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.
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.
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.
Look to

**FLUOROFLEX**

for

**TEFLON**

with the

optimum performance

you're looking for

"Teflon" powder is converted into Fluoroflex-T rod, sheet and tube under rigid control, on specially designed equipment, to develop optimum inertness and stability in this material. You can be sure of ideal, low loss insulation for uhf and microwave applications... components which are impervious to virtually every known chemical... and serviceability through temperatures from \(-90^\circ F\) to \(+500^\circ F\).

Produced in uniform diameters, Fluoroflex-T rods feed properly in automatic screw machines without the costly time and material waste of centerless grinding. Tubes are concentric — permitting easier boring and reaming. Parts are free from internal strain, cracks, or porosity. This means fewer rejects, longer service life.

Mail in the coupon for more data.

**RESISTOFLEX**

*Du Pont trade mark for its tetrafluoroethylene resin. Fluoroflex is a Resistoflex registered trade mark for products made from fluorocarbon resins.*

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 Klembeck, Mgr.
    - 921 E. 8th 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, III.

  - **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. Haines
    - 1430 Gasco Drive, Dayton 3
Several Things Less to Worry About... When You Specify Synkote Coax Cable

Attenuation, impedance, shielding, insulation, velocity of propagation, all the worrisome wire factors affecting your final signal are dependably constant in SYNKOTE Coaxial Cables.

Manufactured to 10 standard specifications, SYNKOTE Coax Cables are available in impedances from 50 to 300 ohms...insure minimum attenuation and maximum dependability at all frequencies and under most conditions.

For specifications other than standard, our engineering service department will be glad to work with you. Write today—your inquiry will be given prompt attention.

SYNkote

DEPENDABLE Coaxial Cable
"Made by the mile — tested by the inch"

PLASTOID CORPORATION
42-61 24th STREET, LONG ISLAND CITY 1, NEW YORK

HOOK-UP WIRE • AIRCRAFT CABLE • TV WIRE • COAXIAL CABLE • NYLON JACKETING • HIGH TEMPERATURE WIRE • MULTI-CONDUCTOR CABLES
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.

ASSAMLY 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.

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.

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.

BELL TELEPHONE LABORATORIES
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
- Permeability Tuning Devices
- Low-Loss Inductors
- Saturable Core Reactors
- Horizontal Output Transformers
- Deflection Yokes
- Telephone Loading Coils

When your drawings call for Ferroxcube 3C cores for your TV deflection yokes and horizontal output transformers, you can forget about procurement problems. These ferrite cores are nickel-free . . . and delivery will be made exactly as scheduled by you!

Improved temperature stability, high saturation flux density, and high permeability are among the other advantages of Ferroxcube 3C.

Complete technical data is yours for the asking in Engineering Bulletin FC-5101A, available on letterhead requests.

Ferroxcube Corporation of America
- A Joint Affiliate of Philips Industries and Sprague Electric Co., Managed by Sprague
SAUGERTIES, NEW YORK
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.

REVERE COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

Executive Offices:
230 Park Avenue, New York 17, N.Y.

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MINIATURIZATION

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Definitely, these remarkably compact controls are the answer to your miniaturization control problems.

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.

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CLAROSTAT MFG. CO., INC., DOVER, NEW HAMPSHIRE
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Export: 25 Warren Street, New York 7, N.Y.

Proceedings of the I.R.E. June, 1952
In maintaining greatest possible accuracy in electron optical experiments, it is imperative that the beam be perfectly aligned axially.

To assure this result it is necessary to eliminate the effects of all stray fields...including the earth's magnetic field.

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.
When it comes to making a Real Saving in space...

Stackpole cup cores with their self-shielding characteristic can be mounted close to the chassis or any other metal part for maximum results in extremely close quarters. In some instances, the high Q circuits made possible through their use permit reduction in the number of tubes needed.

Standard types include numerous shapes and sizes, each available in a wide range of permeability possibilities. Highly specialized types to meet the most critical specifications can be engineered and produced from a broad background of experience in this exacting field.

Electronic Components Division
STACKPOLE CARBON COMPANY
St. Marys, Pa.

STACKPOLE Cup Cores

Other Stackpole core types include:
ALL STANDARD MOLDED IRON CORE TYPES, SIDE-MOLDED, CHoke COIL CORES, SLEEVe TYPES, Threaded TYPES and COIL FORMS . . . also Stackpole CERAMAG® CORES (FERRITES).

Write for Electronic Components Catalog RC-8
Up-to-date news of every British development

**WIRELESS WORLD.** Britain's chief technical magazine in the general field of radio, television and electronics. Founded over 40 years ago, it provides a complete and accurate survey of the newest British techniques in design and manufacture. Articles of a high standard cover every phase of radio and allied technical practice, with news items on the wider aspects of international radio. Theoretical articles by recognised experts deal with new developments, while design data and circuits for every application are published. **WIRELESS WORLD** is indispensable to technicians of all grades and is read in all parts of the world.

*Published monthly, $4.50 a year.*

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10A
ADJUSTABLE RESISTORS

Unaffected by Heat, Cold, Moisture, or Long Use

For circuits requiring a top quality adjustable resistor not affected by moisture, heat, cold, or age ... the Allen-Bradley Type J Bradleyometer is the ideal unit.

The resistor element is molded as a single piece. It is not a film or paint type of resistor. Because of its nature, the resistor can be built up to satisfy any resistance-rotation curve. After molding, the resistor is no longer affected by heat, cold, moisture, or age. There are no rivets . . . no welded or soldered connections . . . and the shaft, cover, faceplate, and other ferrous parts are made of corrosion-resistant metal. Let us send you the latest Bradleyometer data.

Allen-Bradley Co., 114 W. Greenfield Ave.
Milwaukee 4, Wis.

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FIXED & ADJUSTABLE RADIO RESISTORS

Sold exclusively to manufacturers
A NEW BALLANTINE
Sensitive Wide Band Electronic Voltmeter

To measure...
1 millivolt to 1000 volts
from...
15 cycles to 6 megacycles
with accuracy of...
3% to 3 mc; 5% above
with input impedance...
6 mmfd shunted by 11 meg

When used without probe, sensitivity is increased to 100 MICROVOLTS but impedance is reduced to 25 mmfd and 1 megohm

Featuring customary Ballantine
SENSITIVITY — ACCURACY — STABILITY

• Same accuracy at ALL points on a logarithmic voltage scale and a uniform DB scale.
• Only ONE voltage scale to read with decade range switching.
• No “turnover” discrepancy on unsymmetrical waves.
• Easy-to-use probe with self-holding connector tip and unique supporting clamp.
• Low impedance ground return provided by supporting clamp.
• Stabilized by generous use of negative feedback.
• Provides a 60 DB amplifier flat within 1 DB from 50 cycles to 6 MC.

MODEL 314
Price $265

Specifications on other Ballantine Electronic Voltmeters

<table>
<thead>
<tr>
<th>MODEL</th>
<th>FREQUENCY RANGE</th>
<th>VOLTAGE RANGE</th>
<th>INPUT IMPEDANCE</th>
<th>ACCURACY</th>
<th>PRICE</th>
</tr>
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<tbody>
<tr>
<td>300</td>
<td>10 to 150,000 cycles</td>
<td>1 millivolt to 100 volts</td>
<td>1/2 meg. shunted by 30 mmfd.</td>
<td>3% up to 100 KC; 5% above 100 KC</td>
<td>$210</td>
</tr>
<tr>
<td>302B Battery Operated</td>
<td>2 to 150,000 cycles</td>
<td>100 microvolts to 100 volts</td>
<td>2 meg. shunted by 8 mmfd. on high range and 1 meg on low ranges</td>
<td>3% from 5 to 100,000 cycles; 5% elsewhere</td>
<td>$225</td>
</tr>
<tr>
<td>305</td>
<td>Measures peak values of pulses as short as 3 microseconds with repetition rate as low as 20 per sec. Also measures peak values for sine waves from 10 to 150,000 cps.</td>
<td>1 millivolt to 1000 volts Peak to Peak</td>
<td>Same as Model 302B</td>
<td>5% on sine waves; 3% on pulses</td>
<td>$280</td>
</tr>
<tr>
<td>302A</td>
<td>10 cycles to 2 megacycles</td>
<td>100 microvolts to 100 volts</td>
<td>Same as Model 302B</td>
<td>3% below 1 MC; 5% above 1 MC</td>
<td>$235</td>
</tr>
</tbody>
</table>

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

BALLANTINE LABORATORIES, INC.
102 Fanny Road, Boonton, N.J.
Ceramic shafts for tuning condensers are chosen because they are strong and rigidly maintain the initial alignment between rotor and stator blades. Bearing races and metal collars are press fitted to the ceramic shaft, precision ground to tolerance of .0001". Other designs use ceramic shafts, fire metallized for direct soldering of rotor blades.

Miniature ceramic transformer terminal housings take advantage of ceramic stability and are metal coated for shielding. Metallic connectors are permanently bonded to the ceramic by either glaze or by soldering to metallized surface.

Ceramic coil forms have metallized mounting supports for soldering to prevent noise due to corona between insulator and mounting plate.

Our broad experience in metal-ceramic combinations is available to you on request.

50TH YEAR OF CERAMIC LEADERSHIP

AMERICAN LAVA CORPORATION
CHATTANOOGA 5, TENNESSEE
Ever try to price-tag precision?.. 

Absolute precision in a vital instrument—what's it worth?

... to the bomber pilot trusting to Kollsman instruments checked to one-tenth-thousandth of an inch for accuracy.

... to the ship's captain, banking all on the precision of his Kollsman sextant.

At times such as these, can precision ever be price-tagged? Yet its vital presence, or absence, is oft-times the margin between victory or chaos.

Today—to maintain a free, strong America—

Kollsman is devising, developing and manufacturing instruments of utmost precision; dependability and quality in the fields of:

Aircraft Instruments and Controls • Miniature AC Motors for Indicating and Remote Control Applications • Optical Parts and Optical Devices • Radio Communications and Navigation Equipment

And to America's research scientists, seeking the answer to problems of instrumentation and control—the facilities of Kollsman Research Laboratories are immediately available.

Kollsman Instrument Corporation
Elmhurst, New York
Subsidiary of
COIL PRODUCTS CO. INC.
Glendale, California

Proceedings of the I.R.E. June, 1952
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.

Electronics Division
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LONDON, ENGLAND • TORONTO, CANADA
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.

Write on company letterhead for catalog and engineering manual No. 40
Ohmite Manufacturing Co., 4862 Flourney St., Chicago 44, Illinois

... provide DEPENDABLE ELECTRICAL CONTROL
## VOLTAGE REGULATOR AND REFERENCE TUBES

<table>
<thead>
<tr>
<th>TYPE</th>
<th>MAX. DIMENSIONS INCHES</th>
<th>MIN. STARTING VOLTAGE SUPPLY</th>
<th>OPERATING VOLTAGE (Approx.)</th>
<th>MIN. OPERATING CURRENT MA.</th>
<th>MAX. OPERATING CURRENT MA.</th>
<th>MAX. REGULATION VOLTS</th>
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<td></td>
<td>HEIGHT</td>
<td>DIAM.</td>
<td></td>
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<tr>
<td>OA2</td>
<td>2.63</td>
<td>.75</td>
<td></td>
<td>185</td>
<td>150</td>
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<td>133</td>
<td>108</td>
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<td>CK1017</td>
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<td></td>
<td>750</td>
<td>700</td>
<td>.005</td>
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<td>CK1022</td>
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<td></td>
<td>1100</td>
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<td>CK1037</td>
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<td>720</td>
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<td>CK1038</td>
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<td>.005</td>
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<td></td>
<td>1230</td>
<td>1200</td>
<td>.005</td>
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<td>CK5651*</td>
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<td>CK5962</td>
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<td>CK6213</td>
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<td>.40</td>
<td></td>
<td>200</td>
<td>130</td>
<td>1.0</td>
</tr>
</tbody>
</table>

*Voltage Reference Tube

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RAYTHEON MANUFACTURING COMPANY
Receiving Tube Division
Newton, Mass., Chicago, Ill., Atlanta, Ga., Los Angeles, Calif.

Reliable Subminiature and Miniature Tubes • Germanium Diodes and Transistors • Radio Tubes • Receiving and Picture Tubes • Microwave Tubes

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PROCEEDINGS OF THE I.R.E. June, 1952
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JAMES KNIGHTS

Formerly Manufactured by DOOLITTLE RADIO, INC.
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.

COMMUNICATION CRYSTALS for the CRITICAL!
Regardless of model, type, or design, James Knights can provide you with the very finest in stabilized crystals. Today JK crystals are used everywhere communications require the VERY BEST.

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!

ALSO manufacturer of the James Knights Frequency Standard.

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—THINK OF BUSS...

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And a companion line of BUSS Fuse Clips, Fuse
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laboratory and the world's largest fuse pro-
duction capacity.

Each BUSS Fuse Electronically Tested.

To assure proper operation in the field, each
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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.

IF . . . YOU HAVE A
SPECIAL PROBLEM
TURN TO BUSS

USE THIS COUPON — Get All the Facts

BUSSMANN MFG. CO., St. Louis, Mo.
Division of McGraw Electric Company
MANUFACTURERS OF A COMPLETE LINE OF FUSES
FOR HOME, FARM, COMMERCIAL AND INDUSTRIAL USE.

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.

Our staff of fuse engineers is at your service.

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

Please send me Bulletin 5F8 containing complete facts on BUSS
Small Dimension Fuses and Fuse Holders.

Name __________________________________________

Title __________________________________________

Company ________________________________________

Address _________________________________________

City ______________________ State _____________

PROCEEDINGS OF THE I.R.E. June, 1952
MOLYBDENUM PERMALLOY
POWDER CORES*
(New technical data now available)

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.

For high Q in a small volume, characterized by low eddy current and hysteresis losses, ARNOLD Moly Permalloy Powder Toroidal Cores are commercially available to meet high standards of physical and electrical requirements. They provide constant permeability over a wide range of flux density. The 125 Mu cores are recommended for use up to 15 kc, 60 Mu at 10 to 50 kc, 26 Mu at 30 to 75 kc, and 14 Mu at 50 to 200 kc. Many of these cores may be furnished stabilized to provide constant permeability (±0.1%) over a specific temperature range.

MANUFACTURED UNDER LICENSE ARRANGEMENTS WITH WESTERN ELECTRIC COMPANY

THE ARNOLD ENGINEERING COMPANY
SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION
General Office & Plant: Marengo, Illinois

PROCEEDINGS OF THE I.R.E. June, 1942
Leadership is a Habit 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.

Submit your own problems in this highly exacting field to our specialist-engineers. They are eager to be of help. Write for your copy of our new 32-page brochure, the most complete and informative presentation ever made on hermetic seals.

Hermetic Seal Products Co.
29 South Sixth Street, Newark 7, N. J.

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PROCEEDINGS OF THE I.R.E. June, 1932
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low-loss miniature TUBE SOCKETS

7-PIN, 9-PIN and SUBMINIATURES

— available in two grades:

MYCALEX 410 — priced comparable to mica-filled phenolics. Loss factor is only 0.015 at 1 mc., insulation resistance 10,000 megohms. Approved fully as Grade L-48 under N.M.E.S. JAN-1-10 "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 (0.083 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-hygrosopic 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.

MYCALEX Corporation of America
Owners of 'MYCALEX' Patents and Trade-Marks
Executive Office: 30 ROCKEFELLER PLAZA, NEW YORK 20 — Plant & General Offices: CLIFTON, N. J.

PROCEEDINGS OF THE I.R.E.
June, 1952
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!

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.

Insist on Plaskon Alkyd for superior electrical parts.

---

PROCEEDINGS OF THE I.R.E. June, 1952

23A
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, L56.

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.

CAMBRIDGE THERMIonic CORPORATION

custom or standard... the guaranteed components

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 3 C.T.C. ceramic coil form types: LST, LS5, L56, L57, L58. 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 L57 and L58.

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.

SPECIAL!
Rush Service On Small Lots

Now! Your orders for small lots of coils needed for prototype work or for your emergency production needs will be handled on a rush basis by C.T.C.'s new SPECIAL SERVICE DEPARTMENT.

We will furnish single layer or pie wound coils ...tuned or untuned. C.T.C. engineers will also work with you to design coils to fit your specifications, military or civilian.

Plan now to put C.T.C.'S SPECIAL SERVICE DEPARTMENT to work for you. Send prints and specifications for quotation.

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.

C T C

New catalog! Send for your copy now.
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.

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

El-Menco
MOLDED MICA CAPACITORS
MICA TRIMMER
Capacitors

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

New and unique in civil emergency communication systems is the New York City Fire Department's five borough 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.

Write our application engineering department for the latest information and technical data about these and other Eimac tubes.
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.
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 million of a watt (0.000001). The specifications for the Type RD 2517 are:
- Ec -45 volts supply voltage-collector circuit;
- ic 400 ma collector current;
- ie -175 ma 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: 1 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-I 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 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. Stacking only the four components, the user can produce finished connectors on a mass-production basis and yet can make changes in the number of contacts or polarity of any connector as needed.

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.

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 micro 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.

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 394)
Q: Are all brands of Resistors similar in Quality, Specifications and Performance?

A: Naturally, not! That's why DAVEN has earned the right to add a superlative in naming its line SUPER DAVOHM 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.

THE DAVEN CO.
195 CENTRAL AVENUE
NEWARK 4, NEW JERSEY
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 formulas. 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 \( \pm 1\% \).
- 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 \( \pm 5\% \) to 100 mc and to \( \pm 12\% \) 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: \( \pm 0.2 \) mmf to 20 mmf, \( \pm 0.3 \) mmf to 50 mmf, \( \pm 0.5 \) mmf to 100 mmf.

Price: $625.00 F.O.B. Factory

Proceedings of the I.R.E. June, 1952
A Molehill of Difference Can Make a Mountain of Trouble in 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 Titelflex microwave components.

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

Let Our Family of Products Help Yours

Titelflex
511 Frirlinghuyzen Ave
Newark 5, N.J.

Please send me without cost information about the products checked at the left.

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PROCEEDINGS OF THE I.R.E.
June 1952
Don't be "blinded" to your future!

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.

Now is the time for qualified ELECTRONIC, ELECTRICAL and MECHANICAL ENGINEERS . . . PHYSICISTS . . . METALLURGISTS . . . PHYSICAL CHEMISTS and GLASS TECHNOLOGISTS . . . as well as TECHNICAL SALES ENGINEERS to decide to take full advantage of the opportunities now open at RCA to achieve professional success.

Lifelong Career Opportunities

These are not temporary positions. They are independent of national defense requirements. The openings represent a wide choice of long-term government projects as well as challenging work in the permanent expansion of a diversified line of commercial products.

You enjoy these benefits

At RCA, you enjoy professional status, recognition for accomplishments . . . unexcelled research facilities for creative work . . . opportunities for advancement in position and income . . . pleasant surroundings in which to work. You and your families participate in Company-paid hospital, surgical, accident, sickness and life insurance. Modern retirement program. Good suburban or country residential and recreational conditions. Opportunities for graduate study. Investigate opportunities today.

Positions open in the following fields:

TELEVISION DEVELOPMENT—
Receivers, Transmitters and Studio Equipment

ELECTRON TUBE DEVELOPMENT—
Receiving, Transmitting, Cathode-Ray, Phototubes and Magnets

TRANSFORMER and COIL DESIGN

COMMUNICATIONS—
Microwave, Mobile, Aviation, Specialized Military Systems

RADAR—
Circuitry, Antenna Design, Computer, Servo-Systems, Information Display Systems

COMPUTER DEVELOPMENT AND DESIGN—
Digital and Analog Computers, Magnetic Recording, Pulse Circuitry, Storage Components, Systems Design

NAVIGATIONAL AIDS

TECHNICAL SALES

ELECTRONIC EQUIPMENT FIELD SERVICE

Mail Resume

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

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

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.

RCA

Radio Corporation of America

Proceedings of the I.R.E. June, 1952
**International Rectifier Corporation**

**Selenium Diodes**

**D-1224**
- 1/8" diameter
- 1/4" length
- Potted in thermo-setting compound.

**D-1290**
- 5/32" diameter
- 9/32" length
- Potted in thermo-setting compound.

**Specifications**

**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 microamperes
- Reverse leakage at 26V RMS: 3 microamperes
- Frequency max. CPS: 200 KC

**D-1290**
- RMS applied voltage, max.: 26 volts per cell
- Peak inverse voltage: 60 volts per cell
- RMS input current, max.: 375 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.

**International Rectifier Corporation**

**General Offices:**
1521 E. Grand Ave.
El Segundo, Calif.
Phone El Segundo 1890

**Chicago Branch Office:**
205 W. Wacker Dr.
Franklin 2-3889
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 of 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.

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

34A

PROCEEDINGS OF THE I.R.E. Junr. 1952
Symbol of experience

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.

TRUSCON® STEEL COMPANY
Subsidiary of Republic Steel Corporation
1074 Albright Street, YOUNGSTOWN 1, OHIO

TRUSCON ... a name you can build on
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.

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

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

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

PROCEEDINGS OF THE I.R.E. June, 1953
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.

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

BREEZE
Special CONNECTORS

BREEZE CORPORATIONS, INC.
For Critical Applications

Triplet 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 Sintercore.

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 $5\frac{1}{2}$ inches and a maximum diameter of $2\frac{7}{8}$ 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)

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"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 consequent 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

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-µ 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)
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.

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1628-G W. WALNUT STREET
CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

PROCEEDINGS OF THE I.R.E.  June, 1952
"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 Farmers Bank Building
Pittsburgh 22, Pa.

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

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 ½ 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.
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.

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- Corrosion prevention in precise servo amplifier assemblies.
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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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Telephone: FReehold 8-1880

DIPOLE ANTENNAS, SWITCHES, Q-MAX LACQUER AND CEMENT
Oscilloscograms 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 0.035 μsec . . . pulse is found to be of 0.02 μsec rise time from the relation:

\[ T_{\text{rise}} = \sqrt{T_{\text{measured}}^2 - T_{\text{amplifier}}^2} \]

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 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 oscilloscrams is supplied with the instrument as well as suitable filter for visual contrast. A new DuMont 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 (50% 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: 0.01, 1.0, 10, 100 μsec, better than ±5% accuracy.
- Illuminated scale with dimmer control.
- DuMont Type 2592-52 shielded coaxial adapter with 52 ohm termination included.

PRICE $825

DuMont for Oscillography

Allen B. DuMont Laboratories, Inc. Instrument Division, 1500 Main Ave., Clifton, N. J.
You're getting smaller... smaller... and smaller.

"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.

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

SPECIFICATIONS:
SIX DC CURRENT RANGES:
0 to 0.01, 0.1, and 1 microampere; 0 to 10, 100, and 1000 microamperes.

INTERNAL SHUNT RESISTANCE:
- 0.01-amp Range = 50 megohms
- 0.1-amp Range = 5 megohms
- 1-amp Range = 0.5 megohms
- 10-amp Range = 5,000 ohms
- 100-amp Range = 50,000 ohms
- 1000-amp Range = 500 ohms

POWER SUPPLY:
- "A" Batteries........... 2, 0.9 volts (RCA-VS206)
- "B" Batteries........... 2, 2.9 volts (RCA-VS202)

DIMENSIONS:
- 87/8" High, 1/8" Wide, 5/8" Deep

WEIGHT:
- 9 1/2 lbs. (incl. batteries)

RADIO CORPORATION of AMERICA
TEST EQUIPMENT

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 overshwing 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.
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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.
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 2 1/4% 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!
New Materials — New Techniques — New Advantages
Features in 4 New IRC Resistors

IRC Type BOC Boron-Carbon 1/2-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 resistors at savings in space and cost. You'll find it adaptable to a host of critical circuitry needs—in electronics and avionics, communications, telemetry, 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 35°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

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 1/2 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 resistors are too large or expensive.


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.

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 precision... 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 B-3.

New 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.

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:
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**TYPE 43**
(JAN-R-94, Type RV2)

1/4 watt, "1/4" 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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(JAN-R-19, Type RA20)

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

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(JAN-R-19, Type RA30)

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

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1/2 watt 70°C, 1/4" diameter miniaturized variable composition resistor.

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**PROVIDES NEW FLEXIBILITY IN TV RECEIVER CIRCUIT DESIGN**

- 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.

TUNG-SOL ELECTRIC INC., NEWARK 4, NEW JERSEY

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MODEL 260
VOLT-OHM-MILLIAMMETER
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A covers all ranges necessary for Radio and TV set testing
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known the world over for its ruggedness
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D molded recesses for resistors, batteries, etc.
E easy battery replacement
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all components—including case and panels—are specially designed and completely tooled for maximum utility...not merely assembled from stock parts

ranges
20,000 Ohms per Volt DC,
1,000 Ohms per Volt AC
Volts, AC and DC: 2.5, 10, 50,
250, 1000, 5000
Output: 2.5, 10, 50, 250, 1000
Milliamperes, DC: 10, 100, 500
Microamperes, DC: 100
Ampere, DC: 10
Decibels (5 ranges):
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Ohms: 0-2000 (12 ohms center), 0-200,000 (1200 ohms center), 0-20 megohms (120,000 ohms center)

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Model 260 $38.95; With Roll Top $46.90. Complete with test leads and operator’s manual. 25,000 volt DC Probe for use with Model 260, $9.95.

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MICROWAVE is establishing new production facilities to supply a complete semiconductor line including 1N21B, 1N21C, 1N23B, 1N25, 1N26, and special millimeter wave silicon diodes. Production will commence shortly on n-p-n TRANSISTORS under license to The Western Electric Company.

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The seasoned MICROWAVE tube engineering and production teams are now producing several magnetron types including the popular 2J42 and 2J42A. High level TR microwave gas discharge switching tubes are available for several bands. Our research personnel in this field are always ready to assist in special microwave tube design problems.

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with the new RCA-6012 gas thyatron

Expressly designed for industrial control
applications, the new RCA-6012 gas tetrode
features the ruggedness necessary to with-
stand rough industrial usage. It has the addi-
tional advantages of low cost and nation-
wide renewal distribution—both of im-
portance to the end user.

For motor-control, electronic-inverter, and
general relay service at power supply fre-
cuencies, the RCA-6012 is rated to withstand
a maximum peak-inverse-anode voltage of
1300 volts, a maximum peak-cathode current
of 5 amperes, and a maximum average
cathode current of 0.5 amperes.

Operating features of the RCA-6012 in-
clude a negative-control characteristic which
is essentially independent of the ambient
temperature over the range from -75° to
+90° C, low preconduction currents, low
control-grid-to-anode capacitance, and low
control-grid current.

The RCA-6012 is compactly designed, and
employs a structure that increases its resis-
tance to both shock and vibration. A button
stem is used to strengthen the mount struc-
ture and to provide wide inter-lead spacing
as a means of reducing susceptibility to elec-
trolysis and leakage.

For complete technical data on the
RCA-6012, write RCA, Commercial Engi-
neering, Section FR47, Harrison, N. J. . . . or
contact your nearest RCA field office.

FIELD OFFICES: (East) Humboldt 5-3900, 415
S. 5th St., Harrison, N. J. (Midwest) White
hall 4-3900, 589 E. Illinois St., Chicago, Ill
(West) Madison 9-4071, 130 S. San Pedro St.
Los Angeles, Calif.

Another new RCA tube

RCA-6080 is a current-regulator tube for use in regulated dc
power supplies. Similar to the 6A57-G, it features a button-
stem construction for improved resistance to shock and vibra-
tion. The 6082 is a similar tube for aircraft power supplies.

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA
VOLUME 40
June, 1952

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Chairmen of New IRE Professional Groups

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Electronic Computers Group

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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.

GEORGE D. O’NEILL

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 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 be best presented in nonmathematical terms.

In particular, when a problem or concept is put into words, the mind can grapple with its general meaning, unencumbered 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 double the 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.
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 data. 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 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 that 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 materials 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 prevalent 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 more and more scientists 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 lines 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 otherwise have had. Unfortunately, however, some of these reviews are often too 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:

"The population growth, the growth of the entrepreneurship, the increased number of large commercial enterprises, the greater demand for the services of scientists, engineers, and technicians, have increased the problem of dissemination and utilization of information inherent in science and technical research. While it is improbable that these factors could not have been anticipated in advance, they have been certainly unforeseen in the extent to which they have actually emerged in the current situation and the rapidity with which they have been developing."
knowledge of management's 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 importantly, 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 of 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, inverter, 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

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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 plasmotron, 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

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 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 plasmotron 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 plasmotron diode.](image)

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 \( I_1 = 8 \text{ ma} \)) 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 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.

\[ E. O. \text{Johnson, "Controllable gas diode," } \text{Electronics, vol. 24, pp. 107-109, May, 1951.} \]
current and the saturated main current. 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.

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.

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
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.

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

![Image of a plasmatron triode](image_url)
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.

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.

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 \(10^5\) to about \(10^{10}\) 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

\[
\sigma = N_{e} \varepsilon \text{mhos per centimeter},
\]

(1)

where \(N\) is the electron density per cubic centimeter, \(\varepsilon\) is the electron charge in coulombs, and \(\mu\) is the electron mobility in centimeters per second per volt per centimeter. For a typical plasma \(N = 5 \times 10^{10}\) and

\[
\mu = 5 \times 10^{6} \frac{\text{cm}}{\text{sec} \cdot \text{volt}}.
\]

Thus \(\sigma = 4.0 \times 10^{-2}\) 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

\[
\dot{i}_{e} = \frac{1}{4} N_e e, \quad (2)
\]

where \(e\) is the average thermal velocity of the electrons, is given by

\[
e = 1.87 \times 10^{3} \sqrt{\frac{T_e}{m}}.
\]

(2')

\(T_e\) is the electron temperature and \(m\) the electron mass. Choosing as typical values \(N = 5 \times 10^{10}\) cm\(^3\), \(e = 2 \times 10^{7}\) cm/sec, and \(T_e = 1,000\) K., it turns out that \(\dot{i}_e = 40\) ma/cm\(^2\). A similar computation for the positive-ion space-current density yields (for helium) \(\dot{j}_p = 0.27\) 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):

\[
\dot{j}_p = \frac{2.33 \times 10^{-6} V^{3/2}}{\sqrt{M/m}} \frac{1}{d^2}, \quad (3)
\]

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 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

\[
\dot{i}_e = A \dot{j}_p e^{\phi/e}
\]

\[
\phi = \frac{e \phi}{kT_e},
\]

(4)

\[\text{Fig. 13—Plasma cold-electrode characteristic.}\]
where \( j_\text{e} = Ne\tau/4 \) is the random electron space current near the edge of the plasma, \( A_\text{e} \) is the probe area, and \( k \) is the Boltzmann constant. The electron temperature \( T_\text{e} \) can be determined from the semilog plot of \( j_\text{e} \) 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 \( e \) 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 \( V_\text{f} \) 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. Let us now suppose that one of the electrodes is a hot cathode with a total emission density \( j_\text{e} \); 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 \( j_\text{e} \ll j_\text{oa} \), and (2) the cathode is a strong emitter such that \( j_\text{e} \gg j_\text{oa} \). Here, \( j_\text{oa} \) is the random electron current density in the vicinity of the cathode; \( j_\text{oa} \) is the random electron current density in the vicinity of the anode; \( V_\text{s} \) is the plasma-cathode potential difference; \( V_\text{a} \) is the plasma-anode potential difference; \( V_\text{e} \) accounts for the potential drop through the plasma and for the contact potential difference between the cathode and anode; \( V_\text{a} \) 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.

