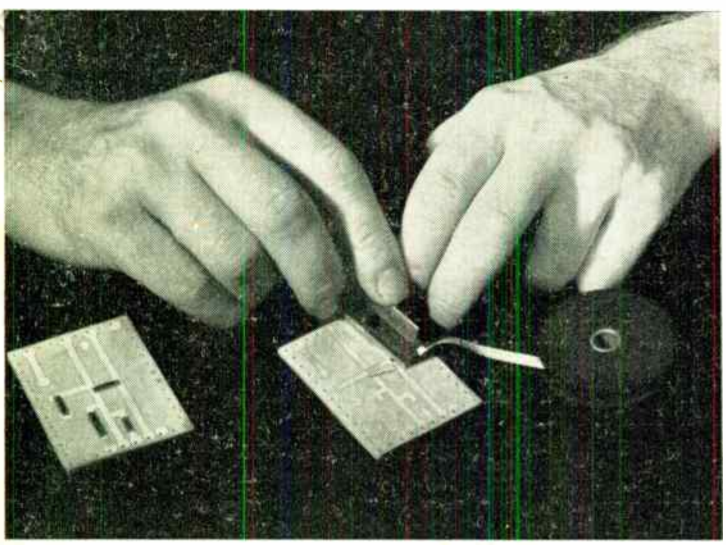
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

December, 1951
Volume 39 Number 12



National Bureau of Standards

UNROLLING MINIATURE RESISTORS

A 1/8-inch wide adhesive-tape resistor is applied from a spool to a silver-printed ceramic chassis, pressed into place, and cut to length. After oven curing, the resistor darkens (see left sample).

PROCEEDINGS OF THE I.R.E.

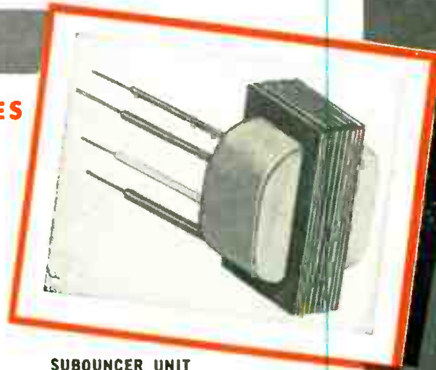
Table of Contents, Indicated by Black-and-White Margin, Follows Page 64A

The Institute of Radio Engineers

MINIATURE COMPONENTS FROM STOCK...

SUBOUNCER UNITS

FOR HEARING AIDS...VEST POCKET RADIOS...MIDGET DEVICES

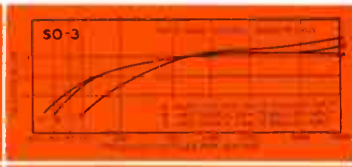


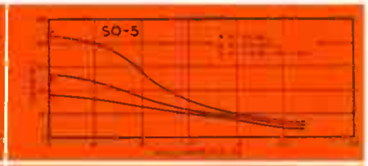
SUBOUNCER UNIT
Dimensions....9/16" x 5/8" x 7/8"
Weight.......................................03 lb.

UTC Sub-Ouncer units fulfill an essential requirement for miniaturized components having relatively high efficiency and wide frequency response. Through the use of special nickel iron core materials and winding methods, these miniature units have performance and dependability characteristics far superior to any other comparable items. They are ideal for hearing aids, miniature radios, and other types of miniature electronic equipment. The coils employ automatic layer windings of double Formex wire . . . in a molded Nylon bobbin. All insulation is of cellulose acetate. Four inch color coded flexible leads are employed, securely anchored mechanically. No mounting facilities are provided, since this would preclude maximum flexibility in location. Units are vacuum impregnated and double (water proof) sealed. The curves below indicate the excellent frequency response available. Alternate curves are shown to indicate operating characteristics in various typical applications.

Type	Application	Level	Pri. Imp.	D.C. in Pri.	Sec. Imp.	Pri. Res.	Sec. Res.	List Price
*SO-1	Input	+ 4 V.U.	200 50	0	250,000 62,500	16	2650	$6.50
SO-2	Interstage/3:1	+ 4 V.U.	10,000	0	90,000	225	1850	6.50
*SO-3	Plate to Line	+ 20 V.U.	10,000 25,000	3 mil. 1.5 mil.	200 500	1300	30	6.50
SO-4	Output	+ 20 V.U.	30,000	1.0 mil.	50	1800	4.3	6.50
SO-5	Reactor 50 HY at 1 mil. D.C. 3000 ohms D.C. Res.							5.50
SO-6	Output	+ 20 V.U.	100,000	.5 mil.	60	3250	3.8	6.50

*Impedance ratio is fixed, 1250:1 for SO-1, 1:50 for SO-3. Any impedance between the values shown may be employed.





SUB-SUBOUNCER UNITS

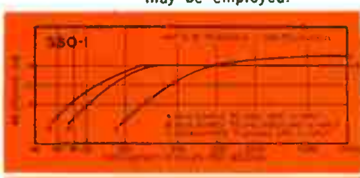
SUB-SUBOUNCER UNIT
Dimensions....7/16" x 3/4" x 5/8"
Weight.......................................02 lb.

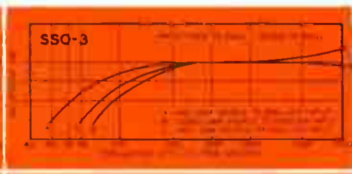
FOR HEARING AIDS AND ULTRA-MINIATURE EQUIPMENT

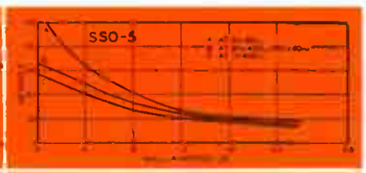
UTC Sub-SubOuncer units have exceptionally high efficiency and frequency range in their ultra-miniature size. This has been effected through the use of specially selected Hiperm-Alloy core material and special winding methods. The constructional details are identical to those of the Sub-Ouncer units described above. The curves below show actual characteristics under typical conditions of application.

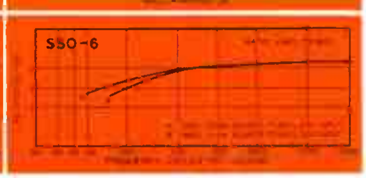
Type	Application	Level	Pri. Imp.	D.C. in Pri.	Sec. Imp.	Pri. Res.	Sec. Res.	List Price
*SSO-1	Input	+ 4 V.U.	200 50	0	250,000 62,500	13.5	3700	$6.50
SSO-2	Interstage/3:1	+ 4 V.U.	10,000	0	90,000	750	3250	6.50
*SSO-3	Plate to Line	+ 20 V.U.	10,000 25,000	3 mil. 1.5 mil.	200 500	2600	35	6.50
SSO-4	Output	+ 20 V.U.	30,000	1.0 mil.	50	2875	4.6	6.50
SSO-5	Reactor 50 HY at 1 mil. D.C. 4400 ohms D.C. Res.							5.50
SSO-6	Ouptut	+ 20 V.U.	100,000	.5 mil.	60	4700	3.3	6.50

*Impedance ratio is fixed, 1250:1 for SSO-1, 1:50 for SSO-3. Any impedance between the values shown may be employed.









United Transformer Co.

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

Forty Years — Sets the Pace . . .

1912 1952

is the theme of the 1952 IRE National Convention and Radio Engineering Show, marking the progress of IRE's first forty years' service.

— What this Annual Event Means —

To the Member and Visitor:

The Conference and Show provide a refresher course on the progress and pace of the science and engineering application of radio in television, broadcasting, electronics and defense.

In no other way can one feel the pulse-beat of a great industry as easily, and learn as much as in the four days of this meeting. From theory in the technical papers, to practice in engineering demonstrations of the exhibits, this is the modern, fast way to keep up-to-date.

So convinced of this is our industry that last March 5,082 different firms sent their engineers and executives to this conference. 15,258 attended sessions of their choice. Of the 22,919 who registered, all but ¼ of 1% visited the exhibits. 7,924 were IRE Members. This means that nearly 30% of the total membership of this engineering society attended its annual national convention.

Skillfully grouped technical papers map the advances in every field of radio, from guided missiles to television, and from communications to industrial electronics. This year, more than 200 papers will be presented, using five lecture halls—giving engineers an equivalent coverage of five conferences in one.

To the Radio Industry:

What is the significance to an industry when 277 exhibitors meet in one place for four days to show their technical products to engineers? It means that IRE has produced through its meeting an economically sound market-place.

These exhibitors are the top 15% of the manufacturers supplying the field, who produce 90% of the products engineers buy! Is it any wonder that the audience comes to such a concentration of equipment, component and instrument exhibits? Think of the time saved!

Here is the record:

1947—177 firms met 12,013 visitors
1948—180 firms met 14,459 visitors
1949—225 firms met 15,710 visitors
1950—253 firms met 17,689 visitors
1951—277 firms met 22,919 visitors
(More than 300 firms in 1952)

When the manufacturer's engineer meets the buyer's engineer, problems are solved and sales are made. In a technical industry, only the engineer can "set the specs" and he is the man usually too busy to see a salesman. But at the Radio Engineering Show, he comes to see what's new.

Audience at Technical Papers Session

Main Floor Exhibits—1951

"Come Again"

The IRE National Convention and Radio Engineering Show

March 3-6, 1952, Waldorf Astoria Hotel and Grand Central Palace, New York

The Institute of Radio Engineers

1 East 79th Street, New York 21, N.Y.

Exhibits Dept., 303 West 42nd Street, New York 18

PROCEEDINGS OF THE I.R.E. December, 1951, Vol. 39, No. 12. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October, 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927. Printed in U.S.A.

Table of Contents will be found following page 64A

Meetings with Exhibits

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

△

IRE National Convention & Radio Engineering Show
March 3-6, 1952
Waldorf-Astoria Hotel and Grand Central Palace
New York City

△

NEREM
Saturday, May 10, 1952
Copeley Plaza Hotel
Boston, Mass.
New England Radio Engineering Meeting
Gen. Chairman: Alfred J. Pote
71 West Squantum St.
N. Quincy, Mass.

△

National Conference on Airborne Electronics
May 12, 13 & 14, 1952
Hotel Biltmore, Dayton, Ohio
Exhibits: Paul D. Hauser
1430 Gascho Drive, Dayton 3.

△

4th Southwestern IRE Conference
May 16, 17, 1952
Rice Hotel, Houston, Texas
Exhibits: Gerald L. K. Miller
1622 W. Alabama
Houston 6, Tex.

△

Western Electronic Show and IRE Regional Convention
August 27, 28 & 29, 1952
Long Beach, Calif.
Exhibits: Heckert Parker
215 American Avenue
Long Beach 2, California

△

National Electronic Conference
Sept. 29, 30, Oct. 1, 1952
Hotel Sherman, Chicago, Ill.

△

Cincinnati Convention on Television
April 19, 1952
Cincinnati, Ohio

You won't find today's most widely used military capacitors listed in JAN!

Joint Army and Navy component specifications were never meant to limit engineering progress—and, with Sprague, they most certainly haven't!

. . . Sprague subminiature capacitors have pioneered size and weight reductions plus high-temperature operation that would have been impossible with conventional capacitors. You'll not find these unique, hermetically-sealed capacitors listed in the current issue of Joint Army-Navy specification JAN-C-25. In every case, however, they have been fully approved for use. The reason is simple: In effect, Sprague subminiature capacitors are *super*-JAN types. They greatly exceed the already high minimum quality limits established by JAN specifications.

As the long-time leader in capacitor development, Sprague clearly recognizes that its engineering obligation extends far beyond conventional standards—*so markedly so that much of today's tremendous production of Sprague components for military use is based on types for which no JAN specifications yet exist!*

Thus, to equipment manufacturers faced with the problems of reducing size and weight or of paving the way to higher temperature operation, Sprague offers help along many lines—from the subminiature capacitors shown here to Vitamin Q☆ photo-flash capacitors to Ceroc☆ 250°C. ceramic-Teflon insulated magnet wire and many others.

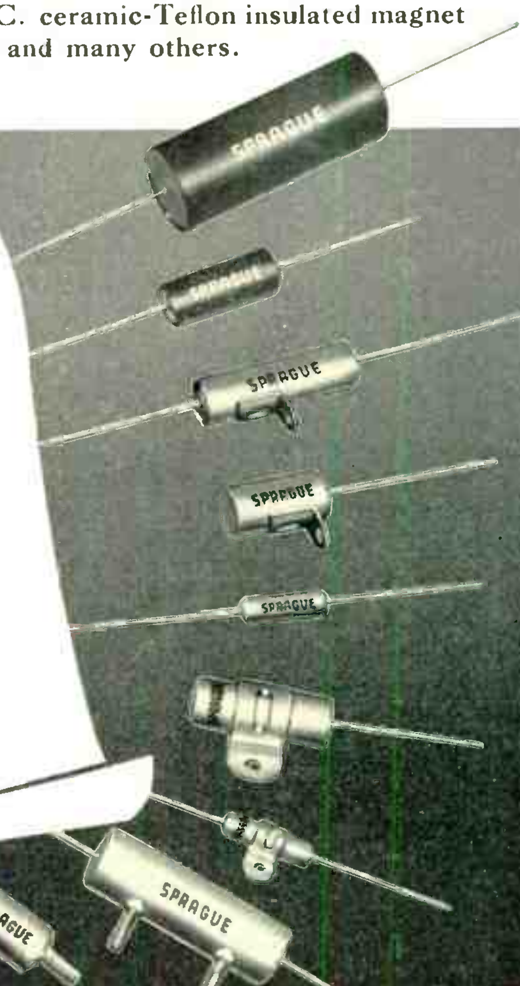
...opening new horizons to critical equipment design

You can do miniaturization jobs with these capacitors that were hitherto impossible!

Conservatively rated in designs for either 85°C. or 125°C. operation and with voltage ratings from 100 to 1000 volts, Sprague subminiature hermetically-sealed paper capacitors are available in physical sizes materially smaller than JAN types. Rigid mounting is greatly simplified by a variety of new mounting designs—also pioneered by Sprague.

For complete information write for Bulletin 213B.

☆Trademark

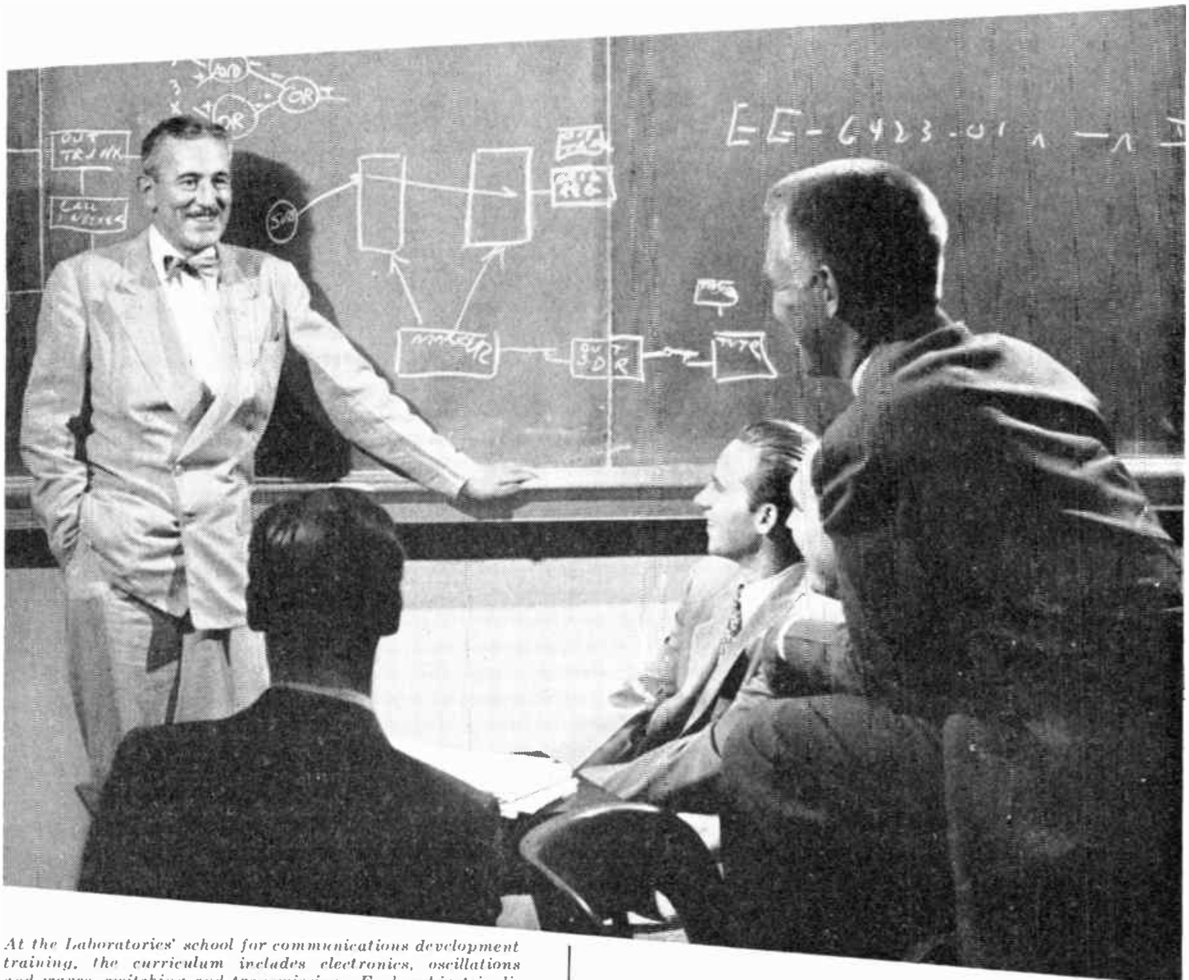


SPRAGUE

PIONEERS IN ELECTRIC AND ELECTRONIC DEVELOPMENT

SPRAGUE ELECTRIC COMPANY • NORTH ADAMS, MASSACHUSETTS

They're headed for new frontiers



At the Laboratories' school for communications development training, the curriculum includes electronics, oscillations and waves, switching and transmission. Each subject is directly keyed to the latest fields of telephone research.

EACH year the Bell System selects hundreds of engineering graduates from technical schools, to find the answers to communications problems through the application of science and technology. A specifically qualified group joins Bell Laboratories to develop *tomorrow's* telephone system — also, in the present emergency, more powerful electronic devices for the armed services.

They come — thanks to the competence of our nation's educators—with an excellent grounding in fundamentals. To equip them still further, the Laboratories operate a school at graduate level for advanced communications.

The new men receive an intensive course in the latest theory and techniques. At the same time they take their places as members of the Technical Staff doing responsible work which, with their classroom instruction, reveals where they can make the most of their aptitudes.

More than ever America's future must depend on men and women who are trained to think far ahead in technology whether for tomorrow's telephones or national defense. By helping them, Bell Telephone Laboratories help make America's telephone system the world's best, help the armed forces keep our country strong.

BELL TELEPHONE LABORATORIES

Exploring and inventing, devising and perfecting, for continued improvements and economies in telephone service.

GENERATORS

10 to 7600 mc

10 to 500 mc *-hp-* 608A SIGNAL GENERATOR

Output range 0.1 μv. to 1.0 v. into 50 ohms. Accuracy ± 1 db. Direct reading frequency and output calibration, no charts or interpolation. CW, pulsed or amplitude modulated output (50 to 1,000,000 cps). Resetability better than 1 mc. Master oscillator power amplifier for widest modulation capabilities. Constant internal impedance 50 ohms. Maximum VSWR 1.2. $850 f.o.b. factory.

-hp- Model 608A

450 to 1,200 mc *-hp-* 610B SIGNAL GENERATOR

Output range 0.1 μv. to 0.1 v. into 50 ohms. Accuracy ± 1 db. Output and frequency directly set and read, no charts or interpolation. Modulation: internal or external pulsed, external amplitude, external square wave. Widely variable pulse length, repetition, and delay features. $925 f.o.b. factory.

-hp- Model 610B

800 to 2,100 mc *-hp-* 614A SIGNAL GENERATOR

Output range 0.1 μv. to 0.223 v. (1 mw). Accuracy ± 1 db. Single dial direct reading frequency and output, no charts or interpolation. CW, pulsed and FM output. Modulation: internal pulsed, FM, external pulsed. Widely variable pulsing, synchronizing, delay and triggering features. Extremely fast rise/decay time 0.1 μsec. Constant internal impedance 50 ohms, SWR 3 db. $1,950 f.o.b. factory.

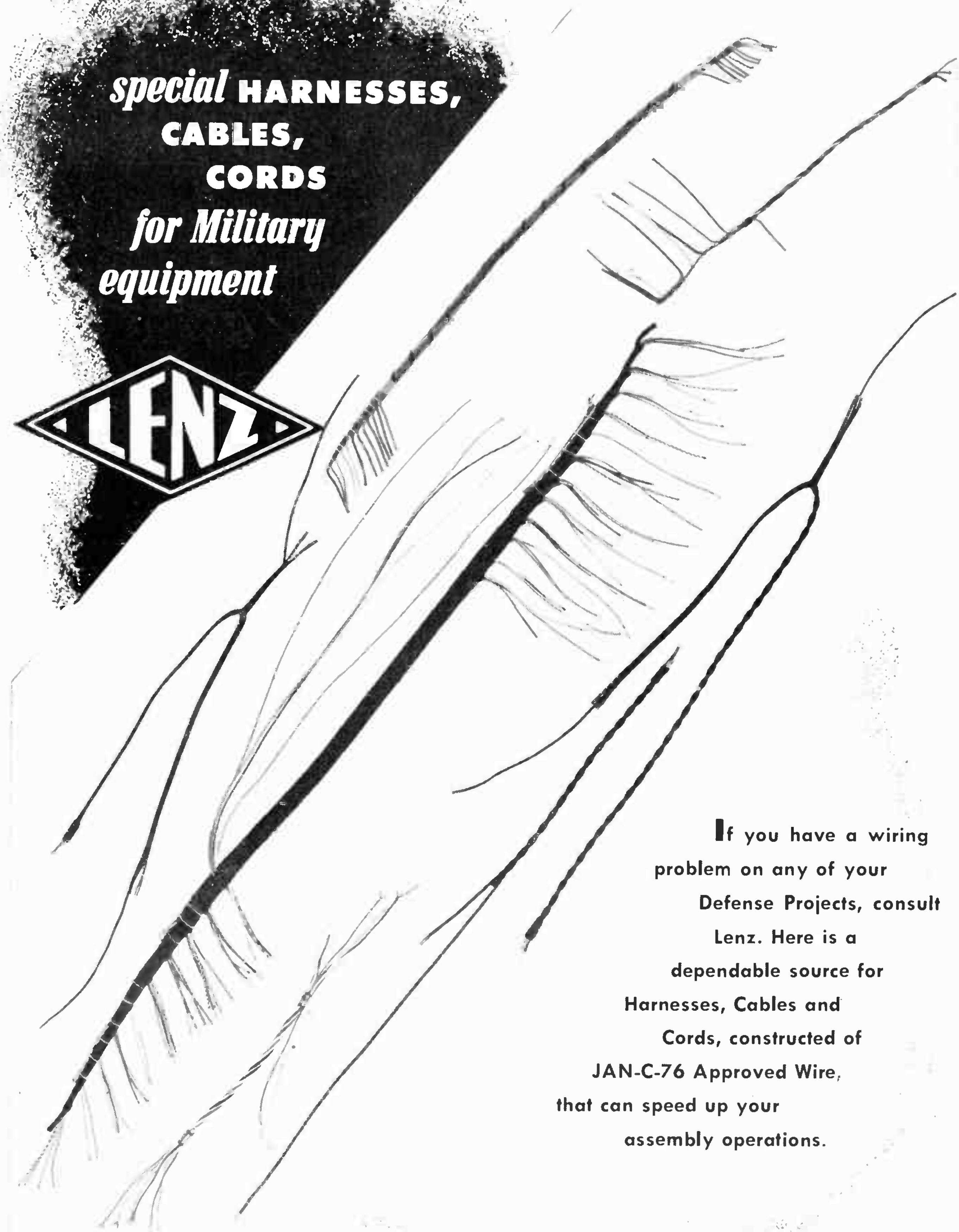
-hp- Model 614A

1,800 to 4,000 mc *-hp-* 616A SIGNAL GENERATOR

Output range 0.1 μv. to 0.223 v. (1 mw). Accuracy ± 1 db. Single dial direct reading frequency and output, no charts or interpolation. Output, modulation, and synchronization features identical with Model 614A. Like Model 614A, instrument automatically tracks frequency changes, requires no voltage adjustment during operation. $1950 f.o.b. factory.

-hp- Model 616A

HEWLETT hp PACKARD



special **HARNESSES,**
CABLES,
CORDS
for Military
equipment

LENZ

If you have a wiring problem on any of your Defense Projects, consult Lenz. Here is a dependable source for Harnesses, Cables and Cords, constructed of JAN-C-76 Approved Wire, that can speed up your assembly operations.

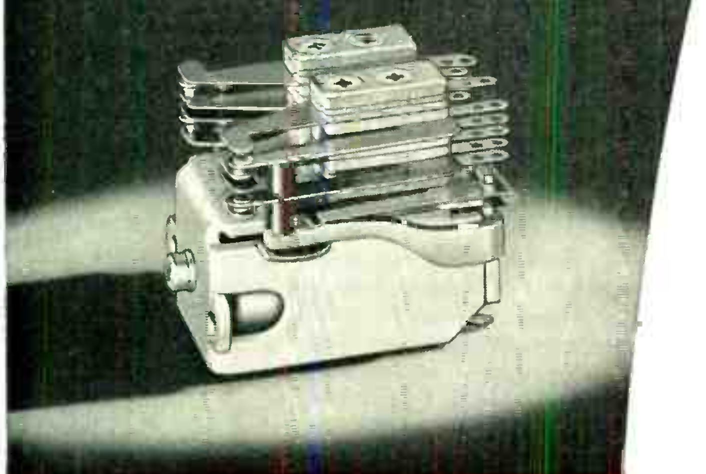
LENZ ELECTRIC MANUFACTURING CO.

IN BUSINESS SINCE 1904 • 1751 North Western Avenue, Chicago 47, Illinois

THE *New* CLARE TYPE "N" RELAY...

a small, highly sensitive relay

designed for efficient operation on low power*



CLARE Type "N" RELAY

CLARE Type "N" Relays are designed for operation on extremely low power. A close-coupled magnetic circuit, generous use of magnetic iron, and unusually efficient coil design give high sensitivity while retaining high contact pressure (minimum 30 grams) and adequate contact gap (minimum 0.015").

Other important advantages include small size, light weight, and especial adaptability to hermetic sealing. Type "N" Relays having not more than 14 terminals for coil and contact springs can be hermetically sealed in enclosures of extremely small size.

For more detailed information on Clare Type "N" Relays you are invited to write for Bulletin No. 109. Clare sales engineers are located in principal cities. Call them or write C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials, Ltd., Toronto 13. Cable Address: CLARELAY.

> *With a 10,000-ohm coil, 1 Form C contact (spdt), and a standard adjustment, this relay will operate on less than 50 milliwatts. With a 450-ohm coil and four Form C contacts (4pdt), it will operate on 7/10 watt, even under conditions of vibration and high ambient temperature.

SPECIFICATIONS of CLARE Type "N" RELAY

COIL
Single- or double-wound. Double wound coils are concentric and have four terminals, two for each winding.

OPERATING VOLTAGE
Up to 180 volts dc.

ARMATURE
Single- or double-arm. Armature has high lever ratio (approx. 3:1) and no air gap.

CONTACT ASSEMBLY
Forms A to C. Maximum of 10 springs per pileup.

CONTACTS
Code No. 4 standard unless load conditions require different contact material.

TERMINALS
All located at armature end of relay.

DIMENSIONS
Length: [illegible]; Width: [illegible]; Height: Min. (with two contact springs): [illegible]; Max. (with ten contact springs): [illegible]

NET WEIGHT
Single-arm, 2½ oz. (approx.); double-arm, 3 oz. (approx.).

MOUNTING
Two #5-40 tapped holes on [illegible]" centers, located so mounting screws cannot foul coil.

Clare Type "SN" Hermetically Sealed Relay

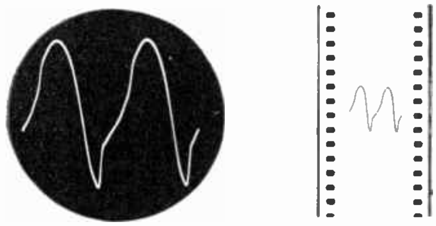
Shown above is one of the hermetically sealed steel enclosures in which the Type "N" Relay can be sealed. Dimensions are: Length: 1-7/16"; Height: 2-1/16"; Width: 1⅝". Net weight of relay having 12 contact springs, six in each pileup, is 5 oz. (approx.). Note connection diagram clearly and permanently imprinted on base of enclosure by silk screen process.

Write for CLARE Bulletin No. 109

CLARE RELAYS

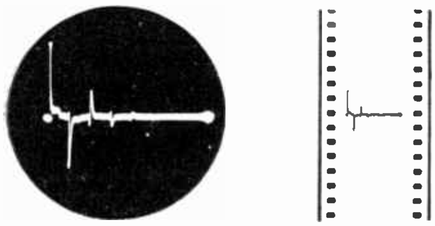
First in the Industrial Field

How much can you expect an oscilloscope camera to do?



