Jr. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 68–90; January, 1952.) Propagation experiments over two line-of-sight paths 22.8 and 17 miles long were designed to reveal the processes which cause the fading observed on clear calm nights. The incident wave fronts were explored with a narrow-beam scanning antenna to determine the angle of arrival of 24-kmcps waves, and the transmission characteristics were observed by a frequency-sweep technique. It was concluded that severe fading is the result of multiple-path transmission in which several components may arrive at the receiver at angles up to about 0.8° above the normal day-time angle of arrival and with path differences up to about 10 feet.

621.396.812.3.029.64:621.3.015.7 1414

Propagation Studies at Microwave Frequencies by means of Very Short Pulses—O. E. DeLange. (*Bell Sys. Tech. Jour.*, vol. 31, pp. 91–103; January, 1952.) Pulses of 4-kmcps radiation, of duration about 3 mμs, were transmitted over a 22-mile path to determine the effects of the transmission medium on the propagation. The results indicate that during fading periods transmission by two or more paths occurs, with path differences up to about 7 feet. The resulting distortion of the pulses puts a lower limit on pulse length and spacing in a pulse transmission system.

RECEPTION

621.396.621 1415

Radio without Valves—(*Elec. Rev.* (Lond.), vol. 149, pp. 599–600; September 21, 1951.) Ge triodes replace thermionic tubes in a simple experimental G. E. C. receiver which combines robustness with sufficient output to operate a loudspeaker.

621.396.621 1416

A General Theory for Frequency Discriminators containing Null Networks—J. L. Stewart. (Proc. I.R.E., vol. 40, pp. 55–57; January, 1952.) Generalized theory is presented which is applicable to the design of discriminators including null networks in which, neglecting second-order effects, coupling capacitors, and the like, a phase difference of a multiple of π radians is maintained at all frequencies.

621.396.621.54 1417

Tracking of Superheterodyne Receivers—(*Radio Tech. Dig., Édn. franç.*, vol. 5, pp. 243–270; 1951.) French adaptation of 482 of March (de Koe). 59 references.

621.396.621.54:621.3.015.7 1418

Pulse Response of A.M. Receiver—R. Kitai. (*Wireless Eng.*, vol. 29, pp. 15–18; January, 1952.) An analysis is made of the transient response of a simple superheterodyne receiver with inductive coupling to an open-ended antenna of effective height about 4 m. The magnitude of the response is shown to depend not only on the amplitude of the input pulses, but also on their duration in relation to the frequency to which the receiver is tuned; hence the receiver can be used to determine the duration of the pulses.

621.396.822 1419

Voltage Peaks of Fluctuation Noise—V. I. Bunimovich. (*Zh. Tekh. Fiz.*, vol. 21, pp. 625–636; June, 1951.) The question, as to how frequently a random value such as a fluctuation noise voltage exceeds a predetermined level, is discussed under the following headings: (*a*) average number of peaks; (*b*) peaks of the amplitude envelope; (*c*) average duration of peaks; (*d*) average frequency and mean square frequency of the spectrum.

621.396.822 1420

Observed Groups of Peaks of Electrical Fluctuations—V. I. Bunimovich. (*Zh. Tekh. Fiz.*, vol. 21, pp. 637–646; June, 1951.) In 1419 above separate random peaks were considered. In the majority of cases peaks occur in groups, and only as such can be observed on the screen of an oscilloscope. Formulas are derived for determining the average number of peaks and groups of peaks, and the average duration of peaks. In the derivation of these formulas it is assumed that the fluctuations have passed through a narrow-band filter, i.e. that the envelope of the fluctuation process is approximately sinusoidal.

621.396.828:621.396.619.13 1421

Reduction of Interference in F.M. Receiver by Feedback across the Limiter—R. M. Wilmotte. (Proc. I.R.E., vol. 40, pp. 34–36; January, 1952.) Vector analysis of the phase relations in a FM receiver with feedback across the limiter shows that interference from an unwanted signal, of intensity differing little from that of the wanted signal, can be substantially suppressed.