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 \( j_\text{oa}A_\text{e} \) leaving the cathode, a current \( j_\text{oa}A_\text{e} \exp(\phi V_\text{e}) \) entering the cathode from the plasma, and a current \( j_\text{oa}A_\text{e} \exp(-\phi V_\text{e}) \), which is equal to the electron current \( I_2 \) in the circuit, entering the anode from the plasma. \( \phi \) is defined in (4). For equilibrium

\[
j_\text{aoA}_\text{e} = j_\text{oaA}_\text{e} \exp(-\phi V_\text{e}) + j_\text{oaA}_\text{e} \exp(\phi V_\text{e}) .
\]

From the potential diagram we see

\[ V_\text{s} + V_\text{e} = V_\text{a} + V_\text{f} \]

Manipulation of the preceding equations yields

\[
I_2 = \frac{A_\text{e}j_\text{oa}}{(1 + \left[A_\text{e}j_\text{oa}/A_\text{e}j_\text{oa}\right] \exp(-\phi V_\text{e}))} .
\]

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,

\[ A_\text{e}j_\text{oa} \exp(-\phi V_\text{e}) = A_\text{e}j_\text{oa} + A_\text{e}j_\text{oa} \exp(-\phi V_\text{e}) \]

\[ I_2 = A_\text{e}j_\text{oa} \exp(-\phi V_\text{e}) , \quad \text{and} \quad V_\text{s} + V_\text{e} + V_\text{a} = V_\text{f} \]

The solution of these equations is

\[
I_2 = \frac{A_\text{e}j_\text{oa}}{2} \left[ \frac{1}{1 + 4 \left(A_\text{e}j_\text{oa}/A_\text{e}j_\text{oa}\right) \exp(-\phi V_\text{e})} - 1 \right] .
\]

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. See Appendix.

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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.

![Graph](image)

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, 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. This value of anode current is approximately one-half the total field-free emission of the cathode.

If the potential $V_n$ 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

\[ I_\text{sat} = A_{sj} j_{\text{sat}} \]  

(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.

![Graph](image)

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

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

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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)

\[
A_{oj} = A_{oj0} + A_{oj0} j,
\]

whence

\[
\frac{A_{oj}}{I_2} = \frac{A_{oj0}}{A_{oj0} + 1}.
\]

(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. Then \( A_{oj0} > A_{oj0} \). Substitution of this inequality into (7a) yields \( A_{oj} > 2 I_2 \). That is, for normal plasmotron operation cathode emission should be at least twice maximum saturated anode current demanded.

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.

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 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.  

![Fig. 20](image)

**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. Nonasymmetries 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

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](image)

**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. 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 \( d i_p \) arising from the ion generation, in a distance \( dx \), by an electron current \( i_e \) is given approximately by

\[
di_p = f(V) i_e e^{-z/\mu} dx.
\]

Here the ionization function \( f(V) \), is to a first approximation, as displayed in Fig. 21, the factor \( e^{-z/\mu} \) is the

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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.

fraction of the original electron current \( i_e \) still able to ionize after traveling the distance \( x \), and \( \lambda_x' \) 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 \( \lambda_x' \), yields

\[ i_p = f(1) i_0 \lambda_x'. \]  
\[ \text{(8)} \]

Since the ionization function varies directly, and \( \lambda_x' \) 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 \( i_0 \) 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

\[ \frac{dN}{dt} = \frac{N}{\tau}, \]  
\[ \text{(9)} \]

where \( N \) is the total number of ions in the plasma and \( \tau \) is the plasma decay constant, in seconds, and is described by

\[ \tau = \frac{\Lambda^2}{D_e}. \]  
\[ \text{(10)} \]

Here \( \Lambda \) is the characteristic diffusion path in centimeters and \( D_e \) is the ambipolar diffusion coefficient in square centimeters per second. The quantities \( N \), \( \tau \), and \( \Lambda \) 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. 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 \( \Lambda \) and \( D_e \) 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. This added complication presents no special problem since the effective value of \( \tau \) 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 \( \tau \) 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

\[ i_p = \frac{N e}{\tau}. \]  
\[ \text{(11)} \]

When (8) is introduced into this, there results

\[ f(1) \lambda_x' i_p = \frac{N e}{\tau}. \]

Probe studies of plasma density and the requirements upon the plasma density indicate that \( N \) can be expressed in terms of the average plasma density \( n_{wa} \), adjacent to the anode, and the active tube volume \( v \), by

\[ X = \psi n_{wa}. \]

The factor \( \psi \), 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

\[ f(1) \lambda_x' i_p = 0.4 \times 10^{-4} \frac{j_{wa}}{\tau}. \]  
\[ \text{(12)} \]

Here \( j_{wa} \) is expressed in ma per square centimeter, \( v \) in cubic centimeters, the electron temperature is taken as 1,000°K, and the ionizing current \( i_p \) has been replaced by its equivalent \( I_p \), the auxiliary discharge current expressed in ma.

This equation, first of all, predicts that the anode current should be proportional to the ionization function \( f(1) \). 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 constriction effect at the aperture, and hence the larger the auxiliary discharge arc drop.

Secondly, (12) indicates that the anode current should be proportional to the auxiliary current \( I_p \). The results in Figs. 4 and 47 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 \( \tau \). Both the qualitative and quantitative behavior shown in Fig. 23 confirm this very nicely.
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: \( A = 8.5 \text{ cm}^2, v = 3.25 \text{ cm}^3, V = 60 \text{ volts}, \) and \( \lambda' = 0.235 \) 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 \( \tau \). For example, tube No. 3 had positive baffles over the end micas; these prevented ion losses to the micas, and so altered \( \tau \).

\[ I_2 = \text{AUX.} \]

![Fig. 22](image)

**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 plasmotron operation and (2) using the same tube without the auxiliary discharge. Now if the necessary auxiliary heater power of

\[ 1.0 \text{ watt} \] (at the conservative rate of 20-\text{ma} emission per watt heating power) is added, the plasmotron 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 plasmotron principle in recifier circuits, where efficiency is a consideration, appears attractive.

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 stratagem, 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.

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

\[ n = n_0 e^{-t/\tau} \]

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 density.

\[ \lambda' = 1/2a \] which is derived by equating the diffusion to the recombination loss. Using the value \( 1.7 \times 10^{-3} \) given by Biondi and Brown, for the volume recombination coefficient \( \alpha \) of helium, and 100 microseconds for \( \tau \) the value of \( n_0 \) turns out to be \( 3 \times 10^{19} \). 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 \( r \) 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.\(^4\) 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.

\( ^4 \) L. Malter and W. M. Webster, "Rapid determination of gas discharge constants from probe data," \textit{RCA Rev.}, vol. 12, p. 191; June, 1951.

The similarity between the two output characteristics of the diode and triode operation stems from the fact that the grid action affects the total quantity \( A_{ij} \) 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.](image-url)
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 and need not be elaborated upon at this point. The plasmatron escapes the difficulties of the thyatron 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 thyatron. 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 thyatron 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.

The electrons enter the plasma at the cathode with a density \( q^- \) and leave at the anode with a flow density \( aq^-/b \). Positive ions are allowed to enter either at anode or cathode with a flow density \( q^+ \), 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

\[
q^+ = nE_n^+ - D^+ \frac{\partial n}{\partial r}
\]

and

\[
q^- = -nE_n^- - D^- \frac{\partial n}{\partial r}
\]

must be satisfied; \( q^+ \) and \( q^- \) are the respective positive ion and electron particle flows, respectively, \( \mu^+ \) and \( \mu^- \) 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

\[
q^+ = - \left[ \frac{D^+ \mu^- + D^- \mu^+}{\mu^- + \mu^+} \right] \frac{\partial n}{\partial r} = - D_{a'} \frac{\partial n}{\partial r},
\]

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 \( \mu^- = \mu^+ \), \( D_{a'} \) 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 re-
suit 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:

\[ \frac{dn}{dr} = \frac{aq_n}{D_a} \left[ \frac{\mu^+}{\mu^-} + \frac{r_0 q_n}{aq_n} \right] n dr . \]  

(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

\[ n = n_b + \frac{aq_n}{D_a} \left[ \frac{\mu^+}{\mu^-} + \frac{r_0 q_n}{aq_n} \right] \ln \left( \frac{b}{r} \right) . \]  

(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) \( q_n^- \) is proportional to \( n_b \) and \( q_n^+/q_n^- \) remains approximately constant, then the density distribution function will be essentially independent of the anode current. Experiment indicates that such is the case. If \( q_n^- \) 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 \( q_n^- \). 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 plasmotron. 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.

In developing the expressions to describe the plasmotron operation, it has been assumed that the current carrying capacity of the plasma is a minimum at the anode, that is, that \( A_{j_b} < A_{j_j} < A_{j_a} \). 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

\[ n_{sb} < n_r \]

or

\[ n_{sb} < n_{sr} + \frac{aq_n}{D_a} \left[ \frac{\mu^+}{\mu^-} + \frac{r_0 q_n}{aq_n} \right] \ln \left( \frac{b}{r} \right) . \]  

(6)

\[ 1 < \frac{r}{b} + \frac{aq_n}{b n_b D_a} \frac{\mu^+}{\mu^-} F \ln \left( \frac{b}{r} \right) . \]

\[ F = \left\{ 1 + \frac{a q_n^-}{a q_n^+} \right\} . \]

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

\[ \frac{a}{b} = q_n^+, \quad \frac{q_n^-}{n_b} = \frac{c^-}{4}, \quad \Theta = \frac{T^-}{T^+} . \]

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

\[ 1 < \frac{r}{b} + \frac{4}{\pi} \frac{\Theta}{1 + \Theta} \frac{r}{L^-} F \ln \left( \frac{b}{r} \right) . \]

or, approximately

\[ 1 < \frac{r}{b} + \frac{r}{L^-} F \ln \left( \frac{b}{r} \right) . \]  

(7)

From (7) it is seen that the necessary conditions for normal plasmotron 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*

F. J. KERR†, SENIOR MEMBER, IRE

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 (1°), 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

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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)3, 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 Allen7 and van de Hulst.8 The existence of a high-electron temperature (about 10⁶ 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.

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. 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 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 \((\mu)\) 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.

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-
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 earth's surface, will be refracted 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 Rs.

The absorption loss for central and oblique rays can be obtained from Smerd's results for the optical depth to the reflecting region, 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 deducting 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 x a/Ro] cps at 30 mc for a point on the sun's disk at a projected distance of a/Ro 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 percent of the echo energy will be contained in a band of 500 cps, and 90 percent 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 \( \pi(1.2 \times R_o)^2 \) 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, \( \sigma \), will be the product of

- \( A \), the equivalent area, \( \pi(1.2 \times R_o)^2 \),
- \( \beta \), 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,
- \( \alpha \), 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 \( \beta = 0.55 \),

\[
\sigma = \pi(1.2 \times R_o)^2 \times 0.55 \times 4 \times 0.45 = 1.5\pi R_o^2
\]

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, \( T_s \), for the quiet sun of 900,000°K.\(^{11}\) The noise power available from a 30-mc plane-polarized antenna of gain 3,000 directed towards the sun would then be

\[
P_s = \frac{kT_s}{\lambda^2} \frac{\Omega_\lambda^2}{4\pi} \text{ watt (cps)}^{-1},
\]

where

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

therefore,

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


\(^{11}\) 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. 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 value, which will be little affected by change of antenna gain, is 17,500°K., equivalent to an available power at the receiver of

\[
P_0 = kT = 2.4 \times 10^{-19} \text{ watt (cps)}^{-1}.
\]

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

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

\[
P_R = \frac{P_T G_T}{4\pi^2R_\odot^2} \frac{1}{4\pi^2} \frac{\sigma R_\odot^2}{4\pi} = 5.2 \times 10^{-16} \text{ w.}
\]

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 \( P_{\text{min}} \) 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

\[
M = \left( \frac{B_n}{2} \right)^{1/2} = 50 = 17 \text{ db.}
\]

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...
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 $f^{-2.4}$ 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 1

<table>
<thead>
<tr>
<th>Frequency (mc)</th>
<th>Transmitter antenna gain (db)</th>
<th>Transmitter power for given cost (db)</th>
<th>Sun. absorption area (db)</th>
<th>Noise (db)</th>
<th>Doppler spread (db)</th>
<th>Relative system sensitivity (db)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>-9.5</td>
<td>0</td>
<td>+3.4</td>
<td>-3.6</td>
<td>-9.3</td>
<td>+1.5</td>
</tr>
<tr>
<td>20</td>
<td>-3.5</td>
<td>0</td>
<td>+1.2</td>
<td>-2.1</td>
<td>-2.2</td>
<td>+0.6</td>
</tr>
<tr>
<td>30</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>40</td>
<td>+2.5</td>
<td>-1</td>
<td>-0.8</td>
<td>-2.6</td>
<td>-0.6</td>
<td>-0.4</td>
</tr>
<tr>
<td>50</td>
<td>+6.0</td>
<td>-8</td>
<td>-1.7</td>
<td>-9.0</td>
<td>-4.0</td>
<td>-1.1</td>
</tr>
<tr>
<td>60</td>
<td>+8.5</td>
<td>-12</td>
<td>-3.3</td>
<td>-11.5</td>
<td>-6.7</td>
<td>-1.5</td>
</tr>
</tbody>
</table>

The effect of frequency variation on all the above factors is given in Table 1. 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, atmospheres, 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.](image)

### 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.](image)

Variation of bandwidth $(B)$ affects the proportion of the reflected energy which is accepted by the receiver $(\beta)$, 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>
<thead>
<tr>
<th>Planet</th>
<th>Distance at closest approach (meters)</th>
<th>Diameter (meters)</th>
<th>IF bandwidth (cps)</th>
<th>Integration time (seconds)</th>
<th>30-mc echo strength relative to minimum detectable echo (db)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Venus</td>
<td>$4.1 \times 10^8$</td>
<td>$12.2 \times 10^4$</td>
<td>500</td>
<td>6</td>
<td>$+12$</td>
</tr>
<tr>
<td>Mercury</td>
<td>$9.2 \times 10^8$</td>
<td>$4.9 \times 10^4$</td>
<td>50</td>
<td>1</td>
<td>$-6$</td>
</tr>
<tr>
<td>Mars</td>
<td>$7.8 \times 10^8$</td>
<td>$6.8 \times 10^4$</td>
<td>96</td>
<td>6</td>
<td>$+18$</td>
</tr>
<tr>
<td>Jupiter</td>
<td>$63 \times 10^8$</td>
<td>$143 \times 10^4$</td>
<td>4,800</td>
<td>4,200</td>
<td>$-16$</td>
</tr>
</tbody>
</table>

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, 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

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.

Müller† and others‡ 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-

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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.

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\(^1\) has shown that the variation of secondary-emission ratio is given quite closely by

\[ R_\theta = R_0 e^{p(1 - \cos \phi)} \]

where \( R_\theta \) is the secondary-emission ratio at angle of incidence \( \phi \), \( R_0 \) 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.

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.

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.

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

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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.

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.

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.

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RF Phase Control in Pulsed Magnetrons*

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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

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.† 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,

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

(1)

where $S$ is a parameter directly related to the locking power and inversely related to the oscillator output power and $Q$, $\omega'$ is the unperturbed oscillator frequency, $\omega_1$ is the perturbing frequency, and $\phi$ 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

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

(2)

Thus the locking phase, $\phi$, 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) 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 $\phi$ 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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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

\[ P(e_n)de_n = \frac{2e_n}{e_n^2} e^{-\frac{e_n^2}{2\sigma^2}}de_n \]

where \( e_n \) is the noise-envelope amplitude and \( \sigma \) 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

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 \( \xi \), \( C \) is the locking signal amplitude, and \( A \) is the resultant of phase \( \phi \). Now \( A \) may be expressed as

\[
Ae^{i\phi} = C + N \cos \xi + jN \sin \xi,
\]

(4)

whose phase is

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

(5)

where \( R = N/C \). Let us first find the statistical properties of \( \phi \) from those of \( \xi \), 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, \( 0 \leq R \leq 1 \); and Case II, \( 1 < R < \infty \).

![Fig. 1 — Vector relationship of \( \phi \) to the noise angle \( \xi \).](image)

**Case I**

If \( R < 1 \), \( \phi \) always lies in either the first or fourth quadrant for any value of \( \xi \). More specifically, \( \phi \) is a double-valued function of \( \xi \), 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 \( \phi \) lying in the interval \( d\phi \) is the sum of the probabilities of \( \xi \) lying in the intervals \( d\xi_1 \) and \( d\xi_2 \). That is,

\[
P(\phi, R)d\phi = P(\xi) \left[ |d\xi_1| + |d\xi_2| \right],
\]

(6)

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

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

(7)

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

\[
R \cos \xi = -\sin^2\phi \pm \cos \phi \sqrt{R^2 - \sin^2\phi}.
\]

(8)

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

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

\[
+ \left[ 1 - \frac{\cos \phi}{\sqrt{R^2 - \sin^2\phi}} \right] d\phi.
\]

(9)

This is the required expression.

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

\[
\cos \xi = -R,
\]

(10)

which, when substituted into (5), yields

\[
\tan (\phi_{\text{max}}) = \frac{R}{\sqrt{1 - R^2}}
\]

(11)

or

\[
\sin (\phi_{\text{max}}) = R.
\]

That is, for any ratio of signal-to-noise greater than \( 1 \), \( \phi \) has a maximum value given by (11). Conversely, for any \( \phi \), 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:

\[
\sin \phi \geq R \geq 1
\]

\[
0 \leq \phi \leq \sin^{-1} R.
\]

(12)

**Case II**

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

\[
P(\phi, R)d\phi = P(\xi)d\xi.
\]

(13)

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

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

(14)

The validity conditions on this expression are simply

\[
1 < R \leq \infty
\]

\[
0 \leq \phi \leq 2\pi
\]

(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 \( \phi \) only since they are even functions of
that variable. Note that the probability of $\phi$ falling in the interval $\Delta \phi$ is

$$\int_0^{\phi+\Delta \phi} P(\phi, R) d\phi,$$

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

$$\int_0^{\sin^{-1} R} P(\phi, R) d\phi = 1/2 \quad \text{for} \quad R \leq 1$$

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 $R$, $\phi$ will certainly lie in the range given by (12) or (15).

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).

$$P(R) dR = 2R/\bar{R}^3 e^{-R^2/\bar{R}^2} dR,$$

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

$$P(\phi) = \int_0^{\infty} P(\phi, R) P(R) dR.$$

This integral may be evaluated by substituting (9), (14), and (16), letting $y = R^3 - \sin 2\phi$, and noting that

$$\int_0^{\pi} e^{-y^2/2} dy = 2\sqrt{\pi} \text{erf} \left( \frac{y}{\sqrt{R}} \right),$$

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

$$P(\phi) = 1/\pi R^3 \left[ R^2/2(e^{-\sin^2 \phi/\bar{R}^2} - e^{-1/\bar{R}^2}) + \sqrt{\pi R^2}/2 \cos \phi e^{-\sin^2 \phi/\bar{R}^2} \text{erf}(\cos \phi/\sqrt{R^2}) \right]$$

$$+ \sqrt{\pi R^2}/2 \cos \phi e^{-\sin^2 \phi/\bar{R}^2} \text{erf}(\cos \phi/\sqrt{R^2})$$

$$- \sqrt{\pi R^2}/2 \cos \phi e^{-\sin^2 \phi/\bar{R}^2} \text{erf}(\cos \phi/\sqrt{R^2})$$

$$+ \sqrt{\pi R^2}/2 \cos \phi e^{-\sin^2 \phi/\bar{R}^2} \text{erf}(\cos \phi/\sqrt{R^2})$$

$$\cdot \left[ 1 - \text{erf}(1/\sqrt{R^2}) \right],$$

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

$$P(\phi) = 1/\pi R^3 \left[ R^2/2 e^{-t/\bar{R}^2} + \sqrt{\pi R^2}/2 \cos \phi e^{-\sin^2 \phi/\bar{R}^2} \right]$$

$$\cdot \left[ 1 - \text{erf}(1/\sqrt{R^2}) \right].$$

Fig. 4 shows these probability densities plotted from (18) and (19) for various values of noise-to-signal ratio.
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 \( \phi \). 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\(^\text{19}\) 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 \( E, R, \beta, \) and \( \alpha \) 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

\[ v_{rf} = V_{R_{ps}}(1 - \eta e^{-kt}) \tag{22} \]

where \( V_{R_{ps}} \) 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

\[ v_{rf} = (1 - \eta)V_{R_{ps}}, \tag{23} \]

Subsequent to this time, the build-up continues exactly as expressed by (22). Actual observations on magnetron rf build-up show \( \eta \) 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.\(^\text{11}\)

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):

\[ g = \frac{E/R}{V_{rf}} - \frac{1}{R} = \frac{E/R}{V_{R_{ps}}(1 - \eta e^{-kt})} - \frac{1}{R} \]

or

\[ g = g_0 + \frac{E\eta}{RV_{R_{ps}}}(1 - \eta e^{-kt}) \tag{25} \]

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

\[ b = b_0 - \frac{E\eta}{RV_{R_{ps}}}{\tan \alpha}(\frac{e^{-kt}}{1 - \eta e^{-kt}}) \tag{26} \]

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

\[ 2(\omega - \omega_0 \omega_0) = -\frac{E\eta}{u \omega CR_{R_{ps}}}{\tan \alpha}(\frac{e^{-kt}}{1 - \eta e^{-kt}}) \tag{27} \]

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

Further simplification yields

\[ \omega = \omega' - \frac{(E\eta)}{2CR_{R_{ps}}}{\tan \alpha}(\frac{e^{-kt}}{1 - \eta e^{-kt}}) \tag{28} \]


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 $\omega'$. Rotman's measurements on steady-state and transient $g - V_e$ 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

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

where $B$ is the passive susceptance, $\phi$ is the locking phase, and

$$\mid \rho \mid = \mid Y_{LP} \mid = \frac{Y_{1}}{V_{i}},$$

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

$$\mid \rho \mid = \mid Y_{LP} \mid \frac{V_{1}}{V_{RFs}(1 - \eta e^{-kt})} = \mid Y_{LP} \mid \frac{V_{1}}{V_{RFs}(1 - \eta e^{-kt})}. \quad (29)$$

This expression is a good approximation so long as $V_{1}/V_{RFs}$ 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 $\mid \rho \mid$ as a function of time. If $V_{1}$ is considered constant, (29) shows that $\mid \rho \mid$ 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 $\mid \rho \mid$ is not as large as indicated, but it may, nevertheless, exceed the steady-state $\mid \rho \mid$ 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

$$2 \left( \omega - \omega' \right) + \frac{E_{n}}{\omega_{0}CRV_{RFs}} \frac{e^{-kt}}{1 - \eta e^{-kt}} \left[ \rho_{0} \left\{ \sin \phi \right\} - 2 \frac{Q_{ext}}{1 - \eta e^{-kt}} \right] = 0, \quad (30)$$

where $\mid \rho_{0} \mid$ is the steady-state reflection factor. Now if $\omega_{1}$ is the frequency of the synchronizing signal,

$$\frac{d\phi}{dt} = \omega_{1} - \omega. \quad (31)$$

If this is substituted into (30), there results

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

where $S = \omega_{0} \mid \rho_{0} \mid Q_{ext}, M = \omega_{1} - \omega'$, and $N = E_{n}/2CRV_{RFs}$ tan $\alpha$. This is the differential equation of starting which was desired. We need only solve the equation in order to find the phase $\phi$ 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 $\phi$, 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 $\phi$ only, so that $\sin \phi$ may be replaced by $\phi$. 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 $\phi$ only. If this is the case, $\sin \phi$ may be replaced by $\phi$, and (32) becomes an ordinary linear differential equation. It may be solved by use of an integrating factor $e^{\rho_{0}t}$, where $P$ is the coefficient of the term in $\phi$. In this case

$$\int P dt = e^{\rho_{0}t} \left( 1 - \eta e^{-kt} \right)^{S/\nu}. \quad (33)$$

If the linear equation is multiplied by this factor and
integrated, there results
\[ e^{\int (1 - \eta e^{-kt}) \frac{S}{k} \phi} = \int [Me^{\int (1 - \eta e^{-kt}) \frac{S}{k} \phi} + Ne^{(\frac{S}{k} - 1)(1 - \eta e^{-kt}) \frac{S}{k}}] \, dt. \tag{33} \]
The indicated integrations are not difficult to carry out if the factors \((1 - \eta e^{-kt}) \frac{S}{k} \) and \((1 - \eta e^{-kt}) \frac{S}{k}\) are expanded in series by the binomial expansion and integrated term by term. Then \(\phi\) may be written

\[
\phi = \frac{1}{(1 - \eta e^{-kt}) \frac{S}{k}} \left[ M \int S \frac{(S - 1)}{k} \cdots \frac{(S + 1 - m)}{k} \eta^m e^{-mt} + \frac{N}{k \frac{S - 1}{k}} e^{-kt} \right] \\
+ \frac{N}{k \frac{S - 1}{k}} e^{-kt} \sum_{m=1}^{\infty} (-1)^m \frac{k^m (S - 1) (S - 2) \cdots (S - m)}{k (S - m - 1)} e^{-st}, \tag{34} \]

where \(\phi'\) is a constant of integration. If the factor

\[(1 - \eta e^{-kt}) \frac{S}{k}\]
is expanded in a series and divided into the bracketed series, and \(\phi'\) evaluated in terms of the initial angle \(\phi_0 (\phi = \phi_0 \text{ when } t = 0)\), there results

\[
\phi = \frac{1}{(1 - \eta e^{-kt}) \frac{S}{k}} \left[ M \int - \frac{N - \eta M}{k \frac{S - 1}{k}} e^{-kt} + \frac{(1 - \eta) \frac{S}{k} \phi_0}{S} \right] \\
+ \frac{N - \eta M}{k \frac{S - 1}{k}} e^{-kt} \sum_{m=1}^{\infty} (-1)^m \frac{m^1 \eta^{m+1}}{k \frac{S - 1}{k} \cdots \frac{S - m - 1}{k}} \right] e^{-st} \right] \tag{35} \]

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. The majority of these data describe the operation of a 3,000-rc magnetron. Nevertheless, the numbers deduced from them are believed representative of such oscillators.

The constant \(S\) is defined as \(S = \omega_0 [\rho_0 / Q_{\text{ext}}]\). Its value ranges approximately from 10^6 to 5 \times 10^7 for typical S-band magnetrons. Note that it is directly proportional to the 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. Now

\[ N = k \eta \tan \alpha = \eta \times 10^{-8} \tan \alpha. \tag{36} \]
The value of \(\tan \alpha\) ranges approximately between the limits 0.1 and 1, and, as previously stated, \(\eta\) lies between 0.7 and 1.

Thus it is seen that the constants \(k\), \(N\), and \(\eta\) 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

\[ \text{See Appendix I.} \]
these concepts in mind, we may now examine our approximate solution.

In almost all practical cases of synchronization, the frequencies \( \omega_1 \) and \( \omega' \) 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, \( \phi_0 \), is zero. The final phase, of course, will also be zero, since \( \phi_{fi} = M/S = 0 \). 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.

![Phase transient](image)

Fig. 6—Phase transient accompanying the starting of a synchronized magnetron with locking power as a parameter for \( M = 0 \), \( N = 2 \times 10^4 \), \( k = 10^3 \), \( \phi_0 = 0 \), 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 with a time constant \( 1/S \). The values of \( \eta \) and \( N^4 \) 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 \( d\phi/dt \mid_{t=0} \) and

\[
\frac{d\phi}{dt} \bigg|_{t=0} = \frac{N}{1 - \eta}
\]

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, \( \phi_0 \), to be different from zero.

![Frequency transient](image)

Fig. 7—Frequency transient accompanying the starting of a synchronized magnetron for \( M = 0 \), \( N = 2 \times 10^4 \), \( k = 10^3 \), \( \phi_0 = 0 \), and \( \eta = 0.8 \).

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 \( \omega' \) and \( M \), and therefore the steady-state phase, \( M/S \), different from zero. Figs. 9 and 10 show

![Phase transient](image)

Fig. 9—Phase transient accompanying the starting of a synchronized magnetron with locking power as a parameter for \( M/S = 0.348 \), \( N = 2 \times 10^4 \), \( k = 10^3 \), \( \phi_0 = 0 \), and \( \eta = 0.8 \).

\[ N^4 \] The values \( N = 2 \times 10^4 \) and \( \eta = 0.8 \) were computed from data on a Raytheon QK-61 magnetron. The corresponding value for \( S \) is \( 5 \times 10^6 \) if \( |p| = 0.1 \).
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).

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, $\omega - \omega'$. This characteristic has been noted very distinctly in experimental circuits.

We have now examined the solution of (32) for small $\phi$. 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. 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(a) shows the phase $\phi$ as a function of time for values of $S$ when the frequency difference, $M = \omega - \omega'$, 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 $M = \omega - \omega'$ is negative.

Fig. 12—Phase and frequency for $M = -5$ megaradians, showing the effect of changes in synchronizing power.

How variations in $|\omega - \omega'|$ 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 $S = 5 \times 10^4$, showing effect of changes in frequency difference $\omega - \omega$. The effect of initial angles $-60^\circ$, $0^\circ$, and $60^\circ$ 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-

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considerably increases the magnitude and duration of the transient. This effect is shown more graphically in Figs. 15 and 16, where \( \phi_0 = \pm 180^\circ \). Under such conditions, the transient becomes quite intense and synchroniza-
tion 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.

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 rela-
tive to the injected frequency, value of the coefficient \( S \), and the values of the design parameters \( N \) and \( \eta \).