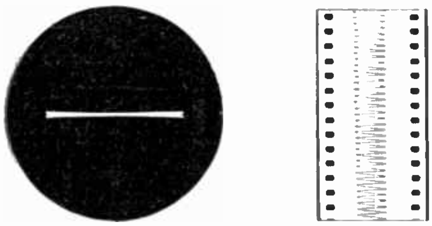
Scope Image | Film Recording

1. Single-frame photography of stationary patterns using a continuously running sweep.



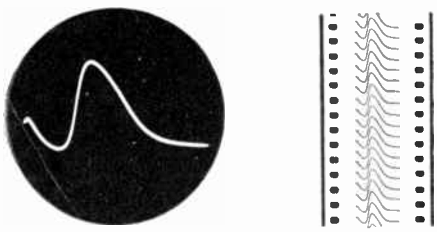
Scope Image | Film Recording

2. Single-frame photography of single transients using a single sweep.



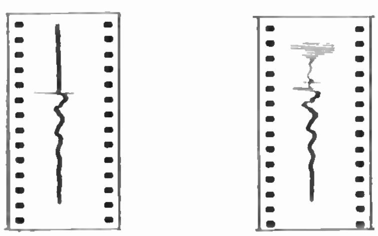
Scope Image | Film Recording

3. Continuous-motion photography employing film motion as a time base.



Scope Image | Film Recording

4. Continuous-motion photography employing oscilloscope sweep as a time base.



FILM MOTION TIME BASE | FILM MOTION & SCOPE SWEEP

5. Continuous-motion photography employing combination of film motion and oscilloscope sweep as a time base.

It's only reasonable that you should expect the oscilloscope camera you buy to record what you see on an oscilloscope screen during any period. But can it be expected to do any more? We think so.

For example, did you know that the *Fairchild Oscillo-Record Camera*—our idea of the most versatile 35-millimeter oscilloscope camera now available—can GREATLY EXTEND THE USEFULNESS OF YOUR OSCILLOSCOPE?

As you know, many non-recurring phenomena occur too rapidly to permit adequate visual study. Others occur so slowly that continuity is lost. Sometimes you have combinations of very slow-speed phenomena and occasional high-speed transients. In any one of these cases, the Fairchild Oscillo-Record Camera will take over where your eye and the oscilloscope leave off.

This extremely versatile instrument is now being used daily by many hundreds of engineers in widely divergent fields. For an idea of what it can do for you, study the five scope images and recordings illustrated at left. Each solves a particular problem.

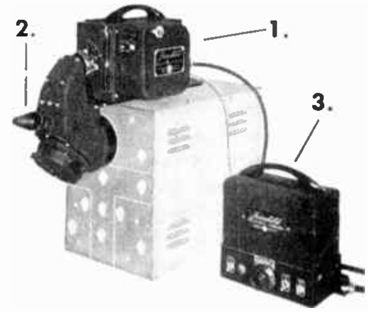
Oscillo-Record users especially like its:

CONTINUOUSLY VARIABLE SPEED CONTROL—1 in./min. to 3600 in./min.

TOP OF SCOPE MOUNTING that leaves controls easily accessible.

PROVISION FOR 3 LENGTHS OF FILM—100, 400, or 1000 feet.

For more data write *Fairchild Camera Instrument Corp., 88-06 Van Wyck Blvd., Jamaica 1, N.Y. Dept. 120-16CI.*

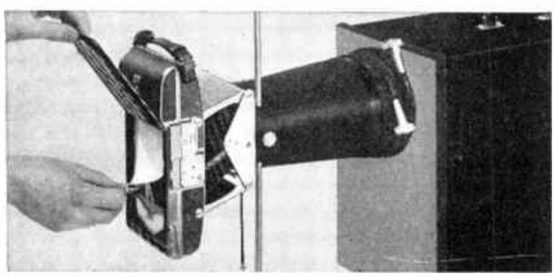


FAIRCHILD OSCILLO-RECORD CAMERA — 1. camera, 2. periscope, 3. electronic control unit. *Available accessories include external 400 and 1000 foot magazines, magazine adaptor and motor, universal mount for camera and periscope, binocular split-beam viewer.*

VALUABLE RECORDS FOR IMMEDIATE EVALUATION

The Fairchild-Polaroid® Oscilloscope Camera produces a photographic print in a minute

Valuable but inexpensive oscillograms for immediate evaluation; automatic one-minute processing without a darkroom; a set up time of two minutes or less—they're just three of the many advantages that are yours when you use the Fairchild-Polaroid Oscilloscope Camera. Wherever individual exposures meet your recording requirements—where you'd like to have permanent records of the traces you're now sketching or carrying in your memory, this is the camera that can bring new speed, ease and economy to your job. Prints are 3¼x4¾ and each records two traces exactly one-half life size. Write today for details.



A minute after you've pulled the tab a finished print is ready for evaluation

FAIRCHILD

OSCILLOSCOPE RECORDING CAMERAS

PRODUCTS OF WORLD-FAMOUS

DAVEN

CREATIVE-ENGINEERING

TYPE T-330

DAVEN ATTENUATORS

Exclusive Features:

1. "Knee-Action" Blade
   - Tamper proof
   - Uniform contact pressure and low contact resistance over the life of every unit.
   - Each rotor blade individually supported to give positive contact in operation under all types of conditions.
2. Brass Case of 2-Piece Construction
   - Greatly reduces clearance space required for removal of cover in rear of unit.
3. "Lock-Tite" Dust Cover
   - Held by positive, bayonet-type lock which prevents cover from falling off under stress of vibration.
4. Enclosed Roller-Type Detent Mechanism
   - For extra long life and positive indexing.
   - Addition of detent does not increase depth of unit.
5. Low-Loss Molded Terminal Board
   - For high resistance to leakage.
6. Rigidly Supported Resistor Strips
   - With air insulation.

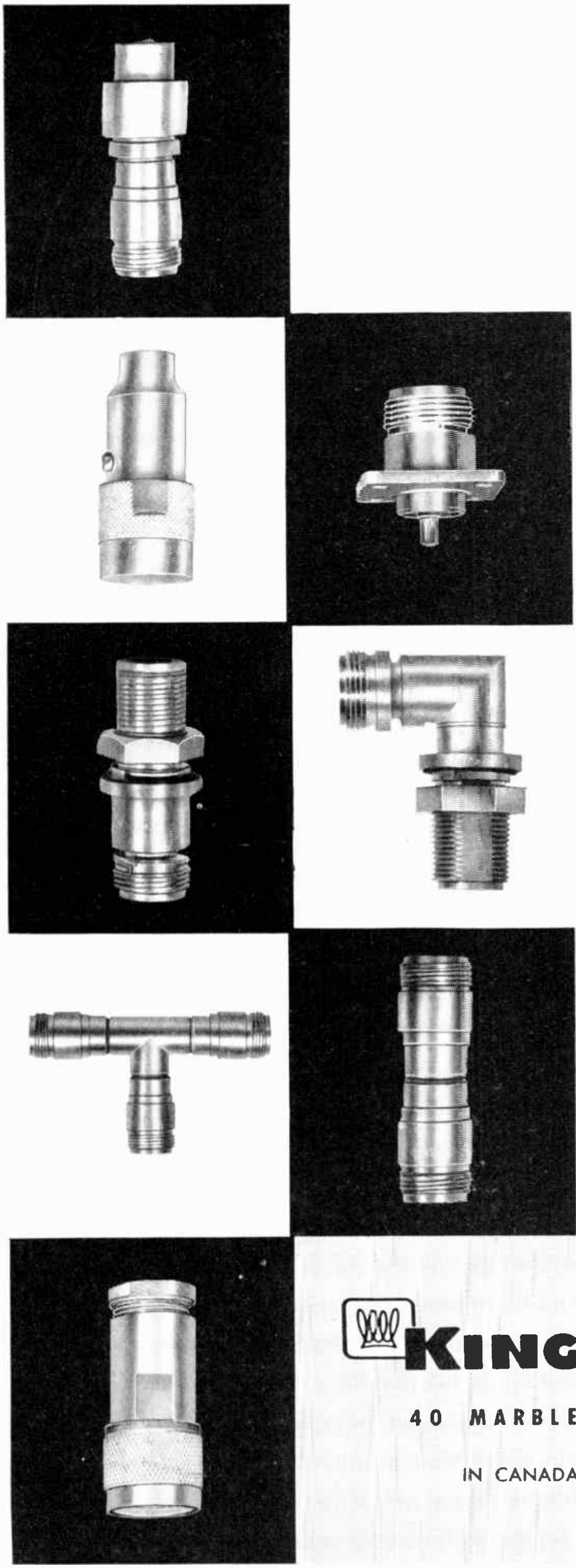
YEARS AHEAD, CREATIVE-ENGINEERING has always characterized Daven's efforts in the production of electronic components and equipment.

The unit illustrated above is only one in a complete line of Daven's world-famous attenuators. Because this Company has pioneered and created so many worth-while improvements in units of this type, you can do no better than to specify Daven Attenuators.

Daven's skilled staff of specialist-engineers is at your "beck and call" to help with your problems in the selection of the right attenuators for the equipment you are designing.

Write for complete catalog data.

THE DAVEN CO.
195 CENTRAL AVENUE
NEWARK 4, NEW JERSEY



"N" SERIES R.F. CONNECTORS BY KINGS

Shown are a few of the "N" Series R.F. Connectors made by Kings. These low voltage connectors are of constant impedance and come in both weather-proof and non-weather-proof types.

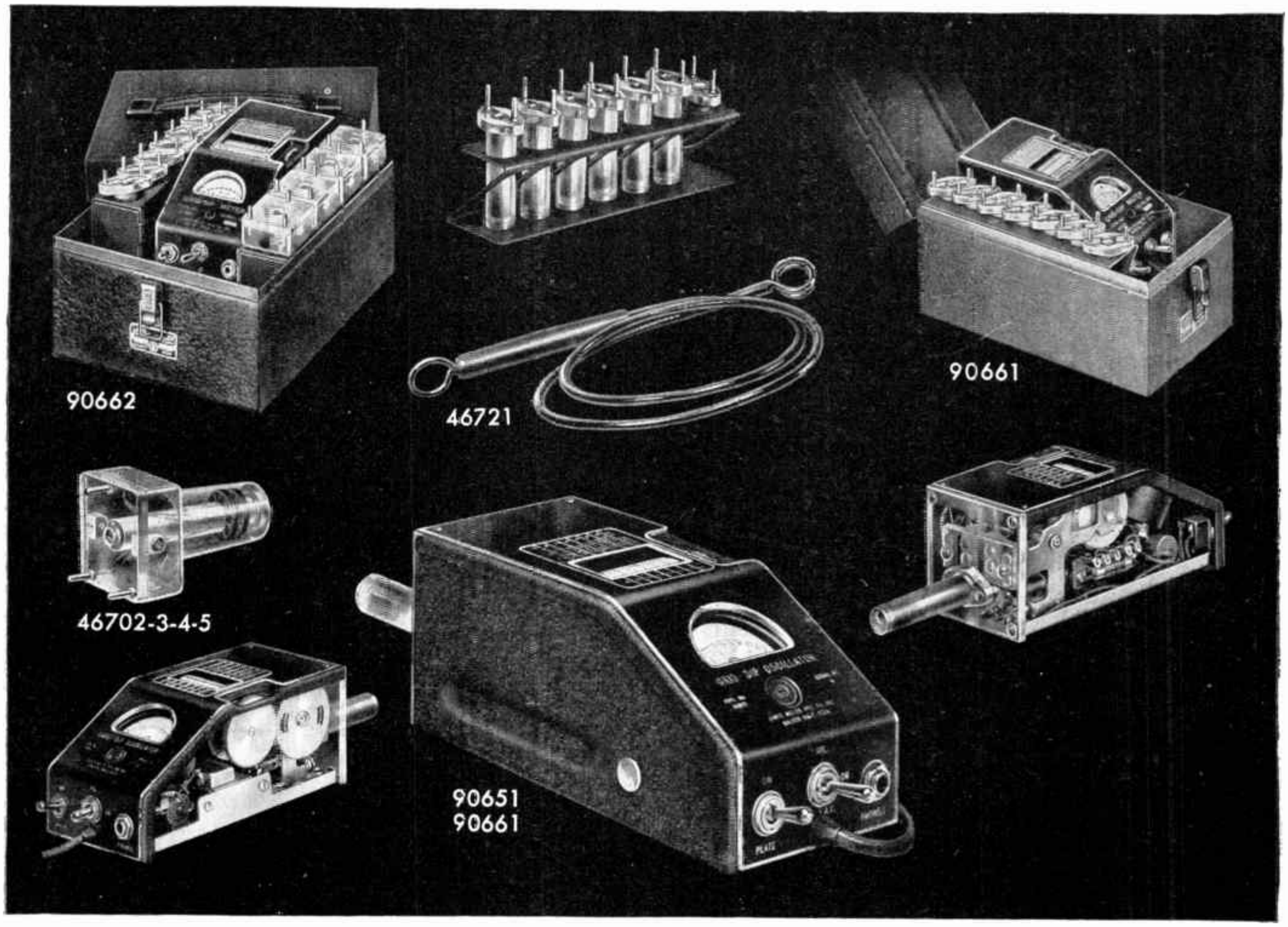
Electronics engineers look to Kings for Connectors. A valued recognition which has been earned by many years of specialized work in this field. When you call on Kings you get the benefit of years of engineering, research and production experience and know-how.

You are invited to write for quotations and delivery dates on all standard and special connectors.

KINGS Electronics CO., INC.

40 MARBLEDALE ROAD, TUCKAHOE, N. Y.

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



Designed for Application

Grid Dip Meters

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

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

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

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

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

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

The Grid Dip Meters may be used as:

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

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

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

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.


125 SERIES
175 SERIES
225 SERIES
90 SERIES
5 BASIC SIZES
70 SERIES
STANDARD CONTROL KNOBS by RAYTHEON
Excellence in Electronics
Now, for the first time . . . standard control knobs worthy of the finest commercial or military equipment . . . available in a complete choice of widely used sizes and functional styles —without the heavy expense of custom designing and tooling.
Made of tough, durable "Tenite II" (cellulose acetate butyrate) with dual setscrews and anodized aluminum inserts drilled for ¼-inch diameter shafts. Miniature (70 Series) round, dial and pointer knobs also available for ⅛-inch shafts.
Black knobs can be furnished with high gloss mirror finish for commercial applications or with non-reflecting matte finish for use on government equipment. Manufacturers interested in colored knobs to match, blend, or contrast with color styled equipment are invited to submit their requirements for quotation.
ALL IN MATTE OR MIRROR FINISH
Write for Complete Information
Address Dept. 6470-KP
RAYTHEON MANUFACTURING COMPANY
Equipment Sales Division — Waltham 54, Mass.
ROUND
SKIRTED ROUND
6 FUNCTIONAL STYLES
POINTER
SKIRTED POINTER
DIAL SKIRTED ROUND
CRANK

FUSITE MULTIPLE TERMINAL PANELS

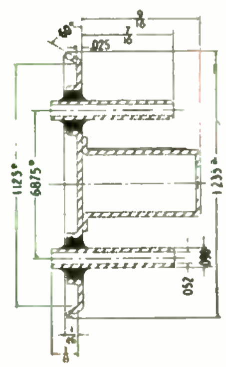
for Plug-in Applications of

HERMETICALLY SEALED ELECTRICAL COMPONENTS

Plug-in type of terminals are becoming increasingly popular because of the easy, fool-proof nature of their use. The tongued centering key fits a correspondingly grooved socket. FUSITE offers a large variety of this style terminal ranging from 2 to 20 electrodes and in several flange sizes and treatments. All terminals are regular FUSITE glass-to-steel construction. Sockets are available for all FUSITE Plug-in terminals.

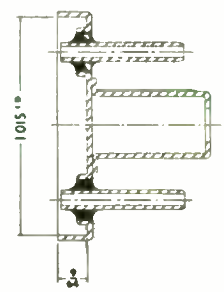
7-900 HTO Series
2000 V (RMS)
Available 2 to 8 hollow tube electrodes.

Available Flange Treatments



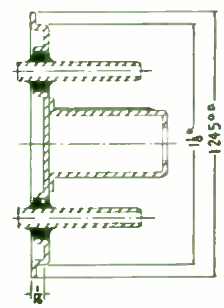
TYPE 1

7-900 HTO Series with standard 60° solder flange.



TYPE 2

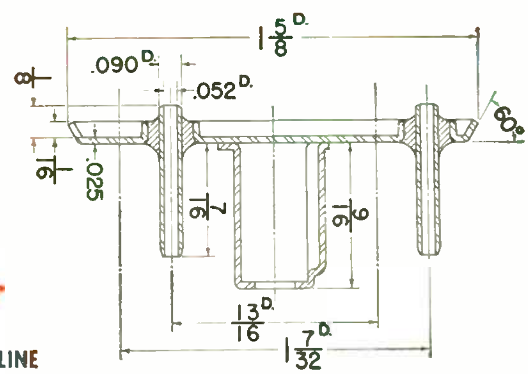
7-900 HTO Series with special 90° solder flange. Caps 1" O.D. can. Also available as Type 2B to cap 1.540" O.D. can. and Type 2C to cap 1 5/8" O.D. can.



TYPE 3

7-900 HTO Series with special offset flat solder flange.

7-2300 HTO Series
2000 V (RMS)
Available 11 to 20 hollow tube electrodes.



Write for Catalog

OF COMPLETE LINE AND ENGINEERING DETAILS -- DEPT. A

INSULATE ELECTRICALLY GLASS TO METAL FUSITE TERMINALS PROTECT PRODUCT PERFORMANCE SEAL HERMETICALLY

THE **FUSITE** CORPORATION

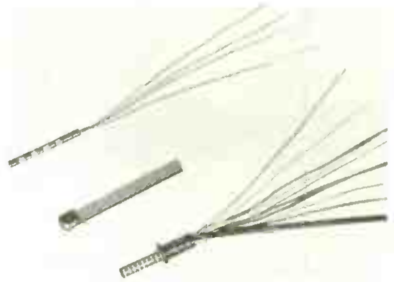
6028 FERNVIEW AVENUE - CINCINNATI 13, OHIO

News—New Products

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

Sub-Miniature Sliprings

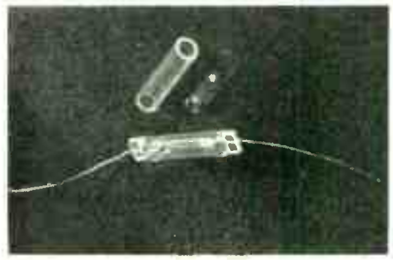
Naer Corp., 631 S. Sepulveda Blvd., W. Los Angeles 49, Calif., has produced a new line of sliprings. These rings are fabricated with Formex magnet wire because of the physical properties of its insulating film, plus a coating of NAER L-45 insulation, making them practical for use in all communications, gear and industrial electronic equipment. It has flexibility, exceptional dielectric strength, heat resistance, abrasion and shock resistance.



These sliprings are molded with a special compound NC-101 which has a Tensile strength, psi 4,000–4,400; flexural strength, psi 6,000–6,500; Rockwell Hardness (M scale) 27.0; heat distortion temperature of 225–235°F. Special molding forms eliminate shrinking, swelling, and temperature effects.

Nylon Tubing for Thermistors

Anchor Plastics Co., Inc., 533–541 Canal St., New York 13, N. Y., announces that extruded Nylon tubing about the diameter of a lead pencil is being employed to encase one type of glass-enclosed bead thermistor used in time-delay circuits.



The thermistor is inserted in a length of tubing and the ends of the tubing pressed flat under heat in an hydraulic press for a given length of time. This gives Nylon the "set" desired.

The tubing provides mechanical protection of the glass bulb, electrical insulation (no metal end caps are required), and protects against straining the glass bulb through flexure of the connecting leads. Nylon readily takes the shape desired and is heat resistant.

Heat resistance is a factor because the connecting leads are soldered during assembly into circuits and the apparatus in which the thermistors are used must operate properly over a wide range of temperatures.

(Continued on page 28A)

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 . . . CHEMICAL and CERAMIC ENGINEERS . . . 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 Magnetrons
- TRANSFORMER and COIL DESIGN
- COMMUNICATIONS— Microwave, Mobile, Aviation, Specialized Military Systems
- RADAR— Circuitry, Antenna Design, Computer, Servo-Systems, Information Display Systems
- INDUSTRIAL ELECTRONICS— Precision Instruments, Digital Circuitry, Magnetic Recording, Industrial Television, Color Measurements
- NAVIGATIONAL AIDS
- TECHNICAL SALES
- ELECTRONIC EQUIPMENT FIELD SERVICE

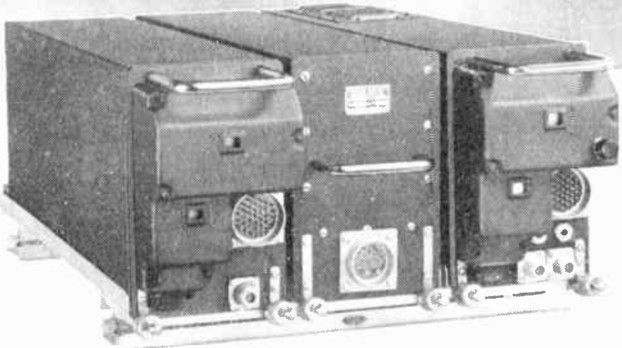
Whatever your plans for the future—you will find the booklet "The Role of the Engineer in RCA" interesting reading. Write for your *free* copy.

MAIL RESUMÉ

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

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

RADIO CORPORATION of AMERICA



WILCOX ...Choice of EASTERN Air Lines

180 Channel WILCOX Communications System Chosen for Eastern's Entire Fleet of SUPER CONSTELLATIONS and MARTIN 4-0-4's

Eastern Air Lines demanded the finest communications equipment available to match the advanced, efficient operation of their modern new fleet. No greater compliment could be paid to Wilcox radio equipment than to be selected for this challenging assignment.

The Wilcox 440A VHF Communications System covers all channels in the 118-136 Mc. band. It is light in weight, small in size, and easy to maintain.

UNIT CONSTRUCTION FOR EASY HANDLING

The 50-watt transmitter, high sensitivity receiver, and compact power supply are each contained in a separate JAN A1-D case. Any unit may be instantly removed frcm the common mount.

FINGER-TIP REMOTE CONTROL

All transmitter and receiver functions are available by remote control. A new channel selector system assures positive operation and minimum maintenance.

DEPENDABILITY AND EASY MAINTENANCE

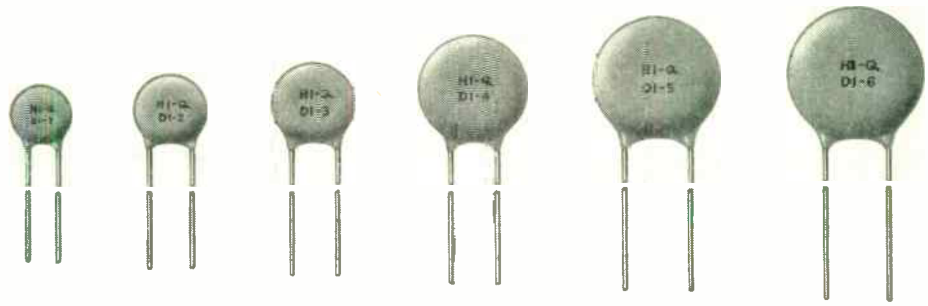
Simple, conventional circuits minimize the number and types of tubes and require no special training, techniques, or test equipment.

Write Today FOR COMPLETE INFORMATION ON THE WILCOX 440A 180 CHANNEL VHF COMMUNICATIONS SYSTEM

WILCOX ELECTRIC COMPANY

FOURTEENTH AND CHESTNUT KANSAS CITY 1, MISSOURI, U.S.A.

NEW from HI-Q*



DI-1 DI-2 DI-3 DI-4 DI-5 DI-6

Illustrations approximately actual size.

Temperature Compensating DISK Capacitors

Capacity range from 475 mmf on the DI-6 N1400 material down to .3 mmf on the DI-1 size with tolerances of ± 5% or greater. Conservatively rated for working voltage at 500 volts DC and flash tested at 1500 volts DC. Insulation resistance at 100 volts is well over 10,000 megohms. Electrodes are fired directly to the low loss dielectric and are coated with a non-hydroscopic phenolic for protection against moisture and high humidities. Conform to RTMA Class I ceramic capacitors.

Extended Temperature Compensating DISK Capacitors

Produced from a recently developed group of extended coefficient ceramics, this type of HI-Q Disk permits a much wider temperature compensating range than was possible on the formerly available normal linear temperature coefficient ceramics. Specifically developed for applications requiring a very large gradient of capacity versus temperature. These new HI-Q Disks exhibit relatively higher dielectric constants permitting capacities in the range intermediate between the high K and linear or normal group of ceramics. The Q (a minimum of 250 at 1 megacycle) is somewhat lower than the Class 1 ceramics. It has, therefore, not been classified by RTMA as Class 1. However, characteristics are superior to by-pass Class 2 ceramics.

ALL HI-Q DISK CAPACITORS COME IN THESE SIX SIZES

Type	Diameter	Lead Width	Thickness
DI-1	5/16″ Max.	3/16″ ± 1/16″	5/32″ Max.
DI-2	3/8″ Max.	1/4″ ± 1/16″	5/32″ Max.
DI-3	7/16″ Max.	1/4″ +1/8″ −0″	5/32″ Max.
DI-4	19/32″ Max.	1/4″ +1/8″ −0″	5/32″ Max.
DI-5	11/16″ Max.	3/8″ ± 1/8″	5/32″ Max.
DI-6	3/4″ Max.	3/8″ ± 1/8″	5/32″ Max.

Companion Lines to the Popular HI-Q By-pass DISK Capacitors

The widely used HI-Q By-pass Disks are fixed ceramic dielectric capacitors which meet RTMA Class 2 specifications. They are available in the complete capacity range of from .3 mmf to 30,000 mmf. Standard tolerances of 5% thru 20% where applicable can be furnished.

Write for Engineering Bulletin Giving Details of all HI-Q DISK Capacitors

HI-Q*

** Trade Mark Registered U. S. Patent Office*

Electrical Reactance Corp.
OLEAN, N. Y.

SALES OFFICES: New York, Philadelphia, Detroit, Chicago, Los Angeles

PLANTS: Olean, N. Y., Franklinville, N. Y. Jessup, Pa., Myrtle Beach, S. C.

News—New Products

(Continued from page 24A)

Pulsed Carrier Generator

A new pulsed-carrier generator called the RADA-PULSER is offered by **Kay Electric Co.**, 14 Maple Ave., Pine Brook, N. J., to give transient response information in laboratories, on production lines, and in the field. The instrument is engineered for military as well as commercial usage.

The Rada-Pulser specifications are as follows: carrier frequencies; 30 and 60 mc; pulse widths; 0.1 and 0.25 microseconds; pulse repetition rate; continuously variable from 500 to 2,000 pps; Maximum rf output, approximately 1 volt at 70 ohms; Attenuators; 20 db, 20 db, 10 db switched 10 db continuously variable.

Pulse output is 50 volts at 70 ohms. A jack is provided to permit use of envelope pulses. External modulation: Input terminals provided to permit modulation by other pulse widths from external source. Trigger pulses: Positive and negative furnished ahead of pulsed carrier to trigger oscilloscope sweep circuit. Regulated power supply is built in.

Price: $595.00, F.O.B. factory. Write to manufacturer for additional information and latest catalog.

Molded-In Selenium Rectifiers

Electronic Devices, Inc., Precision Rectifier Div., 429 12th St., Brooklyn 15, N. Y., announces that all ratings up to 200 ma dc output in the Plastisel line of miniature selenium rectifiers are molded in, similar to small tubular capacitors. The outer case is spiral-wound phenolic tubing filled with wax which shows no sign of softening at 100°C. The thermal conductivity of this wax and the low loss plates compensate for the loss of cooling due to molding-in. These rectifiers are manufactured with bare or insulated tin-copper leads

In ratings from 250 to 500 ma dc the standard open-plate construction is used. However, the plates lead to cooler operation and longer life. All open and closed construction stacks are standard in 380 PIU ratings (130 volts ac into a capacitor). They can also be had in other voltage ratings and as doubler units for special applications. All Plastisel rectifiers are guaranteed for 1,000 hours or 1 year, whichever occurs first.