STATIONS AND COMMUNICATION SYSTEMS

517.512.2:621.39.001.11:621.396.67.012.71 1422

Fourier Analysis and Negative Frequencies—I. J. Shaw. (*Wireless Eng.*, vol. 29, pp. 3–12; January, 1952.) The relation between the time function $f(t)$, expressing a fluctuation, and the frequency components $s(\omega)$ is examined; $f(t)$ is given by an integral in which ω ranges from $-\infty$ to $+\infty$. Neglect of the negative part of the spectrum leads in certain cases to errors in application of the modulation theorem of Fourier analysis. A graphical method is developed for obtaining the frequency spectrum of a finite wave train, in which the contribution of the negative frequencies is taken into account; simplification is achieved by introducing an auxiliary function with rectangular envelope. The method is applicable to the calculation of the polar diagram of antennas by Fourier analysis of the distribution of current across the antenna aperture; negative angles then take the place of negative frequencies.

621.39.001.11 1423

On the Definition of Information—E. Reich. (*Jour. Math. Phys.*, vol. 30, pp. 156–161; October, 1951.) A definition of information is proposed which postulates an invariance under certain types of transformations. For a restricted class of admissible definitions these postulates imply that Shannon's formulation is the only possible one.

621.39.001.11 1424

Prediction and Entropy of Printed English—C. E. Shannon. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 50–64; January, 1951.) A method of estimating the entropy and redundancy of a language is described, based on results of experiments in predicting the next letter of a text from knowledge of the preceding text. Some properties of an ideal predicting system are developed.

621.39.001.11 1425

The Concepts of Information and Transmission-Channel Capacity in Communication Engineering—H. Weber. (*Bull. schweiz. elektrotech. Ver.*, vol. 42, pp. 749–754; September 12, 1951. In German.) A method is shown for calculating the quantity of information in a long text; the discussion is based on Shannon's method (1417) for deriving a text of reduced length which can be used for transmission and expanded to its original length at the receiver.

621.395.44:621.315.052.63 1426

British Developments and Applications of Carrier Current Principles for Operating Requirements of Power Utilities—W. D. Goodman. (*GEC Jour.*, vol. 18, pp. 229–236; October, 1951.) A.I.E.E. Summer General Meeting paper. Details are given of wide band coupling equipment for carrier-current working over power lines; the advantages of interphase coupling over phase-to-ground coupling are explained. Ssb operation is used in order to reduce noise. Recent British types of communication and relaying equipment are described.

621.396.619.13 1427

Investigations of Frequency [-modulation systems with] Negative Feedback—J. Hacks. (*Arch. elekt. Übertragung*, vol. 5, pp. 441–446; October, 1951.) A discussion of the use of negative feedback to reduce distortion in FM transmitters and receivers. By confining consideration to small values of distortion the calculation can be simplified in comparison with earlier methods [see 3120 of 1939 (Chaffee) and 3548 of 1949 (Panter and Dite)] and phase shifts due to the finite group transmission time can be taken into account. The effect on the over-all distortion of nonlinearities in the feedback loop is examined. A circuit for equalizing the feedback over the modulation-frequency range is described. With a feedback factor>2 a significant improvement can be obtained in the effective selectivity of a FM receiver.

621.396.65+621.396.97 1428

Bases of Calculation for the Design of Radio Communication Systems—H. Herzan. (*Radio Tech.* (Vienna), vol. 27, pp. 423–427; October, 1951.) The relative levels of signals and noise desirable for radio broadcasting, telephony and telegraphy systems are discussed. Graphs are reproduced showing noise spectra, ground-wave attenuation, antenna height factor, and space-wave absorption curves for medium and short waves. The use of a logarithmic distance/attenuation scale simplifies the application of the graphs.

621.396.65.029.6 1429

Nonlinear Crosstalk in Multichannel Systems during Transmission over U.S.W. F.M. Links, as compared with the Specifications for Cable Connections—E. Kettel. (*Telefunken Ztg.*, vol. 24, pp. 163–168; October, 1951.) Modulation in carrier-frequency systems is considered, and the distortion in an u.s.w. FM installation is expressed in terms of crosstalk. FM u.s.w. links do not at present equal the performance of carrier-frequency cables, but can bridge useful distances while complying with C.C.I.F. requirements. A comparison between FM and time modulation shows the superiority of the former for a large number of channels from the point of view of signal/noise ratio.