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, 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](image-url)

Returning to Fig. 17, if we write the nodal equation for the voltage \( x \) and differentiate with respect to time, we find
\[
\frac{d^2x}{dt^2} + \mu \left( G_v - g_v - x \frac{dg_v}{dx} \right) \frac{dx}{dt} + \omega_0^2 x = -E \omega^2 \sin \omega t,
\] (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.

\[ G_v = \frac{E}{R} \left( \frac{1}{A} - \frac{E}{R} \right), \]

Fig. 18—Dotted curve shows a possible form of \( g(A) \) as 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

\[ g = G_v + 1 - \frac{A^2}{4}. \] (40)

As a function of the voltage \( x \), (40) may be written as

\[ g_v = G_v + 1 - \frac{x^2}{3}. \] (41)

The details of this transformation will be derived in Section X. Now, substituting (41) into (38), we arrive at

\[ \frac{d^2x}{dt^2} + \mu (x^2 - 1) \frac{dx}{dt} + \omega_0^2 x = -E \omega^2 \sin \omega t. \] (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.

### 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

\[ I = \int \frac{dA}{A} = \omega_0 \left( \frac{g - 1}{2 (\omega_0 C - Q_L)} \right). \]

If \( g \) is constant, the equation integrates directly, giving

\[ A = \omega_0^{2/3}(C(\omega_0 C - 1)/Q_L)^{1/3}. \] (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

\[ \phi = \text{const.} \]

One may obtain an approximate solution to this homogeneous equation by assuming \( x = A \cos (\omega t - \phi) \), where \( A \) and \( \phi \) 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

\[ \left\{ \frac{d^2A}{dt^2} - A \left( \frac{d\phi}{dt} \right) \right\} \left( \omega_0^2 - \frac{d^2}{dt^2} \right) + \mu \left( \frac{3A^2 - 4}{4} \right) \frac{dA}{dt} + \omega_0^2 A^3 = 0. \] (43)
- \left\{ \frac{d}{dt} \left[ \frac{d}{dt} \left[ \frac{d}{dt} A \right] - A \frac{d^2 \phi}{dt^2} \right] \right\} \\
+ \mu \left( \frac{A^3 - 4A}{4} \right) \left( \omega - \frac{d\phi}{dt} \right) \sin (\omega t - \phi) \\
+ \frac{\mu^2}{2} A^2 \cos 3(\omega t - \phi) \\
- \frac{\mu}{4} \left( \omega - \frac{d\phi}{dt} \right) \sin 3(\omega t - \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\(^{-2}\) seconds, containing 300 rf cycles, the assumption is quite well satisfied.

Let us, then, set the coefficient of \( \sin (\omega t - \phi) \) 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

\[
A = \frac{2}{\sqrt{1 + \frac{4 - A_0^2}{A_0^2} e^{-\mu t}}}. \tag{46}
\]

\[
\omega = \omega_0^2 + \frac{\mu^2}{2} \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t} \left[ \frac{2 - \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t}}{1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t}} \right]^{1/2}, \tag{50}
\]

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 \( \mu t \). It is to be noted that the initial amplitude, \( A_0 \), is the critical factor fixing the time of starting, while \( \mu \) enters only as a scale factor. Also, so long as the assumption \( f_0 \gg \mu \) is valid, the value of \( \omega_0 \) does not enter.

Consider (41) which expresses \( g_v \) as a function of \( x \). Rewriting this equation, we have

\[
g_v x = G_v x + x - \frac{x^3}{3}. \tag{47}
\]

If the assumed solution for \( x \) be substituted, there results

\[
g_v A \cos (\omega t - \phi) = G_v A \cos (\omega t - \phi) + A \cos (\omega t - \phi) - \frac{A^3}{3} \left[ \frac{2}{3} \cos (\omega t - \phi) - \frac{1}{3} \cos 3(\omega t - \phi) \right]. \tag{48}
\]

The triple-frequency term may be neglected since the circuit is sharply tuned to the frequency \( \omega_0 \). Under this assumption \( g_v \) reverts to our former \( g \) and

\[
g = G_v + 1 - \frac{A^2}{4}. \tag{49}
\]

Equation (46) shows that the steady-state amplitude, \( A_1 \), 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 \( d^2 A/dt^2 \), we find

\[
\omega = \omega_0^2 + \frac{\mu^2}{2} \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t} \left[ \frac{2 - \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t}}{1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-\mu t}} \right]^{1/2}, \tag{50}
\]

where \( \omega = \omega_0 - d\phi/dt \) and (46) has been substituted. The instantaneous frequency is shown in Fig. 21 for various
values of $A_0$, where the approximation $(1 + x)^{1/2} = 1 + x/2$ when $x \ll 1$ has been made in (50). I 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.

X1. 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 $\phi$. 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

$$
\frac{d^2x}{dt^2} + \mu (x^2 - 1) \frac{dx}{dt} + \omega_s^2 x = -E \omega \left[ \sin \phi \cos (\omega t - \phi) + \cos \phi \sin (\omega t - \phi) \right].
$$

(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

$$
\frac{d^3A}{dt^2} - A \left( \omega - \frac{d\phi}{dt} \right)^2 + \mu \left( \frac{3A^2 - 4}{4} \right) \frac{dA}{dt}
$$

$$
= \omega_0^2 - \frac{E \omega^2}{A} \sin \phi.
$$

As before, this coefficient may be set equal to zero if it is slowly varying. Neglecting the term $d^3A/dt^2$, we can obtain the following:

$$
\frac{d\phi}{dt} = \sqrt{\omega^2 + \mu \left( \frac{3A^2 - 4}{4} \right) \frac{dA}{dt} + (\omega_0^2 - \omega^2) + \frac{E \omega^2}{A}} \sin \phi.
$$

(52)

From the known nature of the solution, $d\phi/dt = \omega$; 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

$$
\frac{d\phi}{dt} = \frac{\mu (3A^2 - 4)}{2 \omega A} \frac{dA}{dt} - \frac{\omega_0^2 - \omega^2}{2 \omega} - \frac{E \omega}{A} \sin \phi.
$$

(53)

We may substitute (46) for $A$, and thus reduce our equation to one in $\phi$ and $t$ alone. We have

$$
\frac{d\phi}{dt} + \frac{E \omega}{4} \left[ 1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-t} \right]^{1/2} \sin \phi = \omega - \omega_0.
$$

where we have used the approximation $(\omega^2 - \omega_0^2)/2\omega \approx \omega - \omega_0$.

We should note here that when $e^{-\phi}$ becomes quite small, (53) reduces to

$$
\frac{d\phi}{dt} + \frac{E \omega}{4} \sin \phi = \omega - \omega_0.
$$

(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

$$
\frac{d\phi}{dt} = \frac{E \omega}{4} \left[ 1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-t} \right]^{1/2} \sin \phi
$$

$$
- \frac{\mu (3A^2 - 4)}{4 \omega_0} \left[ \frac{2}{1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-t}} \right]^{1/2}.
$$

(55)

Values of the constants in this equation corresponding to those used previously are approximately as follows:

$$
S \rightarrow \frac{E \omega}{4} = 5 \times 10^6
$$

$$
k \rightarrow \mu = 10^6
$$

$$
\omega' \rightarrow \omega_0 = 2 \pi \times 3 \times 10^9
$$

$$
1 - \eta \rightarrow A_0/2 = 0.2.
$$

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 $-\mu^2/4\omega_0 \approx -1.33 \times 10^9$ (assuming that $(4 - A_0^2/A_0^2) > 1$). Even if this term remained at its minimum value during the entire starting period, it would correspond to a phase change of only $-1.33 \times 10^9 \times 10^{-12} = -1°$. Hence, we may safely rewrite (55) as

$$
\frac{d\phi}{dt} = \frac{E \omega_0}{4} \left[ 1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-t} \right]^{1/2} \sin \phi.
$$

(56)
By separating the variables and integrating, we obtain

\[
\ln \tan \frac{\phi}{2} = -\frac{E\omega_0}{4} \int \left[ 1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-x^2} \right]^{1/2} dx + C. \tag{57}
\]

The integral may be evaluated by making the substitution

\[
y^2 = 1 + \left( \frac{4 - A_0^2}{A_0^2} \right) e^{-x^2}.
\]

After utilizing the initial condition, \( \phi = \phi_0 \) at \( t = 0 \), the final result is

\[
\ln \tan \frac{\phi}{2} = \ln \tan \frac{\phi_0}{2} + \frac{E\omega_0}{2} \left( \frac{1}{x^2} + 1 \right) - \frac{1}{2} \ln \left[ \frac{(1+x^2)^{1/2} - 1}{(1+x^2)^{1/2} + 1} \right] + \frac{1}{2} \ln \left[ \frac{(1+x^2)^{1/2} - 1}{(1+x^2)^{1/2} + 1} \right] \tag{58}
\]

where \( \chi = 4 - A_0^2/A_0^2 \). The phase, \( \phi \), is plotted as a function of time in Fig. 22 for two values of \( \chi \). It is seen that after the envelope build-up goes to completion, the phase decreases exponentially for small \( \phi \). 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 \( e^{-x^2} \ll 1 \).

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 \( \mu, \omega_0, \omega, \) 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 \( \mu \) 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 \( 1.9 \times 10^7 \), which provides a maximum
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 $A_o$ 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.
necessary to determine the nature and extent of the phase-locked signal. Furthermore, the statistics of the initial phase depend upon the preoscillation noise-to-locking signal ratio. This, then, is the important parameter for determining the performance of the phase-locked loop. 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, \( G + iB \), 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

\[
\frac{g}{\omega_0 C} = \frac{1}{Q_0} + \frac{G}{Q_{ext}}
\]

\[
\frac{b}{\omega_0 C} = \frac{(\omega - \omega_0)}{\omega} + \frac{B}{Q_{ext}}.
\]

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

\[
W = CV_{re}^2
\]

and the rate of energy storage is

\[
dW/dt = 2CV_{re} dV_{re}/dt.
\]

At any instant this rate may be represented as a conductance shunting the circuit of Fig. 25. The magnitude of this shunting effect is

\[
G_S V_{re}^2 = 2CV_{re} dV_{re}/dt.
\]

so that (I-1) becomes

\[
g = \frac{1}{Q_0} + \frac{G}{Q_{ext}} + \frac{2}{\omega_0} \frac{dV_{re}}{dt}.
\]

Now if \( g \) is known as a function of \( V_{re} \), then (I-5a) may be solved to find \( V_{re} \) as a function of time. Assuming the empirical relation

\[
g = \frac{E/R}{V_{re}} - \frac{1}{R},
\]

we find

\[
V_{re} = \frac{E}{RC\omega_0} \left( \frac{1}{1 - e^{-\omega_0 t/(RC\omega_0 + 1/Q_L)}} \right)
\]

where

\[
V_{re} = VRP_0 (1 - e^{-kt}),
\]

and

\[
k = \frac{\omega_0}{2} \left( \frac{1}{RC\omega_0} + \frac{1}{Q_L} \right),
\]

and

\[
N = \frac{E\eta}{2CRV_{RP_0}} \tan \alpha.
\]

If we substitute our expression for \( V_{RP_0} \), we find

\[
N = \frac{\omega_0}{2} \left( \frac{1}{RC\omega_0} + \frac{1}{Q_L} \right) \eta \tan \alpha = k\eta \tan \alpha.
\]

**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 synchronization 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 \( V_0 e^{i\omega t} \) and \( i_0 e^{i\omega t} \). Let there be an externally injected signal whose voltage and current at the oscillator reference plane are \( V_1 e^{i\omega t} \) and \( i_1 e^{i\omega t} \). If \( \omega_1 \) and \( \omega \) are not greatly different, the admittance presented to the oscillator at its plane of reference may be written

\[
Y_L = \frac{ie^{i\omega t} + i_1 e^{i\omega t}}{V_0 e^{i\omega t} + V_1 e^{i\omega t}}
\]

(II-1)
or

\[
Y_L = \frac{i}{V} \left[ \frac{1 + \frac{1}{i} e^{i(\omega_1 - \omega) t}}{1 + \frac{1}{V} e^{i(\omega_1 - \omega) t}} \right]
\]

(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

\[
Y_L \approx G + jB + 2pe^{i(\omega_1 - \omega)t},
\]

(II-3)

where \( \rho = 1/(i(i-1)/V)(G+jB) \) and the approximation is good only when \( V_1/V \ll 1 \). Now \( \rho \) is known as the reflection factor, and may be written

\[
\rho = \frac{Y_{LP}(i_1 - V_1/i)}{2(i_1 + i),}
\]

or

\[
\rho = \frac{Y_{LP} \rho_S}{2(1 - \rho_L^2)} \approx -Y_{LP} \rho_S,
\]

(II-4)

where we have separated the oscillator current and voltage into their incident and reflected components: \( \rho_S = V_1/V_0 = -i_1/i_0 \) = reflection coefficient resulting from the locking signal, and \( \rho_L = V_r/V_0 = -i_r/i_0 \) = 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

\[
Y_L \approx G + jB - 2|\rho| e^{i(\omega_1 - \omega)t}\cos \phi + \rho |e^{i\omega}|
\]

= \( G + jB - 2|\rho| e^{i\omega} \) \quad (II-5)

where \( \phi \) is the phase of the reflection factor. Note that \( \phi \) 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,

\[
\frac{g}{\omega_0 C} = \frac{1}{Q_0 + Q_{ext}} - \frac{2|\rho|}{Q_{ext}} \cos \phi
\]

(II-6a)

\[
\frac{b}{\omega_0 C} \approx 2 \left( \frac{\omega_1 - \omega_0}{\omega_0} \right) + \frac{B}{Q_{ext}} - \frac{2|\rho|}{Q_{ext}} \sin \phi.
\]

(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

\[
\frac{\omega - \omega'}{\omega_0} = \frac{|\rho|}{Q_{ext}} \sin \phi,
\]

(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

\[
\phi = \sin^{-1} \left[ \frac{Q_{ext}(\omega_1 - \omega')}{|\rho| \omega_0} \right].
\]

(II-8)

Now, (II-7) may be written as a differential equation by allowing \( \omega \) and \( \omega_1 \) to differ and noting that

\[
\frac{d\phi}{dt} = \omega_1 - \omega.
\]

(II-9)

This permits us to write

\[
\frac{d\phi}{dt} + \frac{|\rho| \omega_0}{Q_{ext}} \sin \phi = (\omega_1 - \omega').
\]

(II-10)

This equation may be solved directly for \( \phi \) 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 PAPER1 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

Luneberg2 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.

Taking the radius of the disk to be unity, the index \( \mu(r) \) is determined by

\[
\mu (r) = \frac{w(r, r_0)}{r},
\]  

(1)

where the function \( w(r, r_0) \) is determined by

\[
w(r, r_0) = \int_r^1 \frac{\sin t}{\sqrt{t^2 - r^2}} \, dt.
\]  

(2)

When a source of electromagnetic energy moves around the circle \( r = r_0 \), 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 \( r_0 = 1 \) (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 \( \Sigma \) such that the hat-shaped surface consisting of \( \Sigma \) and the annular region of the \( XY \) plane between \( r = 1 \) and \( r = r_0 \) (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 \( \Sigma \) is determined by the equations...
where $r$ and $\mu$ are related by the Luneberg equations (1) and (2).

For each choice of $r_i \geq 1$, (1), (2), and (3) determine a surface which will yield a perfect lens with feed circle $r = r_i$. Equations (1), (2), and (3) are not solvable in terms of elementary functions for the relation between $r$ and $s$, except for the case $r_i = 1$. For any numerical choice of $r_i$, 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 $r = r_i$. However, it is possible to construct an equivalent electro-optical system consisting of the same curved surface $\Sigma$, 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.

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 $\Sigma$ determined by (1), (2), and (3). The only limitation on feed-circle size is the size of the refractive index $\lambda$ 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 $\Sigma$ meets the planar section of the lens system.

Theorem: The surface $\Sigma$ has a vertical tangent plane at $r = 1$.

Proof: We wish to compute $ds/dr$ at $r = 1$. From (3),

$$ds/dr = du/u$$

(3)

$$r = \mu u,$$

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).
\[
\begin{align*}
w(m, r_i) &> \frac{\text{[arcsin } m/r_i]}{\pi} \int_{r_i}^{1} \frac{dt}{\sqrt{t^2 - m^2}} \\
&> \frac{\text{[arcsin } m/r_i]}{\pi} \frac{1}{2} \ln \left(1 + \sqrt{1 - m^2}/m\right), \\
w(m, r_i)/(1 - m) &> \frac{\text{[arcsin } m/r_i]}{2\pi} \frac{1}{1/(1 - m)} \frac{\ln (1 + \sqrt{1 - m^2})}{m}.
\end{align*}
\]

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 \(m \rightarrow 1\) from below, the product \(\rightarrow +\infty\). Hence by (7)

\[\lim [w(m, r_i)/(1 - m)] = +\infty,\]

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.

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**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. 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, 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 potent-

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(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 1.

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 Wiltshire4 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-Peterson7 equations might be directly applied. Unfortunately, the Llewellyn 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 1 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 1 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-Peterson7 equations may then be expressed in the form

$$q(z) = (E^* q_v + F^* v_v)e^{-\beta z}$$
$$v(z) = (H^* q_v + I^* v_v)e^{-\beta z},$$

(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; $q_v$ and $v_v$ are the same quantities at the $s$-plane. With $\omega$ 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 are7 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

$$q(z) = (E q_v + F v_v)e^{-\beta z}$$
$$v(z) = (H q_v + I v_v)e^{-\beta z};$$

(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

$$\begin{pmatrix}
q(z) \\
v(z)
\end{pmatrix} = 
\begin{pmatrix}
E & F \\
H & I
\end{pmatrix} e^{-\beta z} 
\begin{pmatrix}
q_v \\
v_v
\end{pmatrix}.$$  

(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 $e^{-\beta z}$, 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 form4,9

$$\begin{pmatrix}
\cos(hz) & i g \sin(hz) \\
-ig \sin(hz) & \cos(hz)
\end{pmatrix} e^{-\beta z},$$  

(3)

where

$$h = \omega_1/v_0, \quad \omega_1^2 = p_e/e_m,$$
$$g = \omega_0/v_1, \quad \beta = i\omega_2/v_0;$$

e and $m_0$ are respectively the electron charge and electron mass, $v_0$ is the dc velocity of the beam, $p_0$ is the dc charge density, $\xi$ is the distance measured from the entering plane, $\omega_1$ is known as the angular plasma frequency, and $\epsilon$ 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.


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 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

$$
\begin{pmatrix}
q(z) \\
v(z)
\end{pmatrix} =
\begin{pmatrix}
E_m & F_m \\
H_m & I_m
\end{pmatrix}
\times
\begin{pmatrix}
E_{m-1} & F_{m-1} \\
H_{m-1} & I_{m-1}
\end{pmatrix}
\times \cdots
\times
\begin{pmatrix}
E_1 & F_1 \\
H_1 & I_1
\end{pmatrix}
\times
\begin{pmatrix}
\eta/\eta_0 = (\eta_a + \eta_b)\xi/\epsilon^2.
\end{pmatrix}
$$

where $\beta$ is again the total dc transit angle multiplied by the imaginary unit $i$.

$E_m$, $F_m$, $H_m$, and $I_m$ for the $m$th diode can be obtained from the Llewellyn coefficients after eliminating the space-charge factor $\xi$ by the following equation, in which $\eta = e/m_0$:

$$
(\eta/\eta_0) = (\eta_a + \eta_b)\xi/\epsilon^2.
$$

They are

$$
F_m = 1 - \frac{\eta T_m^2 q_0}{2i\epsilon_0 b_m},
$$

$$
H_m = i \frac{\eta}{\epsilon} \frac{T_m}{2\omega_1} \left( 1 + \frac{\eta m_0}{\epsilon m_b} \right) - i \left( \frac{\eta}{\epsilon} \right)^2 \frac{T_m^3 q_0}{4\omega_1 b_m},
$$

$$
I_m = \frac{\eta m_0}{\epsilon m_b} - \frac{\eta T_m^2 q_0}{2i\epsilon_0 b_m},
$$

where, according to the Llewellyn-Peterson notation, $u_{am}$ and $u_{bm}$ are respectively the dc electron velocity at the entering and the exit plane of the $m$th diode, $q_0 = q_0^1$ in reference number 7) is the dc electron current density, and $T_m$, the dc transit time for the $m$th 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 $v = v_0 + v_0 e^{i\omega t}$
- Charge density $\rho = \rho_0 + \rho_0 e^{i\omega t}$
- Convection current density $q = q_0 + q_0 e^{i\omega t}$
- Potential $V = V_0 + V_0 e^{i\omega t}$

where all quantities except $q_0$ are functions of $z$, and $\xi$ is the distance measured from some reference plane denoted as the $o$-plane; $v_0$, $\rho_0$, $q_0$ and $V_0$ are ac quantities, $v$, $q$, $\rho$, and $V$ are ac quantities, $\rho_0^+$ is the dc charge density of the positive ions, and $\rho_0$ 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 $q = q_0 e^{-i\beta}$ and $\beta = i\omega/\epsilon$, the following electronic equation may be obtained:

$$
dQ/dz + dQ/3d\epsilon_0 + Q/\epsilon_0^2 = 0 \tag{7}
$$

$$
v = -i\omega v_0 \frac{dQ}{\epsilon_0 dz} e^{-i\beta}. \tag{8}
$$

The signs of the quantities $e$ and $\rho_0$ for electrons are negative. When electrons move in the direction of posi-

\footnote{10 A similar equation derived by L. D. Smullin, M.I.T. and P. Parzen, Federal Telecommunication Laboratory, Inc. (unpublished).}
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 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 $v_0 = k z^{1/2}$.

Assume the electron stream enters the accelerating space at the $s$-plane. Define a dimensionless quantity which is always positive as

$$x = \left( \frac{-4\eta |q_0| z^{1/2}}{\epsilon k^2} \right)^2. \quad (9)$$

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

$$\begin{vmatrix} E_{ax} & F_{ax} \\ H_{ax} & I_{ax} \end{vmatrix} = \begin{vmatrix} E_{1a} & F_{1a} \\ H_{1a} & I_{1a} \end{vmatrix}^{-1} \times \begin{vmatrix} E_{1x} & F_{1x} \\ H_{1x} & I_{1x} \end{vmatrix}. \quad (10)$$

Where the double subscript is used, the first letter denotes position of initial plane and second letter denotes position of interest. $E_{1a}$, $F_{1a}$, $H_{1a}$, and $I_{1a}$, and $E_{1x}$, $F_{1x}$, $H_{1x}$, and $I_{1x}$ 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.
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

$$
\left[ \begin{array}{c}
E' \\
F' \\
H' \\
I'
\end{array} \right] = \mathbf{E} \cdot (\mathbf{E}^{-1} \cdot \mathbf{F})^{-1} \\
\left[ \begin{array}{c}
H' \\
I'
\end{array} \right]
$$

(11)

$E'$, $F'$, $H'$ and $I'$ 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 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 ease 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 \( \omega_p \) by a reduction factor

\[
\left( \frac{1}{1 + \omega_p^2/\gamma^2} \right)^{1/3}.
\]

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.

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](image)

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 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. 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 derived 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

$$ v_1 = \frac{i}{q_1} q_1 = \frac{i \omega v_0 F_1}{\omega_0 q_0} q_1, \quad (13) $$

where the subscript 1 denotes quantities before the deceleration. The subscript II will be used to denote quantities after deceleration; \( \omega_1 \) 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\(^8\)

$$ v_{11}/v_1 = v_{01}/v_{011}, \quad (14) $$

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 \( v_{01} \) takes on the form of current modulation again, according to (3)

$$ q_{11} = i g_{11} v_{11} = i \frac{\omega_0}{\omega_1 F_{01} F_{11}} v_{11}. \quad (15) $$

The final result for power gain per section from expressions (13), (14) and (15) is thus

$$ \text{power gain} = 20 \log_{10} \left( \frac{V_{01}}{V_{011}} \right)^{3/4} \left( \frac{F_1}{F_{11}} \right) \text{db}, \quad (16) $$

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.

\(^8\) Appendix I.
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.

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.

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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

\[ e(V_0 - V_0) = \frac{1}{2}m_0(v_0 + v_a)^2 - \frac{1}{2}m_0(v_0 + v_a)^2 \]

or

\[ \frac{1}{2}m_0(v_0^2 - v_0^2) = \frac{1}{2}m_0(v_0 + v_a)^2 - \frac{1}{2}m_0(v_0 + v_a)^2. \]

Assuming ac quantities \( v_a \) and \( v_b \) are small compared with dc quantities \( v_0 \) and \( v_a \), and neglecting second order small quantities, we have

\[ \frac{v_0^2}{v_a^2} = \frac{v_0}{v_a}. \]

APPENDIX II

\[ E = \frac{\{ J_1(y)N_1(s) - N_1(y)J_2(s) \}}{\{ J_1(y)N_2(s) - N_1(y)J_2(s) \}} \]

\[ F = i2G \frac{\{ J_1(y)N_1(s) - N_1(y)J_2(s) \}}{\{ J_1(y)N_2(s) - N_1(y)J_2(s) \}} \]

\[ H = \frac{i}{J_1(s)N_2(s) - N_1(s)J_2(s)} \]

\[ I = \frac{J_1(s)N_1(s) - J_2(s)N_1(s)}{J_1(s)N_2(s) - N_1(s)J_2(s)}, \]

where \( G = 2\omega_0/k^2, y = 2x^{1/4}, \) and \( y = \) 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.3,4 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 broadening 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.5

Foster’s theorem, which holds for any lossless, non-active, linear, two-terminal network, may be expressed in the following mathematical form:

\[ X(f) = \frac{1}{B(f)} = \sum_{m=0}^{\infty} \frac{\rho_m}{f - f_m^2} + \frac{qf - r}{f} \]

\[ 0 < f_s < f_1 < \ldots < f_n \]

\[ X = \frac{1}{B} = \frac{4\piMZ_0}{ab\lambda_0} \]

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 \( \rho_m \), \( q \) 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 \( \frac{1}{B} \)

\[ X = \frac{1}{B} = \frac{4\piMZ_0}{ab\lambda_0} \]

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 \( \lambda_0 \) the guide wavelength. Formulas for \( M \) are

4 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.

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† Sperry Gyroscope Company, Great Neck, N. Y.


2 The Representation, Measurement and Calculation of Equivalent Circuits for Waveguide Discontinuities with Application to Rectangular Slots," Polytechnic Institute of Brooklyn, Microwave Research Institute, Brooklyn, N. Y.; 1949.

3 H. A. Bethe, "Lumped constants for small apertures," MIT Radiation Laboratory Rep. 43-22; March 24, 1943.


known for circular and elliptical apertures, and accurate measured values are available for the shapes of Fig. 1. Since \( Z_0 \) is proportional to \( \lambda_n / \lambda \), 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 \( f \to 0 \):

\[
X = \frac{Y_2}{Z_0} = \frac{4\pi M}{ab\lambda (1 - f^2/f_0^2)}.
\]  

(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 \(-\infty \to \infty \):

\[
X = \frac{4\pi M}{ab\lambda} \left( \frac{2f_0}{f} \tan \frac{\pi f}{2f_0} \right).
\]

(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:

\[
l = \frac{\lambda_0}{2} \sqrt{1 + \left(\frac{2aw}{b\lambda y_0}\right)^2}.
\]

(5)

Equation (5) is generally accurate to within a few per cent.

For \( w/l \) small, (5) reduces to \( l = \lambda_0/2 \). 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

\[
l = \frac{\lambda_0}{2} + 0.273w.
\]

(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, and for a "ridged" cross section by the author. 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.

---


The Resonant $Q$ of an Iris

The $Q$ of a resonant iris loaded by matched waveguide terminations may be defined by

$$Q = \frac{f_0 d(B/Y_0)}{2 \left| \frac{d}{df} \right|_{f=f_0}}.$$  \hspace{1cm} (7)

Substitution of (3) in (7) gives

$$Q = \frac{ab\lambda_0}{4\pi M}.$$  \hspace{1cm} (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 $\alpha_0$ introduced by an aperture in an infinitely thin transverse conducting diaphragm in a waveguide is given by

$$\alpha_0 = 10 \log_{10} \left[ \left( \frac{B}{2Y_0} \right)^2 + 1 \right] \text{db.} \hspace{1cm} (9)$$

If the diaphragm has finite thickness and if the frequency is below resonance, the attenuation will be greater than $\alpha_0$. For very large thickness it is obvious that the increase in attenuation will approach asymptotically 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. Hence the following formula for the added attenuation $\alpha_t$ is suggested:

$$\alpha_t = \frac{54.6 t A}{\lambda_c} \sqrt{1 - \left( \frac{\lambda_c}{\lambda} \right)^2} \text{db.} \hspace{1cm} (10)$$

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; $\lambda_c$ is the cutoff wavelength of the waveguide. The total attenuation of the diaphragm is

$$\alpha = \alpha_0 + \alpha_t. \hspace{1cm} (11)$$

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.

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 $\alpha$ for $\alpha$ greater

**Table I**

<table>
<thead>
<tr>
<th>No.</th>
<th>Shape</th>
<th>$l$</th>
<th>$w/l$</th>
<th>$h/l$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Rounded slot</td>
<td>1.565</td>
<td>0.2</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Rounded slot</td>
<td>1.570</td>
<td>0.1</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>Rounded slot</td>
<td>1.785</td>
<td>0.2</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>Dumbbell</td>
<td>1.328</td>
<td>0.4</td>
<td>0.1</td>
</tr>
<tr>
<td>5</td>
<td>Dumbbell</td>
<td>1.410</td>
<td>0.3</td>
<td>0.1</td>
</tr>
<tr>
<td>6</td>
<td>Rosette</td>
<td>1.340</td>
<td>0.2</td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>Rosette</td>
<td>1.340</td>
<td>0.1</td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>Rosette</td>
<td>0.940</td>
<td>0.1</td>
<td></td>
</tr>
</tbody>
</table>

![Fig. 2](image1.png)  
Fig. 2—Attenuation versus thickness for aperture number 2 of Table I.

![Fig. 3](image2.png)  
Fig. 3—Attenuation versus thickness for aperture number 4 of Table I.
than 4.8 db, and the voltage standing-wave ratio $S$ or smaller attenuations. $S$ is convertible to $\alpha$ by

$$\alpha = 10 \log_{10} \left( \frac{S + 1}{4S} \right) \text{ db.} \quad (12)$$

The curves of attenuation versus thickness of Figs. 2 and 3 are typical.

The shunt-susceptance curves of Figs. 4 and 5 were computed, as follows, from the $\alpha$ and $S$ values for zero thickness obtained by extrapolation of the graphical data:

$$\frac{B}{Y_0} = \frac{S - 1}{\sqrt{S}} = 2\sqrt{10^{\alpha/10} - 1}. \quad (13)$$

In Fig. 4, points calculated with the aid of (3) are also shown, and the agreement with the measured points is 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. The resonant wavelengths for numbers 1, 2, 4, 5, 6, and 7 in zero-thickness diaphragms.

![Fig. 5—Normalized susceptance of apertures numbers 1, 2, 4, 5, 6, and 7 in zero-thickness diaphragms.](image)

**Table II**

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<tr>
<th>Aperture 1</th>
<th>Aperture 2</th>
<th>Aperture 3</th>
<th>Aperture 4</th>
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<td>$B$ (Meas)</td>
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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. 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, it is only recently that any careful analysis of this subject has been done. Wang's treatment of the solid beam 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 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.