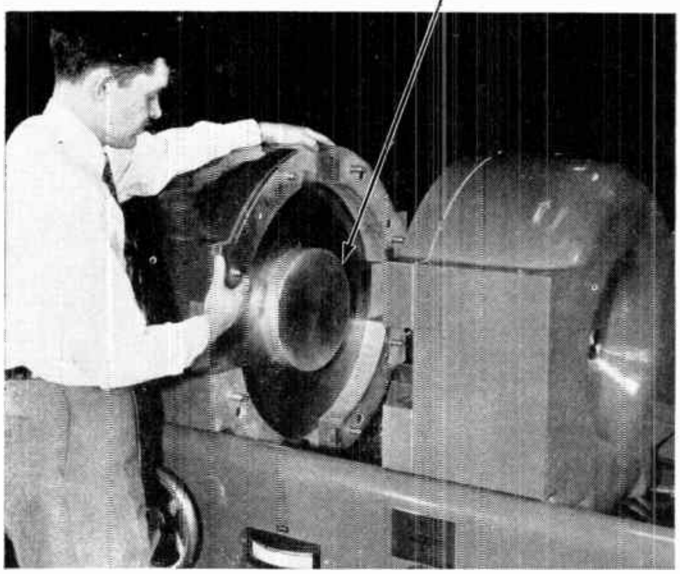
(Continued on pgae 56A)

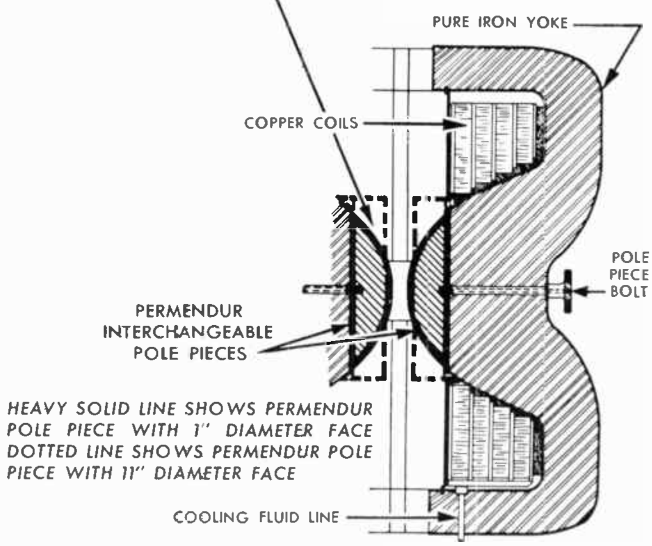
PERMENDUR*

gives you

- ☆ High Magnetic Saturation
- ☆ High Permeability at very high Flux Densities
- ☆ High Value of Positive Magnetostriction
- ☆ Design Possibilities for Savings in Weight, Space and Materials

Typical use of Permendur in the ADL ELECTROMAGNET
(ARTHUR D. LITTLE, Inc.)





Sectional Detail of Yoke Half

Permendur is available in Forgings, Castings, Hot-Rolled Bars and Plates to meet your design needs for form or shape. *Write for information*

*Manufactured under license arrangements with Western Electric Co.

W&D 3929

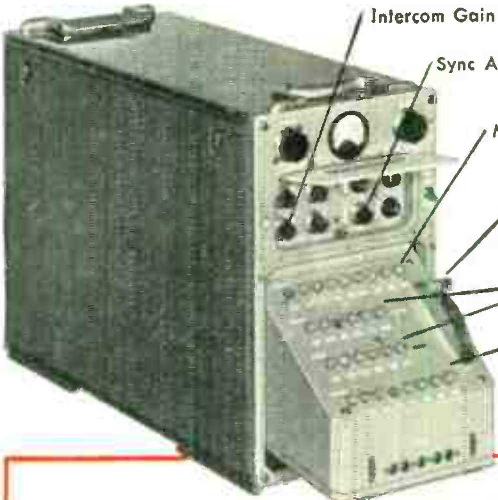
THE ARNOLD ENGINEERING COMPANY

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

General Office & Plant: Marengo, Illinois

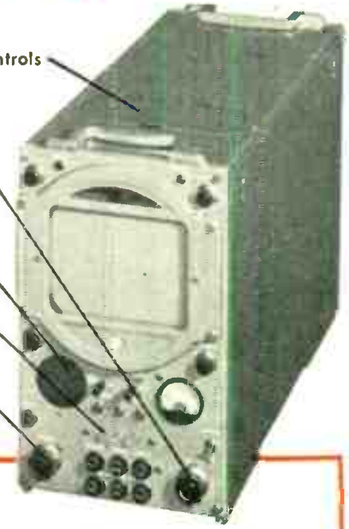
TWO CHAMPIONS

THAT MAKE AN *Unbeatable Team!*



GPL VIDEO SWITCHER

- Studio switching flexibility anywhere
- Panel and active buttons internally illuminated
- Portable, self-contained — Panel enclosed for transit
- Monitor views 5 camera inputs, 2 remotes, outgoing line
- Sound interlock switching for remotes and 2 cameras
- Two open panel positions, 90° and 120°
- Switch panel removable, operable to 5 feet
- Twin fading levers for fades, dissolves
- Two "effects" buses
- Styled to match all GPL TV equipment



GPL MASTER MONITOR

- Selection of 3 pre-set inputs
- 8½" Monitoring tube
- 3" Oscilloscope, also providing test facilities
- Meter readings of line voltage and power supply outputs
- Fast sweep for observing vertical sync block
- Quick-reference calibration voltage button
- Automatic sync of oscilloscope and kinescope sweeps at half-line or half-field frequency
- Regulated pulse high voltage supply isolated from sweep circuits
- High impedance bridging input
- Compact, portable
- Ready accessibility of all controls, tubes, circuits

For the new station, for the expanding station, GPL's champion team of Video Switcher and Master Monitor affords a new high in quality, in field flexibility, in rehearsal and programming control. Both units are packaged for easy portability, with self-contained power supplies. Either can be integrated into your present installation, can accommodate your particular operating conditions.

The Switcher and Monitor team is another example of GPL's unique achievement in the production of *high quality, high utility* equipment for TV stations — another reason why GPL *is* THE INDUSTRY'S LEADING LINE — IN QUALITY, IN DESIGN.

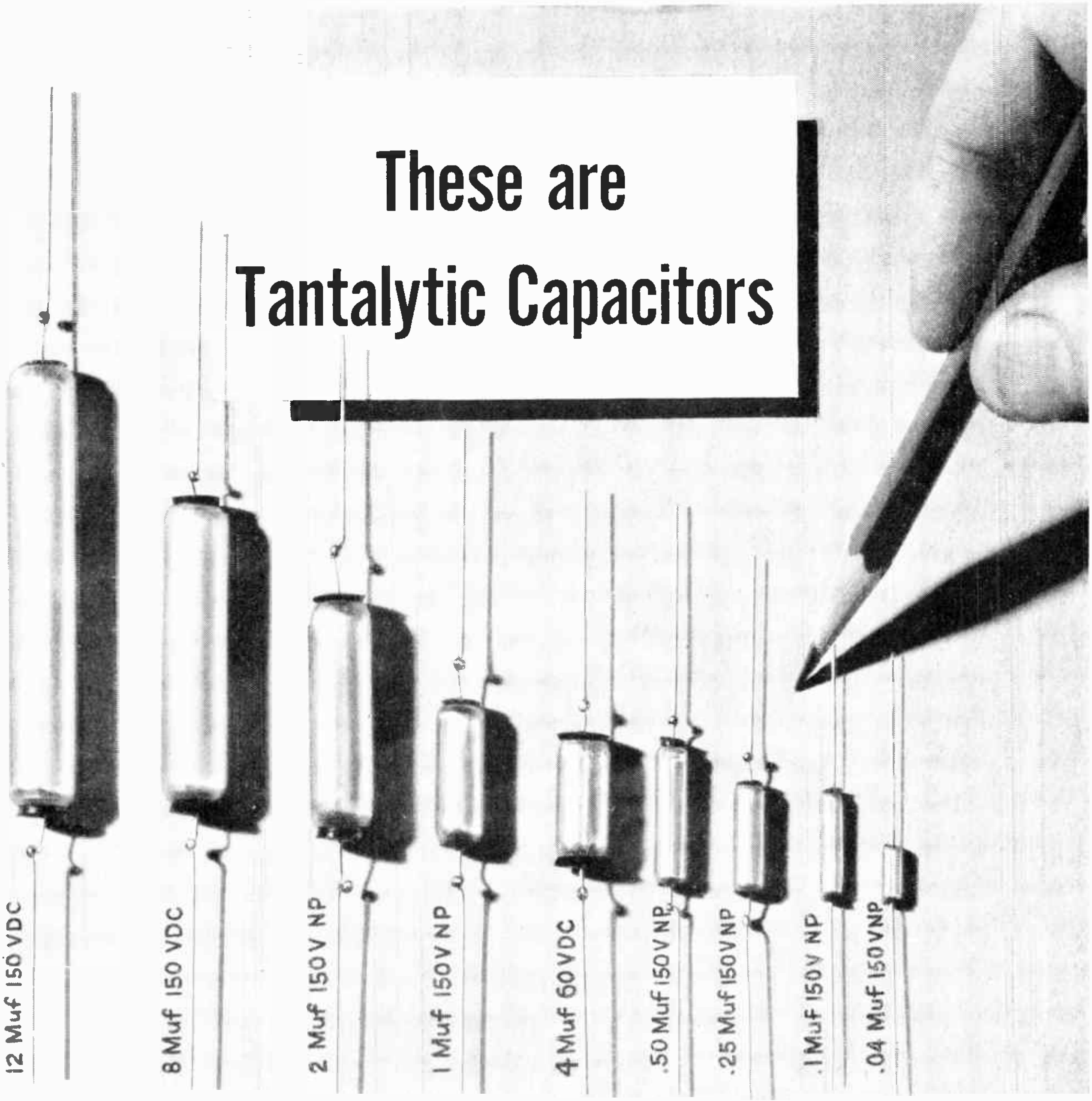
WRITE, WIRE or PHONE FOR DETAILS TODAY

GENERAL PRECISION LABORATORY INCORPORATED

GENERAL PRECISION LABORATORY
INCORPORATED

Pleasantville New York

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



Here is one of the fastest moving developments in recent years—General Electric's new electrolytic-type capacitors. These Tantalytic capacitors with their small size and large capacitance per unit of volume have excellent low temperature characteristics, long operating life and in many cases can replace bulky hermetically-sealed paper capacitors. Ratings presently available for consideration range from .02 muf up to 12 muf at 150 volts dc. Units pictured are representative of these ratings.

Other features of G-E Tantalytic Capacitors include:

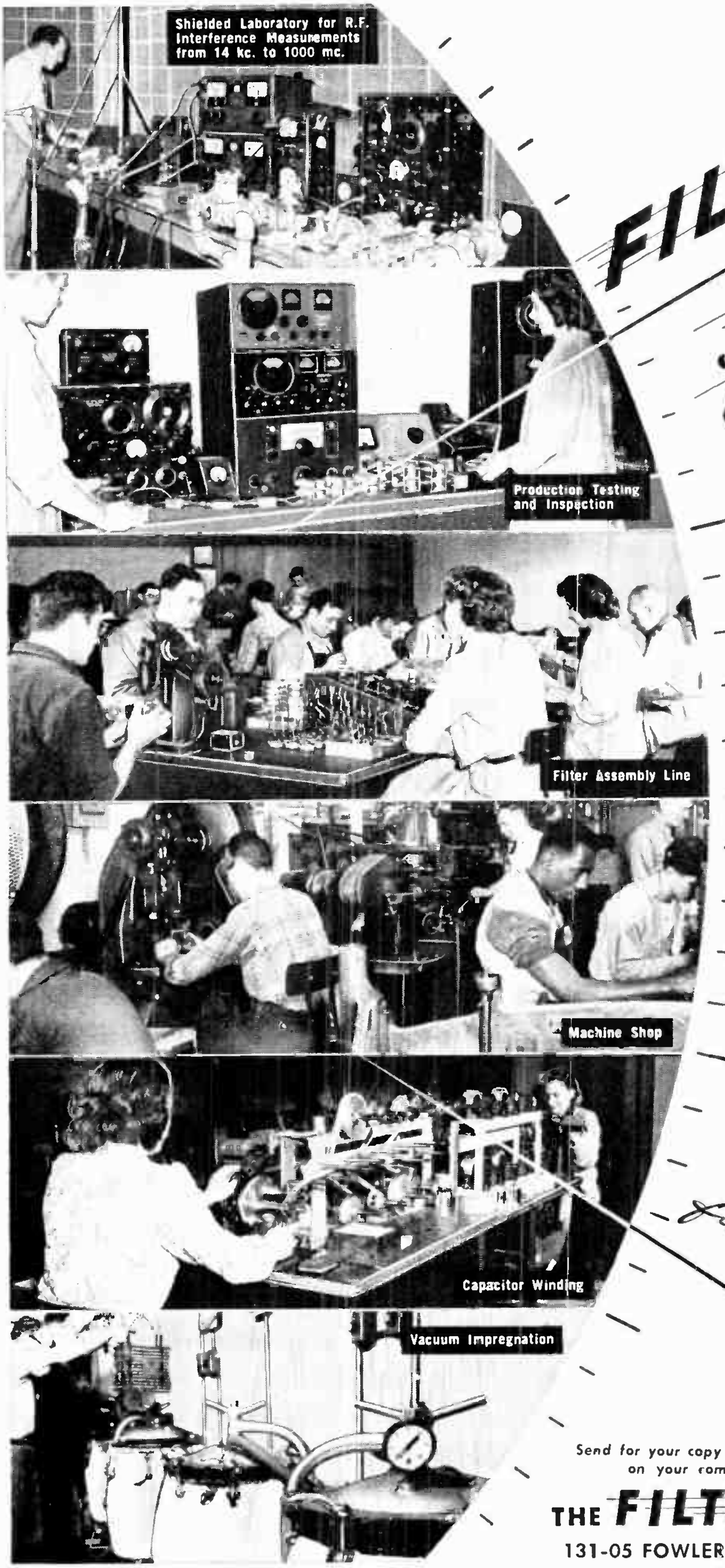
- Extremely long shelf life.
- An operating temperature range from −55°C to +85°C.
- Exceedingly low leakage currents.
- Ability to withstand severe physical shock.
- Completely sealed against contamination.

If you have large-volume applications where a price of 3 to 5 times that of hermetically-sealed paper capacitors is secondary to a combination of small size and superior performance—get in touch with us. Your letter, addressed to Capacitor Sales Division, General Electric Company, Hudson Falls, N. Y., or your nearest Apparatus Sales Office will receive prompt attention.

General Electric Company, Schenectady 5, N. Y.

GENERAL ELECTRIC

407-306



FILTRON'S *Ahead*

in

- QUALITY
- PRODUCTION
- ENGINEERING

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

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

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

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

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

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

RF INTERFERENCE FILTERS FOR:	
Electronic Controls	Signal Systems
Electric Motors	Business Machines
Electric Generators	Electric Appliances
Electronic Equipment	Electronic Signs
Fluorescent Lights	Electronic Heating Equipment
Oil Burners	

Filtered by FILTRON

LOCKHEED XF-90

Send for your copy of our NEW CATALOG on your company letterhead.

THE FILTRON CO., INC.
131-05 FOWLER STREET, FLUSHING, N.Y.

LARGEST EXCLUSIVE MANUFACTURERS OF RF INTERFERENCE FILTERS

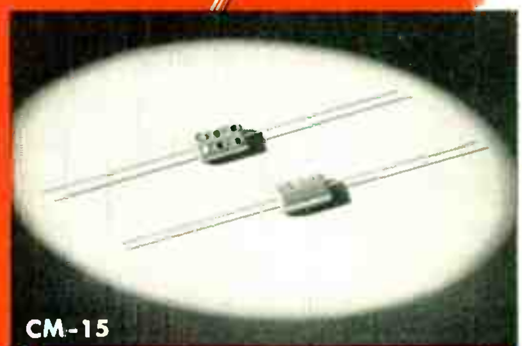
PERFORMANCE...

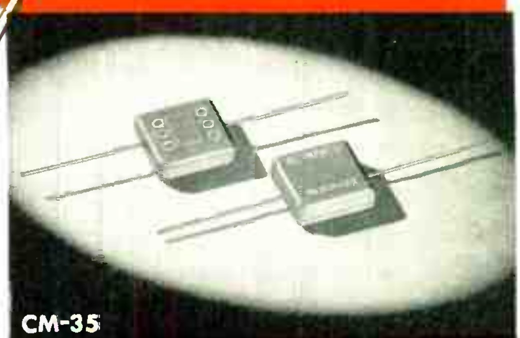
COMES IN...

From the midget champ which spins around a ¼ mile oval in 19 seconds . . . to the Indianapolis winner which clocks 157 m.p.h. on the straightaway . . . performance is the keynote in auto racing. In Electronics El-Menco Silvered-Mica Capacitors set the space. From the tiny CM-15 (2-525 mmf. cap.) to the mighty CM-35 (3300-10000 mmf. cap.) . . . unexcelled performance is paramount.

Precision workmanship, fine materials, careful design . . . these are the qualities which produce peak performance in racing cars and in El-Menco Capacitors. There is an El-Menco Capacitor for every specified military capacity and voltage. Each unit is factory-tested at *double* its working voltage. You are assured of dependability for every application.

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





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

Write on your business letterhead for catalog and samples.

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

CAPACITORS

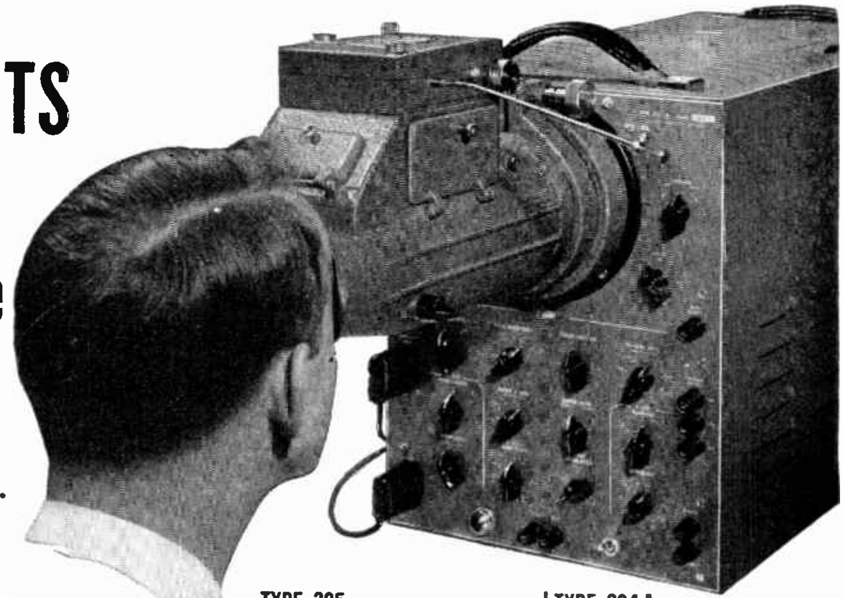
Radio and Television Manufacturers, Domestic and Foreign, Communicate Direct With Factory—

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

A COMPLETE APPROACH TO

SINGLE TRANSIENTS and PULSES of low-repetition rate

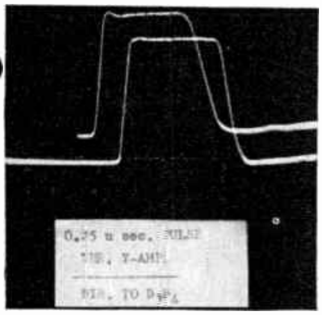
For visual observation of pulses and single transients, the Type 294-A Cathode-ray Oscillograph provides high light-output and wide-band response. For careful study and permanent reference of these signals the Type 295 Oscillograph-record Camera records writing rates as high as 35 inches per microsecond.



TYPE 295	TYPE 294-A
high-speed, single-frame OSCILLOGRAPH-RECORD CAMERA $550.00	high-voltage, high-frequency CATHODE-RAY OSCILLOGRAPH $1320.00

PULSE RESPONSE FREQUENCY RESPONSE SENSITIVITY

This oscillogram illustrates the double exposure technique. Binocular viewing in the Type 295 facilitates proper positioning for close comparison.



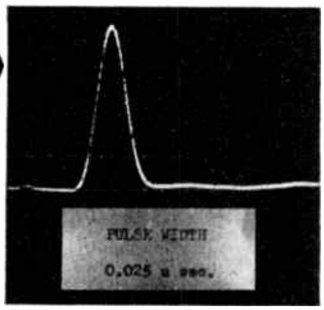
The pulses in the oscillogram at left are identical pulses of 0.25 microsecond width. The first pulse was applied through the Y-axis amplifier of the Type 294-A, and the second, directly to the vertical deflection plates. A comparison of their waveforms illustrates the excellent transient response of the Y-axis of the Type 294-A.

Response of the Y-axis amplifier to a rise time of 0.01 microsecond or less is 0.03 microseconds max. Notice that a minimum of overshoot (less than 2%) is introduced by the amplifier.

For the study of sinusoidal frequencies, the response of the Type 294-A extends from 10 cps. to 12 megacycles (down 30%). Sensitivity of the Y-axis, through the amplifier, is 0.42 peak-to-peak volts per inch.

AVAILABLE DEFLECTION LIGHT OUTPUT SIGNAL DELAY

"Time" and "Bulb" exposures may be taken with the Type 295. And provision is made so that equipment may be triggered simultaneously with shutter opening. With appropriate accelerating potentials, the Type 295 is capable of recording single transients in excess of 280 inches per microsecond.

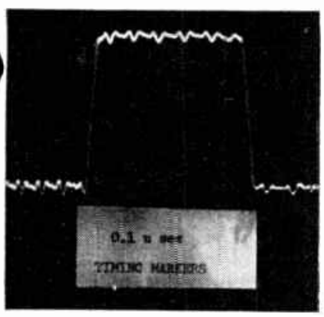


The Type 294-A provides undistorted vertical deflection of 1.3 inches or more for both positive and negative pulses; and 2.75 inches for symmetrical signals. The high light-output of the Type 294-A increases the value of the large, vertical deflection provided by the Y-axis amplifier. This is illustrated by the high visibility of the rise and decay of the pulse shown at left. Here, the Type 5XP— Cathode-ray Tube of the Type 294-A was operated at 12 kv. However, where maximum light output is not required, the accelerating potential may be lowered to 7 kv by means of a switch. At this level of operation, of course, the available undistorted deflection is increased.

To insure the complete display of fast pulses such as those at left, the Y-axis includes a 0.25 microsecond signal-delay line.

SWEEP SPEEDS TIME CALIBRATION

The built-in illuminated data-card of the Type 295 will prove invaluable when making time measurements. A film take-up cassette is arranged so that exposed frames of film may be separated from unexposed film and taken to the darkroom for immediate developing.



Complementing the Y-axis performance of the Type 294-A, sweep durations are continuously variable from 0.1 second to 3 microseconds. By increasing the length of the sweep, speeds greater than 0.25 microsecond per inch may be obtained, thus providing more detail to facilitate the study of short-duration pulses.

Calibration of the sweeps of the Type 294-A is accomplished with vertical marks occurring at intervals of 100, 10, 1, or 0.1 microseconds. In the oscillogram at left, the 0.1 microsecond markers appear mixed with the signal on the vertical axis. Time measurements may also be made by double exposure of first, the signal, and second, the timing markers.

Send requests for information to

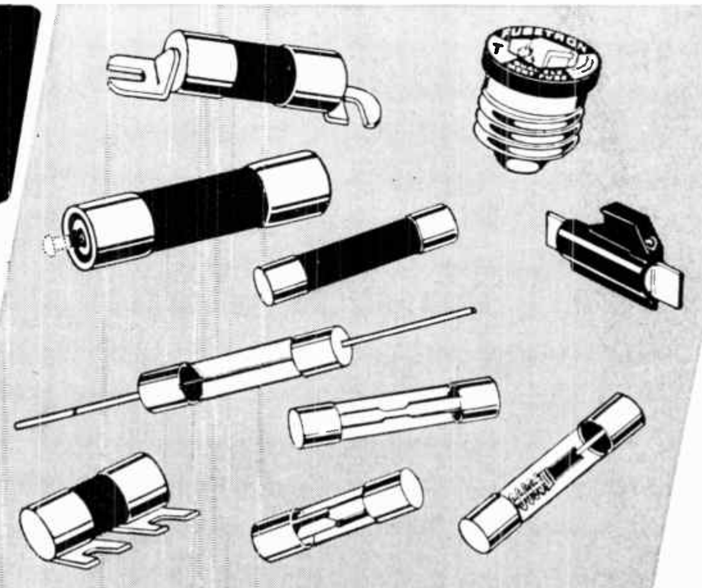
ALLEN B. DU MONT LABORATORIES, INC.
Instrument Division
1000 Main Avenue, Clifton, N. J.

DUMONT *for Oscillography*

WHEN YOU NEED A FUSE —THINK OF BUSS...

Fuse Headquarters for the Electronic Industries

TELEVISION • RADIO • COMMUNICATIONS
CONTROLS • AVIONICS • INSTRUMENTS



DUAL-ELEMENT (SLOW BLOWING) FUSES
RENEWABLE FUSES, ONE TIME FUSES, SPECIAL FUSES

A complete line of fuses is available. Made in Dual-Element (Slow blowing), Renewable and One-Time types. Sizes from 1/500 ampere up.

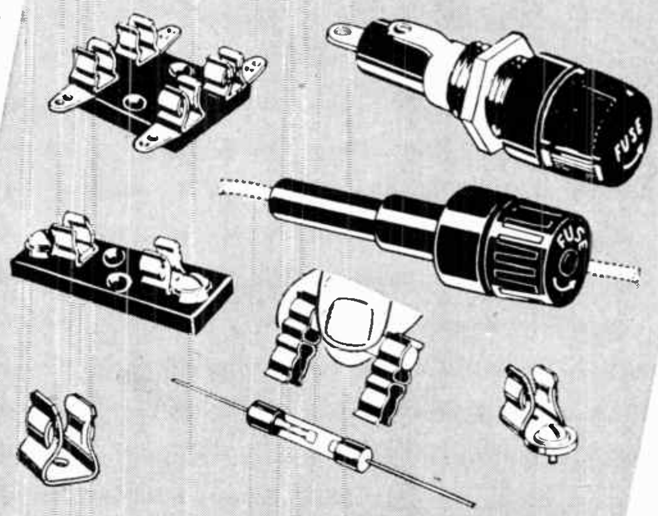
And a companion line of BUSS Fuse Clips, Fuse Blocks and Fuse Holders.

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

Each* BUSS *Fuse Electronically Tested.

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

BUSS *Fuses are made to Protect — not to Blow.*



FUSE CLIPS, FUSE BLOCKS,
FUSE HOLDERS, SPECIAL FUSE MOUNTINGS

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

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.

USE THIS COUPON — *Get All the Facts*

IRE 1251

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

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

Name ______________________

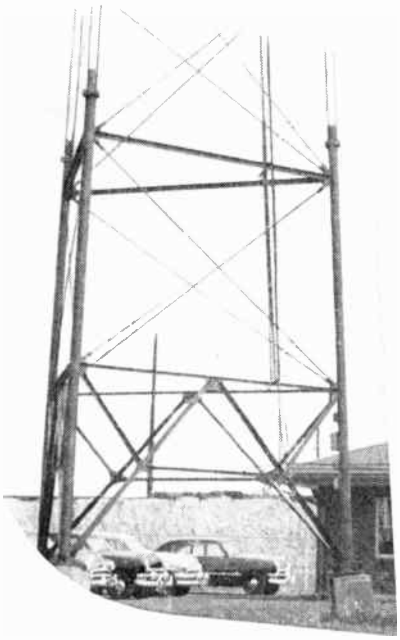
Title ______________________

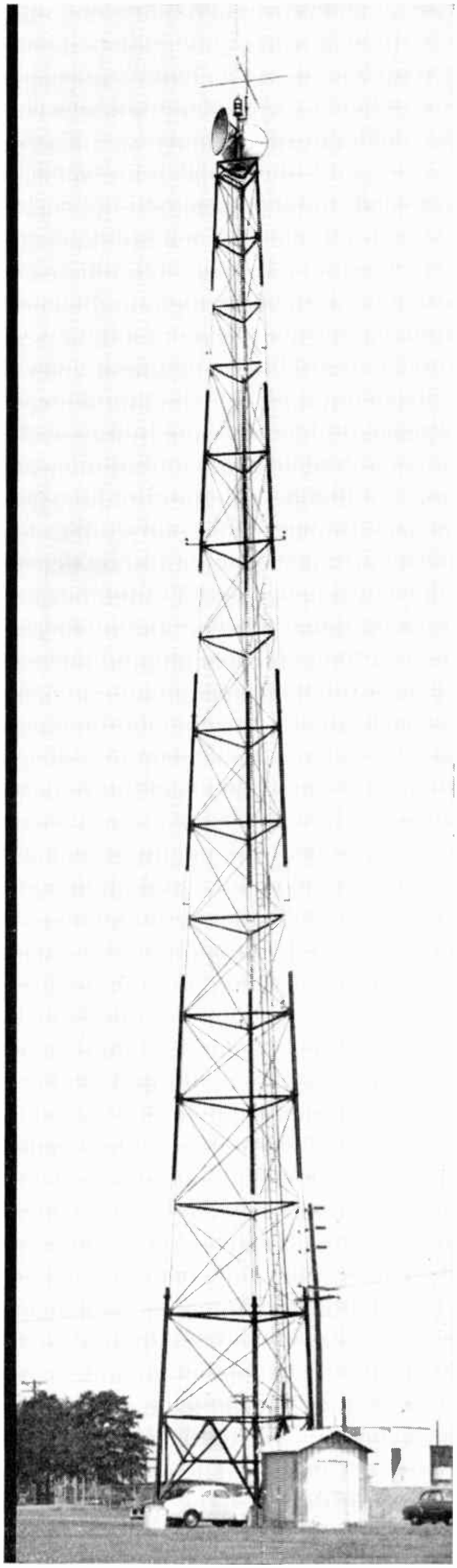
Company ______________________

Address ______________________

City ______________ State ______________

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.