621.396.65.029.62 1430

V.H.F. Radio Multichannel Carrier Telephone Circuits in Colombia—L. C. Simpson, H. J. B. Nevitt and E. J. Eriksen. (*Ericsson Rev.*, vol. 28, pp. 62–72; 1951.) The main features are described of a triple link connecting Bogota and Medellin, distant 250 km apart. FM broadcasting equipment is used; the frequency range is 70–88 mcps.

621.396.65.029.63/.64 1431

The Planning of Beam Links in the Decimetre and Centimetre Wavebands—K. O. Schmidt. (*Telefunken Ztg.*, vol. 24, pp. 129–139; October, 1951.) General considerations and expected future developments are enumerated, and planning details for the FM wide-band 15-cm television link between Hamburg and Cologne are briefly considered. This link is expected to be in experimental operation early in 1952. The importance of relay-station tower construction and suitable power supply arrangements, in this case mainly wind-driven generators, is stressed.

621.396.65.029.63 1432

The IDA 22 Beam-Link Equipment—G. Ulbricht. (*Telefunken Ztg.*, vol. 24, pp. 143–162; October, 1951.) A short survey of German wartime dm-wave links, which covered most of Europe and extended to North Africa, together with an account of the development of this new equipment. A 22-channel installation in experimental operation between Darmstadt and Frankfurt, with an intermediate relay station on the Feldberg, is described in detail. Pulse-phase modulation is used and the transmission frequency lies between 1,900 and 2,100 mcps. To avoid the effects of noise on the pulse flank, an auxiliary hf self-oscillator delivers a voltage low compared with the pulse voltage, but high compared with the noise voltage, to the transmitter. The pulse flank is thus modulated at the self-oscillator frequency but this frequency is rejected by the receiver. Alternative equipment for the modulation circuits uses Ge diodes. A map showing the dm-wave network of the German Post Office in 1951 is also included.

621.396.71 1433

The New Italcable Radio Transmitting Station of Torrenova—M. Suppan. (*Post e Telecomunicazioni*, vol. 19, pp. 473–481; October, 1951.) A general description covering the station lay-out and the equipment available.

621.396.97.029.62 1434
The B.B.C. Scheme for V.H.F. Broadcasting—(*BBC Quart.*, vol. 6, pp. 171–181; Autumn, 1951.) The technical reasons for the proposal to plan a vhf service are briefly outlined. Tests are described in which various forms of modulation and reception were compared from the point of view of noise and interference suppression. Laboratory experiments indicated that the level of receiver hiss and impulsive noise was markedly lower with FM than with AM, either with or without a noise limiter. Similarly, a survey of the quality of reception of transmissions from actual stations gave approximately twice the range for FM as compared with AM for the same quality of service. On this basis a FM broadcasting service has been planned for the United Kingdom; this is discussed with particular reference to common- and adjacent-channel interference and the requirements of suitable commercial receivers.

SUBSIDIARY APPARATUS

621–526 1435
Harmonic Analysis of Nonlinear Servomechanisms—J. Loeb. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 733–735; October 1, 1951.) Continuation of work noted in 502 of March.

621.311.6:621.316.72.076.7 1436
Superregulated Power Supplies—A. W. Vance and C. C. Shumard. (*Electronics*, vol. 24, pp. 109–113; December, 1951.) A description of apparatus giving supplies of 300 v at 20 a, 75 v at 6 a and −500 v at 3 a, all regulated to within 0.001 per cent. Coarse regulation is obtained in stages using grid-controlled thyratrons, which precede fine regulators incorporating servo-stabilized dc amplifiers. A calibrator unit enables comparison of any voltage output with an appropriate reference voltage, and also direct measurement of hum on a cro.

621.352 1437
Small-Size High-Voltage Battery—P. Bristeau. (*Jour. Phys. Radium*, vol. 12, pp. 74A–78A; October, 1951.) Method of manufacture and main factors affecting operation are given for a Zamboni-type battery for low-consump ion portable electronic apparatus. Diamete s 8 mm. length 60 mm. and voltage 200 v.

621.396.68 1438
Power Supply. The U.D.C. number 621.396.68 and its subdivisions, used hitherto for power supply, have been cancelled and replaced by 621.311.6 and subdivisions, which will be used in future in "Abstracts and References."