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.

Fig. 1—Hollow beam in equilibrium, drift tubes, and co-ordinate system.
Conservation of energy supplies the following basic equation:

\[ r^2 + r^2 \dot{\theta}^2 + z^2 = -2 \frac{e}{m} \phi, \tag{1} \]

where the dot notation represents total time derivatives; \( e \) is the electron charge, \(-16 \times 10^{-39}\) coulomb; \( m \) is the electron mass, \( 9.11 \times 10^{-31} \) kg; and \( \phi \) 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 \( \dot{\phi} \) constant has eliminated one variable, leaving \( \dot{\theta} \) and \( \phi \) to be reduced to functions of \( r \).

Conservation of the generalized angular momentum \( p_\theta \), given by

\[ p_\theta = m r^2 \dot{\theta} + e A \theta \tag{2} \]

where \( A \) is the only component of the vector magnetic potential, leads to Busch's theorem:

\[ \dot{\theta} = \frac{e}{2mr^2} (\psi_r - \psi). \tag{3} \]

Here \( \psi \) and \( \psi_r \) 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](image)

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 \( \phi \) 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 \( \alpha \) and \( \gamma \) are functions of \( r_0 \) but not of \( r \) or \( t \), and are determined from the average potential \( \phi \) at \( r_0 \); \( \alpha \) 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

\[ \ddot{r} = -2 \frac{e}{m} \alpha \ln \frac{r}{r_0} - 2 \frac{e}{m} (\gamma - V) \]

\[ -\frac{\Omega^2}{r^2} - \omega_H^2 r^2 + 2 \omega_H \Omega, \tag{5} \]

where

\[ \dot{\psi}_r = -\frac{e}{m} V \]

\[ \psi = \pi r^2 B \]

\[ \omega_H = \frac{e}{m} \frac{B}{2} \]

\[ \Omega = \frac{e}{m} \frac{\psi_r}{2\pi} \tag{6} \]

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 \( \alpha \), 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 \( \psi_r \), we obtain

\[ \psi_r = \pi r_0^2 B \sqrt{1 + K\alpha}, \tag{7} \]

where \( K = e/m/\omega_H^2 r_0^2 \), a negative number. The quantity \( \sqrt{1 + K\alpha} \) is less than or greater than one, depending on the sign of \( \alpha \). For real values of \( \psi_r \), \( K\alpha \) 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 \( (m^2/2) \)) be a minimum at \( r_0 \). The condition that \( \partial^2\psi_r/\partial r^2 \) be negative at \( r_0 \) leads to the inequality

\[ K\alpha > -2. \tag{8} \]

Since \( \psi_r \) 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
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)

\[ \frac{R}{R_0} = 1 + \frac{x^2}{2} + x \sqrt{1 + \frac{x^2}{4} \cos \beta}, \]  

(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 \( \beta \) is

\[ \beta = -2\omega \sqrt{1 + \frac{K\alpha}{2}}. \]  

(10)

This solution is valid only for small oscillations about the equilibrium radius.

It is apparent that as the normalized oscillation amplitude increases the center of oscillation is displaced outward as shown in Fig. 3.

The form of the frequency confirms the previously derived stability condition. Fig. 4 indicates the character of \( \beta \) as a function of flux density \( B \). When \( \alpha \) 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 \( \alpha \) is negative and the electric force is inward, \( \beta \) 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 \( \beta \) be the same for all shells, hence independent of \( r_0 \). Unless \( \alpha \) 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.

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

\[ \tilde{g} = \omega \sqrt{1 + \frac{K\alpha}{2}} - 1. \]  

(11)

Since \( \tilde{g} \) has the same sign as \( \alpha \) 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 \( -\epsilon \gamma \) of which \( -\epsilon V \) is due to axial motion. The difference \( -\epsilon (\gamma - V) \) 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

\[ (\gamma - V)_{\text{opt}} = \frac{(\sqrt{1 + K\alpha} - 1)^2}{2K}, \]  

(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 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 $\psi$ must differ from the flux $\psi_e$. 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.

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 $r_o$, the cathode flux is given by

$$\psi_e = \pi r_o^2 B_c.$$  \hspace{1cm} (13)

Substituting this expression into (7), solving for $\alpha$, and then determining the charge density from Appendix I, we find that

$$\rho = -\frac{e}{r_0} \frac{\partial \alpha}{\partial r} = -\frac{e}{m} \frac{2}{2} (B_c^2 - B^2).$$  \hspace{1cm} (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 $B_{\text{eff}}^2$, it is clear that the greatest economy of magnetic field is obtained when $B_c$ is zero.

Equation 14 may be solved for $B_{\text{eff}}^2$ and expressed in terms of beam voltage and current, resulting in the general design equation

$$B_{\text{eff}}^2 = \frac{e}{m} \frac{2}{2} \frac{I}{\sqrt{V}}$$

$$= 6.9 \times 10^{-7} \frac{I}{r_o^2 - r_i^2} \sqrt{V}.$$  \hspace{1cm} (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.

$$\phi_1 = V + \frac{e}{m} \frac{r_a^2}{4} B^2 \left[ (1 - g_1^2) \ln \frac{r_o}{r_1} + \frac{3}{2} g_1^2 - g_1 - \frac{1}{2} \right],$$  \hspace{1cm} (16)

where

$$g_1 = \sqrt{1 + K \alpha} = \frac{B_e}{B}.$$  \hspace{1cm}

The potential difference between electrodes is given by

$$\phi_2 - \phi_1 = -\frac{e}{m} \frac{r_a^2}{4} B_{\text{eff}}^2 \left[ (\ln \frac{r\alpha}{r_1} - \frac{1}{2}) \right. \left. + \frac{r_b^2}{r_a^2} \left( \ln \frac{r_2}{r_b} + \frac{1}{2} \right) \right].$$  \hspace{1cm} (17)

---

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 $\psi_e$ 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.

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This difference is positive, giving an outward electric force which aggravates the effects of space charge and increases the required focusing magnetic field.

When \( B_e \neq 0 \), 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 \( (\gamma - V)_{\text{opt}} \).

\[
x^2 = 4g_1 \left( \frac{1 - g_1}{1 + g_1^2} \right) \left( 1 - \frac{r_x^2}{r_a^2} \right).
\] (18)

When \( B_e = 0 \), 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

\[
\phi_2 = V - \frac{e}{m} \frac{r_b^2}{4} B^3 \left( \frac{r_a^2}{r_b} + \frac{1}{2} \right).
\] (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 \( \Psi_e \) is the same for all electrons. As before, we solve (7) for \( \alpha \) and then find the charge density \( \rho \).

\[
\rho = \frac{e}{m} \frac{e}{2} \left( \frac{\Psi_e^2}{\pi r_a^2} + B^2 \right).
\] (20)

In contrast to the previous case the cathode-flux linkage now increases the attainable charge density. By integrating this expression for \( \rho \) over the cross section of the beam and solving for the magnetic field quantities, we reproduce exactly (15) if we define

\[
B_{\text{eff}}^2 = \left( \frac{\Psi_e}{\pi r_a^2} \right)^2 + B^2.
\] (21)

The term \( \Psi_e / \pi r_a^2 \) has the form of a geometric mean flux density at the cathode, which suggests the substitution

\[
\frac{\Psi_e}{\pi r_a^2} = g_2 B.
\] (22)

Comparison of the actual and optimum values of \( \gamma - V \) shows that they can be equated if

\[
\phi_1 = V + \frac{e}{m} \frac{r_a^2}{4} B^2 \left[ \ln \frac{r_a}{r_1} \left( 1 - \frac{r_a^2}{r_a^2} g_2^2 \right) \right.
\]

\[
\left. - \frac{1}{2} \left( \frac{r_b}{r_a} \right)^2 \right].
\] (23)

and

\[
\phi_2 - \phi_1 = \frac{e}{m} \frac{r_a^2}{4} B^2 \left[ g_2^2 \left( \ln \frac{r_a}{r_1} + \frac{1}{2} \right) \right.
\]

\[
\left. + g_2^3 \left( \ln \frac{r_b}{r_a} - 1 \right) - \left( \ln \frac{r_a}{r_1} + \frac{1}{2} \right) \right]
\]

\[- \frac{r_b^2}{r_a^2} \left( \ln \frac{r_b}{r_a} + \frac{1}{2} \right). \] (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

\[
g_2 = \left[ \left( \ln \frac{r_a}{r_1} + \frac{1}{2} \right) + \left( \ln \frac{r_a}{r_1} + \frac{1}{2} \right) \right] \left[ \ln \frac{r_b}{r_a} + \frac{1}{2} \right].
\] (25)

Another arrangement may require that the inner drift tube be eliminated. The electric field, hence \( \alpha \), then vanishes, at the inner edge of the beam and

\[
g_2 = \frac{r_a}{r_b}, \quad \psi_e = \pi r_a^2 B. \] (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, Brillouin, and Field, 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:
1. \psi_2 = \pi r_2 B_e

General

\[ B_{011} = B_1 - B_2 = B(1 - \xi_1^2) \quad B_e = \xi_1 B. \]

\[ \phi_1 = 1' + \frac{e}{m} \frac{r_3^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_3} - \frac{1}{2} \right) + \frac{3r_2^2}{r_1 r_3} \left( \ln \frac{r_2}{r_3} - \frac{1}{2} \right) \right\} \]

\[ \phi_2 = \phi_1 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \right) + \frac{3r_2^2}{r_1 r_2} \left( \ln \frac{r_2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ r_2 = 4 \xi_1 (1 - \xi_1) \left( 1 - \frac{r_1^2}{r_1} \right). \]

\[ B_e = 0 \]

\[ B_{011} = B_2^2 \]

\[ \phi_1 = 1' + \frac{e}{m} \frac{r_3^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_3} - \frac{1}{2} \right) \right\} \]

\[ \phi_2 = \phi_1 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \right) \right\} \]

\[ \phi_3 = \phi_1 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \right) + \frac{3r_2^2}{r_1 r_2} \left( \ln \frac{r_2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ \phi_4 = \phi_1 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \right) + \frac{3r_2^2}{r_1 r_2} \left( \ln \frac{r_2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ \phi_5 = \phi_1 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \left( \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \right) + \frac{3r_2^2}{r_1 r_2} \left( \ln \frac{r_2}{r_2} - \frac{1}{2} \right) \right\} \]

**Solid Beam**

\[ B_e = 0 \]

\[ \phi_0 = 1' - \frac{e}{m} \frac{B^2 r_3^2}{4} \left( \ln \frac{r_2}{r_0} + \frac{1}{2} \right). \]

**11. \psi_2 independent of \( r_0 \)**

General

\[ B_{011} = B_1 + \frac{\psi_2}{\pi r_e r_0} \]

\[ \phi_0 = 1' + \frac{e}{m} \frac{r_3^2}{4B^2} \left\{ \ln \frac{r_2 r_3}{r_1 r_3} - \frac{1}{2} \left( \frac{r_3^2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ \phi_0 = \phi_0 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \left( \frac{r_2^2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ \phi_0 = \phi_0 - \frac{e}{m} \frac{r_2^2}{4B^2} \left\{ \ln \frac{r_2 r_3}{r_1 r_2} - \frac{1}{2} \left( \frac{r_2^2}{r_2} - \frac{1}{2} \right) \right\} \]

\[ r_2 = \frac{r_2}{r_2} \left( \ln \frac{r_2}{r_2} + \frac{1}{2} \right) \]

**Inner Electrode Absent**

\[ \xi = \frac{r_3}{r_0} B_{011} = B_1 \left( 1 + \frac{r_3^2}{r_1^2} \right) \]

\[ \phi_0 = 1' + \frac{e}{m} \frac{B^2 r_3^2}{4} \left\{ 1 + \ln \frac{r_2 r_3}{r_1 r_3} - \frac{1}{2} \left( \frac{r_3^2}{r_2} + \frac{r_3^2}{r_3} \right) \right\} \]

\[ B = 0 \]

\[ B_{011} = \left( \frac{\psi_2}{\pi r_e r_0} \right) \]

\[ \phi_0 = 1' - \frac{e}{4m} \left( \frac{\psi_2}{\pi r_e} \right) \left( \ln \frac{r_2}{r_1} + \frac{1}{2} \right) \]

\[ \phi_0 = \phi_0 - \frac{e}{4m} \left( \frac{\psi_2}{\pi r_e} \right) \left( \ln \frac{r_2}{r_1} + \frac{1}{2} \right) + \frac{e}{4m} \left( \frac{\psi_2}{\pi r_e} \right) \left( \ln \frac{r_2}{r_2} - \frac{1}{2} \right) \]

**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 \( r_a = 1.13 \text{ cm} \)
- Outer radius \( r_b = 1.53 \text{ cm} \)
- Beam voltage \( V = 1,000 \text{ volts} \)
- Beam current \( I = 500 \text{ ma} \)
- 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 \( \psi_e \) 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 \( \phi_1 \)
- Outer drift-tube voltage \( \phi_2 \).

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

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**Fig. 6**—Beam focusing as a function of radial electric field.

Cathode in a uniform magnetic field.
The experiments indicated that appearance of design values. 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 \( \rho \) may be represented in the power series form
\[ \rho(r) = \sum_{n=0}^{\infty} \rho_n r^n, \quad n = 0, 1, 2, \ldots \text{ for } r_a \leq r \leq r_b \]

0 for \( r < r_a \) and \( r > r_b \). \hspace{1cm} (26)

Note that only \( \rho_0 \) has the dimensions of charge density. The evaluation of the average potential \( \phi \) consists of integrating Poisson's equation,

\[ \frac{1}{r} \frac{\partial}{\partial r} \left( r \frac{\partial \phi}{\partial r} \right) = -\frac{\rho}{\epsilon}, \hspace{1cm} (27) \]

where \( \epsilon \) 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 by letting

\[ \phi = \phi_s + \phi_L, \hspace{1cm} (28) \]

where \( \phi_s \) is a solution of Poisson's equation and \( \phi_L \) is a solution of Laplace's equation.

The potential \( \phi_s \) due to space charge alone is evaluated first. From (27), integrating twice from \( r_a \) outward,

\[ \phi_s = -\int_{r_a}^{r} \left[ \frac{1}{r} \int_{r_a}^{r} \rho \, dr' \right] \, dr, \hspace{1cm} (29) \]

where we have set \( \phi_s = 0 \) inside \( r_a \).

Before evaluating \( \phi_L \), we shall have to calculate the contribution of \( \phi_s \) at the outer electrode, i.e., at \( r_2 \). At the outer edge of the beam where \( r = r_b \),

\[ \phi_{s_2} = -\int_{r_a}^{r_b} \left[ \frac{1}{r} \int_{r_a}^{r} \rho \, dr' \right] \, dr. \hspace{1cm} (30) \]

Outside the electron beam, the potential \( \phi_s \) is represented by the familiar logarithmic expression for a line charge along the axis,

\[ \phi_s = -\frac{Q}{2\pi \epsilon} \ln \frac{r}{r_b} + \phi_{s_2}, \quad r > r_b, \hspace{1cm} (31) \]

where \( Q \) is the total charge per unit length in the beam.

\[ Q = 2\pi \int_{r_a}^{r} \rho r \, dr. \hspace{1cm} (32) \]

The potential \( \phi_s \) at \( r_b \) is thus

\[ \phi_{s_b} = -\frac{1}{\epsilon} \int_{r_a}^{r_b} \left[ \frac{\rho}{r} \, \ln \frac{r}{r_b} + 1 \int_{r_a}^{r} \rho \, dr' \right] \, dr. \hspace{1cm} (33) \]

Since \( \phi_{s_b} = \phi_L + \phi_{s_1} \), we have

\[ \phi_{s_1} = \phi_s - \phi_{s_b} = \frac{1}{\epsilon} \int_{r_a}^{r_b} \left[ \frac{\rho}{r} \, \ln \frac{r}{r_b} + 1 \int_{r_a}^{r} \rho \, dr' \right] \, dr, \hspace{1cm} (34) \]

and \( \phi_L = \phi_s - \phi_{s_1} \) since we have set \( \phi_{s_1} = 0 \). The solution for \( \phi_L \) is well known:

\[ \phi_L = \frac{\phi_s - \phi_{s_1}}{\ln \frac{r_b}{r_1}} \ln \frac{r}{r_1} \hspace{1cm} (35) \]

With the appropriate substitutions we obtain, for \( r_a \leq r \leq r_b \),

\[ \phi = a \ln \frac{r}{r_1} \hspace{1cm} (36) \]

where

\[ \phi_2 = \phi_1 - \frac{1}{\epsilon} \int_{r_a}^{r} \left[ \frac{1}{r} \int_{r_a}^{r} \rho \, dr' \right] \, dr, \hspace{1cm} (37) \]

Equation (36) may be further simplified by integrating the last term by parts, which gives us

\[ \phi = a \ln \frac{r}{r_1} + \phi_1 - \frac{1}{\epsilon} \int_{r_a}^{r} \left[ \rho \, \ln \frac{r}{r_1} - 1 \int_{r_a}^{r} \rho \, dr' \right] \, dr \]

\[ + \frac{1}{\epsilon} \int_{r_a}^{r} \rho \, \ln \frac{r}{r_1} \, dr. \hspace{1cm} (38) \]

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 \( \rho \) and perform the indicated integrations, we arrive at a general explicit form for \( \phi \), from which the corresponding expressions for several special cases may be extracted quite easily.

\[ \phi = a \ln \frac{r}{r_1} + \phi_1 - \frac{1}{\epsilon} \sum_{n=0}^{\infty} \frac{\rho_n}{(n+2)^2} \left( r^{n+2} - r_a^{n+2} \right) \]

\[ + \frac{1}{\epsilon} \sum_{n=0}^{\infty} \frac{\rho_n}{n+2} \ln \frac{r}{r_a} \hspace{1cm} (39) \]

and

\[ \phi = \frac{a}{\ln \frac{r_b}{r_1}} \ln \frac{r}{r_1} \hspace{1cm} (40) \]

**Evaluation of \( \gamma \) and \( \alpha \)**

When the beam is in equilibrium, i.e., when \( r = r_0 \), the expression derived above for the potential \( \phi \) and the radial field derived from it must agree with the potential and field derived from \( \phi = \alpha \ln \frac{r}{r_0} + \gamma \).

It is clear that \( \gamma \) is just the potential \( \phi \) evaluated at \( r_0 \). Thus,

\[ \gamma = a \ln \frac{r_0}{r_1} + \phi_1 - \frac{1}{\epsilon} \ln \frac{r_0}{r_1} \int_{r_a}^{r_0} \rho \, dr \]

\[ + \frac{1}{\epsilon} \int_{r_a}^{r_0} \rho \, \ln \frac{r}{r_1} \, dr \hspace{1cm} (41) \]

and

\[ \gamma = a \ln \frac{r_0}{r_1} + \phi_1 - \frac{1}{\epsilon} \sum_{n=0}^{\infty} \frac{\rho_n}{(n+2)^2} \left( r^{n+2} - r_a^{n+2} \right) \]

\[ + \frac{1}{\epsilon} \sum_{n=0}^{\infty} \frac{\rho_n}{n+2} \ln \frac{r_0}{r_a} \hspace{1cm} (42) \]
The condition on the radial electric field at \( r_0 \) demands that

\[
\frac{\alpha}{r_0} = \left[ \frac{\partial \Phi}{\partial r} \right]_{r=r_0},
\]

from which

\[
\alpha = a - \frac{1}{e} \int_{r_0}^{r} \rho dr,
\]

and

\[
\alpha = a - \frac{1}{e} \sum_{n=0}^{\infty} \frac{\rho_n}{n+2} \left( r_0^{n+2} - r_0^{n+2} \right) + \alpha_n. \tag{45}
\]

The second term on the right of these last two equations represents the charge per unit length between \( r_0 \) and \( r_0 \). Evidently \( \alpha \) must increase with \( r_0 \) whether it is positive or negative, and at the inner edge of the beam \( a = a \).

From (44), we have

\[
\rho(r_0) = - \frac{e}{r_0} \frac{\partial \alpha}{\partial r_0}, \tag{46}
\]

an equation used in the development of design formulas.

Once the values of \( \alpha \) and \( \gamma \) are determined from the design equation, the requisite electrode voltages may be determined from (41) or (42) and (44) or (45).

At \( r_0 = r_0 \) the integrals in (41) are zero, and we obtain

\[
\phi_1 = \gamma_0 - \sigma_0 \ln \frac{r_0}{r_1}, \tag{47}
\]

since \( a \) is just the value of \( \alpha \) at \( r_0 \), i.e., \( a = \alpha_0 \). Equation (40) may be solved for \( \phi_2 - \phi_1 \),

\[
\phi_2 - \phi_1 = \alpha_0 \ln \frac{r_0}{r_1} - \left[ \frac{Q}{2 \pi e} \ln \frac{r_0}{r_0} \right] + \frac{1}{e} \sum_{n=0}^{\infty} \frac{\rho_n}{n+2} \left( r_0^{n+2} - r_0^{n+2} \right) - \frac{1}{e} \sum_{n=0}^{\infty} \frac{\rho_n}{n+2} \ln \frac{r_0}{r_0}. \tag{48}
\]

and by differentiation

\[
\ddot{r} = - \frac{e}{m} \frac{\alpha}{r_0} + \frac{\Omega^2}{r^2} - \omega_H^2 r^2 \tag{50}
\]

Let

\[
\dot{r} = R \quad \text{and} \quad r_0^2 = R_0. \tag{51}
\]

Then

\[
R = 2\dot{r} \quad \text{and} \quad \dot{R} = 2\dot{r} + 2\dot{r}.\tag{52}
\]

Substituting (49) and (50) into (52), we obtain

\[
\dot{R} = \left[ \frac{2e}{m} \alpha - 4 \frac{e}{m} (\gamma - 1) + 4 \omega_H \right] - 2\frac{e}{m} \alpha \ln \frac{R}{R_0} - 4\omega_H^2 K. \tag{53}
\]

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

\[
\dot{R} = a_0 + a_1 R \tag{55}
\]

where

\[
a_0 = \left[ \frac{2e}{m} (\gamma - 1) + 4 \omega_H \right] - 4 \omega_H^2 \left( 1 + \frac{K\alpha}{2} \right). \tag{56}
\]

We multiply (55) by \( 2R \) and integrate once to obtain

\[
\dot{R} = a_0 R + a_1 R^2 + 2, \tag{57}
\]

obtained by substituting (49) into (56) and setting \( r = r_0 \).

The integral of (56), neglecting the additive phase constant, may be written

\[
\int \frac{R}{R_0} \left[ \frac{\sqrt{1 + \frac{K\alpha}{r} - K(\gamma - 1)}}{1 + \frac{K\alpha}{2}} \right] ^{1/2} \cos \left( 2\omega_H \sqrt{1 + \frac{K\alpha}{2}} \right) \tag{58}
\]

If we define

\[
\left[ \frac{\sqrt{1 + \frac{K\alpha}{r} - K(\gamma - 1)}}{1 + \frac{K\alpha}{2}} \right] = 1 + \frac{x^2}{2}, \tag{59}
\]

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-tuple 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 requiring a very short effective time constant. The effective time constant of an n-tuple iterated structure is shown to be \(2\pi/n(n+1)\), where \(\pi = RC\) 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 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 determine 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 a network 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 \(n\)-tuple iterated structure is the \(n\)th 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

\[
M = \begin{bmatrix} A & jB \\ jC & D \end{bmatrix},
\]

(1)

where \(A\), \(B\), \(C\), and \(D\) are real (if lossless) or complex functions of frequency.

The characteristic equation of \(M\) is obtained by introducing the variable, \(\lambda\), and setting

\[
\begin{vmatrix} A - \lambda & jB \\ jC & D - \lambda \end{vmatrix} = 0,
\]

(3)

where the left-hand side is a determinant.

Expanding (3) and using (2), the characteristic equation of \(M\) is

\[
\lambda^2 - 2\nu\lambda + 1 = 0,
\]

(4)

where we have introduced the parameter

\[
\nu = (A + D)/2.
\]

The solutions of (4) for \(\lambda\) are the “characteristic values” of \(M\).

The Cayley-Hamilton theorem states that, if the matrix \(M\) be substituted for \(\lambda\), in its characteristic equation (4) a valid matrix equation is obtained. It is true, therefore, that

\[
M^2 = 2\nu M - I.
\]

(6)

Now the Tschebyscheff\(^2\) polynomials of the second kind and orders 1 and 2 are

\[
U_1(\nu) = 1
\]

(7)

\[
U_2(\nu) = 2\nu.
\]

(8)

Subsequent orders are related by the recursion formula

\[
U_{n+1}(\nu) = 2\nu U_n(\nu) - U_{n-1}(\nu).
\]

(9)

Substituting (7) and (8) in (6),

\[
M^2 = U_2(\nu)M - U_1(\nu)I.
\]

(10)
Let us assume that
\[ M^n = U_n(\nu)M - U_{n-1}(\nu)I \]
(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),
\[ M^{n+1} = (2\nu U_n - U_{n-1})M - U_nI \]
(12)
\[ = U_{n+1}(\nu)M - U_n(\nu)I \]
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 \( \nu \) 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](image)

**The Unit RC Structure**

The transmission matrix of this unit is
\[ M = \begin{bmatrix} (1 + 1/\nu \tau) & \frac{R}{\nu \tau} \\ \frac{1}{R} & 1 \end{bmatrix}, \]
(13)

where \( \nu = j\omega \), the complex frequency, and \( \tau = RC \), the time constant. Then, from (5)
\[ \nu = 1 + 1/2\nu \tau, \]
(14)
and the transmission matrix of the \( n \)-tuple 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 \( E_0(\nu) \) is the input voltage as a function of \( \nu \), and \( E_\nu(\nu) \) the output, then using the top left-hand term of the matrix \( M^n \) as expanded according to (11),
\[ E_\nu(\nu) = \frac{E_0(\nu)}{(1 + 1/\nu \tau)U_n(\nu) - U_{n-1}(\nu)} \]

(15)

Using (9), we can write this
\[ E_\nu(\nu) = \frac{E_0(\nu)}{U_{n+1}(\nu) - U_n(\nu)}. \]

(16)

To analyze the behavior of this circuit as a differentiator, let us apply a unit step-voltage. The time-domain input is
\[ E_0(t) = 0 \text{ if } t < 0 \]
\[ = 1 \text{ if } t > 0. \]

(17)
The Fourier transform \( F \) of this is
\[ E_0(p) = \frac{1}{p}. \]

(18)
The time-domain response to the step input is then, the inverse Fourier transform of
\[ E_\nu(p) = \frac{1}{p} \]
(19)

This can be evaluated by the usual method\textsuperscript{14,15} for any specific case.

We have, however, defined the "effective time constant," \( \tau_n \), as that operative immediately after the application of voltage.

\[ 1/\tau_n = - \lim_{t \to +} \frac{dE_\nu(t)}{dt}. \]

(20)

It is convenient to remove the discontinuity in \( E(t) \) by subtracting the unit step from it.
\[ E_\nu(t) = E_\nu(t) - E(t). \]

(21)

This does not alter (20)
\[ 1/\tau_n = - \lim_{t \to +} \frac{dE_\nu(t)}{dt}. \]

(22)

We can now apply the limiting and differentiating theorems of the Fourier transform, as follows:
\[ \frac{1}{\tau_n} = - \lim_{t \to +} \frac{d}{dt} F^{-1} \{ E_\nu(p) \} \]
\[ = - \lim_{t \to +} \frac{d}{dt} F^{-1} \{ p E_\nu(p) \} \]
\[ = - \lim_{p \to +} p^2 E_\nu(p), \]

(23)

where \( F^{-1} \{ \_ \} \) means the inverse Fourier transform.

Substituting (19) and (18) in (21) and (23),
\[ \frac{1}{\tau_n} = - \lim_{p \to +} \frac{1}{p^2} \left( \frac{1}{U_{n+1}(\nu) - U_n(\nu)} - 1 \right). \]

(24)
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.

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 similar 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 estimate, 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

\[ F(\lambda) = \frac{a_0 + a_1\lambda + \cdots + a_m\lambda^m}{b_0 + b_1\lambda + \cdots + b_n\lambda^n} = \frac{h(\lambda - \lambda_{o1})(\lambda - \lambda_{o2})\cdots(\lambda - \lambda_{oM})}{(\lambda - \lambda_{p1})(\lambda - \lambda_{p2})\cdots(\lambda - \lambda_{pN})}. \]  

(1)

The quantities \( \lambda_{o1}, \lambda_{o2} \cdots \lambda_{om} \) 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\(^1\) to the ideal low-pass filter characteristics 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 \( \omega \) 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\(^2\) 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.

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.


The analogy between a two-dimensional field and the logarithm of a rational function has been exploited in connection with the approximation problem. 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 \( F \) (if \( F \) is defined in (1)).

\[
\ln F = \ln |F| + j \arg F
\]

\[
\ln |F| = \ln h + \sum_{i=1}^{n} \ln |\lambda - \lambda_{0i}| - \sum_{i=1}^{n} \ln |\lambda - \lambda_{pi}|
\]

\[
\arg F = \sum_{i=1}^{n} \arg (\lambda - \lambda_{0i}) - \sum_{i=1}^{n} \arg (\lambda - \lambda_{pi}).
\]

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)

\[
\ln |F_{c}| = \ln |\lambda - \lambda_{c}| - |\lambda - \chi|.
\]

and

\[
\arg F_{c} = \arg (\lambda - \lambda_{c})(\lambda - \chi),
\]

in which

\[
\lambda_{c} = \sigma_{c} + j\omega_{c}.
\]

If one divides \( F_{e} \) by \( \omega_{c}^{2} \) and replaces \( \lambda \) by \( j\omega \), as he is ordinarily interested only in real frequencies, (5) and (6) may be written

\[
\ln \frac{|F_{e}|}{\omega_{c}^{2}} = \ln \left| \frac{j\omega - \sigma_{c} - j}{\omega_{c}} \right| - \ln \left| \frac{j\omega - \sigma_{c} + j}{\omega_{c}} \right| + j
\]

and

\[
\arg F_{e} = \arg \left( \frac{j\omega - \sigma_{c} - j}{\omega_{c}} \right) \left( \frac{j\omega - \sigma_{c} + j}{\omega_{c}} \right).
\]

Clearly \( \ln |F_{e}| \) and \( \arg F_{e} \) are easily found as functions of \( \omega \) for any \( \lambda_{c} \) if one has simply plots of \( \ln |F_{e}|/\omega_{c}^{2} \) and \( \arg F_{e}/\omega_{c}^{2} \) as functions of \( \omega/\omega_{c} \) for appropriate values of the parameter \( \sigma_{c}/\omega_{c} \). Fig. 3 gives sketches of families of curves corresponding to (8) and (9). An accurate set of curves is given in R.I.E. Report No. 145 corresponding to Fig. 3 as well as sets of curves corresponding to Figs. 4, 7, and 9. Through use of the curves one can readily obtain plots of \( \ln |F_{e}| \) and \( \arg F_{e} \) as functions of \( \omega \) for any complex value of \( \lambda_{c} \). Fig. 3 is instructive.