ERIE miniaturized TUBULAR CERAMICONS®

- New Higher Capacitance Values in Standard Size Tubular Ceramicons
- Available in
  - Radial Lead Non-Insulated (Illustrated)
  - Radial Lead Phenolic Insulated
  - Axial Lead Molded Insulated (Styles comparable to 301 and 302 only)

Compare for size with other Ceramic Capacitors, and specify ERIE GP3 Miniature Ceramicons for Space Saving By-Pass and Coupling Applications.

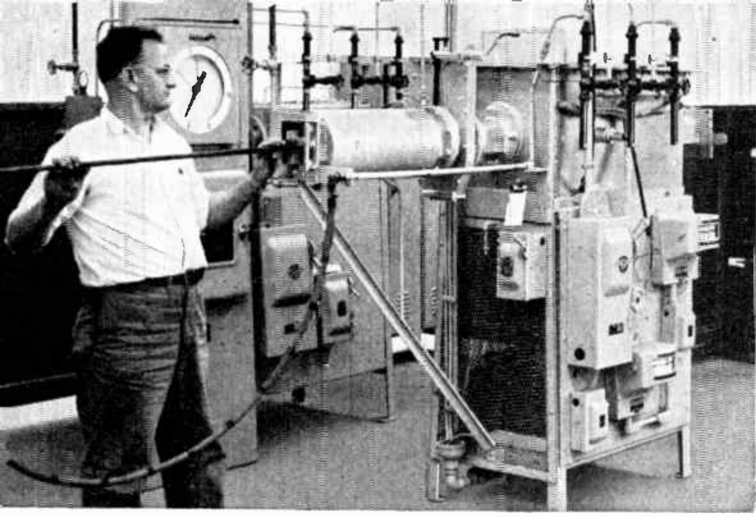
GP3 Ceramicons are manufactured with the use of a high dielectric constant ceramic developed especially for these compact tubular capacitors. Basic development work has been accomplished over the past few years in Erie's engineering laboratories. Since 1949 these units have been made on special order, and they are now available in production quantities.

Write for complete information and samples

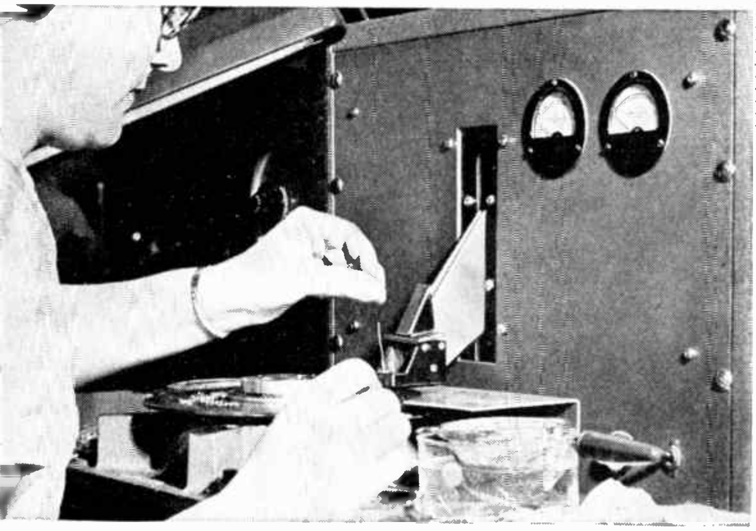
GP3 STYLE	MAXIMUM CAPACITANCE
 STYLE 301 .200 x .400	.002 mfd.
 STYLE 302 .200 x .656	.005 mfd.
 STYLE 307 .230 x .860	.0075 mfd.

Tolerance on Capacitance: +80%, −20%
Hi Pot. Test: 1500 VDC
Life Test: 700 VDC 1000 hours at 85° C

Electronics Division
ERIE RESISTOR CORP., ERIE, PA.
LONDON, ENGLAND · TORONTO, CANADA



1. **MAXIMUM METAL PURITY** is essential in the manufacture of diodes. In photograph above dioxide powder is being reduced to pure germanium metal.



2. **GERMANIUM PELLETS** are mounted to pin assemblies prior to assembly in cases. Precision centering as well as speed are essential to produce these quality units at low cost.

3. **FORMING, SHEARING, AND WELDING WHISKERS** on diode pin assembly calls for careful manipulation under microscope for accurately formed .003 inch diam. whisker.

NEW GENERAL ELECTRIC PLANT PRODUCES RECORD VOLUME OF

WELDED GERMANIUM DIODES

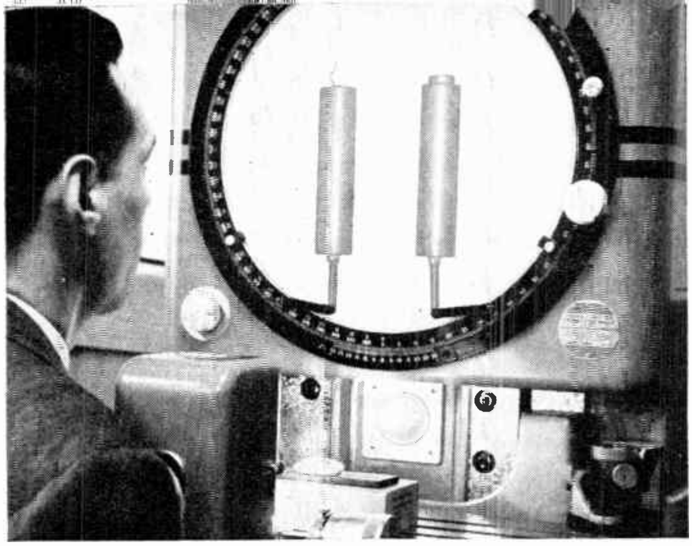
NEW METHODS OF MASS PRODUCTION in G. E.'s germanium products plant* at Clyde, New York are booming diode output to the highest peak in the industry's history. An offspring of the mother plant at Electronics Park, this factory is equipped to produce 12 million diodes a year. Television receivers, computers, communication systems and military requirements present an ever-increasing demand for these minute but vital components.

WITH EXPANDED LABORATORY FACILITIES for the development of new germanium and other semi-conductor items, G. E. is prepared to suggest electronic solutions to your design and manufacturing problems—solutions that may well save you money and improve your product.

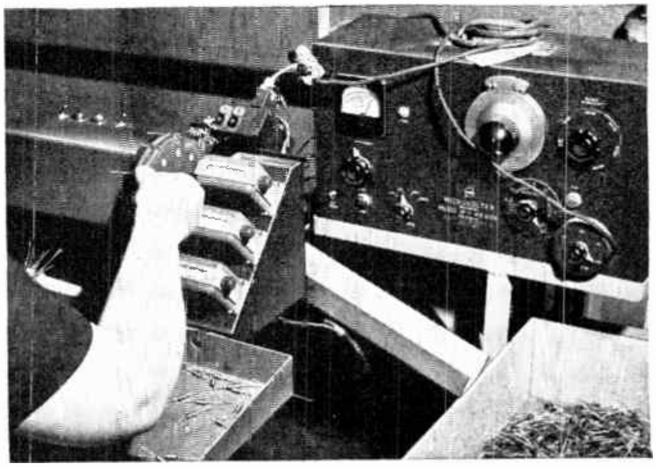
Would you like more information on this? Ask us to call. *General Electric Co., Electronics Park, Syracuse, N. Y.*

*Which you are invited to inspect when in the Syracuse area. Meanwhile, let us send you additional information and specifications on G-E diode products. Write for bulletin #X57-01A.

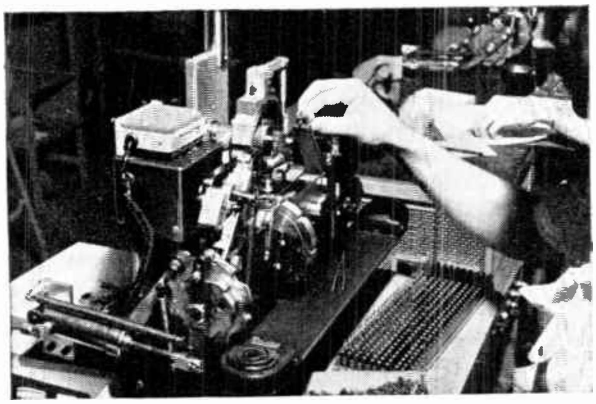
4. **ORDERLY BANK OF WHISKER MACHINES** is typical of modern production facilities in the new G-E plant. Quantities up to 12 million units a year can be produced here.



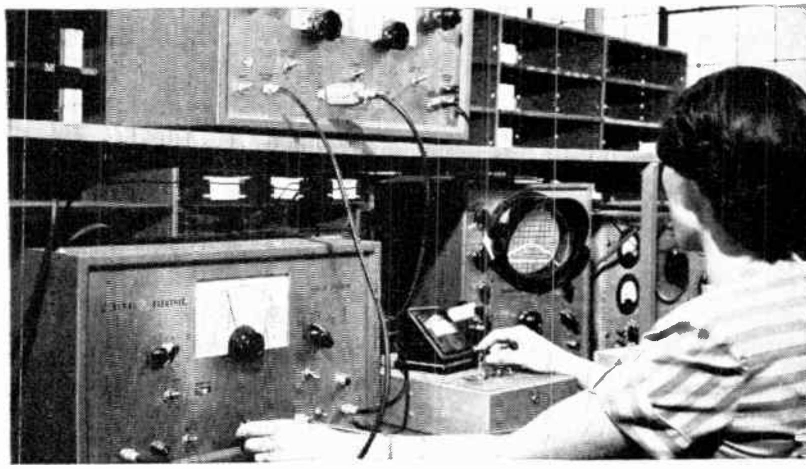
5. CONTOUR PROJECTION of diode parts for microscopic inspection. Pellet and whisker (on screen) must follow rigid specifications. This is typical of quality control processes.



8. AUTOMATIC TEST SEPARATION of diodes by types eliminates costly hand sorting of thousands of units per hour. Every G-E diode is tested many times.

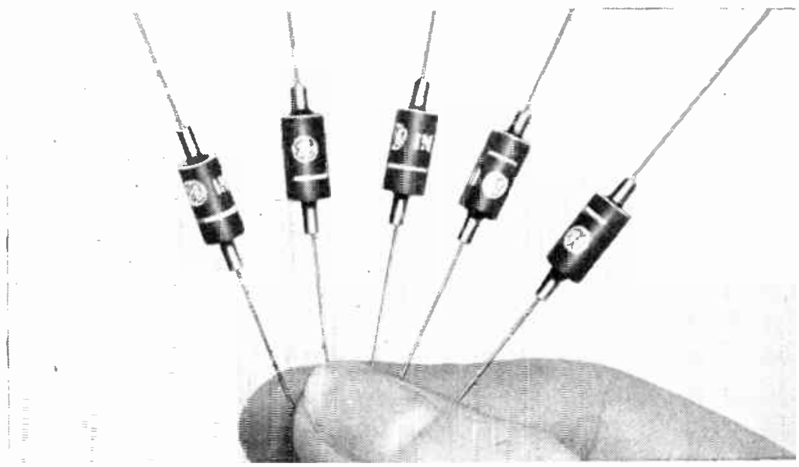


6. FINAL ASSEMBLY of whisker and pellet pins in plastic cases requires special machines designed by G-E engineers for speed and accuracy.



9. HIGH-FREQUENCY TESTING of diodes for television applications has proved successful in supplying over 2 million G-E units to television manufacturers for high efficiency needs.

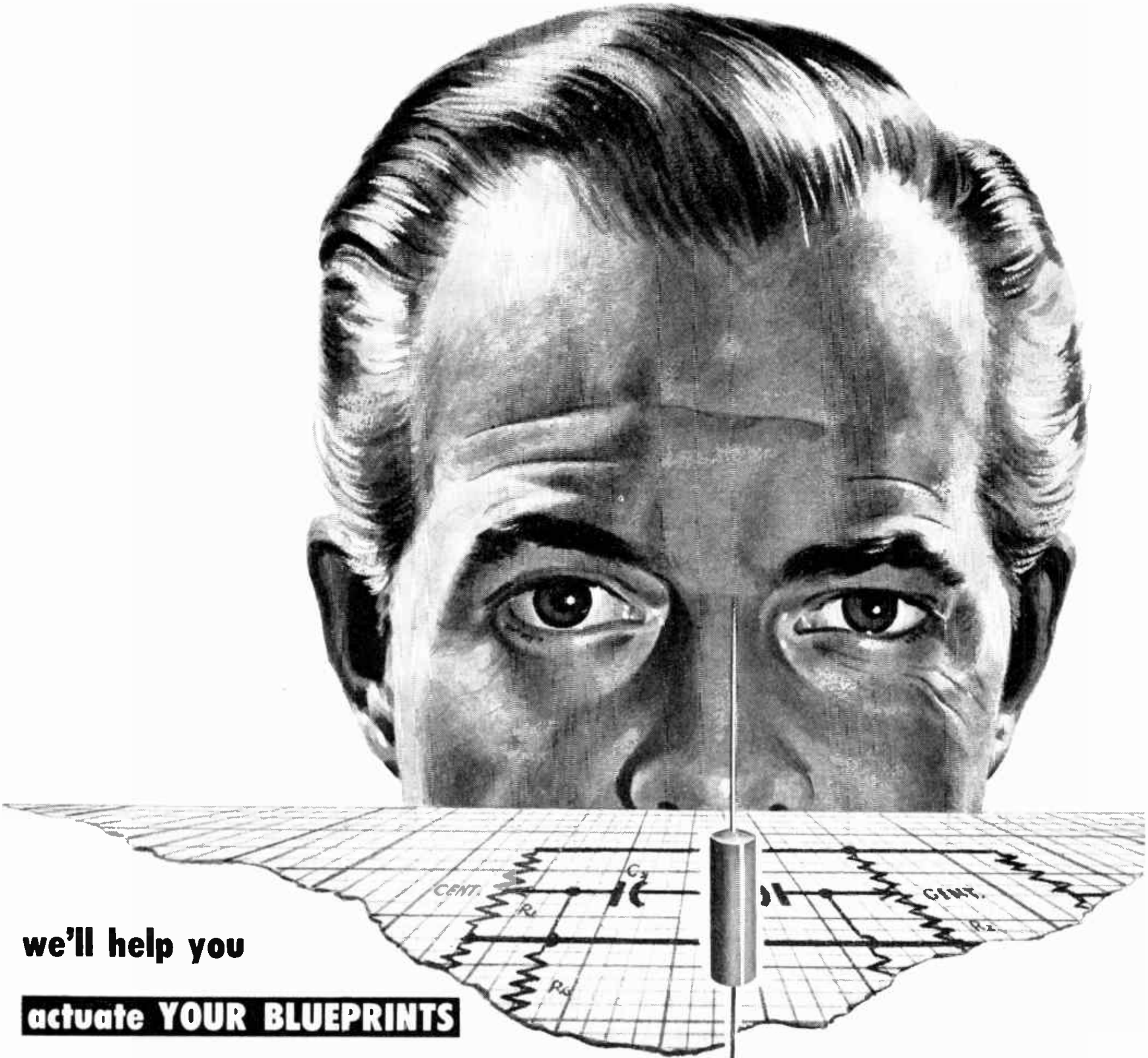
7. ASSEMBLY MACHINES turn out diodes of 12 different varieties. This process represents unusual advancement over former "hand-made" methods.



10. FINISHED G-E DIODES of various types are small, rugged, efficient, and low in cost. These components can replace some categories of vacuum tubes.

You can put your confidence in—

GENERAL ELECTRIC

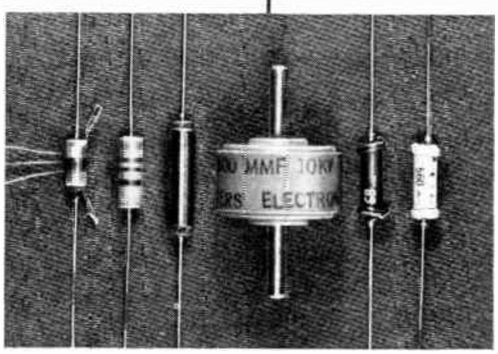


we'll help you

actuate YOUR BLUEPRINTS

...if electronic components such as molded capacitors are your problem

Other electronic components
also built in quantity
to your most exacting specifications
for stability in service



JEFFERS Electronics INC.

A SPEER CARBON COMPANY SUBSIDIARY

DU BOIS, PENNSYLVANIA

Complete your Circuits with Resistors, Coil Forms and Iron Cores by Speer Resistor Corp., St. Marys, Pa. another SPEER CARBON CO. subsidiary

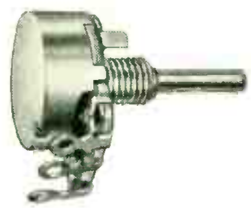
7664

3 New Resistors
−55°C to +150°C

Complete aridity to saturation . . . An *unprecedented* temperature and humidity range

FOR MILITARY APPLICATIONS

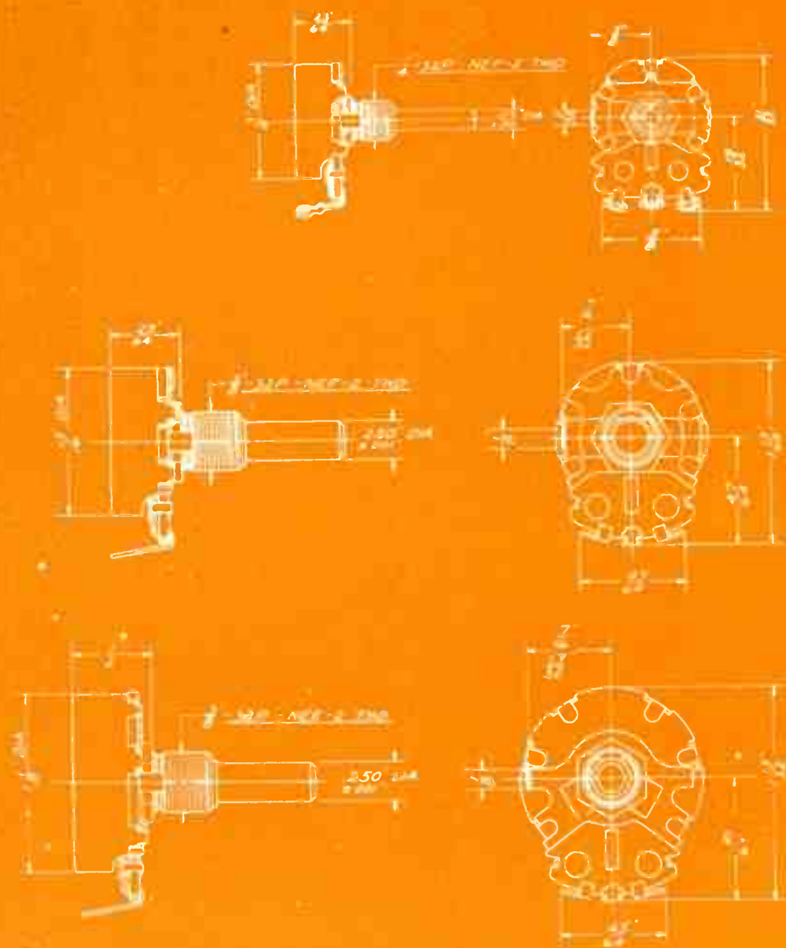
Highly recommended for use in jet and other planes, guided missiles, tanks, ships and submarines, portable or mobile equipment and all other military communications. Manufactured from specially developed materials, these absolutely unique variable resistors are now available in a *complete range of sizes*. (See chart at bottom of page.)



Type 65

Type 90

Type 95

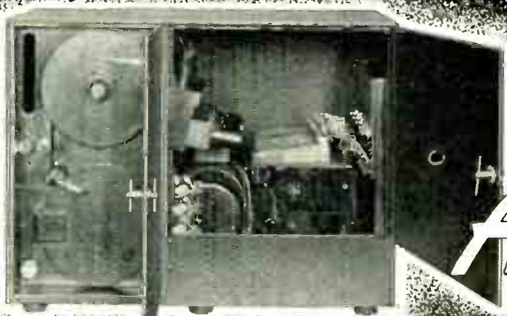


DIAMETER	TYPE 95 1⅛"	TYPE 90 15/16"	TYPE 65 (miniaturized) ¾"
Wattage and Voltage Rating	2 watts @ 70°C with 500 V max. across end terminals	1 watt @ 70°C with 500 V max. across end terminals	½ watt @ 70°C with 350 V max. across end terminals

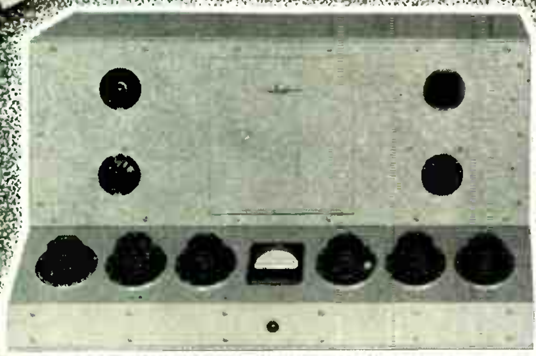
+150°C to −55°C

Specialists in Precision Mass Production of Variable Resistors


Meets Military Specifications
JAN-R-94, Type RV-3A
CTS Type 35, 1 1/8" Diameter
Composition
JAN-R-94, Type RV-2B
CTS Type GC 45 with Switch
Composition
JAN-R-94, Type RV-3B
CTS Type GC 35 with Switch
Composition
JAN-R 94, Type RV-2A
CTS Type 45, 15/16" Diameter
Composition
JAN Type RV-4B
CTS Type FGC 95 with Switch
Composition
JAN Type RV-4A
CTS Type 95, 1 1/8" Diameter
Composition
MEETS ALL
JAN-R-19 SPECIFICATIONS
JAN Type RA 20A
2 Watt (CTS Type 252)
JAN Type RA 20B
2 Watt (CTS Type GC-252)
JAN Type RA 25A or 30A
3 or 4 Watt (CTS Type 25)
JAN Type RA 25B or 30B
3 or 4 Watt (CTS Type GC 25)
EXCEPTIONALLY GOOD DELIVERY CYCLE on military orders due to enormous mass production facilities . . . Please give complete details on your requirements when writing or phoning for further information.
FOUNDED 1896
CTS
CHICAGO TELEPHONE SUPPLY
Corporation
REPRESENTATIVES
Henry E. Sanders
401 North Broad Street
Philadelphia 8, Pennsylvania
Phone: Walnut 2-5369
W. S. Harmon Company
1638 So. La Cienega Blvd.
Los Angeles 35, California
Phone: Bradshaw 2-3321
IN CANADA
C. C. Meredith & Co.
Streetsville, Ontario
SOUTH AMERICA
Jose Luis Pontet
Buenos Aires, Argentina
Montevideo, Uruguay
Rio de Janeiro, Brazil
Sao Paulo, Brazil
OTHER EXPORT
Sylvan Ginsbury
8 West 40th Street
New York 18, N. Y.



RECORDER



REPRODUCER

WHY SEARCH
for technically correct ceramics?

Lavite

CERAMICS Still Available

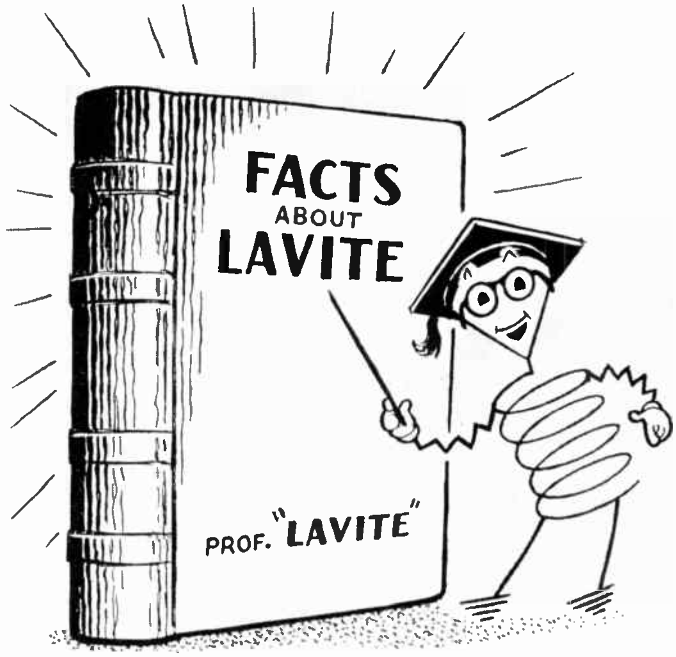
We have again challenged the times, as we have all through our 75 years of developing and producing "Lavite" Technical Ceramics ("Lavite" Steatite, "Lavite" Titanates, "Lavite" Ferrites and others). We have increased our facilities — research, design and manufacturing — to help you meet the drastic demands for ceramics of specific physical characteristics. We invite you to submit your problems to our engineering department for individual study and recommendation — or, perhaps one of the well known and proved standard "Lavite" Ceramics will satisfy your needs. Whichever it be, we will give your requirements immediate attention. Write for general characteristic data on all "Lavite" Ceramics.

D. M. STEWARD MANUFACTURING CO.

3605 Jerome Avenue Chattanooga, Tennessee

Sales Offices in Principal Cities

CELEBRATING OUR 75th ANNIVERSARY 1876 1951



CONSIDER THESE FOR IMMEDIATE DELIVERY

"Lavite" STEATITES

S-10 and S-15 for ordinary radio and general medium-frequency requirements. S-201 and S-300 for ultra low loss — high frequency applications.

"Lavite" SEMI-STEATITES

S-25H with good electrical properties at high temperatures. R-10 for element supports in electrical appliances.

"Lavite" CORDIERITE

S-40H, S-12H and R-15 for special application for extreme thermal change when a low linear coefficient of expansion ceramic is required.

"Lavite" MULLITE

R-20 and R-25 for light duty refractory applications.

"Lavite" ZIRCON PORCELAIN

Z-200 and Z-300 with excellent all around properties including: (1) low electrical loss; (2) high mechanical strength; and (3) thermal shock resistant.

"Lavite" FIRED STONE GRADES

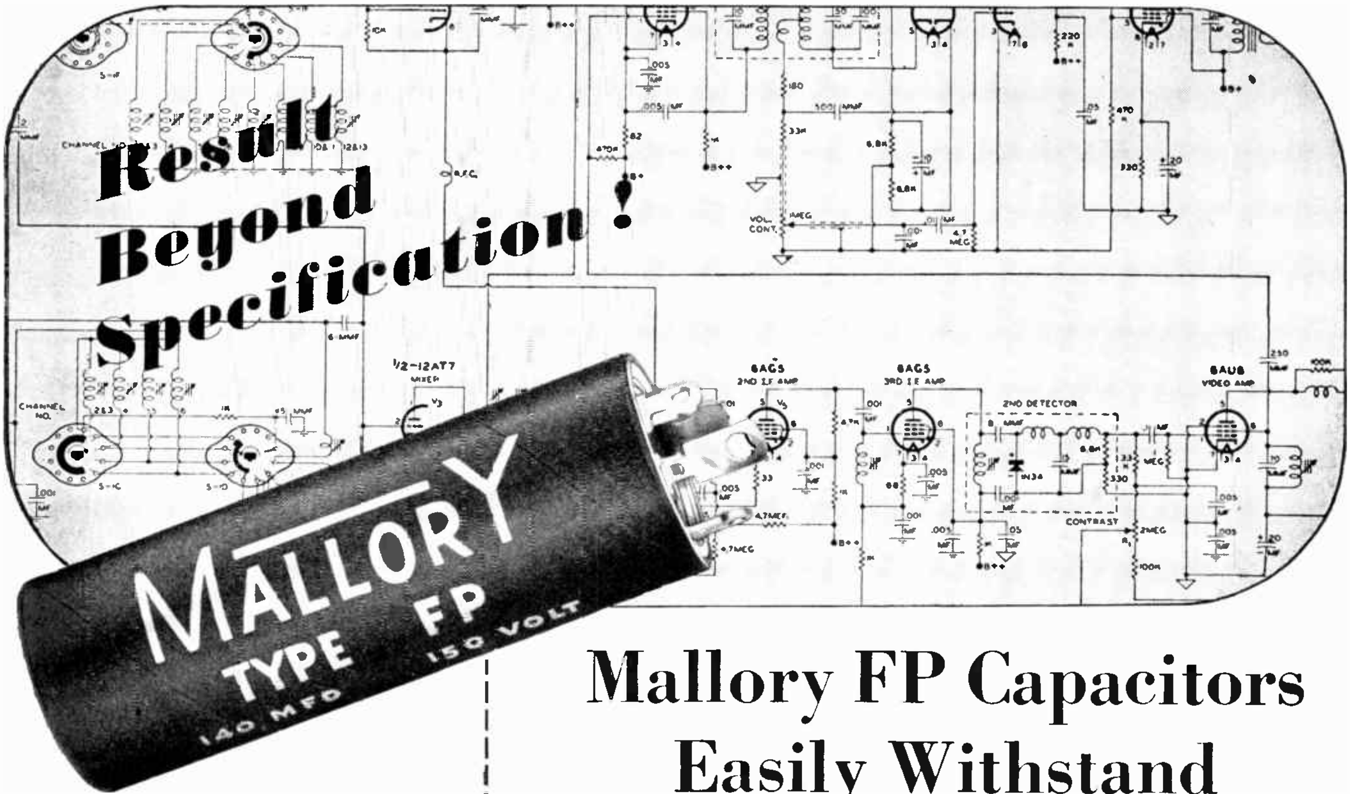
N(M) and A for mechanical parts where special tools are not required and tolerances can be maintained. Especially adaptable for models.