TELEVISION AND PHOTOTELEGRAPHY

621.397.24:621.315.212 1439
Television Transmission Tests on Coaxial Cable—P. Bréant and G. Fuchs. (*Câbles & Trans.* (Paris), vol. 5, pp. 325–336; October, 1951.) A brief description is given of the line and terminal equipment and of the equalization measures adopted for the coaxial-cable circuit between Brive and Montpon. 455-line tests were made over a 73-km and a 36-km loop in this circuit. The conclusion is that low-definition image transmission is feasible over the standard French coaxial cables, provided the transmission system is as nearly as possible linear and the signal/noise ratio is of the order of 5 N.

621.397.24:621.317.34.092 1440
Measurement of Differences of Group Propagation Time on a Non-looped Line—Selz and Iltis. (See 1373.)

621.397.3:517.941.91 1441
A Mathematical Analysis of an Equivalent Scanning Circuit—A. B. McFarlane and J. R. W. Smith. (*Jour. Brit. I.R.E.*, vol. 11, pp. 470–476; October, 1951.) A classical treatment of the switched LCR circuit by means of differential equations, which shows the effects of changes of the Q-factor of the circuit and leads to the formulas usually quoted. Energy fluctuations of the system in the oscillatory and non-oscillatory states are examined and a numerical example is worked out.

621.397.5 1442
The Present Position of Television [in Germany]—H. K. Ibing. (*Z. Ver. Dtsch. Ing.*, vol. 93, pp. 890–895; October 1, 1951.) Principal features of apparatus shown at the German Industries Exhibition, Berlin, 1951, are described.

621.397.5 1443
Electrical and Photographic Compensation in Television Film Reproduction—P. J. Herbst, R. O. Drew and S. W. Johnson. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 57, pp. 289–307; October, 1951.) Defects in film reproduction due to limitations in existing television systems are discussed. Of available methods of compensation, area masking as described by Yule (*Jour. Phot. Soc. Amer.*, vol. 11, pp. 123–132; March, 1945) is recommended. A circuit equivalent for the photographic mask is described.

621.397.5:535.623 1444
A New Technique for Improving the Sharpness of Television Pictures—P. C. Goldmark and J. M. Hollywood. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 57, pp. 382–396; October, 1951.) See 828 of April.

621.397.5:535.88:679.5:535.317 1445
Cast Plastic Lens—(*Mod. Plast.*, vol. 29, pp. 188–189; September, 1951.) Description of the method of producing plastic lenses, up to 22.5 inches in diameter, for the correcting elements of Schmidt large-screen projection systems.

621.397.5(083.74) 1446
Television Standards in Argentina—(*Rec. telegr. Electronica* (Buenos Aires), no. 468, pp. 589–590; September, 1951.) These differ from U. S. standards in two chief respects, the number of scanning lines, which is 625, and the number of frames per second, which is 25. The vision signals are broadcast on 175.25 mcps, the sound on 179.75 mcps.

621.397.5:535.623](083.74) 1447
Color Television—U.S.A. Standard—P. C. Goldmark, J. W. Christensen and J. J. Reeves. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 57 pp. 336–381; October, 1951.) See 831 of April.

621.397.61.029.63 1448
U.H.F. Transmitter uses Beer-Barrel Cavity—W. H. Sayer, Jr. and E. Mehrbach. (*Electronics*, vol. 24, pp. 125–127; December, 1951.) Describes a 700-mcps, 420-w experimental television transmitter incorporating an output stage consisting of six Type-2C39A tubes in parallel used as grounded-grid amplifiers. The cathode-grid and anode-grid tuned circuits of the amplifier are coaxial-line cavities. The design is based on experience with a 600-mcps transmitter in daily use since March 1950. An illustration shows an assembly using eighteen tubes in parallel.

621.397.61.029.63+621.396.67:621.397.6 1449
Ultrahigh Transmitter and Antenna Design and Application—R. G. Peters. (*TV Eng.* (N.Y.), vol. 2, pp. 14–17, 29; October, 1951.) Description of the R.C.A. TTU-1B and TTU-10A television transmitters, together with the antennas and associated equipment, for the frequency range 470–890 mcps.

621.397.611.2 1450
Television Camera Tubes: Part 1—Classification and Method of Operation of Camera Tubes—H. Bruining. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 209–225; September, 1951.) An outline of principles of operation, with individual descriptions of the orthicon, image orthicon, vidicon, iconoscope and image iconoscope.