3 In the following, \( F_{e} \) will always represent the product of a pair of factors with complex conjugate zeros, \( \lambda_{c} \), and \( \lambda_{p} \).

4 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 \( e \); \( \arg \) is the angle in radians unless specifically stated otherwise.

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.

\[
\ln |F_r| = \ln |\lambda - \sigma_r|,
\]

where \( \sigma_r \) is real.

\[
\text{Arg } F_r = \text{Arg } (\lambda - \sigma_r).
\]

If one divides \( F_r \) by \( -\sigma_r \), and sets \( \lambda = j\omega \), the result is

\[
\ln \left| \frac{F_r}{\sigma_r} \right| = \ln \left| \frac{j\omega}{-\sigma_r} + 1 \right|
\]

and

\[
\text{Arg } \frac{F_r}{\sigma_r} = \text{Arg } F_r \quad \text{(if } \sigma_r \text{ is negative)}
\]

\[
= \text{Arg } \left( \frac{j\omega}{-\sigma_r} + 1 \right). \quad (13)
\]

The sketch of Fig. 4 indicates the form of variation of \( \ln |F_r| \) and \( \text{Arg } F_r \) for any real critical frequency, \( \sigma_r \). 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 \( F_r \) or \( F_e \) is unchanged if the sign of \( \sigma_c \) or \( \sigma_r \) is changed, and only the sign of the argument of \( F_e \) or \( F_r \) changes when the sign of \( \sigma_c \) or \( \sigma_r \) is changed.

D. An Illustrative Example

To illustrate the choice of the first approximation, a function of the form

\[
F = \frac{1}{\lambda^3 + a_2\lambda^2 + a_1\lambda + a_0}
\]

will be chosen. The characteristics to be approximated

\[
\ln |F_r| = \ln |\lambda - \sigma|,
\]

where \( \sigma_r \) is real.

\[
\text{Arg } F_r = \text{Arg } (\lambda - \sigma_r).
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If one divides \( F_r \) by \( -\sigma_r \), and sets \( \lambda = j\omega \), the result is

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\]

and

\[
\text{Arg } \frac{F_r}{\sigma_r} = \text{Arg } F_r \quad \text{(if } \sigma_r \text{ is negative)}
\]

\[
= \text{Arg } \left( \frac{j\omega}{-\sigma_r} + 1 \right). \quad (13)
\]

The sketch of Fig. 4 indicates the form of variation of \( \ln |F_r| \) and \( \text{Arg } F_r \) for any real critical frequency, \( \sigma_r \). 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 \( F_r \) or \( F_e \) is unchanged if the sign of \( \sigma_c \) or \( \sigma_r \) is changed, and only the sign of the argument of \( F_e \) or \( F_r \) changes when the sign of \( \sigma_c \) or \( \sigma_r \) is changed.

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\]

will be chosen. The characteristics to be approximated

\[
\ln |F_r| = \ln |\lambda - \sigma|,
\]

where \( \sigma_r \) is real.
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 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, there are three adjustable quantities, \( \sigma_1, \sigma_2, \) and \( \omega_2 \). 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

\[
\frac{\partial \log |F|}{\partial \sigma_1}, \frac{\partial \arg F}{\partial \sigma_1}, \frac{\partial \log |F|}{\partial \sigma_2}, \frac{\partial \arg F}{\partial \sigma_2},
\]

and so on since

\[
\Delta \log |F| = \int_{\partial \sigma_1} \frac{\partial \log |F|}{\partial \sigma_1} d\sigma_1 + \int_{\partial \sigma_2} \frac{\partial \log |F|}{\partial \sigma_2} d\sigma_2 
+ \int_{\partial \omega_2} \frac{\partial \log |F|}{\partial \omega_2} d\omega_2
\]

Similarly,

\[
\Delta \arg F \cong (\text{for small changes}) \frac{\partial \arg F}{\partial \sigma_1} \Delta \sigma_1 + \frac{\partial \arg F}{\partial \sigma_2} \Delta \sigma_2 + \frac{\partial \arg F}{\partial \omega_2} \Delta \omega_2. \quad (17)
\]

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

\[
\frac{\partial \log |F_e|}{\partial \sigma_e}, \frac{\partial \log |F_e|}{\partial \omega_e}, \frac{\partial \log |F_e|}{\partial \omega_e}, \frac{\partial \arg F_e}{\partial \sigma_e}, \frac{\partial \arg F_e}{\partial \omega_e}, \text{ and } \frac{\partial \arg F_e}{\partial \sigma_e},
\]

[see (5), (6), (7), (10), and (11)]. Evaluation of the derivatives is an elementary exercise when one employs the relations

\[
\ln |F_e(\lambda)|_{\text{plane}} = \frac{1}{2} \ln F_e(\lambda) F_e(-\lambda)|_{\text{plane}} \quad (19)
\]

and

\[
\arg F_e(\lambda)|_{\text{plane}} = \frac{1}{2j} \ln F_e(\lambda)|_{\text{plane}}. \quad (20)
\]

Carrying through the steps of differentiating and normalizing gives for \( \partial \log |F_e/\partial \sigma_e \)

\[
\omega_e \frac{\partial \log |F_e|}{\partial \sigma_e} = 2 \frac{\sigma_e}{2.305} \left[ \left( \frac{\sigma_e}{\omega_e} \right)^2 + 1 + \left( \frac{\omega_e}{\omega_e} \right)^2 \right]^{1/2}. \quad (21)
\]

A sketch of \( \omega_e \partial \log |F_e|/\partial \sigma_e \) as a function of \( \omega_e/\omega_e \) for various values of the parameter \( \sigma_e/\omega_e \) 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

Through a similar procedure

\[
\frac{\partial \text{Arg } F_r \text{ (degrees)}}{\partial \sigma_r} = \frac{57.3 \times 3 \frac{\omega}{\omega_r} \frac{\sigma_r}{\omega_r}}{\omega_r^2} \cdot (24)
\]

Fig. 8 shows sketches of \(\omega_r(\partial \text{Arg } F_r/\partial \sigma_r)\) and \(\omega_r(\partial \text{Arg } F_r/\partial \omega_r)\) for various values of the parameter \(\sigma_r/\omega_r\). 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.

In an analogous manner to that just indicated for conjugate critical frequencies, the corresponding derivative functions for real critical frequencies can be determined.

\[
-\frac{\partial \text{Arg } F_r}{\partial \sigma_r} = \frac{-1}{2.305 \left[ 1 + \left( \frac{\omega}{\sigma_r} \right)^2 \right] \frac{F_r}{\omega_r^2}} \cdot (25)
\]

\[
-\frac{\partial \text{Arg } F_r}{\partial \omega_r} = \frac{57.3 \times \frac{\omega_r}{\omega}}{1 + \left( \frac{\omega}{\sigma_r} \right)^2} \cdot (26)
\]
Fig. 9 presents sketches of $-\sigma_{\omega} \left( \frac{d \arg F_1}{d \sigma} \right)$ and $-\sigma_{\omega} \left( \frac{d \log |F_1|}{d \sigma} \right)$ as functions of $\omega = -\sigma$. 

**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 $\sigma_1$ will cause an improvement since the first estimate gives insufficient amplification in the low range. Rough calculations suggest the values $\Delta \omega_1 = 0.2$, $\Delta \sigma_1 = -0.2$ 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.

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 method,...
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 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,

\[
\Delta \log |F| \approx \frac{\partial \log |F|}{\partial \sigma_1} \Delta \sigma_1 + \frac{\partial \log |F|}{\partial \sigma_2} \Delta \sigma_2 \\
+ \frac{\partial \log |F|}{\partial \omega_2} \Delta \omega_2, \\
\Delta \arg F \approx \frac{\partial \arg F}{\partial \sigma_1} \Delta \sigma_1 + \frac{\partial \arg F}{\partial \sigma_2} \Delta \sigma_2 \\
+ \frac{\partial \arg F}{\partial \omega_2} \Delta \omega_2, \\
\tag{16}
\]

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.

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...
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 of 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), $F_1$ a vector in 11 space associated with the ordinates of (a), $F_2$ a vector associated with (b), and $F_2$ a vector associated with (c). For the situation illustrated in Fig. 15, one has the following vectors:

$$
\begin{align*}
F_D &= +0.018, -0.010, +0.004, +0.008, -0.006, -0.006, +0.030, +0.024, -0.004, -0.027, -0.027 \\
F_1 &= +1.08, +0.87, +0.54, +0.33, +0.22, +0.15, -0.29, -0.37, -0.33, -0.29, -0.25 \\
F_2 &= +0.29, +0.33, +0.40, +0.56, +0.76, +0.98, +0.059, +0.11, +0.17, +0.13, -0.055 \\
F_3 &= -0.69, -0.69, -0.71, -0.73, -0.64, -0.33, +0.065, +0.15, +0.26, +0.44, +0.60
\end{align*}
$$

(27)

The problem to be solved, expressed in vector terminology, is to select $\Delta \sigma_1, \Delta \sigma_2$ and $\Delta \omega_3$ so that

$$
F_D \cong \Delta \sigma_1 F_{\sigma_1} + \Delta \sigma_2 F_{\sigma_2} + \Delta \omega_3 F_{\omega_3}.
$$

(28)

The $\Delta$'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, even being larger than three. Accordingly, one must establish a measure of approximation and, on the basis of it, choose the $\Delta$'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 $\Delta$'s are chosen so that

$$
[F_D - \Delta \sigma_1 F_{\sigma_1} - \Delta \sigma_2 F_{\sigma_2} - \Delta \omega_3 F_{\omega_3}] \text{ [Same]} \text{ is minimized.}
$$

(29)

The solution for the $\Delta$'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 $F_{\omega_3}$ and first approximately $F_D$ with them. Finally, one is led to the $\Delta$'s of (28). A simple set of normal orthogonal vectors is

$$
\begin{align*}
F_{n_1} &= a_1 F_{\sigma_1} \\
F_{n_2} &= a_2 F_{\sigma_1} + a_2 F_{\sigma_2} = b_1 F_{\sigma_1} + a_2 F_{\sigma_2} \\
F_{n_3} &= a_3 F_{\sigma_1} + a_3 F_{\sigma_2} + a_2 F_{\omega_3} \\
 &= b_3 F_{n_1} + b_2 F_{n_2} + a_3 F_{\omega_3}
\end{align*}
$$

(30)

The $a$'s are chosen to fulfill the following relationships:

$$
\begin{align*}
F_{n_1} F_{n_1} &= 1 \\
F_{n_1} F_{n_2} &= 0 \\
F_{n_2} F_{n_2} &= 1 \\
F_{n_2} F_{n_3} &= 0 \\
F_{n_3} F_{n_3} &= 0 \\
F_{n_3} F_{n_3} &= 1
\end{align*}
$$

(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 $F_{n_1}$ as found from the first and the expression for $F_{n_2}$ in terms of $F_{n_1}$ and $F_{n_2}$) 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 $a_{22}$), one can evaluate $a_{22}$ and $a_{23}$. The third row of (31) (using $F_{n_1}$ and $F_{n_2}$ and the expression for $F_{n_3}$ in terms of $F_{n_1}, F_{n_2},$ and $F_{n_3}$) 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 involves $a_{33}$ only. Hence by considering the equations in the proper order, one completely avoids the solution of general simultaneous equations.
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:

\[
\begin{align*}
  d_{11} &= 0.5902 \\
  d_{21} &= -0.3608 \\
  d_{31} &= 1.063 \\
  d_{12} &= 0.748 \\
  d_{22} &= 0.6356 \\
  d_{32} &= 1.463 \\
  d_{13} &= d_{23} = d_{33} = 0
\end{align*}
\]

\((32)\)

\[F_{n1} = +0.6374, +0.5135, +0.3187, +0.1948, +0.1298, +0.0885, -0.1712, -0.2184, -0.1948, -0.1712, -0.1476\]

\[F_{n2} = -0.1144, -0.0200, +0.1335, +0.3177, +0.5010, +0.6870, +0.1331, +0.1958, +0.224, +0.1862, +0.1355\]

\[F_{n3} = +0.3228, +0.1250, -0.2103, -0.3609, -0.2191, +0.2997, -0.1756, -0.1038, +0.1376, +0.4180, +0.5768\]

Once the normal orthogonal vectors are selected, one approximates \(F_D\) in terms of them.

\[F_D = c_1F_{n1} + c_2F_{n2} + c_3F_{n3}.\]

\((34)\)

This approximation involves choice of the \(c\)'s so that

\[\left|F_D - (c_1F_{n1} + c_2F_{n2} + c_3F_{n3})\right|\] 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 orthogonality of the vectors, one has

\[\begin{align*}
  \frac{\partial}{\partial c_1} & = 2c_1 - 2c_1F_{d1}F_D = 0 \\
  \frac{\partial}{\partial c_2} & = 2c_2 - 2c_2F_{d2}F_D = 0 \\
  \frac{\partial}{\partial c_3} & = 2c_3 - 2c_3F_{d3}F_D = 0
\end{align*}\]

\((36)\)

To minimize (35), one determines the \(c\)'s by setting the partial derivatives with respect to \(\sigma_1\), \(\sigma_2\), and \(\omega_2\) of (36) equal to zero. This step gives

\[\begin{align*}
  \frac{\partial}{\partial c_1} & = 2c_1 - 2F_{n1}F_D = 0 \\
  \frac{\partial}{\partial c_2} & = 2c_2 - 2F_{n2}F_D = 0 \\
  \frac{\partial}{\partial c_3} & = 2c_3 - 2F_{n3}F_D = 0
\end{align*}\]

\((37)\)

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,

\[\begin{align*}
  c_1 &= F_{n1}F_D \\
  c_2 &= F_{n2}F_D \\
  c_3 &= F_{n3}F_D
\end{align*}\]

\((38)\)

Finally, (38) leads to the optimum pole shifts indicated in (39).

\[\begin{align*}
  \Delta\sigma_1 &= c_1a_{11} + c_2a_{21} + c_3a_{31} \\
  \Delta\sigma_2 &= c_2a_{22} + c_3a_{32} \\
  \Delta\omega_2 &= c_3a_{33}
\end{align*}\]

\((39)\)

In connection with the illustrative example being considered, one has

\[
\begin{align*}
  c_1 &= 0.00175, \quad \Delta\sigma_1 = -0.039, \quad \sigma_1 = -0.439 \\
  c_2 &= 0.00086, \quad \Delta\sigma_2 = -0.024, \quad \sigma_2 = -0.474 \\
  c_3 &= 0.0374, \quad \Delta\omega_2 = -0.055, \quad \omega_2 = 1.011
\end{align*}
\]

\((40)\)

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 \(a\)'s and the \(x\)'s how the final adjustment changed the characteristics. Quantity 35 changed from 0.0003 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 adjustment is very effective in simple cases. However, one can now state definitely that it is impossible to obtain a better approximation than the "a" 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 accompany...
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\textsuperscript{\dagger}, 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 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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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 $\rho'$, $\phi'$, and $z'$, and as co-ordinates above the surface...
where \( \rho, \phi, \) and \( z \), as illustrated in Fig. 2. We will confine the present analysis to fields that have no variation in the \( \phi' \) 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](image)

**Fig. 2**—The co-ordinate system used to describe the annular corrugated-surface antenna.

Borrowing a result from Stratton, we can write down immediately the components of the field existing above the slots as

\[
\begin{align*}
E_x &= -j \beta H_0 \iota^{(1)} \beta \eta e^{j(wt-kz)} \\
E_y &= \beta^2 H_0 \iota^{(1)} \beta \eta e^{j(wt-kz)} \\
H_0 &= \frac{k^2}{j \mu_0} H_0 \iota^{(1)} \beta \eta e^{-j(wt-kz)},
\end{align*}
\]

where

\[ k^2 - \beta^2 = \eta^2. \]

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

\[
\frac{E_x}{H_0} = j \sqrt{\frac{\mu_0}{\epsilon_0}} \tan \eta k d = j \eta \tan kd,
\]

where \( d \) is the slot depth.

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

\[
\frac{E_y}{H_0} \bigg|_{z=z'} = -\frac{\omega \mu_0 \theta}{k^2}. \tag{3}
\]

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 \( E_y = 0 \) along the barrier rings. The expression is

\[
\frac{E_y}{H_0} = j \frac{t}{t + \eta} \eta \tan \eta k d. \tag{4}
\]

where \( t \) is the slot width and \( \omega \) the phase velocity in the 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.

Combining (3) and (4) we find

\[
h = -j k \left( \frac{t}{t + \eta} \right) \tan \eta k d. \tag{5}
\]

Therefore, the field of the annular corrugated-surface antenna decays in an exponential fashion in the positive \( z \) direction for \( m \pi \leq \eta k d \leq m \pi + (\pi/2) \) where \( m = 0, 1, 2 \ldots \). These are the only allowable ranges of slot depths that will permit a surface-guided wave to propagate.

The propagation constant, \( \beta \), of the surface wave in the \( \rho' \) direction is

\[
\beta = k \sqrt{1 + \left( \frac{t}{t + \eta} \right)^2 \tan^2 \eta k d}. \tag{6}
\]

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 \( \rho, \phi, z \) within a volume \( V \) surrounded by a surface \( S \) (with outward directed normal \( \hat{n} \)), and containing no charges or currents is given by Stratton as

\[
\overline{\mathbf{H}}(\rho) = \frac{1}{4\pi} \int_S \left[ \nabla \times \mathbf{E}(r') \right] G(\rho, r') - \left[ \nabla \times \mathbf{H}(r') \right] \delta(\rho - r') \mathrm{d}V, \tag{7}
\]

where \( r' \) is a point on the surface \( S \) and \( G(\rho, r') \) is a Green's function satisfying the wave equation

\[
(\nabla^2 + k^2)\mathbf{E}(r', r') = -\delta(r - r'), \tag{8}
\]

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 \( (r \to \infty) \). It is also assumed that the fields existing at \( z' = 0 \) are excited by some


source 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

$$H(r) = \frac{1}{4\pi} \int_{S} - j \omega \epsilon(r, r') \frac{\partial}{\partial z'} E(r', \xi) \, \mathrm{d}a.$$  \hspace{1cm} (9)

The appropriate Green's function is

$$G(r, r') = \frac{1}{4\pi} \left( \frac{e^{-jkr}}{R_1} + \frac{e^{-jkr'}}{R_2} \right),$$  \hspace{1cm} (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 $\rho > \rho_1$, as is true in the far-zone field,

$$R' = \sqrt{\rho^2 + z^2 \left[ 1 - \frac{\rho \rho' \cos (\phi - \phi')} {\rho^2 + z^2} \right]}.$$  \hspace{1cm} (11)

Transforming to spherical co-ordinates by the relations

$$\rho^2 + z^2 = R^2 \quad \text{and} \quad R \cos \phi = \rho,$$

we find

$$G[K, (\phi - \phi'), \theta] = \frac{1}{2\pi} \frac{e^{-jkr}}{R} e^{jkr'} \cos (\phi - \phi') \cos \theta.$$  \hspace{1cm} (12)

Substituting (12) into (9) we find

$$H_o = \frac{j \omega \epsilon_{0} \beta}{8\pi^2 R} \int_{0}^{\pi} \int_{0}^{2\pi} F_{a}(\rho') e^{jkr'} \cos (\phi - \phi') \cos \theta \cos (\phi - \phi') \rho' \mathrm{d}\rho' \mathrm{d}\phi.$$  \hspace{1cm} (13)

where $a$ and $\rho_1$ are the inner and outer radii of the corrugated surface, respectively.

Integration with respect to $\phi'$ in (13) and substitution of the explicit expression for $F_a(\rho')$ from (1) gives

$$H_o = \frac{-j \omega \epsilon_{0} \beta}{4\pi R} \int_{0}^{\rho_1} \int_{0}^{\phi_1} H_1^{(2)}[\beta \rho] J_1(k \rho' \cos \theta) \rho' \mathrm{d}\rho' \mathrm{d}\phi.$$  \hspace{1cm} (14)

Integration of (14) gives the magnetic component of the far-zone field as

$$H_o = \frac{-j \omega \epsilon_{0} \beta}{\left[ \beta^2 - k^2 \cos^2 \theta \right]} \int_{0}^{\phi_1} \int_{0}^{\rho_1} \left[ k \rho_1 \cos \theta H_1^{(1)}(\beta \rho_1) J_0(k \rho_1 \cos \theta) 
- \beta \rho_1 H_0^{(1)}(\beta \rho_1) J_1(k \rho_1 \cos \theta) 
- k a \cos \theta \beta I_1^{(0)}(\beta \rho_1) J_0(k \rho_1 \cos \theta) 
+ \beta a I_0^{(0)}(\beta \rho_1) J_1(k \rho_1 \cos \theta) \right].$$  \hspace{1cm} (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.

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 $|H_1^{(0)}(1.595)|$, 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
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 $(\beta-k)(\rho_1-a)\sim2\pi$, 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.
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. Reynolds 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:\ 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\ 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. 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: 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

\1 Massachusetts Institute of Technology, Cambridge, Mass.
\3 Synthesis of RC-networks, Jour. of Math. and Physics, vol. 28, no. 1, p. 22; April 1949.
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 $\rho = \rho w$ (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 Seward B. Cohn, see page 161 of the November, 1951, issue of the PROCEEDINGS OF THE I.R.E.

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. 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 acoustic 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.

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.

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, 1925. 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. 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 architect electronic technician of 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.

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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 Radio physics 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.

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.

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.

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.

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.

From 1939 to 1947 Mr. McCandless was employed as a draftsman and designer on electromagnetic 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, Morgantown, N. J. and with Electronic Associates, Inc., Log Branch, N. J., where he was project engineer and chief of quality control, respectively, on various research and development items for USN, USA, and USAF.

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.
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.

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, thyatrons, stuctrolons, 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 1944 to 1947. Since 1937 he has been associated with the mathematics department of Case Institute of Technology where he is now professor of mathematics.

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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.

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

\[ C = W \log \left( 1 + \frac{P}{N} \right) \]  

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)

\[ C = \frac{W}{P} \log \left( 1 + \frac{P}{N} \right) \]  

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

\[ \frac{P}{N} \rightarrow 0, \]

and therefore the upper bound is given by

\[ C/P = \frac{W}{N} \log e. \]  

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

\[ N = \frac{kT}{4R} = kTW \]  

where

- \( k \) = Boltzmann's constant = \( 1.38 \times 10^{-23} \) joule per °K
- \( T \) = absolute temperature.

From this we have the relationship below as the upper bound to \( W/N \)

\[ \left( \frac{W}{N} \right)_{ub} = \frac{1}{kT} \]  

and (5) in (3) gives

\[ \left( \frac{C}{P} \right)_{ub} = \frac{1}{kT} \log e. \]  

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

\[ \left( \frac{C}{P} \right)_{ub} = \frac{P}{kT} \log e. \]
Correspondence

and that the lower bound to the signal power required to transmit \( C \) is

\[ P_{th} = C k T \log_2 \left( \frac{1}{2} \right). \]  

(8)

is 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 an never be less than \( 2.85 \times 10^{-13} \) 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 \( 2.85 \times 10^{-13} \) 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 microvolt of battery power. Its figure of merit would be \( 2.85 \times 10^{-13} \). This shows that \( n \) 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 \( 2.85 \times 10^{-13} \) 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 \( \vec{E} = \vec{E}_0 + \vec{E}_s, \vec{H} = \vec{H}_0 + \vec{H}_s \), where \( \vec{E}_0, \vec{H}_0 \) is the field due to the source \( P \), and \( \vec{E}_s, \vec{H}_s \) 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 \( \vec{E}_C = \vec{E}_0 + \vec{E}_1, \vec{H}_C = \vec{H}_0 + \vec{H}_1 \), where \( \vec{E}_1, \vec{H}_1 \) is the field caused by the source \( P^* \) and \( \vec{E}_s, \vec{H}_s \) is the field resulting from the sources induced in the conducting portions of \( C \).

* Received by the Institute, June 25, 1951.

\[ \vec{H}_s = \vec{E}_0, \]  

(3) can alternatively be expressed as

\[ \frac{\vec{E}_s}{\vec{E}_0} \cdot \frac{\vec{H}_s}{\vec{H}_0} = 1. \]

(1)

The prorated diffracted fields may be defined by

\[ U_1 = \frac{\vec{E}_s}{\vec{E}_0}, \quad U_2 = \frac{\vec{H}_s}{\vec{H}_0} \]

to obtain Booker's extension of Babinet's Principle,

\[ U_1 + U_2 = 1. \]  

(5)

This is a scalar relationship which is only valid if \( \vec{E}_s \) and \( \vec{H}_s \) 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 \( \vec{E}_s \) is not. Since \( \vec{E}_0 = \vec{E}_s + \vec{E}_m \), it follows that \( \vec{E}_s \) and \( \vec{E}_m \) are not collinear for this case; hence, from (2) \( \vec{E}_s \) and \( \vec{H}_s \) are not collinear.

\[ \vec{E}_s + \vec{H}_s = \vec{E}_0 \]  

so that

\[ \vec{E}_s + \vec{H}_s = \vec{E}_0. \]

(2)

Equation (2) is an accurate statement of Babinet's Principle.

If \( \vec{E}_s \) and \( \vec{H}_s \) are collinear at every point, (2) can be written as a scalar equation and divided by \( \vec{E}_0 \) to give

\[ \frac{\vec{E}_s}{\vec{E}_0} + \frac{\vec{H}_s}{\vec{E}_0} = 1; \]

(3)

and since \( P^* \) has been chosen so that

\[ \vec{H}_s = \vec{E}_0, \]  

(3) can alternatively be expressed as

\[ \frac{\vec{E}_s}{\vec{E}_0} \cdot \frac{\vec{H}_s}{\vec{H}_0} = 1. \]

(1)

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

\[ \vec{E}_0 + \vec{H}_s = \vec{E}_s \]  

(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.

The formation of a joint AIEE-IRE Committee has been initiated by CCIR Study Group 14 andards Committee, Chairman Jensen to 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 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 A. G. Jensen on March 20, the Chairman proposed that H. F. Roys be appointed IRE Representative on ASA 257, 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 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.
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 USM, 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


This Group will sponsor a Symposium at the Mackay Radio Transmission 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.

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.

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<td>PGA-7; Editorials, technical papers, and news (48 pages)</td>
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* Public libraries and colleges can purchase copies at IRE Member rates.

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)
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 1945. 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.

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\[
C = W \log \left(1 + \frac{P}{N}\right)
\]  (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."

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)

\[
C = \frac{W}{P} \log (1 + P/N). \quad (2)
\]

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

\[
\left(\frac{C}{P}\right)_{\text{ub}} = \frac{W}{N} \log_2 e. \quad (3)
\]

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

\[
N = \frac{E^2}{4kT} \quad (4)
\]

where

- \(k\) = Boltzmann’s constant = 1.374 \times 10^{-21} \text{ joule per °K}
- \(T\) = absolute temperature.

From this we have the relationship below as the upper bound to \(W/N\)

\[
\left(\frac{W}{N}\right)_{\text{ub}} = \frac{1}{kT}. \quad (5)
\]

and (5) in (3) gives

\[
\left(\frac{C}{P}\right)_{\text{ub}} = \frac{1}{kT} \log_2 e. \quad (6)
\]

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

\[
C_{\text{ub}} = \frac{P}{kT} \log_2 e. \quad (7)
\]
and that the lower bound to the signal power required to transmit $C$ is

$$P_0 = CKT \log_2 2.$$  (8)

As an example of the use of these equations, assume that we wish to transmit $10^6$ bits of information per second (at $300^\circ$K).

Equation (8) says that the signal power can never be less than $2.85 \times 10^{-11}$ 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 $2.85 \times 10^{-11}$ 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 $1 \mu$watt of battery power. Its figure of merit would be $2.85 \times 10^{-4}$. It 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 times gives bits and $P$ times time give energy. Multiplying (8) by one second, we find that at $300^\circ$K, one binary bit of information will always cost more than $2.85 \times 10^{-11}$ 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.

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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, infinitely 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 $E_0 = E_0^r + E_0^s$, $H_0 = H_0^r + H_0^s$, where $(E_0, H_0)$ is the field due to the source $P$ and $(E_0^r, H_0^r)$ 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 the conducting and open regions of $S$ have been interchanged. The total field to the right of $C$ is $E_0 = E_0^r + E_0^s$, $H_0 = H_0^r + H_0^s$, where $(E_0, H_0)$ is the field caused by the source $P'$ and $(E_0^r, H_0^r)$ is the field resulting from the sources induced in the conducting portions of $C$.

If $P'$ is chosen so that $H_0^s = -H_0^r$, it can be shown that

$$H_0^s = -E_0^r, \quad E_0^r = \frac{\mu}{\kappa} H_0^r$$  (1)

so that

$$E_0 + H_0^r = E_0.$$  (2)

Equation (2) is an accurate statement of Babinet's Principle.

If $E_0$ and $H_0^r$ are collinear at every point, then it can be written as a scalar equation and divided by $E_0$ to give

$$E_0^r = -H_0^r; \quad E_0^r = \frac{H_0^r}{E_0} = 1,$$  (3)

and since $P'$ has been chosen so that

$$H_0^s = E_0^r, \quad (3)$$

can alternatively be expressed as

$$E_0^r + H_0^r = 1.$$  (1)

The prorated diffracted fields may be defined by

$$U_1 = E_0^r, \quad U_2 = \frac{H_0^r}{H_0^s},$$

and in this manner, the total diffracted field is

$$U_1 + U_2 = 1.$$  (5)

This is a scalar relationship which is only valid if $E_0$ and $H_0^r$ 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, $E_0$ is horizontally polarized but $H_0^r$ is not. Since $E_0 = E_0^r + E_0^s$, it follows that $E_0$ and $E_0^r$ are not collinear for this case; hence, from (2) $E_0^r$ and $H_0^r$ are not collinear.

* Received by the Institute, June 25, 1951

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

$$E_0 + H_0^r = E_0^r.$$  (2)

valid for thin, perfectly conducting, infinite, plane complementary screens excited by conjugate sources.

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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 antiship submarine warfare in the Caribbean Sea Frontier, and the Medal for Merit for achievements in submarine warfare in the Pacific.

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.

Ping King Tien was born in Chekiang, China, on August 2, 1919. He received the B.S. degree in electrical engineering from the National Central University of China in 1941, and from Stanford University the M.S. degree in 1948 and the Ph.D. degree in 1951.