"Lavite" TITANATES

Available in a range of types and properties including low, intermediate and high "K" groups.

"Lavite" FERRITES

Permeability factor 1000 initial — 3000 maximum and curie point of 250° F. Particularly adapted for television and frequency modulation.



Mallory FP Capacitors Easily Withstand High Ripple Currents In TV Circuits

When you specify Mallory Capacitors for television receivers or other equipment where heat is a problem, you can be sure they will stand the test. Mallory FP Capacitors are designed to give long, trouble-free performance at 85°C.—naturally they give even longer service at normal temperatures. In addition, Mallory FP Capacitors are famous for their long shelf life. Write for your copy of the FP Capacitor Engineering Data Folder.

The superior heat dissipation characteristic of Mallory FP Capacitors—long inherent in our production method—proved to be an important factor in meeting the problem of high ripple currents in TV receivers.

Thorough testing demonstrated that standard Mallory FP Capacitors would stand up in the rugged service involved in the voltage doubling rectifier circuit. The jump from radio ripple currents of about .15 amperes to five or six times this current in TV service places a tremendous burden on capacitors. But Mallory FP Capacitors are giving long, uninterrupted performance . . . at temperatures approximating the boiling point of water.

That's result beyond specification!

Mallory capacitor know-how is at your disposal. What Mallory has done for others can be done for you.

FP is the type designation of the Mallory developed electrolytic capacitor having the characteristic design pictured and famous throughout the industry for dependable performance.

P.R. MALLORY & CO., Inc.
MALLORY

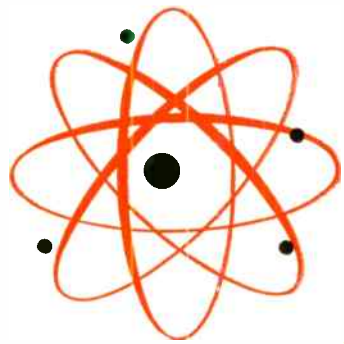
P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA

SERVING INDUSTRY WITH

Electromechanical Products
Resistors *Switches*
TV Tuners *Vibrators*

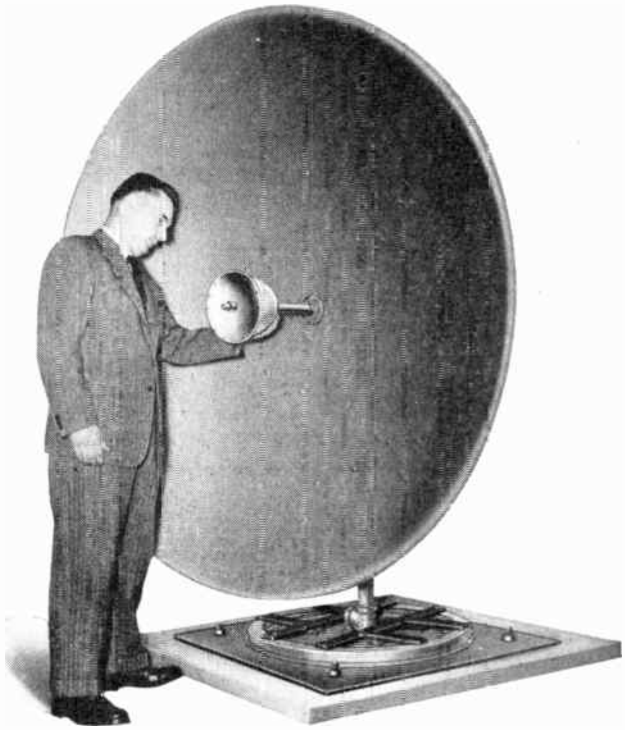
Electrochemical Products
Capacitors *Rectifiers*
Mercury Dry Batteries

Metallurgical Products
Contacts *Special Metals*
Welding Materials



940 MCS

TOP PERFORMANCE



Parabolic Antennas for 940 mcs.

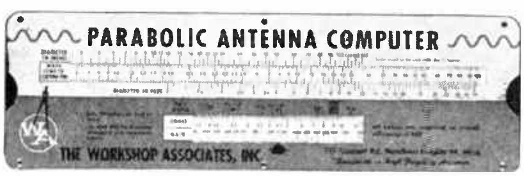
and every other Microwave Frequency

The Workshop was the first manufacturer to bring out a complete line of parabolic antennas for all microwave frequencies. Having specialized in this field for several years, we can supply equipment from our standard line to meet the majority of installations. However, for special requirements, we are equipped to design and supply reflectors in a wide range of sizes and focal lengths.

Model 940

Frequency Range	920 to 960 Mcs.			
Input Impedance	52 ohms nominal			
VSRW	1.20 to 1 over the band			
Power Rating	1 kw. continuous			
Polarization	Either vertical or horizontal available at time of installation.			
Reflector Size	4′	6′	8′	10′
Gain (db, approx., over isotropic radiator)	19	23	26	28
Half Power Angles (H plane)	17.75°	11.75°	8.6°	6.9°
(E plane)	19.75°	12.9°	9.6°	7.8°
Side Lobes	17 db down or better			
Pressurized	Feed can be pressurized to 10 lbs. p.s.i.			
Input Connection	Weatherproof type "N" fitting; special fittings are available for RG-8 U, RG-17 U or 7 8″ copper line. Specify when ordering.			
De-icing	Available for all models. Capacities range from 400 to 4000 watts.			

FREE SLIDE RULE



This pocket size slide rule quickly computes diameter, wavelength, angle and gain for parabolic antennas. Reverse side carries FCC frequency allocations conversion tables and other data. Write for your copy.

OTHER STANDARD MODELS

MODEL NO.	FREQUENCY (MCS.)	GAIN* (DB.)	HALF POWER ANGLE E Plane*	HALF POWER ANGLE H Plane*
2000	1700–2300	27.0–34.5	10.28–3.65	9.2 –3.25
7000	5925–7425	36.0–43.0	3.24–1.36	2.86–1.21

*Gain and Half Power Angles are dependent on size and frequency of parabolas,— 4, 6, 8 or 10 foot diameter.

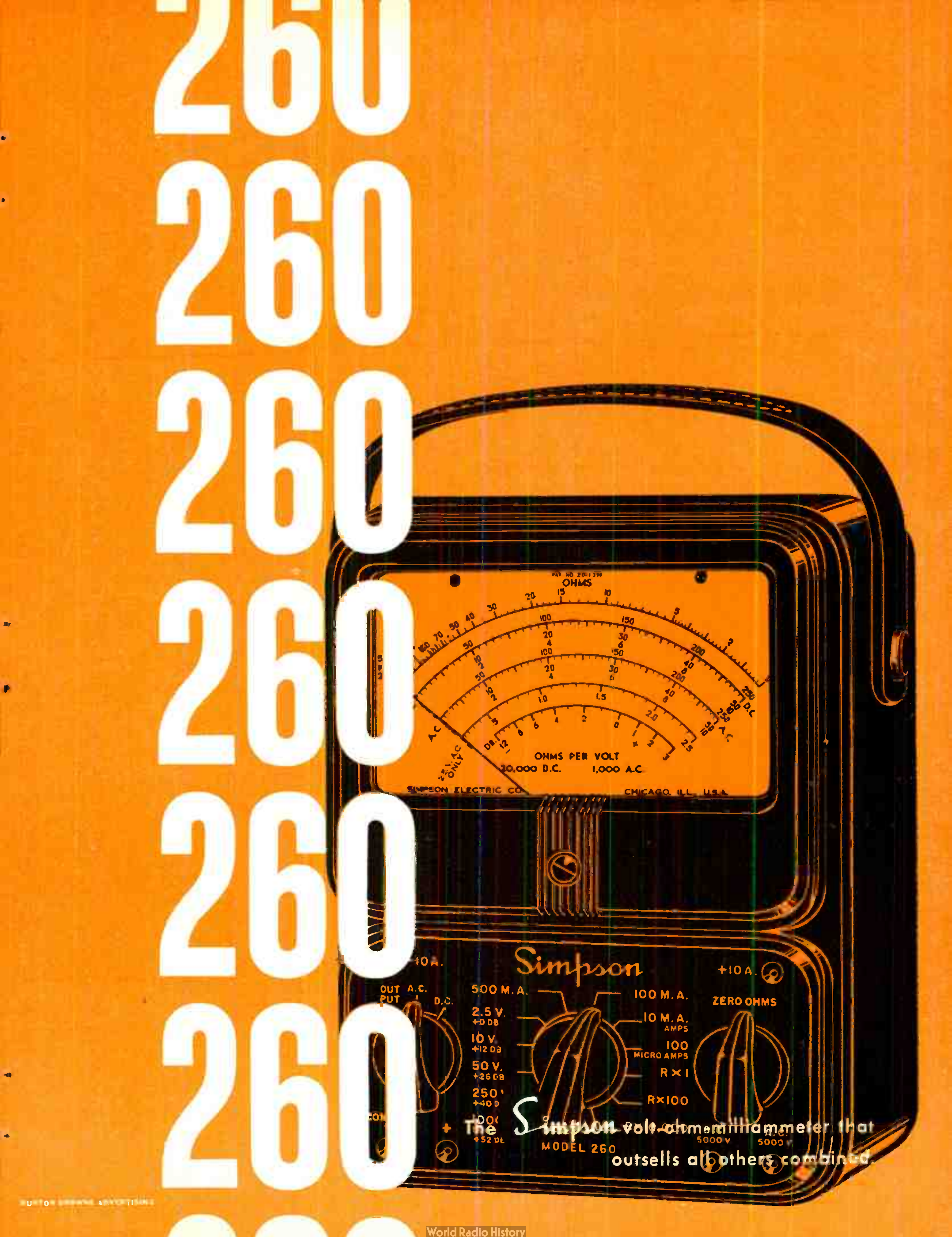
Write for Parabolic Antenna Catalog

The WORKSHOP ASSOCIATES

DIVISION OF THE GABRIEL COMPANY

Specialists in High-Frequency Antennas

135 Crescent Road, Needham Heights 94, Massachusetts


260
260
260
260
260
260
OHMS
OHMS PER VOLT
20,000 D.C.
1,000 A.C.
SIMPSON ELECTRIC CO.
CHICAGO, ILL., U.S.A.
Simpson
500 M.A.
100 M.A.
10 M.A.
100 MICRO AMPS
R×1
R×100
ZERO OHMS
+10 A.
MODEL 260
The Simpson volt-ohm-milliammeter that outsells all others combined.

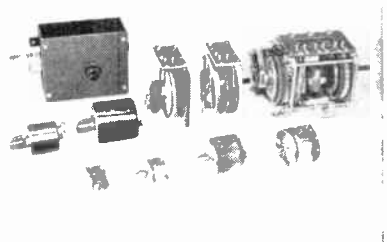
News—New Products

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)

Variable Transformers

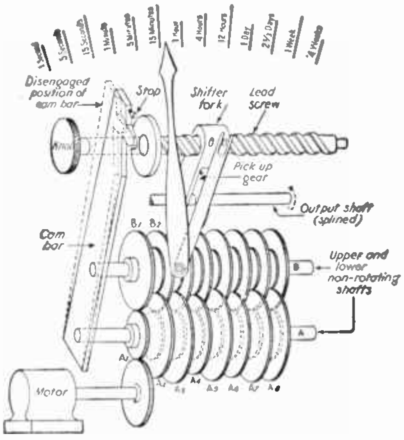
The Superior Electric Co., 83 Laurel St., Bristol, Conn., offers a line of POWERSTAT variable transformers for 400 cps and higher.



Since most of the higher frequency applications involve electrical equipment built to specific requirements, it is difficult to offer a complete line to fulfill every need. A variety of POWERSTAT variable transformers have been built to these specifications. Units are available in a multitude of voltage and current ratings in single and polyphase models.

Multispeed Instrument Drive

Gorrell & Gorrell, Haworth, N. J., have two new instrument drives designated the M-M and S-M units (minutes-to-a-month and seconds-to-a-month speed ranges). They utilize a single unit, motor, gear train, and speed selection mechanism. Applications are apparatus, instrument drives (chart, disc, etc.), chart movements, time and sequence controls.



The S-M unit provides 20 ounce-inch torque with output shaft rotation of 13 speeds, of 1, 5, 15 seconds, 1, 5, 15 minutes, 1, 4, 12 hours, 1, $2\frac{1}{2}$ days, 1 and 4 weeks. M-M unit provides 5 ounce-inch torque at the same speeds except of 1, 5, and 15 seconds. Both are for 115 volts 60 cps ac. Special assemblies for adjacent speed ratios can be had from 5:1 to 1:5, with individual differences as small as 59/61 available. Prices range from $65.00 for single units, to $19.75 for quantity orders.

(Continued on page 60A)

World Radio History

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 56A)

Loading Coils For Mobile Antennas

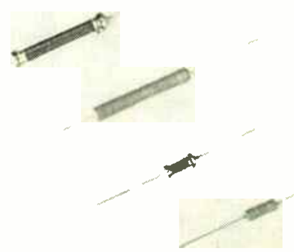
Two new base-loading coils, designed for use with any type mobile-whip antenna, have been announced by the **Mallard Manufacturing Co.,** 6025 N. Keystone Ave., Chicago 30, Ill.

Available for 20 and 75 meter operation, the two new units are designed to fit all standard mounts and whips and, with the adaptor supplied, can be used with non-standard types. Both models are housed in a weather-proof $\frac{1}{8}$ inch thick plexiglass and have removable nylon end caps.

The Hi-Q 20 is wound with plated $\frac{1}{8}$ inch diameter solid copper wire. A flexible copper strap is provided which permits adjustment to exact inductance desired. The Hi-Q 75 features two pie-wound coils over a powdered iron core slug.

Deposited Carbon Resistors

A deposited carbon resistor, known as the "Carb-ohm" is being manufactured by the **Phaostron Co.,** 151 Pasadena Ave., South Pasadena, Calif. For high frequency applications, where high values of resistance are needed, or power dissipations up to 2 watts are required, the "Carb-ohms" may be employed.



These resistors are available hermitically sealed in glass, or clad in a humidity-impervious casing, which provides stability over time and freedom from variations due to climatic changes. "Carb-ohms" are available in a variety of mountings; the axial lead, pictured above, as well as a threaded stud or a tapped hole terminal. Wattage ratings range from $\frac{1}{8}$ to 2 watts with a resistance range of 20 ohms to 200 megohms.

The "Carb-ohm" is manufactured under license arrangements with Western Electric Co., Inc.

An illustrated brochure and complete information are available upon request.

(Continued on page 62A)

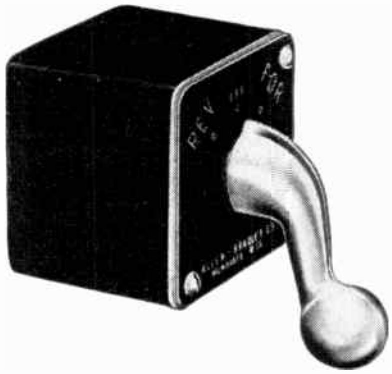

performance insured*

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 60A)

Reversing Drum Switch

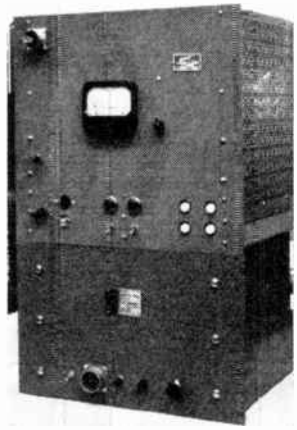


Announcement is made by **Allen-Bradley Co.**, 114 W. Greenfield Ave., Milwaukee 4, Wis., of a compact new reversing drum switch—Bulletin 350 Style A, suitable for a wide variety of mounting arrangements for small workshops and industrial services.

The Bulletin 350 Style A drum switch is the equivalent of a three pole double throw switch. It is small, simply designed, for machine and equipment requiring an economical across-the-line starting, and reversing switch for ac and dc motors rated at two horsepower or less.

Housing of the new unit is bakelite and contains eight fixed contacts, a moving contact assembly, handle-coverplate assembly, and mounting screws. Contacts are cadmium silver alloy, eliminating maintenance. Wide flexibility in mounting is provided.

Magnetic DC Power Supply Regulator



General Magnetics, Inc., 135 Bloomfield Ave., Bloomfield, N. J., has developed a new klystron dc power supply voltage regulator. There are no vacuum tubes, it has an all-magnetic amplifier-type circuit. This regulator maintains power supply output at 18,000 V dc for load current variations from zero to 100 ma, and line voltage variations from 105 to 135 V ac. Regulation accuracy is ±1 per cent. Time delay approximately 0.1 second. The unit may be easily and quickly modified for power supplies of any output rating.

(Continued on page 103A)

The Fountainhead of Modern Tube Development is RCA

Pulse-rated

Four RCA tubes specifically designed for pulsed applications

RCA-5893 "Pencil-Type" UHF Triode

RCA-5893 is a new, medium-mu, "pencil-type" triode employing a double-ended, coaxial-electrode structure. As a plate-pulsed oscillator in grounded-grid service, the 5893 will deliver a useful power output at peak of pulse of 1200 watts at frequencies up to 3300 Mc.

Maximum Ratings as Plate-Pulsed Oscillator

For max. total "on" time, in any 5000-μsec interval, of 5 μsec

Peak Positive-Pulse Plate Supply Voltage	1750 max. volts
Peak Negative-Pulse Grid-Bias Voltage	150 max. volts
Peak Plate Current from Pulse Supply	3 max. amp
Peak Rectified Grid Current	1.3 max. amp
Plate Dissipation	6 max. watts
Pulse Duration	1.5 max. μsec

RCA-5946 UHF Power Triode

RCA-5946 is a new forced-air-cooled power triode for use in circuits of the coaxial-cylinder type. In plate-pulsed service, the 5946 will deliver a useful power output at peak of pulse of 14 kw at a frequency of 1250 Mc.

Maximum Ratings as Plate-Pulsed Oscillator & Amplifier

For max. total "on" time, in any 1000-μsec interval, of	10 μsec	100 μsec
Peak Positive-Pulse Plate Supply Voltage	7500 max.	7500 max. volts
Peak Negative-Pulse Grid-Bias Voltage	600 max.	600 max. volts
Peak Plate Current from Pulse Supply	4.5 max.	3.5 max. amp
Peak Rectified Grid Current	1.0 max.	0.75 max. amp
Plate Dissipation	250 max.	250 max. watts

RCA-4C33 UHF Power Triode

RCA-4C33 is a forced-air-cooled power triode for coaxial-type circuits. For plate-pulsed service, it will provide a power output at peak of pulse of 130 kw at a frequency of 600 Mc.

Maximum Ratings as Plate-Pulsed Oscillator

Peak Positive-Pulse Plate Supply Voltage	13000 max. volts
Peak Negative-Pulse Grid-Bias Voltage	2000 max. volts
Peak Plate Current from Pulse Supply	30 max. amp
Peak Rectified Grid Current	4 max. amp
Plate Dissipation	250 max. watts
Pulse Duration	5 max. μsec

RCA-3E29 Twin-Beam Power Amplifier

RCA-3E29 is a twin-unit, beam power amplifier designed to handle a peak plate current of 10 amp. in pulse modulation service.

Maximum Ratings as Pulse Modulator (both units in parallel)

For pulse length of	1 max.	1.2 max. μsec
DC Plate Supply Voltage	5000 max.	5000 max. volts
DC Grid-No. 2 Supply Voltage	850 max.	850 max. volts
DC Grid-No. 1 Supply Voltage	−200 max.	−200 max. volts
Plate Input	85 max.	60 max. watts
Peak Grid-No. 2 Current	0.5 max.	0.5 max. amp
Plate Dissipation	15 max.	15 max. watts

For further technical data or design assistance on any RCA pulse tube, write RCA, Commercial Engineering, Section 47LR, Harrison, N. J., or contact the RCA Field Office nearest you.

FIELD OFFICES: (EAST) Humboldt 5-3900, 415 S. 5th St., Harrison, N.J. (MIDWEST) Whitehall 4-2900, 589 E. Illinois St., Chicago, Ill. (WEST) Madison 9-3671, 420 S. San Pedro St., Los Angeles, Calif.

RCA

RADIO CORPORATION of AMERICA

ELECTRON TUBES — HARRISON, N.J.

PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 39 — *December, 1951* — NUMBER 12

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

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

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

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

MEMBER AUDIT BUREAU OF CIRCULATIONS

Clyde L. Farrar

Chairman, Oklahoma City Section

Clyde L. Farrar, Chairman of the Oklahoma City Section, was born in Klamath Falls, Ore., in 1896, and received the B.S. degree in electrical engineering and the E. E. degree from the University of Colorado, in 1921, and 1926, respectively. From 1923 to 1929, he was associated with the University of Idaho as assistant professor of electrical engineering in charge of communication, and from 1929 to 1932 he was an associate professor and the technical director of Radio Station KUOA, at the University of Arkansas.

Mr. Farrar has been a member of the faculty of the University of Oklahoma since 1932, in the capacity of associate professor and professor of electrical engineering, and is, at present, the chairman of the School of Electrical Engineering, University of Oklahoma. He has also been the technical director for radio station WNAD and WNAD-FM, University of Oklahoma, from 1932 to the present. Mr. Farrar has been a contributor to various engineering and electronic publications. At one time he was the research supervisor for research projects at the Wright-Patterson Air Force Base.

A Fellow of the American Institute of Electrical Engineers, Mr. Farrar is a member of various professional and honorary societies including, the American Society for Engineering Education, the Oklahoma Society of Professional Engineers, Sigma Xi, Sigma Tau, and Eta Kappa Nu.

Mr. Farrar became a Member of the IRE in 1945, and was elected as Chairman of the newly formed IRE Oklahoma City Section, in October, succeeding Acting Chairman O. A. Nash.

Daniel Silverman

Chairman, Tulsa Section

Daniel Silverman, Chairman of the Tulsa Section, received his B.S. degree in electrical engineering in 1927, at the University of California, where he accepted an assistantship for one year after his graduation. He then attended the Massachusetts Institute of Technology, and received the M.S. and D.Sc. degrees in electrical engineering and physics in 1929, and 1930, respectively. The following four years Dr. Silverman spent with the Westinghouse Company working in various fields such as arcs, gaseous conduction devices, and ignitrons. The next 2 years he worked with the Arma Corporation in Brooklyn in the research, development, and design of navigation and fire-control instruments.

After a year of teaching mechanical engineering at the University of California, from 1936 to 1937, Dr. Silverman was offered a position with the Stanolind Oil and Gas Company in their research department. Since 1941 he has been in charge of the geophysical research work of the Stanolind Company and at present holds the position of Exploration Research Director.

Dr. Silverman is a member of the Society of Exploration Geophysicists, the American Association of Petroleum Geologists, the Seismological Society of America, the American Geophysical Union, the American Institute of Mining and Metallurgical Engineers, the Acoustical Society of America, and the American Physical Society.

An Associate Member of the IRE in 1942, Dr. Silverman became a Senior Member in 1951. He is the first Chairman of the newly formed IRE Tulsa Section.

Concerning Research, Development, and Engineering

J. F. Byrne

Communications and electronic engineering does not end with a theoretically correct (and even highly ingenious) design of needed equipment. Unless the operational aspects, which particularly include reliability under adverse conditions, are equally considered, the engineer's task is only half done.

The following guest editorial is therefore especially timely in a troubled period like the present. It was written by a Fellow of the Institute, who is Director of Engineering in the Communications and Electronics Division of Motorola Inc.

Today economy and dependability of apparatus are particularly at a premium. Hence the following thoughts are especially directed to the attention of the readers of these Proceedings.—*The Editor.*

The rapid advances in the state of the electronics art, particularly in the last decade, are achievements of which we as electronic engineers and physicists are justifiably proud. Even a casual review of the Proceedings of the I.R.E., through this period, will emphasize the progress that has been made. Many trails have been blazed through the wilderness of the unknown. Men, spurred on by critical world conditions and the national consciousness that development and research *must* continue, are almost daily finding new areas of exploration.

Are we sufficiently aware that the fundamental benefits of exploration of the wilderness are not achieved until the highways are paved? Trail blazing, like research and advanced development, is a glamorous activity; paving highways, the reduction of ideas to everyday engineering practice, is plain hard work.

Do we really take sufficient interest in the proper, reliable *application* of new principles? Do we recognize the tremendous importance of engineering work which leads to a truly reliable vacuum tube or perhaps a radar? Are we aware of the fact that to some extent we have built a house of cards that may fall apart at the very time we need it most?

We in the IRE should recognize and point out in all ways at our command the importance of thoroughly engineered contributions as well as the discovery of new areas of scientific exploration. Achievements in production design, quality control, and dependability of either complex systems or a commonplace component are essential to our success. A redoubled effort along these lines is long overdue—let us resolve to encourage the improvement of old highways as well as the blazing of new trails.

Management Aspects of Electronic Systems Engineering*

RALPH I. COLE†, SENIOR MEMBER, IRE

Summary—**This paper cites that in the rapid growth of electronics engineering, management has not always appreciated the full magnitude of the "systems engineering" problem. In setting forth what alert management should attempt to accomplish, the systems engineering influence in research and development is discussed in some detail. In this regard particularly is it significant that more care must be given to the planning of features that make for simpler maintenance in large complicated systems. Installation problems are dealt with briefly and particular emphasis is given to the noise reduction problem within the systems in order that the best optimum *repeatable* results are obtained. Finally, the engineering management phases are treated from the standpoint of management assuming greater responsibility to see that increased engineeting emphasis is given to proper maintenance of installed systems.**

I. Introduction

That "systems engineering" is a challenging field for electronics engineers is hardly open to question. That management can and should contribute greatly to the success of "systems" thinking and operations is not always appreciated. Furthermore, the rapid growth of the electronics field has, in general, caught ever-busy engineering management ill prepared aggressively to plan for the future, and as a result they have been chiefly engaged in solving today's problem today. It is becoming increasingly apparent that the specialization of "engineering management" is essential if efficient utilization is to be made of our nation's manpower and facilities. This is doubly important in the field of electronics systems engineering.

II. Systems Engineering Influence in Research and Development

A. Previous mention of the fact that management should provide for *future planning* is apparent everywhere in the electronic systems field. In general, "planning" may be said to be the cornerstone of "systems" effort. There is rarely any disagreement that upon the soundness of advanced planning will depend in large measure the degree of success, but there is a wide difference of opinion as to how much of this effort should be programmed in advance and by whom. The arising of the need for new electronics systems has stressed the urgency of this problem since, in general, "systems" have cut across many previously considered separate specialties which previously were quite snug and complete within themselves.

B. The need for providing for future growth possibilities of electronic systems stresses the fact that separate component and equipment developments must be closely co-ordinated in order to produce modern up-to-date systems. Furthermore, obsolescence of systems can be avoided only by tying together evolutionary equipment specifications with the encompassing systems ones. These problems point up the fact that the systems engineering influence has produced a need for greater co-ordination than ever before. Likewise, increased activity in deriving new "systems standards" mutually acceptable to large groups of engineers is becoming of increased importance in the entire research and development field. It is causing a revision of many of the concepts previously considered sacred.

C. It is only natural that the tying together of many separate electronic equipments to form systems would markedly reflect in all phases of research and development. Particularly is this true when considering the problem of evaluating the success of the research and development effort. While testing of the individual equipments making up a system provide in some measure an indication of the over-all results to be expected, management should not assume that "systems" merely fit together by assemblage of individual components and equipments. A *comprehensive* systems engineering test program drawn up with greatest care is an absolute must for performance evaluation. The details of this program have previously been discussed in a separate paper.[1]

D. It has become increasingly apparent that one of the greatest influences that systems engineering will have on research and development programs will be a realization of a need for review of the entire "maintenance problem." No longer is it possible to consider that the maintenance practices previously evolved for simpler equipments will suffice for the electronic systems now under research and development. As is usually the case, necessity is the mother of invention, and with encouragement on the part of management the research and development of new maintenance techniques applicable to systems can and should be stressed. In particular, some of the newer maintenance practices in the evolutionary stage are as follows:

(1) Provision for "test points" at various places within the system to enable proper reading of all types of control voltages.

(2) "Supervisory" control from a remote point over the operations of the system including automatic measuring devices and quality evaluation. By way of example, provision can now be provided for judgment of the approximate number of hours of life left in certain elements of a system prior to the need for replacement of any element. Also possible is the provision of remote reading and recording of control parameters to the end that reliability can be judged.

(3) Individual segmentation of circuit elements to permit replacement as a unit. This enables subsequent renewal by an entire separate, similar element without the need to await repair of the original. In some designs, it may even be possible for clever engineering to so reduce the cost of manufacture of the element as to enable discarding rather than repair, in which case the technical maintenance problem is transferred back to the factory.