621.397.611.2 1451
Television Camera Tubes: Part 2—Comparison of Characteristics of Camera Tubes—P. Schagen. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 227–242; September, 1951. Discussion, p. 242.) The types of tube considered are the iconoscope, the image iconoscope, the orthicon and the image orthicon. Characteristics discussed include linearity, effective exposure time, generation of spurious signals, relative ease of operation, signal/noise ratio, spectral response, stability, depth of focus, possibility of black level restoration, and sensitivity. For normal broadcasting operation, the relative sensitivities of the four types in the order given above are 1, 75, 25 and 325.

621.397.611.2 1452
Television Camera Tubes: Part 3—Electronoptical Problems of the Image Iconoscope—J. C. Francken. (*Tijdschr. ned. Radiogenoot.* vol. 16, pp. 243–256; September, 1951. Discussion, p. 257.) Problems involved in the scanning and image-forming processes are discussed. Use of a high-emission cathode such as that described in 773 of 1951 (Lemmens, Jansen and Loosjes) helps to reduce defocusing of the scanning spot due to the inclination of the scanning ray to the target. The electron-image section using a long focusing coil is described; factors causing loss of definition of the electron image are mentioned and methods of reducing their adverse effects are discussed. By using a coil split into three sections and simultaneously varying the currents in these sections, it is possible to obtain continuous variation of magnification over a 1:2 range without rotating the image.

621.397.62 1453
Ultra-speed Theater Television Optics—L. T. Sachtleben and G. L. Allee. (*Jour. Soc. Mot. Pict. Telev. Eng.*, vol. 57, pp. 425–433. Discussion, p. 433; November, 1951) The general properties of a reflection system of the Schmidt type are discussed with reference to practical design considerations and application in the theatre. The relations between focal length, projection distance and screen illumination are demonstrated, and factors affecting resolving power and fine-detail contrast are examined. The design, material and application of the ogee correcting lens are also discussed.

621.397.62 1454
An Ultra-High-Frequency Television Converter—B. F. Tyson. (*Sylvania Technologist*, vol. 4, pp. 78–80; October, 1951.) The design and performance are described of a simple, low-cost converter which enables transmissions in the band 475–890 mcps to be received on existing receivers designed for the band 174–216 mcps. Factors discussed are conversion efficiency, noise figure, frequency stability and possible interference from harmonics of the local oscillator of the vhf receiver.

621.397.621.2:535.371.07 1455
Aluminium-Backed Screens for Cathode-Ray Tubes—R. W. Dudding. (*Jour. Brit. I.R.E.*, vol. 11, pp. 455–462; October, 1951.) The ion-burn screen blemish associated with the electromagnetic cr tube can be prevented by depositing a thin Al layer on the back of the phosphor coating. The required characteristics of such layers, and methods of preparing them, are described. Further advantages of Al-backed screens are the prevention of screen

voltage saturation effects, and improved brightness and contrast.

621.397.7 1456

The Holme Moss Television Transmitting Station—D. B. Weigall. (*BBC Quart.*, vol. 6, pp. 182–192; Autumn, 1951.) A general description is given of the site and layout of buildings and equipment of the North of England television station, and comparison is made with a similar station at Sutton Coldfield [626 (Cork), 834 (Nind and Leyton) and 854 (Bevan and Page) of April]. The sources of power, vision and sound signals are described and also the methods of controlling the transmitters and monitoring the program in the control room. The vision and sound transmissions are combined before passing up the 750-foot mast to the common antenna system. A radiated power of 45 kw is obtained on the vision frequency, the nominal carrier power of the sound transmitter being 12 kw. The corresponding powers of the standby transmitters are 5kw (peak white) and 2kw. A field-strength map of the service area is given and also details of the antenna system, which is basically similar to that at Sutton Coldfield. See also *Wireless World*, vol. 57, pp. 473–474; November, 1951.

621.397.81 1457

Fringe-Area Performance Predictions—E. A. Slusser. (*TV Eng.* (N.Y.), vol. 2, pp. 18–19, 30 and 22–23, 29; October and November, 1951.) Based on data for effective transmitted power, propagation losses, receiving-antenna gain and height and receiver gain, nomograms are constructed for predicting the performance of television receivers in fringe areas.