Dr. Tien was an engineering assistant at the Tien-Sun Industrial Company, Shanghai, China, from 1941 to 1947, and was a research assistant at Rensselaer Polytechnic Institute, Troy, New York, from 1947 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.

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

\[ C = W \log_2 \left( 1 + \frac{P}{N} \right) \]  

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."

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)

\[ C = \frac{W}{P} \log_2 (1 + P/N). \]  

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

\[ P/N \to 0, \]

and therefore the upper bound is given by

\[ \left( \frac{C}{P} \right)_{ub} = \frac{W}{N} \log_2 \frac{1}{kT}. \]  

One may inquire as to how the upper bound to \( W/N \) in a circuit of bandwidth \( W \), the minimum available noise power will be the thermal noise power

\[ N = \frac{P}{4kR} = kTW. \]  

where

- \( k = \text{Boltzmann's constant} = 1.374 \times 10^{-23} \) joule per °K
- \( T = \text{absolute temperature}. \)

From this we have the relationship below as the upper bound to \( W/N \)

\[ \left( \frac{W}{N} \right)_{ub} = \frac{1}{kT} \]  

and (5) in (3) gives

\[ \left( \frac{C}{P} \right)_{ub} = \frac{1}{kT} \log_2 \varepsilon. \]  

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

\[ C_{ub} = \frac{P}{kT} \log_2 (1/e). \]
Correspondence

and that the lower bound to the signal power required to transmit \( C \) is

\[ P_a = C k T \log_2 \frac{1}{P}. \]  

(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 than \( 2.85 \times 10^{-18} \) 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 \( 2.85 \times 10^{-18} \) 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 \( 2.85 \times 10^{-10} \). 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 gives hits and \( P \) times gives energy. Multiplying (8) by one second, we find that at 300°K, one binary bit of information will always cost more than \( 2.85 \times 10^{-10} \) 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 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 \( E = E_s + E_o, H = H_s + H_o \), where \( E_s, H_s \) is the field due to the source \( P \) and \( E_o, H_o \) 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 \( E = E_s + E_o, H = H_s + H_o \), where \( E_s, H_s \) is the field caused by the source \( P' \) and \( E_o, H_o \) is the field resulting from the sources induced in the conducting portions of \( C \).

Fig. 1(a) and (b)—Both screens are infinite in extent. The dotted regions represent conductor.

If \( P' \) is chosen so that \( H_s = -H_o \), it can be shown that

\[ E_s + H_s = E_o, \]

so that

\[ E_s + H_s = E_o. \]  

(2)

Equation (2) is an accurate statement of Babinet's Principle.

If \( E_o \) and \( H_s \) are collinear at every point, (2) can be written as a scalar equation and divided by \( E_o \) to give

\[ \frac{E_s}{E_o} + \frac{H_s}{H_o} = 1; \]

(3)

and since \( P' \) has been chosen so that

\[ H_s = -H_o \]

\( E_s + H_s = E_o \) (3) can alternatively be expressed as

\[ E_s + H_s = 1. \]  

(1)

The prorated diffracted fields may be defined by

\[ U_1 = \frac{E_s}{E_o}, \quad U_2 = \frac{H_s}{H_o} \]

to obtain Booker's extension of Babinet's Principle,

\[ U_1 + U_2 = 1. \]  

(5)

This is a scalar relationship which is only valid if \( E_s \) and \( H_s \) 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, \( E_s \) is horizontally polarized but \( E_o \) is not.

Since \( E_s = E_o + E_o \), it follows that \( E_s \) and \( E_o \) are not collinear for this case; hence, from (2) \( E_s \) and \( H_s \) are not collinear.

Fig. 4

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

\[ \frac{E_s}{E_o} + \frac{H_s}{H_o} = \alpha. \]  

(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.

* Received by the Institute, June 25, 1951.

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 \( E_s = -H_o \), where \( k \) equals 1 ohm and, for simplicity, has been deleted from the expressions above.


August 27-29, and Remote Control, F. W. Lehman, Member.

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: Aerospace Electronics, G. M. Greene; Antennas and Propagation, L. C. Van Atta; Audio, J. K. Hilliard; Broadcast and Television Receiving, H. F. Rice; Broadcast Transmitter Systems, P. G. Caldwell; Circuit Theory, W. R. Abbott; Electron Devices, H. Q. North; Electronic Computers, H. E. Hulse; Information Theory, A. S. Fulton; Instrumentation, W. D. Hersberger; Radio Telemetry and Remote Control, F. W. Lehman. 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 W. W. Fernald 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 was in favor of 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 sub-committee of the Standards Committee. Amending 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 Subcommittees M.9 and 4.9, were agreed upon.

The Electronic Computers Committee met on February 25. Robert Scull took the chair in the absence of Nathaniel Rochester. It was proposed that a second subcommittee on definitions be established in the East to coordinate 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 interchangeable 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. Scull informed the Committee 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, First Chairman of the Standards Committee and Definitions Coordinator, 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 exceeds 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 75.81 "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 largely pending outstanding contributions of 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 regarding 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 L. E. Lempert for his review of the current Standard and the suggestion of much needed changes.

Under the Chairmanship of M. W. Pierce, 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.


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.


IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20-22.

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.

AERONAUTICAL ELECTRONICS

The Group's Aeronautical Electronics Convention held in Dayton, Ohio, May 12-14, was a huge success. The first issue of the Aeronautical Electronics Group's Newsletter, containing 50 pages and issued to the members, was distributed to those attending the Dayton Convention. The TRANSACTIONS (PGAE-3) has been mailed to the Group members.

A Los Angeles Chapter of the Aeronautical Electronics Group has been formed, and plans are being organized for a Chicago Chapter.

AVIATION AND PROPULSION

The Group on Aviation and Propulsion sponsored the Spring Meeting of IRE 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 1953 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


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. I. 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.

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.

<table>
<thead>
<tr>
<th>Sponsoring Group</th>
<th>Publication</th>
<th>Group Members</th>
<th>IRE Members</th>
<th>Non-Members Members</th>
</tr>
</thead>
<tbody>
<tr>
<td>Audio</td>
<td>PGA-7; Editorials, technical papers, and news (48 pages)</td>
<td>$1.10</td>
<td>$1.65</td>
<td>$3.30</td>
</tr>
<tr>
<td>Airborne Electronics</td>
<td>PGAE-3; Symposium on the Integration of Electronic Equipment with Air-Frame Design (28 pages)</td>
<td>$0.85</td>
<td>$1.25</td>
<td>$2.50</td>
</tr>
<tr>
<td>Instrumentation</td>
<td>PGA; Symposium on Subaudio Instrumentation (104 pages)</td>
<td>$0.85</td>
<td>$1.25</td>
<td>$2.50</td>
</tr>
</tbody>
</table>

* Public libraries and colleges can purchase copies at IRE Member rates.
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 prestige of engineering students by bringing to them practicing engineers from the industry.

The arrangements for the speakers has been co-ordinated by A. D. Haddeke, 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. Abraham, A. R. Arsen, R. E. Hansen, J. L. Jones, P. W. Hoxeels, 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.

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 activities of American radio engineers are disseminated through the PROCEEDINGS, and thus giving stimulus to workers in this field everywhere and, in the best possible way, promoting international cooperation.

With the best wishes for the future of The Institute of Radio Engineers and its officers, I am—

Very sincerely yours,

JORGEN RYBNER
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, local 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 100, 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 41 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 past closed leaves it with 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
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 1.4 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.

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. Chatterjee (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)

Actual membership figures for 1949, 1950, and 1951 are shown in Table I. Of the 16,519 nonvoting Associates, 4,151 have been in that grade for more than five years. 3,059

<p>| TABLE I.—TOTAL MEMBERSHIP DISTRIBUTION BY GRADES |
|-------------|-----------------|-----------------|-----------------|</p>
<table>
<thead>
<tr>
<th>Grade</th>
<th>As of Dec. 31, 1951</th>
<th>As of Dec. 31, 1950</th>
<th>As of Dec. 31, 1949</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fellow</td>
<td>Number</td>
<td>Per Cent of Total</td>
<td>Number</td>
</tr>
<tr>
<td>Senior Member</td>
<td>345</td>
<td>1.2</td>
<td>321</td>
</tr>
<tr>
<td>Member</td>
<td>6,070</td>
<td>20.4</td>
<td>6,716</td>
</tr>
<tr>
<td>Associate</td>
<td>17,521***</td>
<td>59.6</td>
<td>18,305***</td>
</tr>
<tr>
<td>Student</td>
<td>4,427</td>
<td>15.1</td>
<td>7,560</td>
</tr>
<tr>
<td>Totals</td>
<td>29,408</td>
<td>100.0</td>
<td>32,602</td>
</tr>
</tbody>
</table>

Members
Arthur A. Bertram (A’44-M’49)
Alfred J. Bremmer (M’46)
Harold M. Brett (M’44)
Theodore B. Buddele (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. Drum (A’51)
Waldo D. Elmer (A’44)
Oscar Hachoguan (A’45)
Peter L. Harbard (S’43-A’45)
Donald M. Harring (A’45)
Henry S. Kaminiski (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 Ossowsky (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.

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 3,052, 1,424 were editorial pages and 1,628 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 1933 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.

PROCEDINGS OF THE I.R.E.

<table>
<thead>
<tr>
<th>TABLE II.—SUMMARY OF INCOME AND EXPENSES, 1951</th>
</tr>
</thead>
<tbody>
<tr>
<td>Income</td>
</tr>
<tr>
<td>Advertising $345,418.45</td>
</tr>
<tr>
<td>Member Dues and Conventions 495,850.95</td>
</tr>
<tr>
<td>Subscriptions 51,695.11</td>
</tr>
<tr>
<td>Sales Items, Binders, Emblems, etc. 20,463.11</td>
</tr>
<tr>
<td>Surplus Income 13,260.88</td>
</tr>
<tr>
<td>Miscellaneous Income 683.36</td>
</tr>
<tr>
<td><strong>TOTAL INCOME</strong> $921,371.88</td>
</tr>
</tbody>
</table>

| Expenses |
| Proceedings Editorial Pages $190,257.11 |
| Advertising Pages 188,025.57 |
| Directory 67,240.06 |
| Section and Student Branch Reasses 33,643.94 |
| Sales Items 13,006.97 |
| Miscellaneous Printing 2,185.37 |
| General Operations Convention Cost 127,080.62 |
| **TOTAL EXPENSES** $822,224.00 |

| Surplus Reserve for Depreciation 6,925.89 |
| **NET SURPLUS** $98,221.90 |

<table>
<thead>
<tr>
<th>TABLE III.—BALANCE SHEET—DECEMBER 31, 1951</th>
</tr>
</thead>
<tbody>
<tr>
<td>Assets</td>
</tr>
<tr>
<td>Cash and Accounts Receivable $276,312.00</td>
</tr>
<tr>
<td>Inventory 13,552.33</td>
</tr>
<tr>
<td>TOTAL CURRENT ASSETS $289,864.32</td>
</tr>
<tr>
<td>Investment in Cost of Building and Land at Cost 601,013.42</td>
</tr>
<tr>
<td>Furniture and Fixtures at Cost 85,803.60</td>
</tr>
<tr>
<td>Other Assets 19,139.20</td>
</tr>
<tr>
<td><strong>TOTAL ASSETS</strong> $1,502,643.52</td>
</tr>
<tr>
<td>Liabilities and Surplus</td>
</tr>
<tr>
<td>Accounts Payable $18,718.46</td>
</tr>
<tr>
<td>Federal Taxes on Emblems, etc. 91.76</td>
</tr>
<tr>
<td><strong>TOTAL CURRENT LIABILITIES</strong> $19,810.22</td>
</tr>
<tr>
<td>Deferred Liabilities 315,896.42</td>
</tr>
<tr>
<td><strong>TOTAL LIABILITIES</strong> $334,706.64</td>
</tr>
<tr>
<td>Surplus—Donated 595,406.44</td>
</tr>
<tr>
<td>Surplus—Earned 390,846.53</td>
</tr>
<tr>
<td><strong>TOTAL SURPLUSES</strong> 980,253.96</td>
</tr>
<tr>
<td><strong>TOTAL LIABILITIES AND SURPLUSES</strong> $1,535,308.14</td>
</tr>
</tbody>
</table>

* 1952 items, PROCEEDINGS for members and subscribers, Advertising, and Convention Service.

<table>
<thead>
<tr>
<th>TABLE IV.—VOLUME OF PROCEEDINGS PAGES</th>
</tr>
</thead>
<tbody>
<tr>
<td>1951</td>
</tr>
<tr>
<td>------------------------------------------------</td>
</tr>
<tr>
<td>Editorial 1,628</td>
</tr>
<tr>
<td>Advertising 1,424</td>
</tr>
<tr>
<td><strong>Total</strong> 3,052</td>
</tr>
</tbody>
</table>

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 remaining being under review. This represenoted 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 stimulate and 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 deposed (with the author’s consent) with the American Documentation Institute at 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 cooperate closely with IRE Professional Groups in 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 X. Goldsmith in matters of editorial policy, content, and format, and by Executive Secretary George W. Bailey in matters of finance and administration. Both functions 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 Com-

June
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 Abbreviations of Radio-Electronic Terms, 1951.
- Standards on Electroacoustics: Definitions of Terms, 1951.

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 Joint IRE/IEE/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 (JICC). 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 Electrophysics, 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 a symposium. 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 12. It was well attended and enthusiastically received.


Joint Technical Advisory Committee. Six meetings of JTAC were held during 1951. Volume VI, Proceedings of the JTAC, was published. This volume included: Section 1, Admissions Committee; Section II, Joint Technical Advisory Committee (IRE-RTMA) with Other Items Pertinent to the Action of the JTAC; and Section III, Approved Minutes of Meetings of the Joint Technical Advisory Committee (IRE-RTMA) both covering the period July 1, 1950-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 Y. L. Hogan as Chairman and Ralph Bow in was elected unanimously as Vice Chairman. On July 1, 1951, Haraden Pratt resigned from membership on JTAC to become Telecommunications Advisor to the President of the United States. A. V. Loughlin 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 coordinating 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
- Tulsa (May), 1951
- Phoenix (January), 1951

The total number of Sections is now 60. There has been an increase in the membership of 25 Sections.

The Subsections of Sections may 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 1, 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 production and research division of Dynamic Electronics-New York, Inc., it was recently announced.

Dr. Bennett was born on June 13, 1917, in New York, N. Y. He received the B.S.E. degree and the M.S.E.E. degree from the College of the City of New York, in 1938, and 1949, 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'S1) has been named president of the recently expanded Joseph Racker Company, a firm of radar consultants and editors.

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. At present, Mr. Racker is working on his doctorate degree.

Herbert A. Elion (A'S2) 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.

Mr. Elion was born on October 16, 1923, in New York, N. Y. He received the B.M.E. degree from the College of the City of New York in 1944, and the M.S. degree from the Brooklyn Polytechnic Institute in 1949. At present, Mr. Elion is working on his doctorate degree.

Paul F. Shuey (A'47-M'46) died recently at his home in Gahion, Ohio. He was 70 years of age.

A native of New Providence, Pa., Mr. Shuey was affiliated with the Northern Electric Manufacturing Company for the past 11 years. He was a member of the American Institute of Electrical Engineers and the Society of Professional Engineers. He was a Fellow at Mellon Institute, Pittsburgh, Pa.
Knox McIlwain (A’31-M’40-SM’43-F’48) was named RTMA representative to the Radio Technical Commission for Aeronautics, recently, by W. R. G. Baker, director of the RTMA Engineering Department. Mr. McIlwain is chief consulting engineer of Hazeline Electronics Corporation.

Robert W. Hellwarth (S’52) has been chosen valedictorian of Princeton University’s graduation exercises in June. He also was recently selected to receive the 1952 RE Student Branch award at Princeton.

The instruction given is of use only to technicians concerned with specific equipment for which the book was written. Without availability of the equipment described, the instruction given is of little value. Other parts of the book are very readable and give useful information on broadcast subjects.

Some of the chapters contain information apparently not related to the rest of the subject matter. This prevents ready reference to these items until the location of all the material in the book is well known. Also, some material appears repetitious as the same subjects are discussed in several places.

A short bibliography is included which will be of interest to the inquiring broadcast technician.

Books


Published (1951) by John F. Rider Publisher, Inc. 408 Canal St., New York 13, N. Y. 434 pages-64-page index-4-page bibliography-225 figures. $5.40.

Harold E. Ennes is a staff engineer of station WIRE and technical director of station WJJC, Indianapolis, Ind.

This book is valuable to the inexperienced broadcast technician. The material is drawn from the author’s experience manufacturers’ instruction books, and publications of the FCC.
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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."
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Mr. McIlwain, who is a native of Philadelphia, Pa., received the B.S. degree from Princeton University in 1918, and the B.S.E.E. and E.E. degree from the University of Pennsylvania in 1921 and 1930, respectively. In 1948, Mr. McIlwain was awarded a Certificate of Commendation by the United States Navy for his achievements during World War II, and the IRE Fellow Award for his "contribution to the technical literature of radio and his activity in the field of radio aids to navigation."

Mr. McIlwain has served on IRE Committees such as Admissions, Board of Editors, Circuits, Education, Standards, and others. In 1935-1936, he was the Chairman of the IRE Philadelphia Section.

Ernst Weber (M'41-SM'43-F'51), formerly with the Office of Scientific Research and Development, has been elected president of Polytechnic Research and Development Co.

Dr. Weber was born in Vienna, Austria, in 1901, and received his education there which included the Sc.D. and Ph.D. degrees. In 1930, he came to the United States as a visiting professor at the Brooklyn Polytechnic Institute where he has remained since that time. In 1945, he was named head of the department of electrical engineering and director of the Microwave Research Institute.

Dr. Weber has served on numerous IRE Committees and is a member of the IRE Board of Directors.

August Hund (M'16-F'27-LM'52), noted electrical engineering consultant and author of numerous papers and technical books in various fields of radio, died recently at his home in Santa Monica, Calif.

Dr. Hund was born at Offenburg, Baden, Germany, in 1887, and received the E.E. degree in 1911, and the D.Eng. degree in 1913, at Carlsruhe, Germany. He later did some graduate work at the University of California. Arriving in the United States in 1912, Dr. Hund was engaged as a research engineer for the General Electric Company until 1914, and from 1915-1917, he was a member of the staff of the University of Southern California, where he taught electrical engineering and physics. He became a consultant research engineer from 1918-1922, and then served with the National Bureau of Standards, from 1922-1929, as senior physicist. From 1929-1934, Dr. Hund was a special research and consultant engineer for the Wired Radio Inc., in New Jersey, and then returned to a practice in California, as a consultant engineer. He was a technical consultant for the United States Naval Radio and Sound Laboratory, in San Diego from 1942-1945, and later returned, although continuing his work on technical books.

Dr. Hund had been granted 45 patents in the fields of piezo-electro oscillators, modulators, amplifiers, and frequency control. His last book, "Short-Wave Radiation Phenomena," was published recently.

Dr. Hund was the acting Chairman of the IRE San Diego Section in 1945, and served on the IRE Committee of Standardization Subcommittee on Electro-Acoustic Devices. He was a member of the American Physical Society, the American Institute of Electrical Engineers, and the International Union of Radio.

Robert W. Hellworth (S'52) has been chosen valedictorian of Princeton University's graduation exercises in June. He also was recently selected to receive the 1952 IRE Student Branch Award at Princeton.

Books


Published (1951) by John F. Rider Publisher, Inc., 480 Canal St., New York 13, N. Y. 434 pages (+1 page index +1-page bibliography + viii pages. 35 x 23. $5.40

Harold E. Ennes is a staff engineer of station WRE, and was former director of station WJJC, Indianapolis, Ind.

This book is valuable to the inexperienced broadcast technician. The material is drawn from the author's experience manufacturers' instruction books, and publications of the FCC.

The instruction given is of use only to technicians concerned with specific equipment for which the book was written. Without availability of the equipment described, the instruction given is of little value. Other parts of the book are very readable and give useful information on broadcast subjects.

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Orrin W. Towner WJAS, Inc. Louisville 2, Ky.
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NTSC. In addition to IRE committee meetings, the National Television Systems 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 35 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, 68 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.

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 1929-1942, 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 research 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.

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 project engineer at Freed Radio Corporation and the M. W. Kellogg Company.

Mr. Elion was born on October 16, 1924, in New York, N.Y. He received the B.M.E. degree from the College of the City of New York in 1944, and the M.S. degree from the Brooklyn Polytechnic Institute in 1949. At present, Mr. Elion is working on his doctorate degree.

Paul F. Shuey (A’37-M’46) died recently at his home in Gahon, Ohio. He was 70 years of age.

A native of New Providence, Pa., Mr. Shuey was affiliated with the Northern Electric Manufacturing Company for the past 11 years. He was a member of the American Institute of Electrical Engineers and the Society of Professional Engineers. He was a Fellow at Mellon Institute, Pittsburgh, Pa.
August Hund (M’16–F’27–LM’52), noted electrical engineer consultant and author of numerous papers and technical books in various fields of radio, died recently at his home in Santa Monica, Calif.

Dr. Hund was born at Offenburg, Baden, Germany, in 1887, and received the E.E. degree in 1911, and the D.Eng. degree in 1913, at Carlsruhe, Germany. He later did some graduate work at the University of California. Arriving in the United States in 1912, Dr. Hund was engaged as a research engineer for the General Electric Company until 1914, and from 1915–1917, he was a member of the staff of the University of Southern California, where he taught electrical engineering and physics. He became a consultant research engineer from 1918–1922, and then served with the National Bureau of Standards, from 1922–1929, as senior physicist. From 1929–1934, Dr. Hund was a special research and consultant engineer for the Wired Radio Inc., in New Jersey, and then returned to a practice in California, as a consultant engineer. He was a technical consultant for the United States Naval Radio and Sound Laboratory, in San Diego from 1942–1945, and later retired, although continuing his work on technical books.

Dr. Hund had been granted 45 patents in the fields of piezo-electric oscillators, modulators, amplifiers, and frequency control. His last book, “Short-Wave Radiation Phenomena,” was published recently.

Dr. Hund was the acting Chairman of the IRE San Diego Section in 1945, and served on the IRE Committee of Standardization Subcommittee on Electro-Acoustic Devices. He was a member of the American Physical Society, the American Institute of Electrical Engineers, and the International Union of Radio.

Robert W. Hellwarth (S’52) has been chosen valedictorian of Princeton University’s graduation exercises in June. He also was recently selected to receive the 1952 IRE Student Branch award at Princeton.

Knox McIlwain (A’31–M’40–SM’43–F’48) was named RTMA representative to the Radio Technical Commission for Aeronautics, recently, by W. R. G. Baker, director of the RTMA Engineering Department. Mr. McIlwain is chief consulting engineer of Hazeline Electronics Corporation.

Mr. McIlwain, who is a native of Philadelphia, Pa., received the B.S. degree from Princeton University in 1918, and the B.S.E.E. and E.E. degree from the University of Pennsylvania in 1921 and 1930, respectively. In 1948, Mr. McIlwain was awarded a Certificate of Commendation by the United States Navy for his achievements during World War II, and the IRE Fellow Award for his “contribution to the technical literature of radio and his activity in the field of radio aids to navigation.”

Mr. McIlwain has served on IRE Committees such as Admissions, Board of Editors, Circuits, Education, Standards, and others. In 1935–1936, he was the Chairman of the IRE Philadelphia Section.

Ernst Weber (N’41–SM’42–F’51), formerly with the Office of Scientific Research and Development, has been elected president of Polytechnic Research and Development Co.

Dr. Weber was born in Vienna, Austria, in 1901, and received his education there which included the Sc.D. and Ph.D. degrees. In 1930, he came to the United States as a visiting professor at the Brooklyn Polytechnic Institute where he has remained since that time. In 1945, he was named head of the department of electrical engineering and director of the Microwave Research Institute.

Dr. Weber has served on numerous IRE Committees and is a member of the IRE Board of Directors.

Louis G. Pacent (A’12–M’15–F’27), pioneer engineer, president of the Pacent Engineering Corporation, died recently in New York, N. Y. He was 58 years old.

Mr. Pacent started experimenting with wireless transmission in 1906, and had his own amateur station in 1909. At 16, he joined the Naval Militia as a communicator in 1913, and served aboard the U.S.S. Gloucester until 1917.

During World War I, Mr. Pacent worked on the development of communications equipment for the Army and Navy. Later he formed his own business, the Pacent Electric Company, Inc., which designed and produced electrical and radio apparatus for General Electric, Westinghouse, Western Electric, RCA, and the Federal government. In 1921, the first short-wave transatlantic message was sent and received in Scotland, on the 200-meter transmission, as Mr. Pacent had formerly suggested.

In the 1920’s and 1930’s, Mr. Pacent increased his research interests to the new talking motion picture field. He designed the first all-power-operated talking-motion-picture sound equipment, while a consultant for Warner Brothers Pictures, which was installed in 1928. In 1933 he formed the engineering company which he headed at his death.

Mr. Pacent received the Certificate of Appreciation award from the War Department in 1946 in recognition of his services to the department during World War II. In 1951 he received a Marconi Memorial Medal of Achievement.

Mr. Pacent served on various IRE Committees, and was a fellow of the Society of Motion Picture Engineers, a member of the American Institute of Electrical Engineers, and the president of the Radio Club of America.

Books


Published (1951) by John F. Rider Publisher, Inc., 460 Canal St., New York 13, N. Y. 434 pages + 1 page index + 1-page bibliography + 6 pages + 27 figures. $3.00. $3.50.

Harold E. Ennes is a staff engineer of station WIFE, and the technical director of station WAFJ, Indianapolis, Ind.

This book is valuable to the inexperienced broadcast technician. The material is drawn from the author’s experience manufacturers’ instruction books, and publications of the FCC.

The instruction given is of use only to technicians concerned with specific equipment for which the book was written. Without availability of the equipment described, the instruction given is of little value. Other parts of the book are very readable and give useful information on broadcast subjects.

It is unfortunate that this edition does not include information made available within the last seven years on studio design, new microphones, and the CAA requirements; however, other subjects discussed are satisfactorily up to date.

Some of the chapters contain information apparently not related to the rest of the subject matter. This prevents ready reference to these items until the location of all the material in the book is well known. Also, some material appears repetitions as the same subjects are discussed in several places. A short bibliography is included which will be of interest to the inquiring broadcast technician.

ORNIE W. TOWNER WIIAS, Inc. Louisville 2, Ky.
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 "wire-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 practical 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 between the 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.—M. Nicot, The Pennsylvania State College (On leave from Royal Meteorological Institute of Belgium, Brussels, Belgium).—Recent physical data concerning the solar radiation 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. Diffuse 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 fathomed the entire range of frequencies. 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 effect 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.—W. Findley, 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, h') 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 h, whereas the maximum electron density Ne is not simply related to h. The well-known anomalies when Ne 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 Ne. 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 of 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 fading 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 development 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.—W. A. 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 the microwave region and the sensitivity is approximately 10 μw. 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 barrier
is mounted. One terminal is used only as a reference point from which the temperature rise of the other is measured. A 20-junction, copper-to-constantan thermople 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-substituting calibrator 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±1 percent.

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 multihem 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 about 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. Greisenkemper 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 due to the metal 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 10⁴ to 10⁸ 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 kc and 24 kc 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 of the signal can be 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. Hime, 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 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 is the 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 fading of cw emissions shows that observed characteristics of these fading distributions are functions of the relative amplitudes of the received "e" and "z" 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. Feinberg, 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. Mallinkrodt, W. Snyder, R. A. Hellwell, 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 established that the effect of 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 IONOSPHERIC 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, pulse 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 1949.

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 of the system of "generalized transmission-line" equations is then solved by an extension of the Peano-Baker matrixing method. The solution is a uniformly convergent series which is interpreted as four "waves" which undergo continuous multiple reflection. All orders, each wave contributing to all of the others. While this 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 slot, and the associated resistance 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 antennas.

* This work sponsored by the Office of Naval Research and the Bureau of Ships under contract N7on529.

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 gain 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 symmetry 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 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.—E. W. Moore, J. Moore, and E. T. Kuck, U.S. Navy Bureau of Aeronautics.—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 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 (658).7550.

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 index of refraction (r) of the atmosphere and for determining the refractive scale of refractive 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 r, 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 120-foot tower at the Brookhaven National Laboratory, L. I., N. Y. It has been found from a preliminary data analysis that r variations were caused mostly by variations in water-vapor density. The r data at 410 feet show a wide range of different components, corresponding to wavelengths of roughly 10 to 1,000 meters.

* Now with the Johns Hopkins University Computer Laboratory, 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 fluctuations over the Pacific Ocean and the refractive-index variability 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. Sides of the equipment installation in an aircraft are shown.

* This work was sponsored by Office of Naval Research under Contract No. N60-1-P-01.

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 tropospheric refractive index; the field propagates smoothly and continuously with height. A simple estimate of such reflection from the normal troposphere gives a 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 reflection 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 whole atmosphere or a unit index above 30,000 feet, or an exponential model which falls asymptotically to unity at great heights, gives more accurate results. The contributions of these partial reflections in the propagation of electromagnetic waves in the visible spectrum are considered.

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 closely to the partial reflection. This method 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 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 scattering antenna height influence. 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 theory. 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.—I. J. Anderson and J. F. Colwell, Navy Electronics Laboratory.—The phenomenon of reflection of plane waves from a continuously varying medium of indices of refraction has been investigated from a theoretical standpoint for some time. Recently, it has been suggested that, depending on the position and type of receiving and transmitting antenna, such reflections may account for the large field strength observed beyond the optical horizon. 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.

26. ATMOSPHERIC NOISE AT VERY LOW FREQUENCIES.—John 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 vlf waveform is flat from 180 to 20,000 cycles, falling off at lower and higher frequencies. The power spectrum of the day-time week 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 non-typical flash. Our current work is toward an improved flash analysis consistent with the phenomenological 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 atmospheric noise power usually lies at 6 to 15 kc. Kosak 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 CORRELATION TO ATMOSPHERIC RADIO NOISE PREDICTION.—Edna Shults, 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. Nicholien, The University of Pennsylvania.—When using noise meters for measuring electromagnetic noise, 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
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-nc 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-nc radiation. With but a few 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 very frequently on 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 simultaneously. Such delayed events increase in intensity as the flare fades and often attain greatest intensity when the flare has faded completely.

3. SPACE-CHARGE WAVES IN MAGNETICALLY FOCUSED BEAMS. M. Chodorow and L. Ziteelli, 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 character 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 common types 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 surf ace, 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 surface layers and two thirds in the volume 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 its 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 onedimensional 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 is 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 has suggested that it is possible to account for the effect of 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 is equal to the temperature of 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 with a large amount of numerical labor. There are, however, numerical differences which may be attributed to lack of constancy of the electron temperature.