E. While one often hears the term "industrial designer" (as, for instance, in the automotive field), his potential influence in the research and development of electronic systems is not always realized. In one particular field, however, namely, that of radio and television broadcasting, the industrial designer entering the electronic systems field has greatly simplified control and at the same time made for a more pleasing appearance of the elements making up the complete installation. What has been achieved there is undoubtedly possible in other fields, and the electronic engineer who is not familiar with the industrial engineering practices should, before undertaking systems development, be made conversant with what can be achieved by this type of designer.

III. Engineering Problems Concerning Systems Installation and Initial Operation

A. It has previously been pointed out that "electronic systems" must be subjected to a careful evaluation program during the research and development stages. During this period, performance limits will be established and the proof of the degree of success will be the *repeatibility* of this data under actual operating conditions after the initial installation.

B. The performance of electronic systems irrespective of the exact nature or type is, in general, a function of the noise level either internally generated or induced within the system from outside radiation. In either case, one finds that the noise level at the output terminals may prove unexpectedly large if the planners have not foreseen the problems and have recognized the meticulous detail which must be followed in electronics installation work. Management should appraise these problems as having a distinct bearing upon operating success, and should in this regard be guided by the following:

(1) Keep all lengths of leads as short as possible.

(2) Design all transfer circuits to be of the lowest possible impedance since by so doing the noise pickup factor is greatly reduced.

(3) Plan for the suppression of direct radiation pickup insofar as possible by such methods as shielding and by employing as great a physical separation from the disturbing element as is possible.

* Decimal classification: R005. Original manuscript received by the Institute, April 6, 1951. Presented, 1951 IRE National Convention, New York, N. Y., March 19, 1951.

† 3151st Electronics Group, Griffiss Air Force Base, Rome, N. Y.

[1] R. I. Cole, "Management's role in research and development of electronic systems," Proc. I.R.E. vol. 38, pp. 1252–1253; November, 1950.

(4) Insure that where interconnections between equipments of a system are required that the connecting "paths" be examined, particularly when it is planned that many wires are to be run in the same cable without shielding.

(5) Insure that provision is made for adequate *grounding* of all elements of the system in order to reduce circulating currents and hence to reduce the noise level.

(6) Make adequate allowances in the planning to provide for filtering of expected noise and to provide for *extra filter elements* so that they may be on hand when final testing is to be made.

C. No discussions of systems installation and, in particular, electronic systems installation would be complete without mention of the fact that provision must be made for special operator and technician training to carry on the various specialized functions. Often this action requires management to set up special training schools as well as insuring the preparation of special handbooks covering installation practices which may be placed at the disposal of the operator and technician.

IV. Engineering Management of Installed Systems

A. Having placed an electronic system into operation, the problem of how to keep all elements functioning properly day after day now assumes greatest importance. Engineering managements must not relax in their diligence of frequently re-examining the problem of what could be called "maintenance engineering." This specialty is one of the most neglected fields and strong measures must be taken to insure greatly improved operation at lowest possible cost. The following are a few of the many problems which management should stress:

(1) Preparation of adequate systems maintenance manuals to insure that operator and maintenance personnel can understand the systems operation and can spot trouble as it develops.

(2) Establishing a system of routine preventive maintenance to insure that minimum outages will result in any given period.

(3) Establishment of a reserve supply of components which are most likely to fail during given periods of operation, often based upon an interval of time of one year.

(4) Establishment of daily or hourly log sheets to indicate performance being achieved, and thereby being able to compare at any time with that achieved previously.

V. Conclusions

The guide lines and criteria set forth herein are but a few of the many that should be used by engineering management in evaluating electronic systems engineering effort. It is apparent that there is room for tremendous growth in this particular field. Engineering management must be prepared to solve ever more complicated tasks if the great strides promised by new inventions are to become a reality and consequently be incorporated into electronic systems.

High-Frequency Gas-Discharge Breakdown*

SANBORN C. BROWN†

This paper has been secured by the Tutorial Papers Subcommittee of the IRE Committee on Education as a part of a planned program of publication of valuable tutorial material. It is here presented with the approval of that Subcommittee.—*The Editor.*

Summary—**A phenomenological description of high-frequency gas-discharge breakdown is given, describing the similarities and differences between these discharges and the more familiar dc type. High-frequency discharge breakdown is controlled by the process of electron diffusion and, besides the theory of its behavior, the physical limitations of tube size, gas pressure, and frequency for this type of breakdown are given. The particular case of hydrogen is cited. The effects of superimposing a small dc field and a magnetic field on the ac field are also discussed.**

Introduction

Knowledge of the breakdown of a gas discharge under the action of a high-frequency field has greatly increased in the last few years. The present paper is written to describe the similarities and differences between these discharges and the more familiar dc type of breakdown, discussing the physical phenomena that occur, and showing the type of information that may be obtained from these relatively simple types of breakdown.

In a high-frequency gas-discharge breakdown, the primary ionization due to the electron motion is the only production phenomenon which controls the breakdown. For this reason, high-frequency studies are much simpler than the dc type of breakdown which must always have a source of electrons present (often supplied by secondary electron effects) to make up for the electrons which are continuously swept out by the field. If one calculates the maximum kinetic energy in the oscillatory motion of an electron at the minimum field intensities for breakdown experimentally determined, one finds that this energy corresponds to about 10^{-3} electron volt. Therefore, it is obvious that the energy of oscillation is insufficient to account for breakdown.

A free electron in a vacuum under the action of an alternating field oscillates with its velocity 90 degrees out of phase with the field. It thus takes no power, on the average, from the applied field. The electrons can gain energy from the field only by suffering collisions with the gas atoms, and they do so by having their ordered oscillatory motion changed to random motion on collision. The electrons gain random energy on each collision until they are able to make inelastic (exciting or ionizing) collisions with gas atoms. The fact that the electron can continue to gain energy in the field, on

* Decimal classification: R139. Original manuscript received by the Institute, April 23, 1951.

† Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass. This work has been supported in part by the Signal Corps, the Air Material Command, and O.N.R.

the average, despite the fact that the field changes sign, can be seen by showing that the energy absorbed is proportional to the square of the electric field and, hence, independent of its sign. The rate of gain of energy of the electron from the electric field E_{rms} is $P=eEv$, where e is the electronic charge and v is the average drift velocity resulting from the applied electric field.

If the electrons are acted on by an electric field $E=E_0e^{j\omega t}$, oscillating with a radian frequency ω, in the presence of a gas in which collisions act as a viscous damping force, Lorentz[1] showed that the equation of motion may be written

$$m(dv/dt) + (m\nu_c)v = -eE_0e^{j\omega t}. \tag{1}$$

The collision frequency ν_c is the reciprocal of the mean free time between collisions and is considered to be constant. The electron velocity determined from this equation of motion can be written

$$v = -eE_0e^{j\omega t}/(j\omega m + m\nu_c). \tag{2}$$

The average drift velocity of an electron is proportional to the electric field in which it travels and the proportionality constant is called the mobility μ. Thus $v=-\mu E$, which may be written in the form of an ac mobility.

$$\mu = e/m(j\omega + \nu_c). \tag{3}$$

This expression can be compared with the case of dc electric fields where $\omega=0$, or with the case of high-pressure discharges where the collision frequency is much larger than the radian frequency of an oscillating electric field, in which one of the standard expressions for the mobility of the form $\mu=e/m\nu_c$.

The current density of n electrons per unit volume is $J=-nev$ where the velocity is given by (2). Separating the real and imaginary parts of the current, we may write E for E_{rms}, and

$$J = \frac{ne^2E}{m\omega}\frac{\nu_c/\omega}{(\nu_c/\omega)^2+1} - j\frac{ne^2E}{m\omega}\frac{1}{(\nu_c/\omega)^2+1}\cdot \tag{4}$$

The rate of energy gain of the electrons from the field is the real power going into the gas per unit volume,

$$P = J_rE = (ne^2E^2/m\nu_c)\left[\nu_c^2/(\nu_c^2+\omega^2)\right]. \tag{5}$$

It is convenient to write the electric field in terms of an effective field which would produce the same energy transfer as a dc field

$$E_e^2 = E^2\left[\nu_c^2/(\nu_c^2+\omega^2)\right], \tag{6}$$

so that the rate of energy gain of the electrons from the field is

$$P = ne^2E_e^2/m\nu_c. \tag{7}$$

Thus we see that the rate of energy gain of the electrons in the field is proportional to the square of the electric field, and therefore in a manner independent of the sign of that field.

Diffusion of Electrons

A gas-discharge breakdown occurs when the gain in electron density due to the ionization of the gas becomes equal to the loss of electrons. In a dc discharge, the electron loss is due primarily to the mobility motion caused by the steady electric field. In the ac case, the electrons are not thus swept from the discharge and their loss can be due to such phenomena as diffusion, recombination, and attachment. Under all experimental conditions so far studied of high-frequency breakdown, the loss has been due entirely to diffusion.

Diffusion occurs whenever a particle concentration gradient exists. The total flow of particles out of a volume of high concentration may be written from ordinary kinetic theory considerations[2] as

$$\Gamma = -D\nabla n, \tag{8}$$

where D is the diffusion coefficient for electrons, n the electron density, and Γ the electron current density in electrons per second per unit area. The diffusion coefficient is proportional to the average velocity of the electrons and in terms of the mean free path, l, may be written as $D=lv/3$.

We will develop the breakdown conditions for a region bounded by walls which absorb electrons. A radioactive source near the discharge tube provides a small amount of ionization S in the tube. A detailed study of the build-up of the discharge is obtained from considering the continuity equation for electrons

$$\partial n/\partial t = \nu_i n + S - \nabla\cdot\Gamma \tag{9}$$

or in terms of (8)

$$\partial n/\partial t = D\nabla^2 n + \nu_i n + S. \tag{10}$$

In this equation the term $D\nabla^2 n$ describes the loss of electrons by diffusion. The term $\nu_i n$ is the rate of gain of electrons by ionization and S is the rate at which electrons are produced by an external source. For the case of infinite parallel plates

$$\partial n/\partial t = D(\partial^2 n/\partial x^2) + \nu_i n + S. \tag{11}$$

Assuming that the approach to breakdown is so slow in time that $\partial n/\partial t$ may be neglected,

$$-S = D(\partial^2 n/\partial x^2) + \nu_i n. \tag{12}$$

This is a characteristic value problem which may be solved under the conditions that S is uniform throughout the cavity and that the boundary condition on the electron density is zero at the walls. The solution of this equation tells us that the electron density before breakdown, at any distance x from the median plane between

[1] H. A. Lorentz, "The Theory of Electrons," Leipzig, Germany; 1909.

[2] E. H. Kennard, "Kinetic Theory of Gases," p. 188, McGraw-Hill Book Co., Inc., New York, N. Y.; 1938.

parallel plates of separation L, may be written

$$n = (4S/\pi) \cos (\pi x/L)/[D(\pi/L)^2 - \nu_i]. \quad (13)$$

Breakdown can be defined by the condition that the electron density goes to infinity, which occurs when $\nu_i = D(\pi/L)^2$.

If we write $\nu_i/D = (\pi/L)^2 = 1/\Lambda^2$, for parallel plates, the quantity Λ is known as the characteristic diffusion length and is very useful in describing the shape of the gas container when discussing diffusion problems. One other example of a very useful boundary condition is the case of a cylinder of height h and radius r, with flat ends. This geometry leads to the relation that $1/\Lambda^2 = (\pi/h)^2 + (2.4/r)^2$, where the diffusion to the end plates is given by the first term on the right and the second term describes the diffusion to the cylindrical walls.

Ionization Coefficients

Gas-discharge phenomena under the action of dc fields are often described in terms of Townsend ionization coefficients. If one considers that electrons in a dc field create α new electrons in a path one centimeter long in the field direction, the increase of electrons, dn, produced by n electrons in a distance dx will be $dn = \alpha n\, dx$, and $n = n_0 e^{\alpha x}$ where n_0 is the initial electron concentration. The quantity α is called the first Townsend coefficient. This first Townsend coefficient may also be written as the ionization produced by an electron falling through a potential difference of one volt rather than travelling one centimeter. This coefficient is given the symbol η and is related to α by $\eta = \alpha/E$.

These Townsend first ionization coefficients may be given in terms of an "ionization" collision frequency. Since α is the number of electrons produced by the collisions of the primary electrons travelling one centimeter, one can write $\alpha = \nu_i/v$, where ν_i is the frequency of ionization and v is the average drift velocity of the electrons in the field. The average drift velocity $v = |\mu E|$, and one may write $\alpha = \nu_i/\mu E$ or

$$\eta = \nu_i/\mu E^2. \quad (14)$$

By analogy with the first Townsend coefficient for dc ionization where the electron loss is controlled by mobility, we may define a coefficient for high-frequency discharges[3] where the loss is by diffusion as:

$$\zeta = \nu_i/DE^2. \quad (15)$$

From our previous discussion of diffusion we saw that at breakdown $\nu_i/D = 1/\Lambda^2$. Thus we may measure the ac ionization coefficient by measuring the breakdown field in tubes of known size since $\zeta = 1/\Lambda^2E^2$. Measurements of ζ as a function of E/p, the ratio of the rms electric field, and the pressure for various gases are shown in Fig. 1.

[3] M. A. Herlin and S. C. Brown, "Breakdown of a gas at microwave frequencies," *Phys. Rev.*, vol. 74, p. 291; August, 1948.

There is a very close physical relation between the ac and the dc ionization coefficients. If one divides (14) by (15), there results the relation $\eta/\zeta = D/\mu$. Townsend showed that the ratio of D/μ was a measure of the average energy of the electrons[4] and determined this average energy as a function of E/p, experimentally.

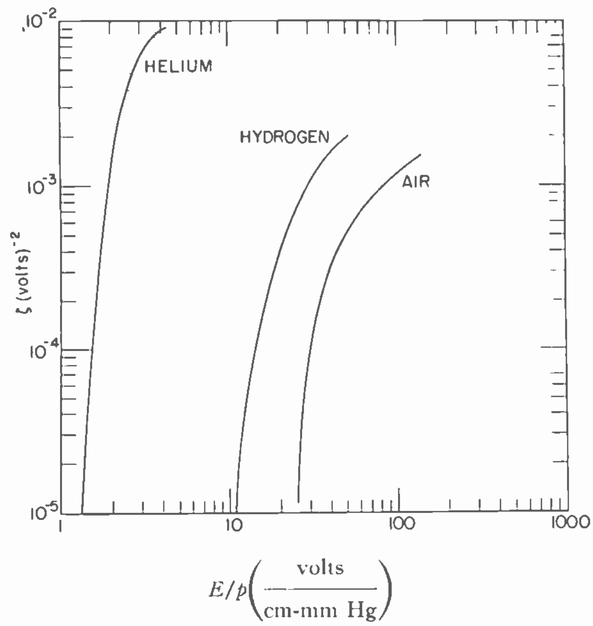


Fig. 1—AC ionization coefficient for helium, hydrogen, and air.

Thus, in principle, one can determine η from ζ or vice versa from these Townsend-like measurements. There is difficulty in carrying this out exactly, however, because the actual values depend on how the averaging of the energy is carried out. Since the electron energy-distribution functions are different for the ac and the dc cases, one might expect mathematical complications to arise if this were tried in actual cases. For the one case of hydrogen gas, where considerable simplifications are possible in the mathematics, it has been shown[5] that the dc Townsend coefficient can be determined from the ac ionization coefficient.

High-Frequency Breakdown

Typical of the behavior of the breakdown field at high frequency with changes in gas pressure are the curves shown in Fig. 2. At first sight these curves look similar to corresponding data taken with dc fields, that is, as the pressure is decreased the breakdown field first decreases and then increases again at low pressures. In the low-pressure region, the rising breakdown field with decreasing pressure in high-frequency discharges corresponds to the increasing loss of efficiency in the transfer of energy from the field to the electrons. We saw in the introduction that the electron only gained energy insofar as it made collisions with the gas atoms

[4] J. S. Townsend, "Electricity in Gases," Clarendon Press, Oxford, England, p. 166; 1915.

[5] L. J. Varnerin, Jr. and S. C. Brown, "Microwave determination of average electron energies and the first Townsend coefficient in hydrogen," *Phys. Rev.*, vol. 79, p. 946; September, 1950.

and that between collisions it oscillated out of phase with the field and hence gained no energy. Thus, as the pressure is reduced, one must increase the field to make up for the loss of efficiency by just the factor of the effective field given in (6). In the high-pressure region,

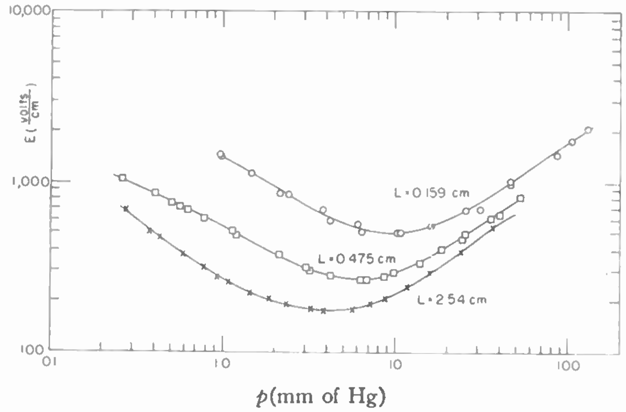


Fig. 2—Typical breakdown curves for hydrogen at a frequency of 3,000 mc.

the reason for the rising breakdown field with increasing pressure in high-frequency discharges is the same as in the dc case. As the pressure increases, the electron mean free path decreases, and the energy per mean free path decreases. Since to cause ionization the electron must gain energy corresponding to the ionization potential, more and more electric field must be applied to do this as the pressure increases. The minimum corresponds essentially to the point where the frequency of collision between electrons and gas atoms is equal to the frequency of the applied rf field. Thus, at low pressure where the electron makes many oscillations per collision, its behavior is governed by strictly ac considerations. At high pressure, where the electrons make many collisions per oscillation, their behavior is the same as in a dc discharge.

The remarkable feature of the breakdown curves, for those used to dc phenomena, is the fact that the greater the electrode spacing, the easier it becomes to cause a breakdown. This, of course, is a necessary result of the breakdown condition of the balance between energy gained from the field and electron loss by diffusion. As the electrode spacing becomes less, the diffusion loss becomes greater; therefore, the field must increase to make up for the increased loss.

Curves of gas-discharge breakdown as a function of pressure are often plotted in dc work as Paschen curves in which, for a particular gas, the breakdown voltage V is found to be a function of pd independent of the magnitude of the electrode spacing d. The same type of quantities may be introduced in the high-frequency case, where for the breakdown voltage we write $E\Lambda$, the field times the diffusion length, and for pd we use $p\Lambda$. In the case of high-frequency phenomena we have one more variable than for the dc case, namely, the frequency, and this may be introduced as the variable $p\lambda$ where λ is the wavelength of the applied field. In terms of these three variables, all high-frequency breakdown data may be put on a three-dimensional surface. This is illustrated for the case of hydrogen in Fig. 3. In this kind of a plot, breakdown data taken for a given size tube at constant applied frequency as that of Fig. 2,

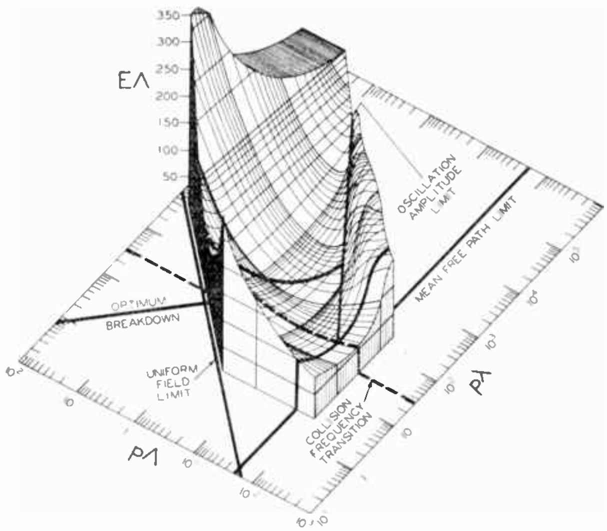


Fig. 3—A surface constructed from the experimental determinations indicated in Fig. 2, correlating the breakdown voltage measurements in terms of variables derived from dimensional analysis.

fall on lines at 45 degrees in the $p\Lambda - p\lambda$ plane. All the higher-pressure data taken above the minimum breakdown field correspond to the case of an electron making many collisions per oscillation. In this case the frequency of the applied field makes no difference, that is, the data are independent of the variable $p\lambda$ and all the $E\Lambda$ versus $p\Lambda$ curves have the same shape, corresponding to the Paschen Law curves for dc breakdown. This is shown in Fig. 4.

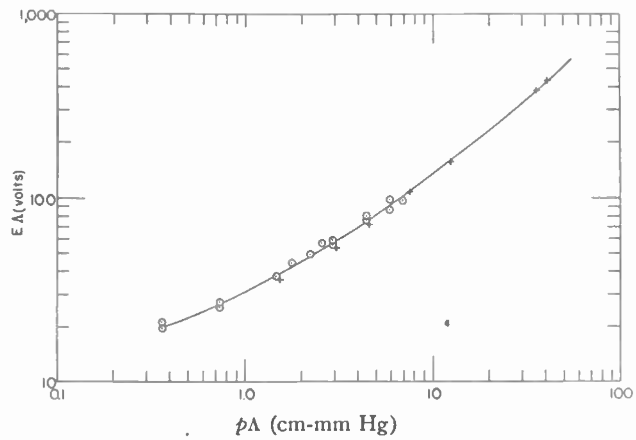


Fig. 4—High-pressure ac breakdown data showing similarity to dc Paschen law when electron makes many collisions per oscillation. Crosses are data at 10 cm wavelength (7) and circles are wavelengths of 16,700; 10,600; 8,050; 4,740; 460; and 389 cm (10).

Breakdown Theory

To derive a more quantitative description of the breakdown behavior and avoid mathematical complexity, one must choose a gas in which simplifying assumptions are most likely to apply. The two most important of these assumptions are, first, that the frequency of collision be constant in energy, an assumption we have already used in (1), and, second, that all inelastic collisions be ionizing, that is, that there be no excitation. There turns out to be no such gas in nature, but it is possible to make such a gas by adding small amounts of mercury vapor to helium. Let us coin a name of "Heg" for this gas. The helium gas by itself has the property, in the region of electron energy in which breakdown occurs, of behaving such that ν_c = constant. Helium has a metastable level at 19.8 volts since transitions from this level to the ground state by radiation are forbidden. Since the metastable states have mean lives of the order of thousands of microseconds, practically every helium atom which reaches an energy of 19.8 volts will collide with a mercury atom and lose its energy by ionizing the mercury. Therefore, each inelastic collision will produce an ionization and the effective ionization potential will be $u_i = 19.8$ volts.

Having thus found a gas to which we can apply a simple theory, let us first consider what happens at high pressure. Here the power which goes into the electrons from the electric field is dissipated in elastic collisions between the electrons and the gas molecules. This region corresponds to the lowest values of E/p measured experimentally. The data on either the Townsend first ionization coefficient or the ac ionization coefficient, such as in Fig. 1, show that here E/p is nearly constant for a wide variation of these coefficients and is equal in Heg to 1 rms volt/cm/mm Hg. Thus, for an Heg discharge, in which nearly all the loss goes into nonionizing collisions, the field and pressure are related by the equation $E = p$. It can be seen that the high-pressure breakdown measurements tend to approach this line. This is shown in Fig. 5.

In the low-pressure region, the electrons make many oscillations per collision and the breakdown field may be determined by equating the number of oscillations to ionize to the number of collisions to diffuse out of the tube. Since all inelastic collisions are ionizing ones, all the input power (which is the rate of transfer of energy) goes into ionization, this is to say that the power is the frequency of ionizing collisions, ν_i, times the energy to ionize a gas atom, eu_i. Thus we may write from (7),

$$\nu_i = P/eu_i = eE_e^2/mu_i\nu_c. \qquad (16)$$

Since we are discussing the low-pressure region, we may assume that $\omega^2 \gg \nu_c^2$ in (6) for E_e^2. In the diffusion section we found that $\nu_i = D/\Lambda^2$ and that $D = lv/3$. This leads to a relation for the frequency of ionization of the form $\nu_i = lv/3\Lambda^2$. If we multiply numerator and denominator of this expression for ν_i by the velocity and combine with (16) we obtain

$$\nu_i = v^2/3\Lambda^2\nu_c = eE^2\nu_c/u_im\omega^2. \qquad (17)$$

We solve this expression for the electric field. For ω we may write $2\pi c/\lambda$, where c is the velocity of light. The electron energy is related to the velocity by the equation $eu = mv^2/2$. Combining these, we obtain

$$E = 2\pi c\sqrt{2uu_i/3}/\Lambda\lambda\nu_c. \qquad (18)$$

The collision frequency ν_c has been measured for many gases and the results have been summarized by Brode.[6] Brode gives his results in terms of the probability of collision P_c, which may be defined as $1/pl$. The relation between the probability of collision and the frequency of collision is $\nu_c = pvP_c$, since elementary considerations show that ν_c may be written as v/l. From Brode's data we find that $\nu_c = 2.37 \times 10^9 p$. For Heg the ionization potential $u_i = 19.8$ volts; since we are assuming a very low pressure where all the power goes into ionization, the average energy u also equals 19.8 volts. Calculating the electric field under these approximations leads to a relation $E = 1284/p\Lambda\lambda$. Calculating the electric field from this relation for the two different size cavities, for which data are given in Fig. 5, shows that the experimental values do approach these theoretical lines at very low pressures.

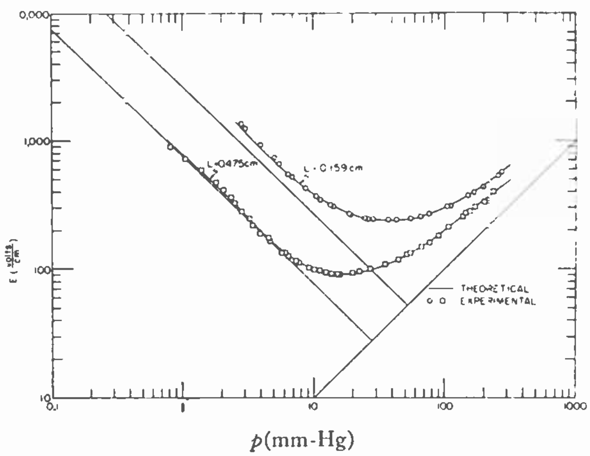


Fig. 5—Experimental breakdown electric fields compared with a simplified theory.

In this paper we have treated the breakdown problem in a qualitative manner, discussing the behavior of an average electron. This gives us less accurate agreement in detail than a more rigorous treatment, but allows us to discuss the mechanisms in a correct and more easily understood fashion. An accurate description of the breakdown phenomena can be given theoretically only by taking into account the electron energy distribution. The electron energy-distribution function may be determined by setting up the electron continuity equation, accounting for production and loss of electrons by ionization and diffusion, and expressing the results in terms of the high-frequency ionization coefficient ζ. The precision with which such a treatment predicts the experimental observations gives us confidence in the validity of the approach, but the mathematical complexity of this type of theory largely obscures a clear picture of the mechanisms involved. For

[6] R. B. Brode, "The quantitative study of the collisions of elections with atoms," *Rev. Mod. Phys.*, vol. 5, p. 257, October, 1933.

the more rigorous treatment, the reader is referred to the original papers on the subject.[3,7]

The Limits of Diffusion-Controlled Breakdown

Certain basic assumptions have been made in the calculations of breakdown as a balance between the ionization rate and the loss of electrons by diffusion. Let us examine the limits of experimental parameters beyond which diffusion is no longer the controlling phenomenon. These limits can be discussed in terms of the variables $p\Lambda$ and $p\lambda$, and we will choose the case of hydrogen since more quantitative work has been done with this gas than any other.

At low frequencies the experimental measurements of breakdown are always taken in vessels whose dimensions are small compared to the wavelength of the exciting power. For this case, an assumption of uniform field between the electrodes is very good. At very-high frequencies there exists a limit to the size of the discharge tube consistent with the assumption of the diffusion theory that there be no region of zero electric field except at the walls of the tube. This can be written in terms of the size of the tube allowable to sustain a single loop of a standing wave of the electric field. The relation between the parallel plate separation L and the wavelength is that $\lambda/2=L$. Since in terms of the diffusion length $L=\pi\Lambda$, the Uniform Field Limit shown in Fig. 3 is calculated from

$$p\lambda = 2\pi(p\Lambda). \tag{19}$$

The diffusion theory will not apply where the electron mean free path becomes comparable to the tube size. In the limiting case, this can be expressed as the mean free path l being equal to the diffusion length Λ. We have seen that the probability of collision P_c is equal to $1/pl$. To plot this condition in Fig. 3, we write

$$p\Lambda = 1/P_c. \tag{20}$$

The value of P_c is not a constant, but depends upon the electron energy. Assuming that the average electron has an energy equal to one third of the ionization potential, the average electron energy would be 5 volts for hydrogen. Using Brode's measured value for the probability of collision in hydrogen for the average electron, $P_c=49$(cm-mm Hg)$^{-1}$. With this value, we obtain the horizontal line in Fig. 3 marked *mean free path limit*.