621.397.828 1458

A Versatile TVI-less 40-Watt Transmitter—S. Fisher. (*CQ*, vol. 7, pp. 11–20; September, 1951.) Precautions described against interference with television reception by a 10–160-mcps amateur transmitter include shielding and filtering at the transmitter and various trap and filter arrangements for connection to the receiver.

621.397.828:621.365.5 1459

Curing R.F.-Heater Television Interference—P. S. Rand, J. J. Lamb and A. J. Riley. (*Mod. Plast.*, vol. 29, pp. 101–110; September, 1951.) A general survey of the problem, with a description of a sensitive wide-range wavemeter for measuring the frequency of stray radiation. Methods of preventing leakage from heating sets are described; careful attention to the design of the cabinet is necessary. Alternatively a screened room may be used for the equipment, with filter circuits in the power leads.

621.397.5 1460

Bases Techniques de la Télévision. [Book Review]—H. Delaby. Publishers: Eyrolles, Paris, 340 pp. (*Radio Tech. Dig., Éd. franç.*, vol. 5, pp. 293–294; 1951.) Companion volume to "Principes Fondamentaux de Télévision" (1218 of 1949). Deals with all parts of the television transmission chain from the camera to the picture tube; recommended to radio engineers.

TRANSMISSION

621.396.61:621.397.97 1461

Description of a 40-kW Broadcast Transmitter—A. G. Robeer and B. Swets. (*Commun. News*, vol. 12, pp. 16–32; October, 1951.) Detailed description of a transmitter for six preset frequencies between 6 and 24 mcps, incorporating "instantuners" [click gear; 2941 of 1949 (Vervest)] for all tuning elements, continuously variable inductors, and triple contacts for heavy currents.

621.396.619.13 1462

On the Operating Conditions of Reactance Circuits for realizing Very Large Reactance Variations—H. Fricke. (*Fernmeldetech. Z.*, vol. 4, pp. 458–461; October, 1951.) A pentode rectance tube is operated about a point of zero slope on the Ia/Vg characteristic. A suitable characteristic is obtained by applying a high screen voltage so that a space charge develops between screen and anode. With an EF14 tube, frequency deviations of +11.7 per cent at 1.46 mcps and ±7.9 per cent at 27.5 mcps were obtained.

621.396.619.14+621.314.26]:621.385.029.64 1463

The Travelling-Wave Valve as a Microwave Phase Modulator and Frequency Shifter—W. J. Bray. (*Proc. IEE*, Part III, vol. 99, pp. 15–20; January, 1952.) The principles of operation are discussed. It is shown theoretically and confirmed by measurement that phase modulation of about ±2.5 radians can be achieved with a typical tube operating at 4 kmcps, by injecting the modulating signal in series with the beam voltage. This principle when applied in a frequency shifter gives gain and power output only 6 db below those obtainable with the same tubes used as amplifiers. Possible applications in the transmitters and repeaters of radio relay systems are outlined.

621.396.619.232 1464

High-Efficiency Grid Modulation—L. A. Moxon. (*RSGB Bull.*, vol. 27, pp. 144–147 and 193–196; October and November, 1951.) The Taylor and Terman-Woodyard grid-modulation circuits are analyzed. The two systems share a common principle, the use of a quiescent tube to supply extra power during the positive half-cycle and at the same time to alter the load impedance seen by the power tube in such a way that the over-all efficiency is high except during the negative half-cycle, for which ordinary grid modulation is used. An important difference between the systems is that in the Taylor circuit the quiescent tube has to supply all the power at the modulation peak, whereas, in the Terman-Woodyard system, impedance-inverting networks are used to bring about an equal sharing of the load between the two tubes. Experiments carried out with both systems indicate that the Terman-Woodyard circuit is not only practicable, but is in no way inferior to the Taylor circuit. Tested with a lamp load, the Terman-Woodyard circuit gave the higher efficiency and also much better linearity. Circuit adjustments are not unduly critical.