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 due to field distortion. 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 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 circuit. 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. Arvid and J. Eleftheriou, 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 transmission line. The authors have studied conditions and broadening of the transducer are studied. This method gives very easily the insertion losses of the transducer and the losses of the line. The scattering matrix coefficients of various types of susceptances on 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 Lobachevsky geometry illustrate this point. They form a normal 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 axioms 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 geometrical 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. Queathem and A. Haehnel, 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 necessary to use a selective spectrum generator, for example, a non-linear 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 restricted band of the entire spectrum is used. In such a case it would be desirable to generate a restricted band of frequencies and to 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 the same as 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 this variation is that it is shown how the different design parameters influence the envelope of the frequency spectrum. Experiments conducted on such a circuit show agreement within 0.3% 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. Shmos, 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, except for a change 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 normally at the level where \( \omega_n = \omega \), i.e., there is no net energy flow. 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.—K. T. Chang and E. H. Liu, Pennsylvania State College.—This paper is...
By the method of Briggs-Phillips (1950), the "angular spreading" of the downcom ing 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 diffraction screen at long wave-lengths.

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 these speeds obtained by the Booker- Ratecliffe-Shinn (1950) autocorrelation method and Ratecliffe's (1948) speed of fading method.

A study of the facing rate of the first reflected echo indicates the presence of fades of 1/4 to 15 minutes duration with an average facing rate of 1.3 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 KCS. \( \text{R. E. Jones, G. H. Millman, and R. J. Neury.} \)

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 cloud which produces 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 2X10\(^{-8}\) to 8X10\(^{-8}\) sec\(^{-1}\). This would correspond to a height range of about 270 km for our ionospheric model. These results appear to be in good agreement with recent work by Kellogg (1951) from the meteorological viewpoint.

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44. SPORADIC-E STRATIFICATION AND CORRELATION WITH LOW-FREQUENCY SOUNDINGS. \( \text{R. A. Hill, all, Stanford University.} \)

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 frequencies on the month's data 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 picture, the height of the slower trace (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 correlation between the two frequencies, 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 varied 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. \( \text{Marcel J. E. Golay, Signal Corps Engineering Laboratories.} \)

A criticism is made of the emphasis recently placed by communication engineers on correlation concepts and devices. Specific suggestions, illustrated by an example borrowed from radiolocation, 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. \( \text{David A. Huffman, Massachusetts Institute of Technology.} \)

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. \( \text{William B. Wrigley, Georgia Institute of Technology and Air Force Cambridge Research Center.} \)

Since poles of the complex impedance of a linear system represent the decrments and frequencies of rotating modes in the linear time domain, it is suggested that a nonlinear system may 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 oscillating generators of the second order which is described by the differential equation, \( x - N(X, x) = 0 \).
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 input intelligence is imparted to the pulses. 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.

50. RANDOM PROCESSES IN WAVE PROPAGATION.—W. S. Amyx, 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 scattering formula. This is shown by two new derivations of the propagation constant of a "bloppy" 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 Lévy 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 between trends of 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 a suitable portion of the terrain between the transmitting and receiving sites. The mean-square residual height variation from such a polynomial provides a measure, through Bayley's criterion, of the "visibility" of the terrain, 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 terrain. 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 of 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. Kirkby, 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 directional 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, 289 mc, and 460 mc with both horizontally and 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 \( A = 790 K/m^2 \) and \( B = 8400 K^2 \) in the expression for the refractive index of moist air \( n = 1 - \frac{A}{T} + \frac{B}{T^2} \) 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 this is not the reflection 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 and how far it is worth while to specify 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 Hitesfield, 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 of 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 scattering from spherical ice on the attenuation 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 INO-SHORE FIELD STRENGTH RECORDING FOR LOW FREQUENCIES.—J. C. Blair, J. N. Brown,
and J. M. Watts, National Bureau of Standards.—Application of the techniques here-tofore used in high-frequency ionosphere recorders is described. The heat-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. C. Willard, 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 increases enormously 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 (4) 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 mud for either F-layer or sporadic-E transmission. It is especially useful 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 t circuit is described. Antenna rotation and ppi display of scatter echoes shows a glance to the areas 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 range of ionospheric conditions by 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 used with normal 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 records, determines the equivalent height of reflection for any backscattered echo information. 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, occurred 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 1DT, 1CW for the eclipse observations. Locations at Charlottesville, Virginia; Derwood, Maryland; and Chincoteague, Virginia established a west-east path 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 detected.

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. Lurie, 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 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 the plane wave impinging on such a medium is subject to various angular conditions 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 variational methods in 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 numbers 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 signal 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 in transmission overcoast to coast fading is also 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.

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. Cribb, National Bureau of Standards, Ralph Showers, University of Pennsylvania, and others.—In measuring atmospheric noise one is generally interested in obtaining 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 protection 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 factors 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 obtaining the statistics 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. Pomerene, 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 a considerable difference 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 IONOSPHERIC 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. Observations 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, while the others showed marked differences from the monthly median. In general, the deviation from the monthly median was carried over a considerable distance 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° 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 Inverness, 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. The 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 display was to the North, the greater 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 a combination of the incoming auroral protons. The scattering of each column may be treated in the same manner as the "whistles" from meteorionization, 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 distance traveled by each echo 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 column size is not known, but indications are that this size may possibly be of the order of 1 meters in diameter.

66. THE LENGTH OF IONIZED MEETEOR TRAILS.—L. A. Manning, O. G. Villard, Jr., and A. M. Peterson, Stanford University.—The work was done with the aid of a coincidence counter and a coincidence recording oscilloscope. For instrumentation 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 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 describing the individual modal behavior. In relatively simple problems, involving scattering or arbitrarily oriented dipole antenna in a stratified region, the transmission-line problem is solved readily since there is no coupling between modes. In general differential problems in which mode 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 contributions of individual 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 spherical stratified region (the general plane and spherical earth problem) and (b) the diffraction of a plane wave by two obliquely disposed unbounded semi-infinite plane parallel plates. 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 worked out, and experiments have been made, to determine the effect of these slots on the transmission of the wave with respect to frequency and phase, and some experimental checks have been obtained. The calculated patterns show that in the semi-circular arc 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 semi-circle over which the slot is optically invisible and particularly near the endpoint of the slot the field is very well represented by $E_r \cos (\rho - \phi)$ where $E_r$ is the value of the pattern at $\phi = \pi$ (the point opposite the slot) and $\rho$ 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 $\sum \frac{A_n}{\rho} \cos n \phi (\rho - \delta)$, where $A_n$ is complex. Near $\rho = \infty$, 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 AP 19 (122)-454.
finding the envelope of the excitation coefficients—and hence the coefficients themselves—are outlined.

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 π/2 in the case of a focal point, but which may differ from π/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.

71. INVESTIGATION OF A SURFACE-WAVE LINE FOR LONG-DISTANCE COMMUNICATION.—G. Goubau, C. Sharp, and S. W. Atwood, 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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ACOUSTICS AND AUDIO FREQUENCIES

534.231-1:621.396.671
M. 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. Tokohu Univ., vol. 14, pp. 79-93; 1956.) 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

534.321.9:534.232
1182 Ultrasonic Generators and their Applications—N. Azumii. (Radio Tech. Dig. (France), vol. 5, pp. 271-278 and 290-325; 1951.) A survey paper with 165 references.

534.321.9:538.052

534.341:534.386.11
1190 Influence of Chandeliers on the Acoustics of Theatres and Concert Halls—L. Villard. (Ball. Tech. suisse romande, vol. 77, pp. 243-255; September, 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 non-diffuse sound absorption and also act as sound diffusers; their replacement by modern lighting fittings may have an adverse effect.

534.351:621.305.813
1191 A New Evaluation of the Transmission Quality of a Telephone System—G. Fontanella. (Tech. Mitt. Schweiz. Telegr.-Teleph. Verw., vol. 29, pp. 384-399; 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.

534.355.625.3

534.355.92.001.4
1193 Testing of Components of Hearing Aids—F. Müller. (Punk 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

534.613.018.781:621.315.321
1194 Distortion of a Signal Transmitted by a Perfectly Homogeneous Coaxial Line—K. Czarnage. (Cables & Trans. (Paris), vol. 5, pp. 271-283; October 1, 1951.) For frequencies up to about 10 kc the telegraphy equation represents coaxial-line propagation sufficiently closely, provided the coefficients are treated as functions of frequency. By considering separately the high and low 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
Abstracts and References

1952

1821.306.70: 621.307.6

1209

Television in Buenos Aires—(Rev. elec. Electron. Buenos Aires, no. 408, pp. 62-63, 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 the building, each made up of three horizontal folded dipoles.

1952.306.67: 621.307.6

1210

Fourier Analysis and Negative Frequencies—(Shaw. (See 1422.)

1952.306.677

1211


1952.306.677.5

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

1952.310.157: 621.387.14

1214


1952.310.187: 621.392.512

1215

High-Power F.M. Antenna Design—M. B. Sleeper. (F.M.-TV, vol. 11, pp. 11-12, October, 1951.) Description of the modified doughnut-type antenna and feeder system at station WMJ, on Cloistoga, Pea. N.C. Adjustable directive pattern. Simple array theory is developed for prediction of the main features of the radiation patterns.

621.306.67: 621.392.43

1206

Automatic Impedance Matcher—(True. (See 1224.)

621.306.671: 621.306.019.13

1207

High-Power F.M. Antenna Design—M. B. Sleeper. (F.M.-TV, vol. 11, pp. 11-12, October, 1951.) Description of the modified doughnut-type antenna and feeder system at station WMJ, on Cloistoga, Pea. N.C. Adjustable directive pattern. Simple array theory is developed for prediction of the main features of the radiation patterns.

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High-Power F.M. Antenna Design—M. B. Sleeper. (F.M.-TV, vol. 11, pp. 11-12, October, 1951.) Description of the modified doughnut-type antenna and feeder system at station WMJ, on Cloistoga, Pea. N.C. Adjustable directive pattern. Simple array theory is developed for prediction of the main features of the radiation patterns.
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 maximum practical independence of the frequency swing of the plant, in this design. They are based on analysis of samples of ordinary-grade transformer iron and matal, and simplify the determination of all necessary design parameters.

621.314:2.012.3  
2126 Transformer Iron Losses—N. II. Crowhurst. (Electron. Eng., vol. 23, pp. 396–403; October, 1951.) Four charts are presented for application of transformer design. They are based on the use of samples of ordinary-grade transformer iron and matal, and simplify the determination of all necessary design parameters.

621.314:6.011.1  
1227 Theory of the Linear R.M.S.—Value Rectifiers—O. Schmid. (Arch. Elekt. Ubertragung, 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/peak-voltage is found by a graphical method. Curves are plotted from which the best time-constants ratio and the waveform errors of first and second type can be determined. See also 39 of March.

621.314:7:540.289  
1218 Transistors: Circuit Design—G. Kuoisek. (Electronics, vol. 24, pp. 128-132, 134; December, 1951.) An explanation of the operation of duality technique to the design of transistor amplifier, oscillator, modulator and multi-vibrators circuits, starting from the corresponding tube circuits.

621.310:80  
1219 Nonlinear Resistors with Sintered-Semiconductor Base—N. Guyen Thien-Chi and J. Suechel. (Ann. Radiol., vol. 6. pp. 291–298; October, 1951.) The resistors are made of a material with a sintered-carbournium base, having the same property of resistance and a voltage/current relation of the form \( V = kI^x \). Two types are available, type D (0.8-w rating) for operation at a few tens of milliamperes, and type II (5-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  

621.318:572:621.3:85.8:001.8  
1221 Electron-Beam Switches—F. Schütz. (Telefunken Z., pp. 171–186; October, 1951.) A well illustrated review of various types and their applications.

621.318:45  
1222 Capacitors with Ceramic Dielectric: Performance and Operating Characteristics—A. Dazin. (Ondr. Elekt., vol. 31, pp. 342–356 and 357–362; 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 very low dielectric-constant; (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. Poston. (Elect. Rev., vol. 54, pp. 30–38; October, 1951.) Description of electrolytic capacitors using tantalyum-foil electrodes. A capacitance of approximately 0.01 mmf/m² has been obtained. The Ta₂O₅ insulating film can operate at an electrical stress of over 19,000 v/m². Tantalytic capacitors are considerably smaller than corresponding ones of other types.

621.302:43:621.3:68.67  
1224 Automatic Impedance Matcher—V. True. (Electronics, vol. 24, pp. 98–102; December, 1951.) Description of equipment for matching 35-foot whip antenna to a 50-ohm coaxial feeder over the frequency range 2–18 mhz. Matching is performed by a ballener circuit in which the capacitance of the matching circuit is adjusted by a servomechanism controlled by a circuit which determines the phase angle between feeder current and voltage: the capacitance of the circuit is controlled by a circuit which measures the total load impedance of the feeder. The use in no case exceeds 1.25. With minor adjustments the equipment can be adapted for different frequencies, power levels, types of load, and feeder characteristic impedances.

621.392.5  
1225 General Input-Output Relations for Linear Networks—L. A. Zadel. (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 dissipated power is divided into 'the content' and 'co-content'—which are due of each other. The dissipation itself has stationary properties in linear but not in (general) in nonlinear network. It is 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 cost 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 concept of a 'system' (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 'rectangular representation' of a circuit of linear elements is explained and extended to the nonlinear case, including reactive elements. It is shown that the equations of motion of a nonlinear system, possessing resistance, may be expressed in Lagrangian form, thus emphasizing the importance of energy and also showing that the Principle of Duality is applicable. Finally, systems are considered to move among moving magnetic circuits (as in rotating machines)." See also 1226 above.

621.392.5  
1228 Conditions of Validity of Matrix Analysis for Quadrupole Assemblies: Applications to Feed-Forward Networks—A. Kaufman. (Ondr. Elekt., vol. 31, pp. 496–405 and 446–452; October and November, 1951.) The basic method of interconnecting two quadrupole so that they may be represented by a single matrix, and the representation of any 4-terminal network so that matrix elements for a matrix of order 4 are outlined. Impedance relations are derived which must be satisfied for the matrix calculation to be valid in the case of (a) parallel-pauld; (b) series-pauld, and (c) parallel-series combination of input and output of two quadrupole. Particular examples of each case are noted. Different networks and their corresponding relations are given for each matrix of order 4. 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"—H. J. 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, relative to the time-dependent and frequency-dependent parts of the system.

621.392.5:517.755  

621.392.5:546.431.824—621.392.5:519.241.1  
1231 Particulars of Substitution of Tatarian Delay Lines—L. Al Orman and L. G. Callouï. (Electronics, vol. 24, pp. 224, 248; December, 1951.) Mechanical vibration waves may be excited in BaTiO₃ piezoelectric material by the application of an electric field, provided that the material between the electrodes has 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. Ehlrich. (Toule la Radio, no. 10, p. 2; June, 1951.) The flash bulb is operated by a current energized from a cathode-coupled double-triode 'flip-flop' circuit, the time constant of which is adjustable by a potentiometer with scale calibrated in seconds or milliseconds.

621.392.5:509.046  
1233 The Ferromagnetic Faraday Effect at Microwave Frequencies and its Applications—The Microwave Gyror—C. L. Hogan. (Bell Syst. Tech. Jour., vol. 31, pp. 1–31; January, 1952.) A new type of gyror [590 of 1951 (Trellegen and Kraus)] dependent on the Faraday rotation of the plane of polarization of an em wave has been developed. An extraordinary wave propagating in a ferromagnetic material with dielectric and magnetic loss gives a formula for the Faraday rotation which indicates that 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. Gyror 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.

621.392.5:621.3:14.2  
Abstracts and References

I line. An attenuator using similar principles to that of the actual load resistance represents the transformation ratio of the general quadrupole and nearly corresponds to the mutual inductance of a transformer on open circuit. The ratio of output to input current being 1/π, but the phase difference between output and input is about 90°. Doherty's modulation circuit makes use of the transformer properties of choke filters.

62.139.52: 621.390.611.21 1235
The Maximum Bandwidth of Narrow-Band Quartz-Lattice Filters—W. Rave. (Arch. elektr. Ubertragung, vol. 5, pp. 455-554; 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 calculated bandwidth 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.

62.139.52: 621.390.612.3 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 approximate equations, are presented from which the required data can be read off to within about 1 per cent.

62.139.52: 621.390.612.4 1236
U.F. Oscillator Attenuator—F. Regaglia. (F.T.T.V., 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 ($37$ of $60$).

621.390.645: 621.390.97 1238
The Broadcasting-Network Amplifier-R. Pavl, H. v. Schau and W. Schwenk. (Fernmeldetechn., vol. 4, pp. 452-457; October, 1951.) Description of the high-fidelity line equipment for program transmission on four incoming or outgoing lines.

621.390.611.21 1239
Quartz Crystal Vibrators as Circuit Elements—H. E. Pearson (P.O. Elect. Eng. Jour., vol. 44, pp. 922-926; 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.

621.390.611.3 1240

621.390.611.3: 621.301.21 1241

621.390.611.4 1242
Representation of the Complete System of Natural Oscillations of Cylindrical Cavities Resonators with Horizontally Stratified Dielectrics—Legiedeg and P. Urban. (Acta Phys. austriae, vol. 3, pp. 320-341; March, 1950.) Results previously obtained by Legiedeg (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.

621.390.611.4: 538.506 1243

621.390.615.17 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 1000 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.

621.390.645 1245

621.390.645 1246
New Miniature Intermediate-Frequency Amplifiers—R. Tolleck. (IEE, vol. 35, pp. 143-145; October, 1951.) Description of the National Bureau of Standards Model VI, a 7-tube amplifier for the frequency range 200-1000 kcps, constructed by providing separate subassemblies for (a) all the inductors, (b) all the capacitors and (c) all the tube shields.

621.390.645 1247
The Series Amplifier—E. L. Crosby, Jr. (Radio and Telecomm. 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 each 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.

621.390.645: 018.424: 621.377.92 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 Thoma (109 of 1951). It is located at the antenna, away from the rest of the apparatus, and is modified by inter-modulation between received signals is reduced by restricting the response to the required frequency band of 2.5-20 mcps by careful design of the signal path. The signal gain is made 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.

621.390.645: 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 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. This theory is discussed and illustrated by an example giving component values and measured performance.

621.390.645: 029.3 1250

621.390.645: 35 1251

621.390.622 1253
Noise, Resistance Fluctuations and the Flicker Effect—M. Surdin. (Physica, "Grav., vol. 17, pp. 538-550; May, 1951.) In French.) Published experimental results on resistance fluctuations and flicker published in 1947 (3035 of 1950) that in oxide cathodes the 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 flicker effect is due to fluctuations of the conductivity of the oxide layer, the mechanism involved being identical with that responsible for resistance fluctuations.

621.381.4 1254

621.390.615 1255
Theory and Design of Valve Oscillators. [Book Review]—H. A. Thomas. Publishers: Ciapman & Hall, London, 2nd ed., 317 pp., 366 (Electric. Eng., vol. 29, pp. 1027; October, 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 material on frequency stabilization has taken place.

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
The Alternating-Current-Maintained Pendulum—N. Minorovsky. (Compt. Rend. Acad. Sci. (Paris), vol. 233, pp. 728-729; October 1, 1951.) The differential equations are developed and the solution of an oscillating 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

535.361.2
Scattering of Electromagnetic Waves from Two Concentric Spheres—A. L. Aden and M. Kerker. (Jour. Appl. Phys., vol. 42, pp. 2424-2426; 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; for certain appropriate conditions is reduced to the well-known solution for scattering from a single sphere.

535.37+537.311.3
Radiationless Transitions of Electrons in Crystals—F. Stockmann. (Z. Phys., vol. 130, pp. 477-474; 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 path in valence crystals. Observations of luminescence and inhibition in phosphors and of conduction in semiconductor devices are assisted in support of the theory.

535.43

537.33
General Solutions of the Equations of Electromagnetics and Magnetostatics—E. 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 analog of Green's scalar identity.

537.221
Contact Potential Differences—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
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 of metals, provided Fermi-Dirac rather than Maxwell statistics are used.

537.311.302.832.210
P-N Transition of an Oxide-Coated Cathode—Isikawa, Sato, Okumura and Sasaki. (See 1472.)

537.311.37
A General Formula for the Conductivity of a Gas Containing Free Electrons—L. G. Hill and E. J. Hall (Proc. Phys. Soc., vol. 64, pp. 851-861; October 1, 1951.) Electron drift in gases is discussed in terms of the method of free paths, for constant and alternating electric fields, and without applied magnetic fields. The 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 given 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.

537.50

537.58

537.581-13:537.569
The Ionization in the Incandescent Gases of Jet-Propulsion Mechanisms—C. Klein. (Ann. Telecommunication, vol. 6, pp. 287-289; October, 1951.) Anomalous 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 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. observations. The results are very different from Greeke's experimental results (657 of April). Physicochemical study of the combustion process shows that in most cases Saha's formula is inapplicable thermal equilibrium is not attained; calculation based on Sauer'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

538.12+538.65
Magnetic Field and Torque for a Magnetic Ellipsoid in a Permeable Medium and an External Field—J. Diedrichsath (Ann. Phys. (Lpz.), vol. 9, pp. 316-324; October 15, 1951.)

538.221
The Mechanism of Discontinuities in Magnetization—Sauer 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 result is related to that produced by hf biasing in magnetic sound recording.

538.241

538.566
Wave Packets, the Poynting Vector, and Energy Flow: Part 2—Poynting and Macdonald Velocities in Dissipative Anisotropic Media—C. O. Hillmer. (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 realistic results, but even then the direction obtained for the energy flow differs from that found by wave-packet methods, which probably give the best results. Part 4: 2607 of 1951.

538.613

538.145

538 Modern Magnetism. [Book Review]—L. F. Bates, Publishers: Cambridge University Press, London, 3rd ed. p. 506, pp. 308 (Electron. vol. 147, pp. 1207-1208; October 19, 1951.) "A fresh material is incorporated, including h.f. 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

523.621.306.822
Radio Telescopes and Astronomy—H. Klinger. (Frank u. Tim, 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.

522.72+523.854:621.306.822

523.74+1951.07/09

523.75+1281
A New Radio Method for Measuring the Electron Density in the Solar Corona—K. E. Macchi and F. G. Smithi. (Nature (London), vol. 168, pp. 599-600; October 6, 1951.) The method involves the "occultation" of the radiation from radio sources situated in directions near that of the sun. The solar radiation is prevented from masking the weaker radiation from the stars 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 from similarities as low as $10^{14}/	ext{cm}^2$. 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. R. B. P. Denis, É. L. Doroux and J. L. Steinberg. (Compt. Rend. Acad. Sci. (Paris), vol. 231, pp. 917-919 (October, 1951). Measurements are reported of the intensity of radiation wavelengths of 3.12 and 178 cm received during the eclipse at a location close to Markak, India, in the French observation on 31.15 cm confirming 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.

523.80 1283

A Diffraction Theory of the Scintillation of Stars on Optical and Radio Waves—C. G. Lingle. (Mon. Not. R. Astr. Soc., vol. 111, pp. 139-162 (October, 1951). The theory of refraction at optical wavelengths requires excessive atmospheric density gradients, and fails to explain the observed scintillations in solar, sidereal, and other large optical opticals. Theoretical calculations are insufficient to explain all the observations. The effects are explained by Friedel diffraction at a nonhomogenous gradient layer, which requires much smaller density gradients. A similar theory explains the observed intensity fluctuations of radio waves from discrete extraterrestrial sources.

523.82 1284

The Diffraction of Radio Waves in Passing through a Phase-changing Ionosphere—A. Hewitt. (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 resulting 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 variations in phase and constant variations in phase are considered. Knowledge of the amplitude and phase variations at the ground enables the average magnitude of the phase deviations produced in 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 diffraction 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.

523.85 1285


550.385 1286

International Data on Magnetic Disturbances of the same order as previously published nighttime values. The rate of production of charge carriers in the F2 layer at the start of the disturbance was twice as great as normal.

550.510 1296


550.510 1297

Spot-Frequency Ionospheric Recording—A Combination of Sweep and Fixed-Frequency Techniques—H. West. (Jour. Geophys. Res., vol. 66, pp. 1039-1049, December, 1951.) A combination of sweep and fixed-frequency techniques is proposed. The instrumental recording is used instead of photographing the complete echo trace, only one selected frequency being recorded. The photographic paper is moved on slightly after each frequency in the cycle. The resulting record is a combination of three tracings at different frequencies. Changes in structure with time are much more easily observed than on a series of full echo traces.

550.510 1300

Contribution to the Study of the Electron Distribution in the Ionosphere and of the Absorption of Short Waves—E. Argenzio, M. Mayot and K. Rawer. (Ann. Geophys., vol. 6, pp. 618-619, December, 1951.) A panoramic recorder is used instead of photographing the complete echo trace, only one selected frequency being recorded. The photographic paper is moved on slightly after each frequency in the cycle. The resulting record is a combination of three tracings at different frequencies. Changes in structure with time are much more easily observed than on a series of full echo traces.

550.510 1301

Group Velocities and Group Heights from the Magneto-ionic Theory—Shinn and Whale. (See 1405.)

550.510 1302

Radio-Wave Propagation at Oblique Incidence including the Lorentz Polarization
551.594.21  1303
The Distribution of Electricity in Thunderclouds—D. J. Malan and B. E. Schonland.
(Acoust. Roy. Soc., A, vol. 209, pp. 158-172; October 23, 1951.) Various independent methods of measuring the heights of initiation of successive lightning strokes are discussed; the values obtained support the hypothesis of a columnar distribution of negative charge.

551.594.22  1304
The Use of 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 at the summit of the Hebrides 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 Atmospheric on Board the Commandant Charcot, 1950-1951 Cruise—K. Bureau and J. J. Vaury. (Compl. Res. Inst. Sci. (Paris), vol. 233, pp. 1039-1051; October 29, 1951.) The third anticratic cruise of the Commandant Charcot was followed by one across the Pacific, during which further centers of atmospericism were located. The apparatus was the same as that used on the second cruise [see 637 of 1951 (Bureau and Barre)]. The whole of the Pacific west of 120° W seems to be a direct source of atmosperic oscillations. A typical recording for a 24-hour period on 18th-19th April 1951 shows the approximately linear rise of the level of atmospheric starting at sunset. Maximum activity for the Pacific center occurs at a later hour than for known continental centers, at 1700-1800 local time.

LOCATION AND AIDS TO NAVIGATION

621.306.9  1306
Radar Technique—E. Michelsen. (Freq. zvez., vol. 5, pp. 258-259; September, 1951.) A brief historical note of developments in Germany during the prewar and early war periods.

621.306.9:621.306.822  1307
The Detection of Pulse Signals near the Noise Threshold—R. E. Spencer. (Jour. Brit. IRE, vol. 1, 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 single pulse and then of repeated pulses. Matching the IF response to the frequency characteristics of the pulse is considered. The probability that the presence of a signal can be distinguished is shown to be dependent on the noise level of the detector law provided that it is given a monotonically increasing output with increased input. When signal and noise are of comparable magnitude, there is less of performance if detection is incorrectly distributed between the IF and video-frequency stages. Methods of presentation are described and the main practical conclusions are summarized.

621.390.63:621.390.812  1308
Disturbances caused by the Atmosphere in Aids to Navigation—E. Vasse. (Inde Elect., vol. 31, pp. 370-383; October, 1951.) Sources of error in direction-finding systems are discussed, including the effects of tilts and asymmetry in the higher ionized layers, and of lateral refraction and diffus-reflection in the troposphere.

621.396.32-1/2  1309
The Requirements for Radio Aids at Sea—J. R. N. White. (Jour. Inst. Nav., vol. 4, pp. 357-344; 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, conse, Dcena 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.

621.396.933  1310
Hydrobolic Navigation Systems in Germany—E. Roessler. (Elektrotech. Z., vol. 72, pp. 567-572; 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 loran assembly to the remote display unit might be adapted to provide displays in vessels in the service area of the radar equipment.

621.396.933  1311
Hydrobolic Navigation Systems in Germany—E. Roessler. (Elektrotech. Z., vol. 72, pp. 567-572; 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 loran assembly to the remote display unit might be adapted to provide displays in vessels in the service area of the radar equipment.

621.396.933  1312

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 of the radar control system at Farnborough, which is designed primarily for speed and flexibility. The approach consists of the controlling radar and extends into 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. Elect. Eng. (Aust.), vol. 23, pp. 167-181; September, 1951.) The basic aid adopted for use in Australia. The Australian visual aural radio range system (V. A. R.) in conjunction with distance-measuring equipment. The V. A. R., operating at about 170 m in the atmosphere of the American p-8 system (1138 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.

MODERN RADIO AIDS TO AIR NAVIGATION—THE D.M.E. PROJECT OF THE DEPARTMENT OF AIRPLANE NAVIGATION—H. F. Gander. (Proc. IRE, vol. 12, pp. 275-282; September, 1951.) A study of the D.M.E. project is reviewed and the main technical features of the equipment are briefly described. The chief radio-beacon sites discussed together with the proposed program of installation testing.

621.396.933(08)  1316
Radar as an Aid to Air Navigation in the Arctic—K. R. Grooman. (Jour. Inst. Nav., vol. 4, pp. 399-401; October, 1951.) Summary of paper read at meeting of Canadian Institute of Navigation, 1951. The use of radar for the determination of drift and groundspeeds, and for position-finding, 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 flight in the Arctic.

MATERIALS AND SUBSIDIARY TECHNIQUES

534.311.0001:534.315.615:537.5  1317
Mechanical Properties of Polymers at Ultrasonic Frequencies—W. Mason and H. J. Mecklenburg. (Bell Tech. Jour., vol. 31, pp. 112-171; January, 1952.) Different types of measuring 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

537.37  1319
Size Effects in the Luminescence of Na2S—K. Wallis. (Phys. Rev., vol. 84, pp. 375; October 15, 1951.) Thermoluminescence measurements for crystals of Na2S.CdS.CuZn and ZnS were made by saturation excitation at -160°C followed by a uniform increase of temperature to 100°C at a rate of 2°C per minute. Evidence is given of many more shallow traps in sodium sulphide (stage diameter 5.5a) than in large ones (average diameter 14a). The decay of phosphorescence with time was, however, found to be essentially independent of crystal size.

536.20  1320
A Fluorescent Cadmium-Iodide Screen under the Influence of Visible and Ultraviolet Radiation, Cathode Rays, X Rays and X Rays—S. Schlichtig. (Compl. Res. Acad. Sci. (Paris), vol. 243, pp. 1012-1024; October 29, 1951.) A strongly luminescent screen is obtained by mixing liquid iodide and cadmium iodide. When used as a cathode 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
538.221


538.221


538.221

The Value of the Spontaneous Magnetization of Binary Nickel Alloys as a Function of Temperature—J. W. Went. (Physica, 5, 81, pp. 596–600; June, 1951.)