Within the limits of experimental conditions in which diffusion theory adequately explains the breakdown behavior, several different phenomena may occur. One of the important phenomenological changes is the transition from many collisions per oscillation of the electron to many oscillations per collision. This can be written as the condition that $\nu_c=\omega$, where ν_c, the collision frequency, is the ratio of the average velocity to the mean free path, and ω is the radian frequency of the applied field. From Brode's data, we can obtain the relation $\nu_c=5.93\times10^9 p$. Putting this in terms of the proper variables, we obtain

$$p\lambda = 32. \tag{21}$$

This relation is plotted in Fig. 3 as the dotted line marked *collision frequency transition*.

We can calculate a line on the $p\Lambda-p\lambda$ plane corresponding to the minimum breakdown field for any given container size. At low pressure we have seen that the breakdown field approaches the condition given by (18). For hydrogen $\nu_c=5.93\times10^9 p$, $u_i=15.4$ volts and we again consider $u=u_i$. This leads to a value of $E=400/\Lambda\lambda p$. The limiting value of E/p for the ionization coefficient in hydrogen is 8 volts/cm/mm Hg. Therefore at high pressures $E=8p$. Eliminating the field between these two equations will allow us to calculate the pressure at which breakdown will occur most easily. In terms of the variables of Fig. 3 this leads to the equation

$$p\lambda = 50/p\Lambda. \tag{22}$$

This relation is plotted in Fig. 3 as the line marked *optimum breakdown*.

When the amplitude of the electron oscillation[8] in the electric field is sufficiently high, the electrons can travel completely across the tube and collide with the walls on every half cycle. We have already seen that under the action of the field, an electron attains a velocity $v=eE/m\nu_c$. Putting in the sinusoidal variation of the field with time,

$$\begin{aligned} v &= -(eE_p/m\nu_c)\sin\omega t \\ x &= (eE_p/m\omega\nu_c)\cos\omega t, \end{aligned} \tag{23}$$

where E_p is the peak value of the field. The limiting case on the diffusion mechanism in which all of the electrons will hit the walls would be calculated by setting the oscillation amplitude equal to one half the electrode separation. Thus the oscillation amplitude becomes equal to $eE_p/m\omega\nu_c=L/2$. Substituting λ in terms of ω, v/l in place of ν_c, and $1/pP_c$ for l we obtain

$$p\lambda = (\pi mcvP_c/e)pL/(E_p/p). \tag{24}$$

From Brode's data again we use $vP_c=5.93\times10^9 p$(sec-mm Hg)$^{-1}$, and putting in this numerical value with the parallel plate relation that $L=\pi\Lambda$, one has

$$p\lambda = 10^6 p\Lambda/(E_p/p). \tag{25}$$

This equation can be solved numerically where the experimental values of the breakdown field are available. These have been measured, and the calculation yields the *oscillation amplitude limit* of Fig. 3.

[7] A. D. McDonald, and S. C. Brown, "High frequency gas discharge breakdown," *Phys. Rev.*, vol. 75, p. 411; February, 1949; vol. 76, p. 1634; December, 1949.

[8] E. W. B. Gill and R. H. Donaldson, "The sparking potential of air for high-frequency discharges," *Phil. Mag.*, vol. 12, p. 719; September, 1931.

Phenomena Outside Diffusion-Controlled Breakdown

Many experimenters have studied rf breakdown outside the limits set by the diffusion theory. The phenomena in these regions are complicated by secondary effects similar to that found in dc discharges. For example, when the oscillation amplitude limit is crossed, the work of Githens[9] and Thomson[10] gives the sudden increase in the breakdown voltage $E\Lambda$ shown in Fig. 3. The increase probably results from the increase in electron loss because of the amplitude of the oscillation forcing the electrons against the walls and hence out of the discharge. They found that this hump, shown in Fig. 3, flattened off, and later work by Gill and von Engel[11] showed that the breakdown field thereafter decreased. This decrease, being a function of electrode material, is due to a new source of electrons entering the discharge from secondary-electron production by the electron bombardment.

Some work has also been carried out measuring the breakdown at high frequencies below the mean free-path limit.[12] Here again the phenomena become complicated by the increased loss and the introduction of secondary effects, and the problem again approaches the difficulty of a dc discharge.

Effect of a Superimposed DC Field

The gas in a tube will break down when the losses of electrons to the walls of the tube are replaced by ionization in the body of the gas. When an ac field alone is applied, electrons are lost by diffusion. When a small dc sweeping field is applied, electrons are lost both by diffusion and mobility and the ac breakdown field will increase. The breakdown condition can be formulated mathematically by a consideration of these processes.[5]

Equation (8) described the particle current flowing through a unit area due to the phenomenon of diffusion. When the electron current is due not only to diffusion but also to the motion due to an electric field, one may write similar equations for the electron loss by adding a mobility motion term. Thus we can have:

$$\Gamma = - D\nabla n - \mu E_{dc} n \tag{26}$$

and

$$\nabla \cdot \Gamma = - D\nabla^2 n - \mu E_{dc}(\partial n/\partial z). \tag{27}$$

Equivalent to (10) we may write

$$\partial n/\partial t = D\nabla^2 n + \mu E_{dc}(\partial n/\partial z) + \nu_i n + S. \tag{28}$$

In a similar fashion to the solution obtained for the electron density in (13), (28) may be solved for the case of a cylinder of axial height L and axial coordinate z, radius R and radial co-ordinate r by the method of separation of variables to yield

$$n = (\text{constant}) J_0(2.4r/R)[\sin \pi z/L] e^{(-\mu E_{dc} z/2D)}. \tag{29}$$

In this expression the term in the zero-order Bessel function represents diffusion to the cylindrical walls of the tube; the sine function represents diffusion to the end walls; and the exponential represents the deformation of the sine caused by the sweeping of electrons by the dc field. This solution is subject to the condition $\nu_i/D = 1/\Lambda_{dc}^2$, where Λ_{dc} defines a modified diffusion length given by the relation

$$1/\Lambda_{dc}^2 = 1/\Lambda^2 + [E_{dc}/(2D/\mu)]^2. \tag{30}$$

For this case, the characteristic diffusion length is given by $1/\Lambda^2 = (\pi/L)^2 + (2.4/R)^2$, which is the diffusion length previously discussed for cylindrical tubes.

The only difference between the breakdown condition in the ac-dc case and the pure ac case is the substitution of a modified diffusion length Λ_{dc} for the characteristic diffusion length Λ. It will be noted that the modified diffusion length of a cavity is smaller than the characteristic diffusion length. A cavity whose electron losses are increased by a dc sweeping field is equivalent to a smaller cavity without a sweeping field.[5]

Superimposed Magnetic Field

The discharge breakdown at high frequencies in the presence of a constant magnetic field has been studied by a number of workers[13] in the diffusion-controlled region. Let us consider the motion of an electron between collisions under the influence of an electric field along the x axis, $E = E_0{}^{j\omega t}$, and a constant magnetic field B along the z axis. The equation of motion similar to (1) is then

$$m(d\vec{v}/dt) + e\vec{v} \times B = - eE_0{}^{j\omega T}. \tag{31}$$

The solution of this equation corresponds to the superposition of a circular helical motion and a plane elliptical motion. For the helical motion whose axis is along the magnetic field, the velocity oscillates at the cyclotron frequency $\omega_b = eB/m$ and the energy of this motion is constant. For the elliptical motion the velocity oscillates at the frequency of the applied field and the energy is determined by the magnitude and frequency of the applied field.

The mean energy gain between collisions of the electrons with the gas atoms can be obtained as before, con-

[9] S. Githens, "The influence of discharge chamber structure upon the initiating mechanism of the high frequency discharge," *Phys. Rev.*, vol. 57, 822; May, 1940.

[10] J. Thomson, "Sparking potentials at ultra-high frequencies," *Phil. Mag.*, vol. 23, p. 1; January, 1937.

[11] E. W. B. Gill and A. Von Engel, "Starting potentials of electrodeless discharges," *Proc. Roy. Soc.*, vol. 197A, p. 107; May, 1949.

[12] E. W. B. Gill and A. Von Engel, "Starting potentials of high-frequency gas discharges at low pressure," *Proc. Roy. Soc.*, vol. 192A, p. 446; February, 1948.

[13] J. S. Townsend and E. W. B. Gill, "Generalization of the theory of electrical discharges," *Phil. Mag.*, vol. 26, p. 290; August, 1938.
A. E. Brown, "The effect of a magnetic force on high frequency discharge in pure gases," *Phil. Mag.*, vol. 29, p. 302; March, 1940.
B. Lax, W. P. Allis, and S. C. Brown, "The effect of magnetic field on the breakdown of gases at microwave frequencies," *Jour. Applied Phys.*, vol. 21, p. 1297; December, 1950.

sidering the input power to an electron as $P=eEv$. By a suitable averaging of the velocity and averaging the energies over the collision times, the mean energy gain between collisions can be determined as

$$u=(eE_0^2/4m)\{1/[(\omega+\omega_b)^2+\nu_c^2]+1/[(\omega-\omega_b)^2+\nu_c^2]\}. \quad (32)$$

At low pressures where the collision frequency approaches zero, the energy approaches twice the energy of the elliptical motion of the electron. At higher pressures, such that there are many collisions per oscillation, the energy of the elliptic motion loses its meaning and the collision energy becomes $eE_0^2/2m\nu_c^2$. One can use (32) to define an effective field E_e which is the rms field at high pressure, and (32) may be written $u=eE_e^2/m\nu_c^2$. This concept is useful when the collision frequency ν_c is independent of velocity since this single function takes into account the effects of frequency and magnetic field on the energy. At low pressures the effective field has a maximum at resonance with the cyclotron frequency, as shown in Fig. 6.

The electrons produced by ionization have initially very little energy, but this increases in steps of u until the energy reaches the ionization energy u_i, disregarding excitation collisions. The number N of free times to

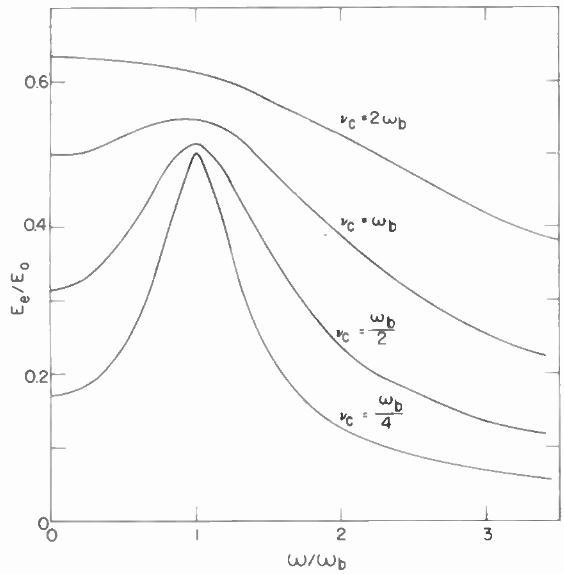


Fig. 6—The effective field as a function of the frequency showing the resonance in the presence of a magnetic field.

ionize is $N=u_i/u$ when ν_c is constant. The electrons thus double their number by ionization every N collisions; unless some equally effective process exists which removes electrons, their number will increase exponentially. Here we are considering electron diffusion to the walls of the discharge tube to be the balancing process. In the absence of the magnetic field, the random-walk theory[14] gives the mean square distance $\Lambda^2=Nl^2/3$ reached in N free paths of mean square length l^2, so that if the average electron reaches the wall in a distance Λ the diffusion process will just balance ionization. This is the condition for breakdown, and it may be written

$$u/u_i = l^2/3\Lambda^2 = v^2/3\Lambda^2\nu_c^2. \quad (33)$$

If there is a magnetic field, u will be altered according to (32). At the same time the diffusion theory must be altered to take into account the curved paths between collisions. This may be done by appropriately decreasing the mean free path or increasing the diffusion length. We shall adopt the latter and denote the new length by Λ_e, where

$$\Lambda_e^2 = [(\omega_b^2+\nu_c^2)/\nu_c^2]\Lambda^2. \quad (34)$$

The effect of a magnetic field is to make the dimensions of the tube at right angles to the field appear larger to an electron.

When the mean free path is much smaller than Λ_e, the intercollision energy gain u is correspondingly smaller than the ionization potential. From (33) we see that breakdown should occur at the same effective field if the ratio of the mean free path to the effective diffusion length is the same. That is, the effective field for breakdown is a function of $p\Lambda_e$ only. Combining (32) and (33) we obtain

$$E_e^2 = 2uu_i/3\Lambda_e^2. \quad (35)$$

We saw in (34) that the effect of a magnetic field is to make the dimensions of the tube at right angles to the field appear larger to an electron. Equation (35) shows that this should reduce the effective field for breakdown in the same proportion.

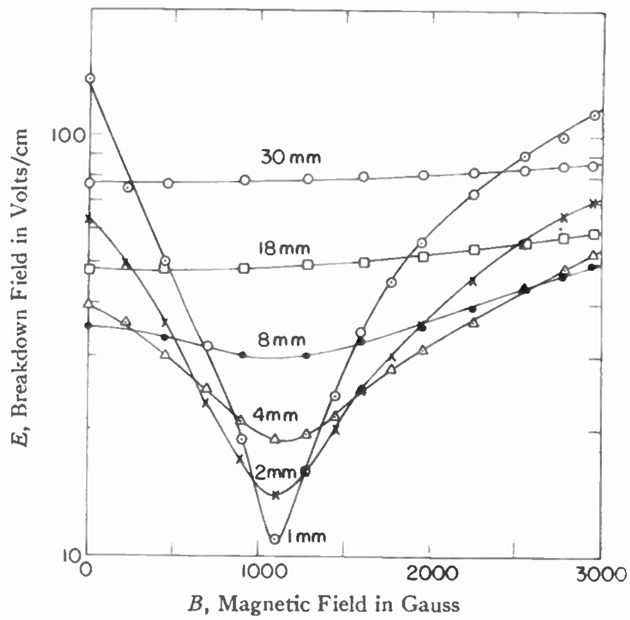


Fig. 7—Breakdown of helium in transverse electric and magnetic fields.

Experimental data for breakdown in He*g*, which again agree with our simplifying assumptions of constant ν_c and no loss of energy to excitation, are shown in Fig. 7. The result of plotting the effective field in place of the actual field would be to remove the resonance effect of the magnetic field.

[14] E. H. Kennard, "Kinetic Theory of Gases," McGraw-Hill Book Co., Inc., New York, N. Y., p. 271; 1938.

We have just been discussing the case in which the electric field and the magnetic field have been mutually perpendicular. If the two fields are oriented in the same direction, the effect is one of changing the diffusion only. Experimental breakdown curves in Heg gas for this case are shown in Fig. 8.

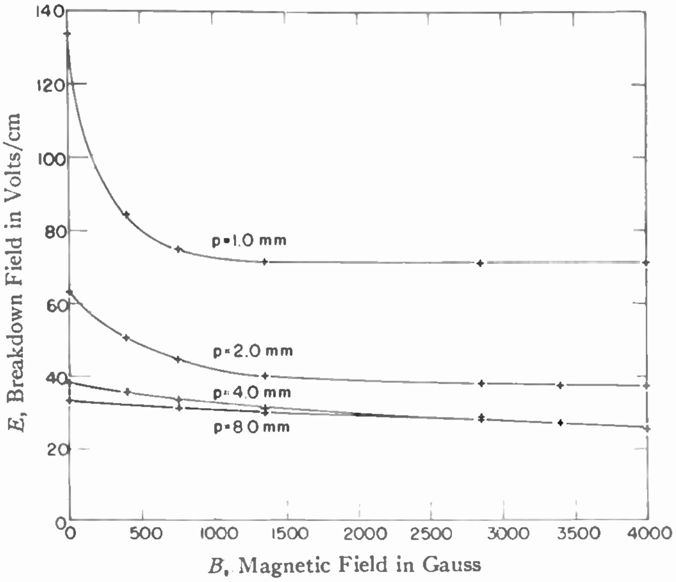


Fig. 8—Breakdown of helium in parallel electric and magnetic fields in a cylindrical cavity.

Conclusion

In summary it can be said that high-frequency gas-discharge breakdown is much simpler than dc-discharge breakdown because of the absence of necessary secondary effects. In all physical cases so far studied, the governing loss mechanism has been diffusion of electrons to the boundaries of the tubes. Although detailed calculations of the diffusion processes lead to considerable mathematical complexity, simplified theories based on the behavior of the average electron are quite adequate for many cases. Superposition of dc and magnetic fields on the hf field affect the breakdown in general as they modify the diffusion loss of the electrons.

Limits can be set on the application of the diffusion process; when breakdown is studied outside these limits, the discharge becomes complicated by the same secondary phenomena as are present in a dc discharge, even if the electric field oscillates at a radio frequency.

Simulation—Its Place in System Design*

HARRY H. GOODE†

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

Summary—**The place of simulation in system design is developed in relation to all other steps in the design process. From this, the relation of simulation to both the analytical attack, on the one hand, and to a final operational system, on the other, is developed. This leads naturally to the program to be followed in a simulation setup. In this discussion the pitfalls to be avoided are emphasized. Then simulation of a human link in a system is discussed. Finally, a critique of simulation is given with relation to cost, time, difficulty of execution, treatment of nonlinearities, and noisy inputs.**

Introduction

With the advent of the large, fast computer many extreme statements were made, pro and con, about the use of simulation in system design. Yet, like any other tool, simulation is what its user makes it. It can be a powerful aid to the skillful worker.

It is the purpose of this paper to describe this tool, expound and exemplify its use, and give it its proper place in design procedure. To avoid spending time disagreeing as to the nature of simulation, let us adopt a viewpoint to serve for this discussion alone. Simulation, we will agree, is a tool for the study of a system by the cut-and-try examination of its mathematical representation by means of a large analog or digital computer. The mathematical representation is assumed to be made up of separate expressions, each concerned with a system component. This definition will be too broad for some, too narrow for others. However, it will serve to delimit the present discussion.

Recently, in the course of the work of a Research and Development Board panel concerned with simulation, a program for system design was stated which, with some modification, will serve as an outline for this discussion. This modified program consists of

1. Description of system inputs
2. First-order design
3. Component analysis
4. Experiment for required parameters
5. System analysis
6. System simulation
7. Modification toward an optimum system by repeating steps (2) through (6).

This ordering of the design steps gives a proper perspective to the position of simulation in the scheme of design.

* Decimal classification: R004×621.375.2. Original manuscript received by the Institute, April 9, 1951. Presented, IRE Meeting at the Waldorf Astoria, New York, N. Y., March 22, 1951.

† Engineering Research Institute, Willow Run Research Center, University of Michigan, Ypsilanti, Mich.

These steps require explanation. The statement of system inputs is, of course, worth a greater expenditure of time than most give to it. Too frequently the system finally designed operates best for inputs which may never occur in practice. The expected ranges of input variables, the frequency spectra, and the multiplicity of inputs should be explicitly stated. If chance disturbances of the inputs are expected, an attempt should be made to determine the power spectrum and amplitude of the noise.

Passing now to the creation of a first-order design, this step displays the essential difference between the experienced designer and the novice. The former, for a stated set of inputs, can usually choose a reasonably adequate-looking block diagram of a system which will meet these inputs. The ability to do this is the intangible, "know-how." At Michigan we have found it desirable to stop at the first-order design and attempt a description of the proposed system operation based on a quick look at the first-order design. This may lead to obvious modifications since the system may be unable to carry out steps which seemed obvious at first.

At this point component analysis begins in earnest. The tools for such an investigation are mainly analytical. The questions to be answered are mainly subsystem questions. What input may be expected at Component X, and what will its resultant output be? If it is linear, what is its transfer funcion? If it is not linear, suppose it is linearized and describe its performance. Are tests available on this type of component; if so, how did it perform?

In the course of component analysis, experimental values will frequently be lacking. An extensive experimental program may be needed to obtain the values required to describe component performance.

While small-scale simulations may be useful at any time, advocates of the use of system simulation have often urged its use at many of the points along the route just followed. However, in the author's opinion one is not yet ready to pay the price of a system simulation. A system analysis using linear analytic methods will profitably eliminate obviously bad system setups and indicate regions in which further investigation is desirable. The analyst, if he arrives at an expression for the system at all, will be able to investigate complicated parameter regions and may even be able to approximate an optimum set of parameters. But with all this done, the system must be tried, and no complicated system can, with economical reason, be built, tried, and discarded. This latter technique, by the way, is too often used even today. Here is where simulation pays its way; its cost will be discussed later.

In the course of the above steps, the system was, of necessity, represented mathematically. Using the definition of simulation given above (i.e., that it is a tool for the cut-and-try examination of such a system's mathematical representation) the system can be set up and its performance observed. Simulation makes this possible at a reasonable cost.

Assuming the simulation completed, it is possible to designate the places at which the system fails, or performs badly. The designer may then return to the first-order design and make the necessary modifications in preparation for repeating his design steps. Of course, in practice the time order of these design steps is not as clear-cut as we have made it here. It is indeed a mythical engineering staff that has the leisure to proceed in such an orderly fashion.

It is desirable to discuss in more detail the relationship of simulation to its two chief competitors in the system-design process. On the one hand, simulation must be compared with the methods of analysis (i.e., mathematical analysis); on the other hand, it must be viewed as a substitute for designing a system by the cut-and-try or build-and-discard method.

As pointed out above, analysis is powerful and decisive in an environment which allows acceptable linearization. In some relatively simple cases it may even deal with noisy system inputs. But with growing complications in the form of nonlinearities, and even discontinuities, with a long path through many boxes for noisy inputs and with human beings in the system, the analyst must surrender. Because it deals with nonlinearities, probability, and human beings in a somewhat more realistic fashion, simulation must supersede analysis in an environment containing such elements.

At the other extreme, the construction of an operational system allows test under conditions as they actually are simply because it is the system. However, for even moderately complicated systems the cost of construction is prohibitive of trial. Moreover, the time consumed in producing the system limits the number of trials. And finally, the control of observations when the system is built is difficult.

It should be noted that with increasing complication simulation as a substitute for the two extreme attacks stands out more and more prominently as the preferred method. Analysis must be relegated to a preliminary role and operational construction to the final product since the reasons above increase in importance with complication.

When the decision to use simulation has been made, it is necessary to know something about available automatic computing machines. When this paper was planned, it was intended that a short summary of available machines would be given. However, much more on the subject may be gained from the April, 1950 issue of the PROCEEDINGS OF THE I.R.E., where a bibliography on computing machines appears.[1]

We go on, therefore, to examine in detail the execution of a program of simulation. It follows from the previous steps of system design that a detailed mathematical statement of the system has evolved. The simulation program will entail the following steps, each of which will be discussed further:

1. Choice of computer
2. Programming
3. Choice of cases to be treated

4. Coding (or circuit mapping)
5. Machine setup and debugging
6. Running
7. Data reduction
8. Analysis of results
9. Reporting.

I. Choice of Computer

In choosing a machine it will be necessary to consider the type of problem to be solved. The first choice is between analog and digital computers. Those problems which are essentially the solution of a large set of differential equations fall to the lot of the analog machine. Thus, mechanical systems, aircraft-systems designs, and energy-supply systems are all naturally suited to the analog machine. On the other hand, problems which require many discrete steps and logical choices fall into the digital computation category. Telephone systems, some systems planned for warfare, and traffic systems are digital problem types.

Suppose the machine chosen is of the analog type. The next problem will be the capacity required, a question answered directly from an examination of the mathematical statement of the system with regard to the number of operations, such as integrations, additions, etc., required in the problem. The capacities of various machines, of course, differ with regard to the type of mathematical operations which they can perform.

Having determined the capacity of the machine necessary, a joint consideration of accuracy and speed of response will be necessary. If the accuracy requirements are high, it will be necessary to pay in time of a run in order to stay within the speed of response of the machine chosen. If, on the other hand, accuracy requirements are not great, the speed of response will not be a factor and one may run as fast as the machine allows. In case the first situation holds, in which accuracy requirements are high, it may be necessary to rule out using machines with slow response because of the time which has been added to a run. It will be seen later that the number of runs will be chosen on a quite different basis. If this number is large, some interplay between these factors and the number of runs will be necessary in deciding upon a machine.

If the problem is one for the digital computer, it will be necessary to consider the problem from the point of view of initial data versus computation. If the problem is one in which a small amount of data is fed to the machine and a large amount of computation is necessary, the output being relatively small, any of the machines designed for scientific application are satisfactory. However, if the problem requires a large amount of data to be put in, a small number of computations to be performed, and a large amount of information to be put out, some of the business machines will be more to the point. Finally, a type of problem which has been developing in our work requires that a large amount of data be stored, a large amount of computation be performed, and a large amount of information be put out. The machine suited to this type does not exist at present, but it will probably be available within the next year.

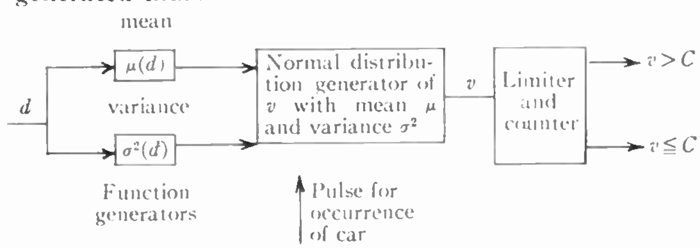
Having decided on the machine to be used from the standpoint of capacity, it will be necessary to consider accuracy requirements and speed of response, as for the analog machine. However, for the digital machine, this consideration falls most naturally under "Programming."

II. Programming

After the mathematical statement of the problem has been made in great detail, it may be found necessary to resort to simplifications. It is difficult to say without discussing a particular problem what these simplifications may consist of. They crop up in various contexts. Perhaps in later papers some account will be given of these simplifying methods. One broad principle in connection with those problems containing probability considerations is the following:

Where a set of boxes, to which the input is a given probability distribution, can be represented as an operator in the mathematical sense so as to provide the distribution of the output, the entire set may be substituted by directly generating, equipment-wise, the probability of the output.

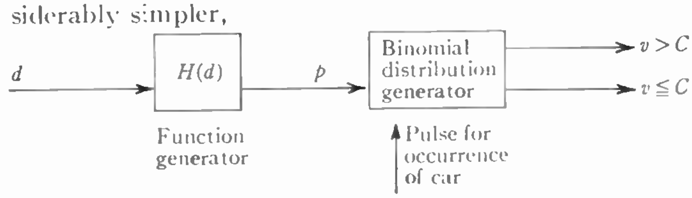
Suppose, for example, a traffic problem is being studied in which the occurrence of motorists' speed (v) is normally distributed with a mean and variance dependent upon the distance from a traffic intersection. Suppose further that the problem requires that each motorist traveling above a speed C will be stopped. Equipment-wise, this element of the problem may be generated thus:



However, a simple analysis show that $v\leq C$ will occur with probability

$$p = \int_{-\infty}^{C} N(\sigma,\mu)dv = \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{(C-\mu)/\sigma} e^{-y^2/2}dy = f(\mu,\sigma) = H(d)$$

so that we may substitute this diagram, which is considerably simpler,



When the mathematical setup has been simplified as much as possible, it will be necessary to consider various computational methods. In the analog computer this is generally a straightforward affair with components corresponding to mathematical operations. In the digital computer there will, in general, be alternatives for a

[1] "Radio progress during 1949," Proc. I.R.E., p. 373; April, 1950.

particular computational method. Thus, for an integration one may resort to an extremely short interval and use a simple tangent approximation of a function to obtain required accuracy. On the other hand, more sophisticated methods entailing high-degree polynomials and iteration may be used, resulting in a larger interval being permissible but requiring a larger number of computation steps within the interval. It is something of a problem to balance interval size and sophisticated methods against one another so that an approximation to the minimum number of computations for a given accuracy may be obtained.

Finally, with regard to programming, it is to be noted that the detailed consideration of the computational methods to be employed will suggest to the designer questions about his original design and sometimes make it possible for him to better his original system merely from a consideration of these methods. This will occur most frequently when the system itself contains many computational processes. The attempt to simplify the simulation computation will result in a simplification of the actual system computation.

III. Choice of Cases

The next steps in the execution of the program are the choice of the cases to be run, measures of effectiveness to be used, methods of reduction of data, and the determination of the total number of runs to be made. In choosing the cases to be run, nothing will substitute for experience in doing a problem. However, some general notes are as follows:

The designer should permit great flexibility in changing these runs when, in the course of examining the outputs, necessary changes become evident. There is some tendency to set up a program of runs and drive it to the bitter end in spite of the fact that the designer knows, early in the game, that he will have to change the direction of his investigation.

Again, before running on a full program it is desirable to sample a few points. This may result in a complete shift of the emphasis, which was chosen on an *a priori* basis as the result of analysis. In choosing values in a given variable range it is desirable to have at least three values. These, as nearly as possible, should be located at the minimum and maximum reasonable values and at a guessed optimum. From these values, one obtains some knowledge of the curvature of the function being used as a measure of effectiveness.

The smaller the number of design variables permitted the easier will be the execution of the program. Where possible, consideration of some design variables should be eliminated on the basis of the preliminary analysis. Otherwise the number of runs required becomes excessive, even for today's fast machines.