TUBES AND THERMIONICS

537.525:621.396.822 1465

Determination of the Electron Temperature in Gas Discharges by Noise Measurements—K. S. Knol. (*Philips Res. Rep.*, vol. 6, pp. 288–302; August, 1951.) The electron temperatures of discharges in He, Ne, Ar and Xe, as deduced from measurements of available noise power and from experiments using a probe, agree with values deduced theoretically from the gas pressure and the radius of the discharge tube. Precautions necessary to match the discharge to the waveguide are described.

537.525.92:537.533.7 1466

Nullification of Space-Charge Effects in a Converging Electron Beam by a Magnetic Field—M. E. Hines. (Proc. I.R.E., vol. 40, pp. 61–64; January, 1952.) Discussion of the conditions necessary for maintaining uniform convergence of a conical electron beam in the presence of space charge. Brillouin's focusing condition (3101 of 1945) is extended to conical flow. A converging magnetic field is found necessary.

537.533 1467

Theory of Electron Emission from a Metal in an Electric Field—A. E. Glauberman and I. I. Tal'yanski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 78, pp. 661–664; June 1, 1951.)

621.383:546.289 1468

A Photovoltaic Germanium Cell—B. J. Rothlein. (*Sylvania Technologist*, vol. 4, pp. 86–88; October, 1951.) Practical determinations of the photovoltaic characteristics of experimental cells are discussed. Factors affecting speed of response and sensitivity are shown to be related to the lifetime of the hole-electron pairs created.

621.385.029.62/.63 1469

Traveling-Wave Amplification by means of Coupled Transmission Lines—W. E. Mathews. (Proc. I.R.E., vol. 40, p. 64; January, 1952.) Corrections to paper abstracted in 867 of April.

621.385.029.62/.63:621.396.822 1470

Traveling-Wave Tube Noise Figure—D. A. Watkins. (Proc. I.R.E., vol. 40, pp. 65–70; January, 1952.) The optimum positions of the traveling-wave-tube circuit entrance with respect to the space-charge noise waves, and the corresponding minimum noise figures, are depicted graphically as functions of the space charge and the circuit loss. The noise figure is reduced if the electron stream is accelerated through a short nonresonant gap placed at a noise-convection-current minimum. A similar reduction is given by two velocity jumps, a deceleration at a noise-current maximum followed by an acceleration at a noise-velocity maximum. Measurements taken on an experimental tube of this type agree with the theoretical predictions.

621.385.032.216 1471

The Electronic Temperature of Oxide Cathodes: Interpretation of Experimental Results—H. Dormont. (*Jour. Phys. Radium*, vol. 12, pp. 710–716; July/September, 1951.) Experiments reported by Champeix (2068 of 1950) indicating a difference between thermodynamic and electronic temperature are discussed. The results are consistent with the existence at the cathode surface of a potential barrier of step form; such a barrier would also account for the numerical value of the coefficient A in Richardson's equation.

621.385.032.216:537.311.33 1472

***P-N* Transition of an Oxide-Coated Cathode**—Y. Ishikawa, T. Sato, K. Okumura and T. Saski. (*Phys. Rev.*, vol. 84, pp. 371–372; October 15, 1951.) *P*-type conduction was found for a (BaSr)O cathode in an oxygen atmosphere, as opposed to *n*-type conduction in vacuo. Hall effect and conductivity measurements for a wide range of oxygen pressures show that (BaSr)O has the characteristics of a semiconductor.

621.385.032.42 1473

The "Grip-O-Matic" Water Jacket for Large Water-Cooled Transmitting Valves—A. G. Robeer and W. L. Vervest. (*Commun. News*, vol. 12, pp. 10–14; October, 1951.) Description of a water jacket designed so that no tools are required for mounting or demounting the tube.

621.385.3 1474

Theory and Calculation of the E. S. Penetration Factor of Planar and Cylindrical Triode Valves, assuming Two-Dimensional Potential Distributions—M. Landsberg. (*Z. angew. Math. Phys.*, vol. 2, pp. 375–393; September 15, 1951.) General formulas are derived using the Weierstrass functions; the calculation leads to an infinite system of linear equations, which can be solved by using certain power series. The relation between the formulas deduced here and approximations previously derived is shown.

621.385.3:546.289 1475

Transistors—The U.D.C. number 621.585.3: 546.289 used hitherto for transistors will be replaced by 621.314.7.

621.385.8:621.318.572].001.8 1476

Electron-Beam Switches—F. Schröter. (*Telefunken Ztg.*, vol. 24, pp. 171–186; October, 1951.) A well illustrated review of various types and their applications.