538.221


538.234

Notes on the Metallization of Surfaces by Evaporation in Vacuum. I. Demayer (Comp. Rend. Acad. Sci. USSR, vol. 79, 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.

540.23:540.55:537.311.33

Electrical Properties of Selenium: Part 2—Microcrystalline Selenium—H. H. W. Henks. (J. Phys. Radium, vol. 22, pp. 1268–1278; October 1, 1951.) The magnetic properties of selenium have been studied in dependence on history of preparation, i.e., as functions of initial temperature of the liquid selenium, quenching procedure, nucleation procedure, temperature-crystallization, and time of crystallization. Single crystals and also microcrystalline matrix of these crystals were investigated. The effect of small quantities of sodium, oxygen, and iodine was determined. The frequency dependence of resistivity was also studied. Carrier densities and effective mobilities are estimated from the dependence on ferromagnetic power and resistivity. The results obtained, and their relation to other observations and to theories of semiconductors, are discussed. Part 1: 968 of April.

540.23:03


540.28:537.314.6

Properties of Ionic Bombarded Silicon—R. S. Ohl. (Bell Syst. 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 allows little rectification can be modified by bombardment to make its rectifying properties better than those of most unbombarded materials.

546.289:537.312.67

The Electrical Conductivity of Liquid Germanium—W. H. Keys. (Phys. Rev., vol. 84, pp. 86–87; October 15, 1951.) The resistivity of liquid Ge near the melting point was found to be 60 μm cm (about 1/15 that of solid Ge at the melting point), with a position of the transition of about 100°C. The purity of the Ge was stated to be over 99.9 per cent.

546.321:85:021.301.5

Difference Between the Dielectric Constants of Free and Clamped KH2PO4 Crystals—H. Baumgartner. (Helv. Phys. Acta, vol. 24, pp. 326–329; September 20, 1951. In German.) Measurements in the temperature range 150°C to ~200°C confirm the Curie-Weiss behavior of the lattice, but the dielectric constant and the temperature, both for a crystal effectively clamped by exciting it at its 201st harmonic (10 mcp) and for a free crystal excited at its respective frequency of 1 kep. The graph for the free crystal is displaced about 4°C, toward higher temperatures, relative to that for the clamped crystal. The results are in complete agreement with Müller's theory.

546.431.824–31


546.431.824–31

Theory of Barium Titanate: Part 2—F. Debye. (Phys. Rev., vol. 42, pp. 1065–1070; October, 1951.) The theory given in part 1 (663 of 1950) is extended, and expressions are given for the piezoelectric constants, the elastic coefficients and derivatives with respect to the temperature, 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 frequency, and comparison is made with data obtained 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


546.817.221:621.3.014.6

The Rectifying Properties of Lead Sulphide Antimonide and Lead Sulfide Antimonide. I. Adachi and I. S. Sekiya. (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 on 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.

Electric and Thermoelectric Properties of Particles of Tetrahedral Titanium Disilicide—B. I. Bolkatsa, F. I. Vaunin and A. K. Salitina, (Zh. Tekh. Fiz., vol. 21, pp. 532–540; May, 1951.) Experiments were conducted for determining, within a wide range of temperature, the specific conductivity and thermoelectric emf of samples of TiO₂ at various degrees of reduction. The results obtained are explained from the standpoint of modern theories on the conductivity of semiconductors.

Dielectric Properties of Various Preparations of the Osmium-Bismuth—J. N. Belyavskii, N. V. Novosel'tsev, A. I. 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 TiO₂ has been studied on six different samples. The results obtained are shown in curves and tables.

Growing Piezoelectric Crystals—R. Mosaner and M. Wurl. (Arch. elek. 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.

Interatomic Distances and Ferromagnetism in Spinel—R. S. Weim. (Phys. Rev., vol. 84, pp. 1001–1007; October 15, 1951.) Discussion with particular reference to the study of the spin interaction is inversely proportional to the distance from a metal ion to a nearest neighbor (i.e., an oxygen ion) and hence to another metal ion.


Synthetic Materials—P. Domin. (Ferrmeldtech. Z., vol. 4, pp. 461–467; October, 1951.) Review of various plastics and their derivation. Mechanical and electrical properties are tabulated.

Frequency Shift of Piezoelectric Oscillations of Quartz under High Pressure—Michel and J. P. Perez. (Physica, τ Grau., vol. 17, pp. 563–564; May, 1951.) Measurements are reported on two quartz resonators, an AT cut with frequency 942 kec and a BT cut with frequency 826 kec at atmospherical pressure. The slope of the frequency/pressure curve is positive for the AT and negative for the BT cut.

Technical Control in Glass Manufacture—J. H. Partridge and E. Preston. (GEC Jour., vol. 14, pp. 212–220; October, 1951.) Descriptions are given of an automatic method of controlling the temperature of a large glass-making tank furnace, a photoelectric pyrometer for measuring 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.

The Impulse Integral, a Counterpart of the Duhamel Step Integral—R. Lurie, M. M. Vider and W. Reichardt (Ann. Phys. (Lpz.), vol. 9, pp. 307–315; October 15, 1951.) If a time-variable disturbance force acts on a linear system and 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 (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.


The Logarithmic-Complex Number Plane and the Complex-Number Calculator—W. de Beauchair. (Z. Ver. Deut. Ing., vol. 93, pp. 953–957; October 15, 1951.) The complex-number plane with rectangular and polar-coordinate networks can so be transformed that multiplication and division of complex numbers reduce to simple addition and subtraction. This forms the basis of a simple calculator.

The Binc—A. A. Auerbach, J. F. Eckert, J. R. Shaw, J. R. Weinler and L. D. Wilson. (Proc. I.R.E., vol. 40, pp. 291–299; January, 1952.) A comprehensive description of a high speed electronic digital computer. The Binc 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 relaying gates; the latter allow a "pulse" to pass through when the information is correct. 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 an electronic tape or magnetic type writer.


Magnetostriiction Storage Systems for a High-Speed Digital Computer—Millerhurst, Robbins and D. Barr. (See 1200.)

Given by which a straightforward quantitative analysis of a wave can be effected. Numerical calculations are made for a square wave.

Highly Stable V-T Voltmeter—M. G. Scroggie. (Electronics, vol. 24, pp. 142, 224; December, 1951) See Table of MT.

A method differs from those of Benoit (395 of 1950) and Roberts and von Hippel (178 of 1947) is presented for the graphical solution of the equation $2\pi a + \lambda = 0$, as well as involved in the determination of dielectric constants by the method of standing waves in wave guides.


Measurement of the Electrical Properties of Highly Absorbing Dielectrics on Centimetre Waves by the Infinite Layer Method—N. N. Markovitch. (J. Tech. Fis., vol. 21, pp. 647-661, June, 1951) The dielectric under test is found in turn, using a calibrated AF, and applied to a $\mu$m-2 klystron multiplier. At a frequency of 100 kHz, the klystron is used for calibrating waveguides, for frequency measurement, and for estimation of the depth of parasitic modulation.

A Triple Check of Each Narrow-Loaded Crystal. G. Beadle. (J. Tech. Fis., vol. 40, pp. 514-517; October, 1951.) Details are given of an electrical relay mounted on the control tube.

A Review of Progress in Electric Furnaces—D. M. Dovoy and J. Jenkins. (GEC Jour., vol. 18, pp. 194-211; October, 1951.) Descriptions are given of both well-known and novel types of furnace operated by arc and resistor heating.

The Linear Electron Accelerator as a Pulsed Neutral Source. (Proc IEEE, vol. 2, pp. 295-300, October, 1951.) Descriptions are given of the principles of process control are discussed, and a method of converting the physical quantities into dc is described. Details are given of a pulsed electron source which operates an electro-pneumatic relay mounted on the control tube.

A Simple and Uneleveled, of Eighteenth Harmonic (Odd Harmonics only)—P. Kemp. (Electronic Eng., vol. 23, pp. 390-393; October, 1951.) Grouping of selected or litaves in the wave equations is applied, as in earlier analysis (465 of 1943), to obtain the coefficients of the sum and difference components. Tables are given by which a straightforward quantitative analysis of a wave can be effected. Numerical calculations are made for a square wave.

3.2170.20:621.39 1370
Quartz-Crystal Measurement at 10 to 180 Mc/s—E. H. Bewer. (Proc. I.R.E., vol. 40, pp. 36-40; January, 1952.) A method is described for measuring the equivalent parameters of the main and spurious modes. The crystal is mounted on a slide between the anode and cathode of an amplifier tube, and the voltage across it is recorded as a function of frequency. The required parameters can then be derived.

3.2170.20:621.36.67.029.64 1380
Automatic Antenna Wave-Front Plotter—R. M. Garrett and H. Barnes. (Electronics, vol. 25, pp. 120-125; January, 1952.) Description of equipment which scans a plane, 30 x 36 inches, in front of a microwave antenna and plots either phase or amplitude contours.

3.2170.79:621.36.645.018.424 1381
Wideband Pre-amplifier—Horner. (See 1280.)

3.2165.611.15:534.844.1 1382
Equipment for Acoustic Measurements: Part 2—A Portable Tube Source Developed for Use in Room Acoustics—C. G. Mayo and D. G. Beadle. (Electronic Eng., vol. 23, pp. 386-373; October, 1951.) Description of a unit which provides an adjustable output of from 0.1 to 1 mw into a 6000 load. The effective incremental capacitance of the tuning capacitor in the variable oscillator is dependent on the circuit gain. This is varied by a signal from a transmission relay to effect a 10 per cent variable at 75ms with negligible amplitude modulation. A synchronous motor and gear-box effect a frequency sweep of 20 cps-20 kcps in 4, 8, 16 or 32 minutes. Detailed circuit diagrams and performance curves are shown.

Wide-Band Converter for Signal Generator—D. M. Hill. (Electronic Eng., vol. 24, pp. 118-121; December, 1951) FM and AM signals are provided over the whole frequency range 100 kcps-20 mks by a vih signal generator operating in conjunction with a wide-band converter.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.706 1384
Self-Adjusting Timer for Bullet Photography—H. W. Burlock. (Rev Sci. Instr., vol. 22, pp. 741-745, October, 1951.) A reference voltage is produced proportional to the time interval taken for the bullet to travel a given distance. This voltage is used to measure off a second interval equal to the first and immediately following it, at the end of which the light source is flashed. The system is useful for measuring the velocity of the bullet is not known in advance.

538.509.047 1385
The Dielectric Behaviour of some Types of Human Tissues at Microwave Frequencies—H. A. Wilton. (Phys. Med. Biol., vol. 1, pp. 336-349; May, 1951.) Measurements of the complex dielectric constant were made over the wavelength range 6.5-17 cm, using a coaxial line method described by Roberts and von Hippel (178 of 1947). Results are discussed in relation to dielectric theory.
the current decreases and then changes its sign. An explanation of the phenomena is advanced.

546.8-24: 513.61.02.4
Dielectric and Thermoelectric Properties of Partially Reduced (Blue) Titanium Dioxide—B. I. Boltake, F. I. Vavilov and A. E. Salutina. (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 TiO₂ at various degrees of reduction. The results are obtained and explained on the standpoint of modern theories on the conductivity of semiconductors.

546.31: 021.315.61.24
Applications of Various Preparations of Titanium Dioxide—I. N. Belverov, N. S. Novosiltsev, A. L. Khokhlov and M. S. Shul’man. (Zh. Tekh. Fiz., vol. 21, pp. 547–551; May, 1951.) The effect of impurities on the dielectric properties of TiO₂ has been studied on six different samples. The results obtained are shown in curves and tables.

548.0: 537.228.1
Growing Piezoelectric Crystals—M. Moaner and M. Wuri. (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
Interatomic Distances and Ferromagnetism in Spinels—R. S. Weiss. (Phys. Rev., vol. 84, p. 379; April, 1951.) Discussion of a 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.

621.315.61.2: 537.11

621.315.616

621.396.611: 549.514.51
Frequency Shift of Piezoelectric Oscillations of a Quartz Tank under High Pressure—A. V. Michels and J. P. Pérez. (Physica, § Grav., vol. 17, pp. 563–564; May, 1951.) Measurements are reported on two quartz resonators, an AT-cut frequency 942 kc/s and a BT-cut with frequency 604 mps at atmospheric pressure. The shape of the frequency/pressure curve is positive for the AT and negative for the BT cut.

666.1
Technical Control in Glass Manufacture—J. H. Higginbotham. (GEC Jour., vol. 18, pp. 212–220; October, 1951.) Description of a machine that is used for measuring 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: 021.396.611
The Impulse Integral, a Counterpart of the Duhamel Step Integral—R. Lucx, M. Püdder 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

517.9

518.55
The Logarithmic-Complex-Number Plane and the Complex-Number Calculator—W. de Beauchârt. (Z. Ver. Dtsch. Ing., vol. 93, pp. 955–957; October 21, 1951.) 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
The Bina—A. A. Auerbach, J. P. Eckert, Jr., R. F. Shaw, J. R. Weiner and I. D. Wilson. (Proc. I.R.E., vol. 40, pp. 12–29; January, 1952.) A comprehensive description of a high-speed electronic digital computer. The Bina 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 rescaling gates; the latter allow a "substitution" or transformation of the circuit structure. 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
Logical Description of Some Digital Computers (October 15, 1951). A digital computer is represented by a circuit diagram in the EDVAC computer.

681.142: 517.942.9

681.142: 021.392.5
Magnetostriiction Storage Systems for a High-Speed Digital Computer—Millsorh, Robbins and De Barr. (See 1200.)

517.6: 021.396.611
Abstracts and References

1052


021.373.35+J38.535.4-C3:3.31 1368


021.373.3.020.63 1369


021.373.3.020.04 1370


021.373.3.020.04 1371


021.373.3.092.06:521.379.24 1373

Measurement of Differences of Group Propagation Time on a Non-looped Line—J. N. See and J. H. Hiss (Can. Soc. Trans. Paris), vol. 10, pp. 142-146, October 1951.) Two X-rays are simultaneously applied to the cable, one of fixed frequency 100 kcps modulated at 10 kcps, the other varying from 250 kcps to 10.5 mcs and modulated at 50 kcps. The 10-kcps modulation is obtained by division from the 50-kcps modulation, so that there is a constant though unknown phase relation between them. The two carriers were modulated in the presence of saw-tooth modulated modulator, and the phasing of the two-carrier waves was varied. The receiving end, and the phases, after multiplication of the 10-kcps frequency, compared. The choice of frequencies is based on minimum requirements for the transmission of the unique definition television signals. The equipment, details of which are given, has an accuracy of within ±0.02 µs for group-propagation time measurements, the sensitivity being 0.005 µs.

021.373.35 1374

Harmonic Analysis of Waves up to Eleventh Harmonic (Odd Harmonics only)—P. Kemp (Electronics Eng. vol. 23, pp. 390-391, October, 1951.) Grouping of selected or injected in the wave equations is, as earlier analyses (466 of 1943), to obtain the co-efficients of the sum and difference components. Tables are given by which a straightforward and quantitative analysis of a wave can be effected. Numerical calculations are made for a square wave, 021.373.725 1375

'Stromdämmung'—a New [German] Term for Rating Voltmeters—M. New (Proc. F. W. E., vol. 5, pp. 514-517, October 1951.) The term 'stromdämmung' and subsequent to the noun 'stromdämmung' in 'stromdämmung,' (literally, a current reduction), defined as the ratio of total meter resistance to voltage for full scale deflection.

021.373.725 1376

Highly Stable V.T. Voltmeter—G. Sebastian (Radio electron., vol. 24, pp. 142, 224, December, 1951.) See Table 1041 of Volt.

021.373.76 1377

The Secondary-Standard U. H. F. Generator of the Companie générale de T. S. F.—M. Denis and H. W. Turner (11 T. S. F., vol. 6, pp. 148-149, October 1951.) A 1750 mcps quartz crystal controlled frequency is multiplied to 2700 mcps and applied to a type 14 KV 14 kV multiple. No distortion is achieved and the frequency is 1st with its intermodulations 100%. Available where 24 < S 50. Further subdivision is obtained by interpolation, using a standard variable frequency oscillator. The normal range of 2400-430 kc can be extended to 0-60 kc by use of multiplication methods. Accuracy is within 5 parts in 10 kc. It is suggested that the output may be used for dual tuning meter, for frequency measurement, and for estimation of the depth of passive modification.

021.373.3.020.04 540.217 1378


021.373.36.092:062.379.24 1379

Navy Primary Standard Frequency Meter—W. Carr (Electronics, vol. 24, pp. 142, 1951.) Description of the new cavity resonator equipment for frequency measurements from 15 kc to 27 mcps with accuracy to within 1 cm in 10 mcps. This equipment is used by frequency measurement units at large naval radio stations on shore.

021.373.76 1380

Navy Primary Standard Frequency Meter—W. Carr (Electronics, vol. 24, pp. 142, 1951.) Description of the new cavity resonator equipment for frequency measurements from 15 kc to 27 mcps with accuracy to within 1 cm in 10 mcps. This equipment is used by frequency measurement units at large naval radio stations on shore.

021.373.76 1381

A Wideband Converter for Signal Generator—D. M. Hill (Electronics, vol. 24, pp. 118-121, December, 1951.) FM and AM signals are provided over the whole frequency range 100 kc to 216 mcps by a wideband generator operating in conjunction with a wide-band converter.

021.373.76 1382

Other Applications of Radio and Electronics

551.756 1383

Self-Adjusting Timers for Bullet Photographic

551.756 1384

Self-Adjusting Timers for Bullet Photographic

R. H. Lucas (Rev. Sci. Inst., vol. 22, pp. 741-745, October 1951.) A reference voltage is produced proportional to the time interval taken for the bullet to travel a given distance at this voltage is used to measure off a second interval equal to the first and immediately following it at the end of which the light source is flashed. This system may be used for a high-speed photography of the bullet not known in advance.

538.509.24 1385

The Dielectric Behaviour of Some Types of Human Tissues at Microwave Frequencies—11. F. B. L. and A. Bergman (Proc. IEE., Part II, vol. 69 pp. 669-670, October 1951.) Measurements of the complex dielectric constant were made over the wavelength range 5-91 cm, using a novel low-noise method described by Roberts and von Hippel (178 of 1947). Results are discussed in relation to dielectric theory.

021.52-521.389 1386

An Electronic Process-Controller—J. R. Bounias and A. Bergman (Proc. IEE., Part I, vol. 69, pp. 669-670, October 1951.) Description of the new cavity resonator equipment for frequency measurements from 15 kc to 27 mcps with accuracy to within 1 cm in 10 mcps. This equipment is used by frequency measurement units at large naval radio stations on shore.

021.365 1387

A Review of Progress in Electric Furnaces—D. M. Doye and J. Jenkins (GBJ., vol. 18, pp. 194-211, October, 1951.) Descriptions are given of well-known and novel types of furnace operated by rf, arc and resistance heating.

021.384.6 1388

The Linear Electron Accelerator as a Pulsed Neutron Source—T. F. Ferris (Nuclear Phys. Rev., vol. 3, pp. 9-19, 1951.) In laminar principles of process control are discussed and a method of converting the physical quantities into dc is described. The power of an electron beam which operates a fast pneumatic relay mounted on the control tube.

021.385 1389


021.385.83 1390


021.385.83 1391

Theory of the Elliptical Electrostatic
June

PROCEEDINGS OF THE I.R.E.  

621.386.11  1403  A Geometric Interpretation of the II-Wave and Coupling Factor in Ionospheric Long-Wave Theory—N. N. Davis. (Journ. Geophys. Rev., vol. 56, pp. 611-612; December, 1951.) Analysis shows that the II waves are simply the combination 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 real and imaginary parts of the Appleton-Hartree equations is found to represent the space rate of twist of the polarization axes.

621.386.11  1404  Spanish Method for the Prediction of Optimum Working Frequencies at Any Distance—H. C. Phillips, Jr. (Journ. Appl. Phys., vol. 22, pp. 1051-1052; October, 1951.) The group velocity in the ionosphere above southern England is calculated for the ordinary and extraordinary waves 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 velocity/frequency curves for various magnetic latitudes, for the ordinary ray incident vertically on a layer with parabolic distribution of refractive index. It is shown that estimates of layer thickness neglecting the magnetic field are incorrect for the magnetic equator but are too high for other latitudes and by amount varying from 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.

621.391.11  1405  Group Velocities and Group Heights from the Magnetosphere—D. A. Sherman and H. A. White. (Journ. Atmos. Terr. Phys., vol. 2, pp. 85-105; February, 1951.) The group velocity in the ionosphere above southern England is calculated for the ordinary and extraordinary waves 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 velocity/frequency curves for various magnetic latitudes, for the ordinary ray incident vertically on a layer with parabolic distribution of refractive index. It is shown that estimates of layer thickness neglecting the magnetic field are incorrect for the magnetic equator but are too high for other latitudes and by amount varying from 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.

621.391.11  1406  Contribution to the Study of the Electron Distribution in the Ionosphere and of the Absorption of Short Waves—Argence, Mayot and Rawer. (See 1408.)

621.391.11  1407  Radio-Wave Propagation at Oblique Incidence including Lorenz Polarization Term—Kelso. (See 1302.)

621.391.11  1408  Tropospheric Propagation beyond the Horizon—J. Feinstein. (Journ. Appl. Phys., vol. 22, pp. 1292-1293; October, 1951.) Recent calculations of the field strength of high waves, which are the result of a combination of ordinary and extraordinary waves, 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 the reflection of light at the ground is given, and results deduced from it are shown graphically.

Abstracts and References

621.396.621

621.416
A General Theory for Frequency Discriminators containing Null Networks—J. L. Stew- art. (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 radians is maintained at all frequencies.

621.461.54

621.461.54:621.3015.7
Pulse Response of A.M. Receiver—R. Kitai. (Wireless Eng., vol. 29, pp. 15-18; January, 1952.) An analysis made is of the transient response of an A.M. 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 only on the amplitude of the input pulse, 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.

621.476.822
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 m-square frequency of the spectrum.

621.482 Observed Groups of Peaks of Electrical Fluctuations—V. I. Bunimovich. (Zh. Tekh. Fiz., vol. 21, pp. 637-646; June, 1951.) In 149 above separate random peaks were considered. In the majority of cases peaks occur in groups, and one of which can be observed on a printout 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.

621.482.826:621.6019.13

STATIONS AND COMMUNICATION SYSTEMS

517.512:621.390.011:621.396.67.0127.1
Fourier Analysis and Negative Frequencies—J. L. Shaw. (Wireless Eng., vol. 29, pp. 3-12; January, 1952.) The relation between the time function \( \phi \), expressing a fluctuation, and the frequency components \( s \) of \( \phi \) is examined; if \( \phi \) is given by an integral in which \( s \) ranges from \( -\infty \) to \( +\infty \), the negative part of the spectrum (carrier) 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 distribution 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.

621.390.011.11
On the Definition of Information—E. Reich. (Jour. Math. Phy.., 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.

621.391.11
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 in the next letter of a text from knowledge of the preceding text. Some properties of an ideal predicting system are developed.

621.390.011

621.395.44:621.315.052.63
British Broadcast Commission and Applications of Carrier Current Principles for Operating Requirements of Power Utilities—W. D. Goodman. (G.E.C. Jour., vol. 18, pp. 229-236; October, 1951.) A.I.E.E. Summation General Meeting paper. Details are given of wide band coupling equipment for carrier-current working over power lines; the advantages of interphase coupling over monophasic coupling are explained. Sal operation is used in order to reduce noise. Recent British types of communication and relaying equipment are described.

621.396.019.13
Investigations of Frequency [modulation systems with] Negative Feedback—J. Hacks. (Arch. elektr. Ubertrag., vol. 5, pp. 441-450; October, 1951.) A discussion of the use of negative feedback to reduce distortion in F.M. transmitters and receivers. By examining considerable to small values of the distortion the calculation can be simplified in comparison with earlier methods [see 3120 of 1939 (Chaffee) and 3544 of 1949 (Panch and Dito) and phase shifts due to the finite group transmission time can be taken into account. The effect on the-all-over distortion of nonlinearities in the feedback loop is examined. A circuit for equalizing the output of modulators-fundamental frequency range is described. With a feedback factor > 2 a significant improvement can be obtained in the effective selectivity of a F.M. receiver.

621.396.65:621.390.97
Bases of Calculation for the Design of Radio Communication Systems—H. Herzan. (Radio Tech. (Vienna), vol. 27, pp. 423-427; October, 1951.) The theoretical concepts developed for radio broadcasting, telephony and telegraphy systems are discussed. Graphs are reproduced showing noise spectra, ground-wave attenuation, antenna loss, and frequency-related 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
Nonlinear Crosstalk in Multichannel Systems during Transmission over U.S.W. F.M. Links compared with the Specifications for Cable Connections—E. Ketel. (Telefunken Ztg., vol. 24, pp. 163-168; October, 1951.) Modulation in carrier-frequency systems is considered, and the distortion in an all-s.m. F.M. installation is expressed in terms of crosstalk. F.M. 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 F.M. 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
V.H.F. Radio Multichannel Carrier Telephone Circuits in Colombia—L. C. Simpson, H. B. Nevitt and E. J. Eriksen. (Eriksen Rev., vol. 28, pp. 199-206; February 1950.) A description of a triple link connecting Bogota and Medellin, distant 250 km apart. F.M. broadcasting equipment is used; the frequency range is 70-88 mcs.

621.396.65.029.63/04
The Planning of Beam Links in the Deci- meter and Centimeter Regions—K. Schmidt. (Telefunken Ztg., vol. 24, pp. 129-139; October, 1951.) General considerations and expected future developments are enumerated, and planning details for the F.M. 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 power construction and suitable power supply arrangements, in this case mainly wind-driven generators, is stressed.

621.396.65.029.63
The IDA 22-Beam Link Equipment—G. Ulbricht. (Telefunken Ztg., vol. 24, pp. 143-149; October, 1951.) A description of the wartime U.S.W. 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. F.M. with modulator and demodulator construction and suitable power supply arrangements, in this case mainly wind-driven generators, is stressed.

621.396.65.029.63


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, and for correcting the elements of Schmidt large-screen projection systems.

Television Standards in Argentina—(Rec. tele. 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.


An Ultra-High-Frequency Television Converter—B. T. Tyson, (N.Y. Technologist, vol. 4, pp. 74–81; October, 1951). The design and performance are described of a simple, low-cost converter which enables transmissions in the band 475–990 mcps to be received on existing receivers using the girl of television receivers. Further advantages of Al back are the improved reception of weak signals.


Television Camera Tubes: Part 2—Comparison of Characteristics of Camera Tubes—P. Schagen, (Jour. Brit. I.R.E., vol. 16, pp. 227–242; September, 1951). Classification, p. 242.) The types of tubes considered are the iconoscope, the image iconoscope, the orthicon and the image orthicon. Characteristics discussed include linearity, effective exposure time, distortion of the image and the degree of flare, and degree of flare in the image.


621.396.619.13

621.396.619.13 1462 On the Operating Conditions of Reactance Circuits for rectification of electron waves. Variations—H. Fricke. (Fermendelechez. Z., vol. 4, pp. 458-461; October, 1951.) A pentode reactance tube is operated about a point of zero slope on the \( I / V \) 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 +1.77 per cent at 1.46 mcps and -7.9 per cent at 27.5 mcps were obtained.

621.396.619.14 + 621.314.26 1463 The Travelling-Wave Valve as a Microwave Phase Modulator and Frequency Shifter—W. J. Fontana. Proc. I.E.E., 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 mcps, by injecting the modulating signal in series with the beam voltage. This principle applied to high frequency shifter gives gain and output power 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.

621.396.219.33

621.396.219.33 1464 High-Efficiency Grid Modulation—L. A. Matthews. (R.S.G.B. Bull., vol. 21, pp. 144-147 and 193-196; October and November, 1951.) The Taylor and Terman-wooddy grid-modulation circuits are analysed. 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 tube in such a way that the output 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-Woodward system, impedance-inverting networks are used to bring about an equal shunting of the load between the two tubes. Experiments carried out with both systems indicate that the Terman-Woodward circuit is not only practicable, but is the way in which the electrostatic circuit. Tested with a lamp load, the Terman-Woodward circuit gave the higher efficiency and also much better linearity. Circuit adjustments are not unduly critical.

621.396.822

621.396.822 1465 Determination of the Electron Temperature in Gas Discharges by Noise Measurements—K. S. Knohl. (Philips Res. Rep., vol. 6, pp. 288-302; August, 1951.) The electron temperatures of discharges in He, Ne, Ar and Xe, as deduced from measured noise power levels, and from experiments using a probe, agree with values deduced theoretically from the gas pressure and the radius of the discharge tube. Precisions have been achieved to match the discharge to the waveguide are described.

621.396.512: 537.533.7

621.396.512: 537.533.7 1466 Noise of the Space-Charge Effects in 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.


621.383.546.289

621.383.546.289 1468 A Photovoltaic Germanium Cell—B. J. Rothstein. (Sylvania Technol., vol. 4, pp. 88-100; December, 1951.) Description 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: 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 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-conversion-current minimum. A similar reduction is given by two velocity jumps, a deacceleration at a certain 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

621.385.032.216 1471 P-N Transition of an Oxide-Coated Cathode—Y. Ishikawa, F. Shuku and I. Okumura. (Phys. Rev., vol. 84, pp. 371-372; October 15, 1951.) P-type conductor was found for a (BaSr)O cathode in an oxygen atmosphere, as opposed to n-type conductor 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 1472 The "Grip-O-Marie" Water Jacket for Large-Water-Cooled-Transmitting Tubes—A. G. Robere and W. L. Verveer. (Commun. New., 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.

621.385.032.42 1473 The "Grip-O-Marie" Water Jacket for Large-Water-Cooled-Transmitting Tubes—A. G. Robere and W. L. Verveer. (Commun. New., 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.

621.385.032.42 1473 The "Grip-O-Marie" Water Jacket for Large-Water-Cooled-Transmitting Tubes—A. G. Robere and W. L. Verveer. (Commun. New., 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.

621.385.032.42 1473 The "Grip-O-Marie" Water Jacket for Large-Water-Cooled-Transmitting Tubes—A. G. Robere and W. L. Verveer. (Commun. New., 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.
here and approximations previously derived is shown.

621.385.1:56.289

Transistors—The U.I.C. number 621.585.3: 56.289 used hitherto for transistors will be replaced by 621.314.7.

621.385.8:621.318.572.001.8


621.385.831

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 9002 and the gain-bandwidth product about 320 mcps.

621.396.615.141.2

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

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 coordinate s. Thus the system can be used for amplification of microwave signals.

621.396.615.141.2

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 x-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

Equivalent Networks of Klystrons—S. Uda and J. Ikekuchi. (Tech. Rep. Tohoku Univ., vol. 14, pp. 117–120; 1956.) 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.