621.385.831 1477

An Experimental High-Transconductance Tube using Space-Charge Deflection of the Electron Beam—J. T. Wallmark. (Proc. I.R.E., vol. 40, pp. 41–48; January, 1952.) 1951 I.R.E. National Convention paper. The tube uses a new principle, combining current-density and deflection control; a conventional grid controls the space charge, which produces a displacement of the electron beam. Transconductances of 25 millimhos have been obtained with only 3 ma output current using an orbital-beam construction with one stage of electron multiplication. The measured equivalent noise resistance is about 900Ω and the gain-bandwidth product about 320 mcps.

621.396.615.141.2 1478

The Electron Theory of the Planar Magnetron—V. M. Lopukhin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 505–515; May, 1951.) A theoretical investigation is presented of the interaction between the em field and the electron currents in a magnetron. The presence of the electron stream causes a rise in the resonance frequencies. The necessary conditions for the excitation of the magnetron are considered.

621.396.615.141.2 1479

The Electron Theory of a Centimetre-Wave Decelerator—V. A. Lopukhin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 516–526; May, 1951.) The properties of a magnetron are considered for the limiting case when the spacing of the slots tends to equality with the slot width. This system, in the presence of an electron stream, acts as a complex filter with alternate pass and stop bands. Under certain conditions the direct wave passing through the system may be split up into three components, of which one will have an amplitude increasing exponentially with co-ordinate z. Thus the system can be used for amplification of microwave signals.

621.396.615.141.2 1480

Self-Excitation of a Decelerating System—E. I. Vasil'ev and V. M. Lopukhin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 527–531; May, 1951.) The decelerating effect of the split-anode magnetron is considered; discussion is limited to the case of π-oscillations with the electron stream occupying the whole space between cathode and anode (Fig. 1). An equation (6) is derived for determining the frequencies of oscillations. The solution of this equation gives the necessary conditions for the excitation of the system and also determines the variation of the frequencies depending on the parameters of the magnetron and of the electron stream.

621.396.615.142.2:621.3.012.8 1481

Equivalent Networks of Klystrons—S. Uda and J. Ikeuchi. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 117–129; 1950.) The induced current due to an electron stream flowing through a gap is calculated, taking account of the voltage which appears across the gap. From this the internal admittance of a klystron is deduced and equivalent circuits are derived for double-cavity and reflex klystrons, which assist in explaining their operation and also in their design.

621.396.615.142.2.029.64 1482

Ground-Transmitter Klystron for Air Navigation—V. Learned. (*Electronics*, vol. 25, pp. 136–165; January, 1952.) Description of a tube for operation at 9.3 kmcps, with peak power of 7.5 kw and average power output of over 200w. A permanent magnet is used for beam focusing.

621.385 1483

Radio Valve Data. [Book Review]—Publishers: Iliffe & Sons, London, 1951, 80 pp., 3s 6d. (*Electronic Eng.*, vol. 23, p. 410; October, 1951.) New edition of the *Wireless World* reference book giving the main characteristics of over 2,000 types of British and American tubes, and over 100 cr tubes.

621.385.029.6 1484

Rundfunkröhren. Band 1951, Eigenschaften und Anwendung der neuen UKW-Röhren (Broadcasting Valves. 1951 Volume. Properties and Application of New U.S.W. Valves). [Book Review]—L. Ratheiser. Publishers: Regeliens Verlag, Berlin 1951, 128 pp., 16 DM. (*Z. angew. Phys.*, vol. 3, p. 400; October, 1951.) Details of receiver tubes and circuits; complementary to the earlier volume (1835 of 1950).

MISCELLANEOUS

061.3 1485

I.R.E.-U.R.S.I. Fall Meeting, Cornell University, Ithaca, N. Y., October 8–10, 1951—(Proc. I.R.E., vol. 40, pp. 106–110; January, 1952.) Summaries are given of 43 technical papers presented.

621.39.001 1486

Post Office Research—(*Electrician*, vol. 147, pp. 1017–1019; October 5, 1951. *Elec. Times*, vol. 120, pp. 609–610; October 4, 1951.) A brief account of the work undertaken at the Dollis Hill Research Station.