With regard to measures of effectiveness, it is desirable to use as few as possible for the evaluation of the results. To a large extent, the choice of these measures is inherent in the physical problem. However, cases have occurred where redundant measures of effectiveness have been used. It is worth the time necessary to eliminate such redundancies since they serve only to confuse the designer when he evaluates his results. It is desirable that measures of effectiveness be dimensionless. The purpose of this is to enable the designer to extrapolate from the results of cases run to those cases which have not been run, but whose parameters in dimensionless form are the same as those already run. Where possible, the quantity used should have a readily apparent meaning.

It will pay the designer to know what his final answers will look like. The effort expended toward this end will minimize the number of unnecessary steps taken in the course of simulation. Moreover, knowing the required answer leads immediately to a choice of the variables to be recorded, which in any machine of modest size is by no means a trivial problem. Recording facilities will, in general, be found to be limited, and a well-considered choice of variables will be profitable. At this point, it is well to consider also the possibility of cutting down human computation by making the outputs of the machine as nearly like the final form as possible. This will not always be permissible because of the amount of equipment used up in relatively unnecessary computations; however, where such equipment is available, much will be gained by putting out final results directly from the machine.

The number of runs to be made at a parameter space point in a system subject to chance occurrence is a difficult question to answer without some experience with the problem at hand. The designer usually has no time for a detailed statistical analysis by which the significance of the values he is obtaining may be determined. Such a statistical analysis is frequently a larger problem than the system analysis. However, there are some compensating factors. It is found that in the ordinary run of problems convergence toward a stable value for the measure of effectiveness is fairly rapid. In our experience about 70 runs at a point determine the required function at that point within adequately close limits. If there are no probability considerations in the problem, only one correct run will, of course, be necessary at any point.

IV. Coding (or Circuit Mapping)

The definition of "circuit mapping," as used here, applies to an analog machine, and is that step in the computational procedure concerned with the choice of the particular circuit to be followed through the machine. There is a choice of using highly trained or technical personnel to execute this step, with a resultant saving in equipment for performing a particular operation, or of using personnel without such technical qualifications and thereby absorbing equipment at a much higher rate. The designer will have to make this choice on the basis of the personnel and facilities available to him.

A careful look at the circuit map will prevent hasty conclusions concerning what the machine is actually computing. A detailed examination of the circuits being used may prevent many hours of search and wonder. A technique which has been found useful in our work has been the maintenance of a one-to-one relationship between the boxes in the actual system and the diagram of the computational components used in the machine. This has two useful results: It enables the trouble sites to be located more easily and it allows the designer to visualize, with greater ease, what is happening in the simulator.

The corresponding operation to circuit mapping for the analog machine is coding for the digital machine. Here too, time spent in checking and rechecking the coding will pay off in less time consumed for false runs.

With regard to probability operations in the digital machine, a number of techniques have been evolved for the generation of random numbers. There is a much better control of the distribution generated in a digital machine than on the analog machines where one is at the mercy of physical processes for generating such distributions.

V. Setting up the Machine

Where possible, a removable plug board, such as that which the Reeves people have recently added to the REAC, should be employed on an analog machine. This arrangement is desirable from the point of view of an organization concerned with several problems. It makes possible having two or more problems set up on the machine. Time may then be shared among the problems. It is then possible for engineers on one problem to examine results while another problem is being run.

The choice of display from the machine is of some importance. We have found it useful to make this display as graphic as possible in terms of the geometry of the real system.

VI. Running the Problem

Notes about running the problem are as follows:

When possible, data reduction should be carried on as the machine completes each run; the results may then be monitored for any required re-orientation of the program. The design engineer can frequently spot an obvious difficulty or need for change, thus reducing the number of runs and, therefore, time and cost expended.

The time per run may sometimes be cut by running the problem on a fast time scale. In case such a procedure is adopted, it is necessary to consider carefully the operations involved so as to be certain that the actual system conditions are not being violated.

It will be necessary to have some kind of check to determine if the machine is turning out correct results. In cases where probability is not a consideration, hand-computed solutions for relatively simple conditions will furnish a clue to whether the machine is operating properly. Where probability enters into the system, it will be necessary to check the problem with a "no noise" condition and to accept the results of probability runs as correct when those with no noise have turned out correctly. For this purpose, the outputs of the original probability distribution generators should be recorded and frequently tested.

VII. Data Reduction

As pointed out above, it is desirable to know what the final answers will look like before the runs begin. It is even more desirable to choose the methods of data reduction in advance. These methods may run all the way from a simple tabulation of the result turned out by the machine to a complicated graphical and analytical method of obtaining the measure of effectiveness.

In large organizations, it is administratively desirable to clarify the responsibility involved in following the simulation through the steps outlined above.

For completeness, the steps of analysis and reporting have been added to the outline given above. These two steps will require the direct attention of the system designer since they lead directly to the future course of action in the system design.

A subject deserving special discussion in connection with simulation is the human being treated as a simulated member. Several levels of realism required may be distinguished. For those systems in which decision making is his function, it is frequently satisfactory to represent the human being's response by a time delay. In air traffic control, for example, the response of a pilot to a voiced command will be delayed. A normal distribution of delay times about a measured average may be adequate.

If his response is functional, the problem becomes difficult and attempts to represent his presence have been weak. The reasons for this are threefold: The human being is a nonlinear device whose functional response is difficult to obtain; his functional response is dependent upon his recent past history; even where response is a reasonably well-known functional form, numerical values are difficult to obtain.

This state of affairs usually leads to an attempt at simulating the system and putting a real human being into it—offering him the proper operational stimuli and furnishing him with operational controls for response. It is frequently not realized, after all this trouble is taken, that one or two human beings do not adequately represent the population from which the ultimate system operators will be drawn. It is desirable in those cases in which the human being is a factor to consult with both the psychologist and the statistician.

Finally, it is desirable to mention some of the frequently quoted remarks about simulation and discuss their implications:

1. "Simulation is costly": Simulation costs have been estimated at a fraction of a cent per run, and an actual cost of $15,000 a run has occurred in a recent government-sponsored experiment. But the correct answer on cost is a complicated one. It depends on the problem, on the extent of realism required, on the familiarity with the system, on the reliability of the equipment used, etc.

However, no one now doing simulation would suggest an order of magnitude under dollars per run, and probably no one would accept as necessary over $100 per run. The way to measure cost is not in absolute number of dollars but in the cost of doing without. At its worst, this may mean the building of a useless system.

Trajectory calculation is a fair measure of run cost on some typically complicated current systems. Rough estimates of these costs for a few machines are: differential analyzer at M.I.T.—$30, Cyclone at Reeves Instrument Corporation—$15, Meteor at M.I.T.—$25, and Mk I at Harvard—$30.

2. "Simulation is easy": Simulation is not easy; it requires deep understanding of the problems being solved, the booster of simulation notwithstanding. Just as the physicist must decide what factors are *a priori*, irrelevant to his experiment, so must the system designer be able to point out simplifications which will lessen the detail in setting up the system, but will not vitiate the results.

In this connection, it has been found desirable in our work to state explicitly the ways in which the simulation we undertake differs from the real system. This has two useful purposes: a. The limitations of the results are expressly stated and conclusions are, therefore, warily drawn. b. When these limitations are too great, a solution to the obvious problem of the need for increased scope, complexity, and realism of the simulation is sought.

3. "Simulation complexity vitiates results": This can certainly be true if care is not exercised. However, two notes are of importance here: First, the complicated group of components may frequently be substituted by the result of a previous analysis and simulation of the group of components. The greatest exercise of both care and ingenuity are necessary in this regard. Secondly, the performance of analog gear in this respect has amazed those using it for large and complicated systems. It is true that an inherent accuracy of 0.1 per cent of full-scale reading for any one component is all that the average user can afford. It is also true that errors in such components, if used in an open-ended chain, will pile up according to known laws. However, in most systems of complexity many closed loops occur, and the system itself is usually a large closed loop. This results in our experience showing an error of less than 0.5 per cent of full scale maximum allowable answers due to computation for problems involving as many as 90 operational amplifiers (or, put another way, in computations with upwards of 100 fundamental mathematical operations).

4. "Simulation is fast": To the engineer waiting for results on a system simulation, it is certainly not fast. Yet the true measure of speed lies in what comparable operational tests of equipment, if available, would require in time. Then, no matter how slow the computer, no matter how many bugs occur in the setting up, and no matter how many human computers make errors in reducing data, simulation is fast—at least tenfold over operational attempts.

Of course, accuracy may be paid for with time, as usual. In the analog setup, if the system input contains high frequencies, a slow time scale is necessary to allow computational components time to respond. With the digital computer, the payment is again in time, with the necessary iterations for integration accuracy, or the choice of a small-time increment for high-frequency response. But over-all simulation is fast.

It should now be clear that, without minimizing its difficulties, simulation is a powerful tool.

The Civil Aeronautics Administration VHF Omnirange*

H. C. HURLEY†, ASSOCIATE, IRE, S. R. ANDERSON†, ASSOCIATE, IRE, AND H. F. KEARY†, ASSOCIATE, IRE

***Summary*—This report describes the development of a vhf-omnirange system, operating in the frequency band of 112 and 118 mc. The range furnishes magnetic-bearing information with respect to the range station, and provides definite track guidance between the station and any point within its service area.**

The omnirange produces two 30-cps signals: one is constant in phase and independent of the aircraft's position, and the other varies in relative phase directly in accordance with the magnetic bearing of the aircraft from the station. A phase-measuring device in the receiver enables the pilot to determine his magnetic bearing with respect to the station, and to select and fly a range course on any desired magnetic bearing.

The accuracy of the system, including the receiver plus ground equipment, in operation at the Civil Aeronautics Administration Experimental Station, Indianapolis, Indiana, is approximately 1.5 degrees. The ground equipment has been in continuous operation for more than 3 years.

The vhf omnirange has been selected as part of the common Civil-Military System, Transition Program, as recommended by the Radio Technical Commission for Aeronautics.[1]

* Decimal classification: R526.12. Original manuscript received by the Institute, October 2, 1950; revised manuscript received, April 16, 1951.

† Civil Aeronautics Administration Technical Development and Evaluation Center, Indianapolis, Ind.

[1] Report on Air Traffic control, prepared by Special Committee 31.

Introduction

In PLANNING a system of airways as a means of navigation for all aircraft and facilities for the separation and maintenance of a continuous flow of a large volume of traffic, it becomes apparent that radio ranges defining courses or tracks in only a few directions are inadequate.

The CAA has carried out the development of two-course visual, two-course aural vhf ranges in accordance with recommendations made by the Radio Technical Commission for Aeronautics in the summer of 1939. Development was completed early in 1942, but the conversion of the airways to vhf-range operation was interrupted by World War II. Just prior to 1945, however, the conversion program was resumed on a limited scale with the establishment of the Chicago-New York vhf airway. Simultaneously, the CAA had been working on the development of a vhf omnidirectional radio range, along the principles proposed by D. G. C. Luck of RCA, at the Kansas City meeting of the RTCA in the summer of 1939. Progress on this development has taken such a promising turn that two-course vhf ranges were rapidly superseded by omnidirectional vhf ranges, more commonly referred to as "omniranges."

The general principle of the omnirange dates back to Lee de Forest, who proposed rotating a radio beam keyed to identify the sectors forming the 360-degree sweep of the beam.[2] Another omnirange was developed, using a rotating beam; however, an omni-identification signal was transmitted each time the beam passed through north.[3] The beam rotated clockwise so that the time interval between the omnisignal and the beam alignment with the observer, measured by means of a stop watch, would be a measure of the observer's azimuth. A practical form of this type of range was developed in England by rotating a loop antenna.[4] Later this rotation was accomplished electrically.[5] A purely electrical means of rotating a field pattern for use in an omnirange has been proposed.[6,7,8] The omnirange developed by RCA was described in a paper by Luck.[9] In a paper[10] presented before the RTCA annual meeting in January, 1945, Stuart, Director of the Office of Technical Development, described an early model of the CAA omnirange.

The present omnirange has elements in common with other types of radio ranges and direction-finding systems, and represents a refinement in practice rather than a fundamental departure in principle. It may be likened to the standard four-course aural radio range, with a means provided for precisely determining the ratio of the two signals at any point and thereby the azimuth from the station at any point, rather than at points lying only on certain specific courses where the signal ratio is unity.

The antenna system, which is so widely employed in direction-finding and radio-range systems, is the familiar arrangement of five radiating elements located at the corners and the center of a square. Opposite pairs of antennas are operated 180 degrees out of phase, and the electrical spacing between the elements is small compared to the wavelength, resulting in a figure-of-eight field pattern. This field pattern is rotated, so to speak, by means of a capacity goniometer, which is driven by a synchronous, 1,800-rpm motor. The rotating goniometer acts as a balanced modulator, eliminating the carrier frequency and supplying sideband energy at a carrier frequency ± 30 cps to the two antenna pairs. Since the entire field is, in effect, rotated once for each rotation of the goniometer, it is obvious that each direction in space will have a certain phase of the rotational frequency associated with it, and that this phase will change degree for degree with a change in azimuth relative to the station. If we have a signal supplying a 30-cps voltage of reference phase which is independent of the azimuth, we can determine the azimuth from the station by comparing the phase of the two 30-cps signals.

The purpose of this report is to present a description of the latest CAA omnirange system and to indicate what improvements have been made over the earlier model.

Ground-Station Equipment

General

A block diagram of the essential components of the ground-station transmitting equipment is shown in Fig. 1. A conventional vhf transmitter has its final stage

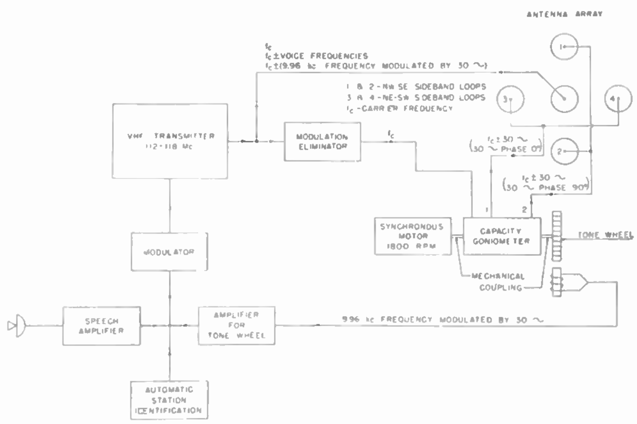


Fig. 1—Block diagram of vhf omnirange transmitting equipment.

amplitude modulated by a subcarrier and simultaneous voice. The subcarrier itself is frequency modulated at 30 cps. The output from the final stage is fed to the center antenna of the five-loop array and also to the input of the modulation eliminator. A goniometer functions as a mechanical sideband generator and delivers 30-cps sideband energy to its output circuits. Output

[2] United States Patent No. 833,034; 1906.
[3] J. Zenneck (Trans. Seelig), "Wireless Telegraphy," McGraw-Hill Book Co., New York, N. Y., p. 368; 1915.
[4] T. H. Gill and N. F. S. Hecht, "Rotating loop radio transmitters and their applications to direction-finding and navigation," *Jour. Inst. Elec. Engrs.*, vol. 66, p. 256; 1928.
[5] H. A. Thomas, "A method of exciting the aerial system of a rotating radio beacon," *Jour. Inst. Elec. Engrs.*, vol. 77, p. 285; 1935.
[6] P. H. Evans and J. W. Grieg, United States Patent No. 1,933,248; 1933.
[7] J. W. Grieg, United States Patent No. 1,988,006; 1935.
[8] G. H. Brown and D. G. C. Luck, United States Patent No. 2,112,824; 1938.
[9] D. G. C. Luck, "An omnidirectional radio-range system," *RCA Rev.*, pt. I, vol. VI, no. 1; July, 1941; *RCA Rev.*, pt. II, vol. VI, no. 3; January, 1942; *RCA Rev.*, pt. III, vol. VII, no. 1; March, 1946.
[10] D. M. Stuart, "The omnidirectional range," Presented before the RTCA Annual Meeting; January, 1945; later published in *Aero Digest;* June 15, 1945.

Number 1 of the goniometer delivers 30-cps sideband energy of zero-degree audio phase to one pair of diagonally opposite loop antennas. Output Number 2 of the goniometer feeds 30-cps sideband energy of 90-degrees audio phase to a second pair of diagonally opposite loops. The four antennas are arranged in the form of a square around the center loop antenna and are fed in such a manner as to produce two figure-of-eight field patterns. The two figure-of-eight patterns are in phase quadrature in both space and time, and can be visualized as a single figure-of-eight pattern with a positive lobe and negative lobe rotating at an audio rate of 30 cps, which produces amplitude modulation (in space) of the main rf carrier transmitted from the center antenna. This energy, when detected in a suitably designed receiver, is known as the "variable-phase" signal. The 30-cps signal produced in the receiver from the frequency-modulated subcarrier is termed the "reference-phase" signal. The purpose of the subcarrier modulation is to provide a means of discriminating between the two 30-cps voltages in the receiver.

Fig. 2—Omnirange ground facility.

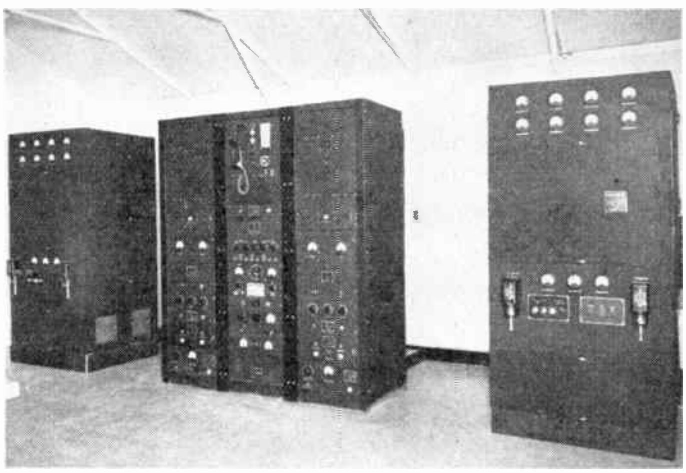


Fig. 3—Omnirange transmitting equipment.

Fig. 2 is a view of a CAA omnirange ground facility using a 30-foot tower. The antenna array is situated inside the cylindrical housing on the tower. The transmitters and associated equipment located in the building on the ground are shown in Fig. 3. Approximately half the equipment is "standby" to permit maintenance while providing continuous service.

Antenna Array

The polarization most desirable for an omnirange operating in the 100-mc region of the radio spectrum was determined by experiment. The maximum amount of scalloping[11] was ± 3 degrees for the vertically polarized arrays and ± 0.7 degree for the horizontally polarized array. The site used for these tests was considered to be about average. Trees probably caused the scalloping which was observed.

The polarization used in the present omnirange system is horizontal, as a result of conclusions drawn from the early comparison tests. Fig. 4 shows the arrangement of the loop radiators and the approximate shape of the horizontal-plane field patterns. The pair of antennas with currents I_1 produces the figure-of-eight pattern E_{SB1}. Similarly, the pair of antennas with currents I_2 produces the pattern E_{SB2}. The center antenna with current I_A radiates the E_c pattern. The three patterns combine in space to produce an amplitude-modulated wave whose modulation frequency is $\rho/2\pi$. The most significant characteristic about this amplitude-modulated wave is the phase angle of the modulating frequency, which varies as the azimuth angle, ϕ, for small values of spacing, S.[12]

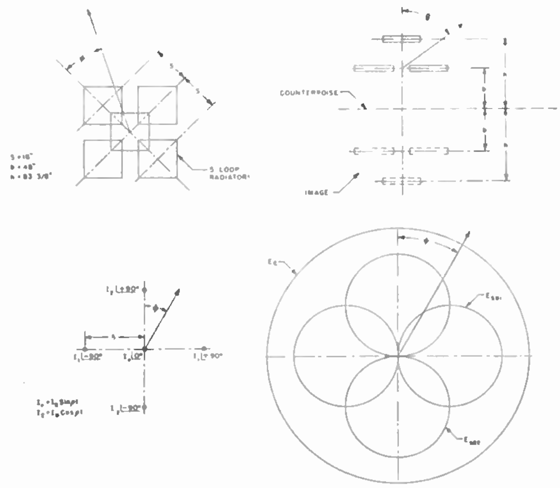


Fig. 4—Mechanical layout, currents of radiators, and theoretical horizontal-plane field patterns.

The diagonal spacing between each pair of sideband antennas is 112 degrees at midband, 115 mc. Each sideband antenna is mounted on top of a pedestal 168 degrees above the counterpoise. The carrier antenna also is supported by a pedestal 222 degrees above the counterpoise. The spacing between adjacent sideband an-

[11] Scalloping is defined as the amplitude of the course-deviation indicator's fluctuations due to the effect of reflections from trees, buildings, wires, the earth's surface, and the like.

[12] H. C. Hurley, S. R. Anderson, and H. F. Keary, "The CAA vhf omnirange," CAA Technical Development Report No. 113.

tenna sides is 10.5 degrees, and the sideband antennas are symmetrically disposed around the carrier antenna. Three rf receptacles are mounted directly under each loop transposition. Two of the receptacles are connected directly to the transposition. RF power is fed to the loop antennas through Type RG 8/U, solid, dielectric coaxial cable. Since the loop antenna is a balanced circuit, it is necessary to convert from unbalanced to balanced line in order to feed the loop in the proper manner. The conversion is accomplished by the use of an 180-degree section of unbalanced line.

Diagonally opposite pairs of sideband antennas are connected through 350-degree lengths of cable and are excited 180 degrees out of phase. This line length was found to be optimum at 115 mc for minimum parasitic currents in the sideband antennas, induced therein by the center antenna. The 180-degree phase relation is obtained by reversing the connections at the transposition in one loop antenna of each diagonal pair. One pair of sideband loops is fed from the goniometer output Number 1 and the second pair of sideband antennas from output Number 2. The center carrier antenna is fed directly from the transmitter output stage by coaxial cable.

Tests conducted to determine the operating characteristics of the vhf loop antennas included field measurements to obtain the horizontal-plane patterns, impedance measurements to ascertain impedance matching requirements, rf-current measurements to determine current distribution and balance around the periphery of the loop, and measurements of parasitic currents due to interaction between loops grouped together under normal operating conditions.

During the development of the omnirange it was found that the antenna shelter will produce errors if it is not cylindrical in shape (see Fig. 5). A square house was found to produce an octantal error curve of ±1.5 degrees.

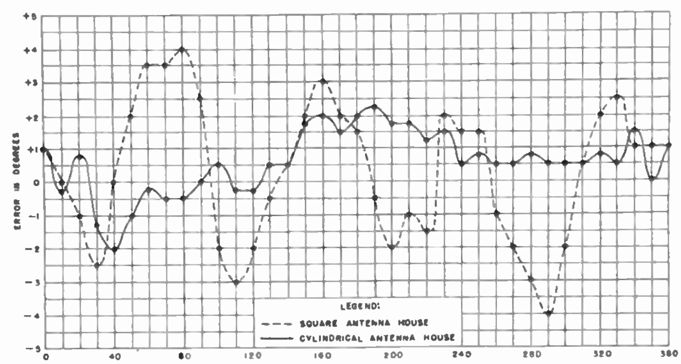


Fig. 5—Flight-calibration error with cylindrical and square antenna houses. Bearing selector in degrees.

Fig. 6 shows the vertical-plane field-strength pattern of the present omnirange antenna array mounted on 15- and 30-foot high counterpoises. The measurements were made from an airplane as it flew away from the station, and include the pattern of a horizontal-V receiving antenna located on top of the vertical stabilizer of a DC-3 airplane. It will be seen that the lobes, at small values of θ, are essentially independent of counterpoise height because the energy is reflected from the counterpoise instead of the ground. On the other hand, the lobes at large values of θ are dependent upon counterpoise height since little energy is reflected from the counterpoise at these angles and the earth reflects a considerable signal.

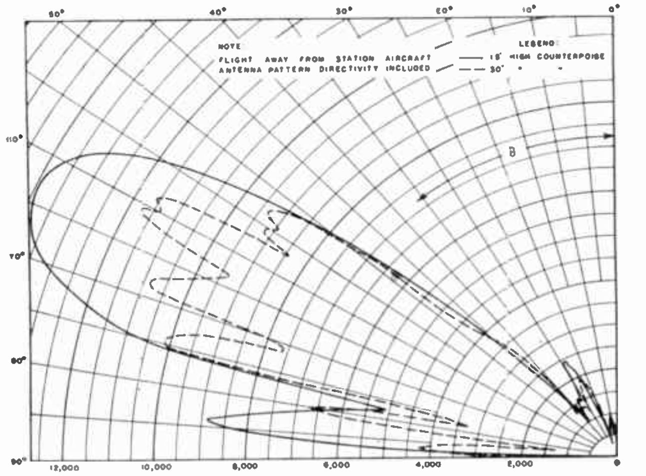


Fig. 6—Measured vertical-plane field patterns of antenna array mounted on 15- and 30-foot high counterpoise. Relative field strength.

Goniometer and Subcarrier Generator

The goniometer consists of two mechanical sideband generators on a common drive shaft, rotating at 1,800 rpm. One sideband generator is oriented 90 degrees with respect to the other so that the electrical outputs of the two generators are in phase quadrature at the modulation frequency, as shown in Fig. 7(a). The important

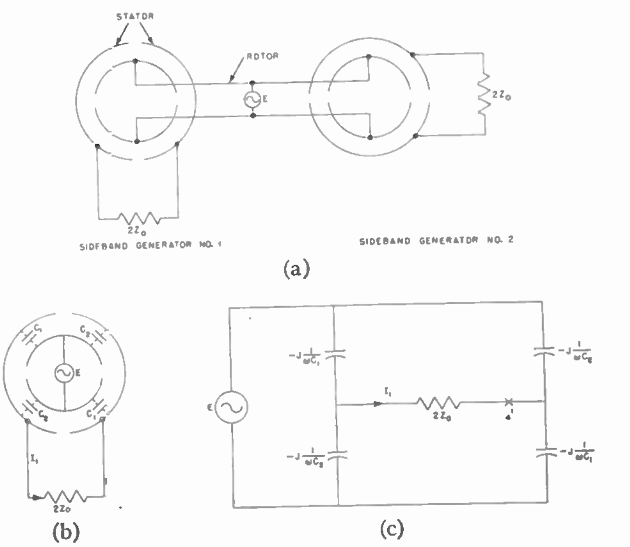


Fig. 7—(a) Arrangement of stator and rotor plates of goniometer. (b) A view of one-sideband generator of the goniometer showing its capacities. (c) Schematic diagram of (b).

capacities of one of the sideband generators are shown schematically in Fig. 7(b). Fig. 7(c) is a schematic diagram of sideband generator Number 1.

It is demonstrated in another report[12] that the currents flowing in the loads $2Z_0$ of Fig. 7 are of two frequencies, one higher and the other lower than the carrier frequency by the goniometer rotational frequency. Also, the Number-1 and Number-2 outputs consist only of sidebands that are 90 degrees out of phase at the modulation frequency.

Fig. 8 shows a view of the goniometer. Two rotors, each having four plates, are mounted one at each end of a longitudinally split rotor shaft to form one section of the sine-wave modulating capacitors. The rotor is driven at a speed of 1,800 rpm by means of a synchronous motor. The stators consist of five plates each, angularly displaced by 90 degrees. The goniometer receives unmodulated rf power from the modulation eliminator and delivers two rf outputs having amplitudes proportional to the sine and cosine of the rotational angle of the goniometer. Transmission-line balance sections are required to permit operation of the balanced goniometer from the concentric coaxial cable. The medium for coupling the rf carrier from the modulation eliminator to the sine-wave modulating capacitors consists of a pair of rotor plates, located near the center of the rotor shaft and in close relation to a set of stators, each consisting of two plates. Faraday plates prevent capacitive coupling between the modulating stators and the coupling stators. A 360-degree dial with a vernier indicator makes it possible to determine the rotor position. The dial is adjusted to indicate zero when the rf power from one of the two output circuits is a minimum.

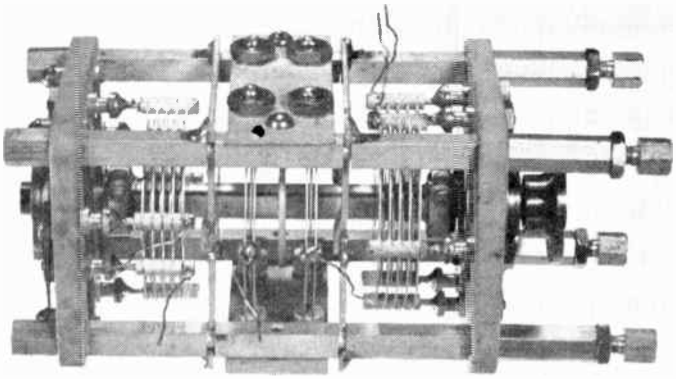


Fig. 8—View of goniometer with shield plates removed.

The input impedance to the goniometer is 40 to 60 ohms resistive, and is -10 to $+10$ ohms reactive, over the frequency range of 112 to 118 mc. The input is resonant at 115 mc. The rated rf power input is approximately 100 watts. Practice has shown that sufficient accuracy can be obtained if the null locations are 90 degrees apart ± 2 degrees, and the maxima are equal to within ± 5 per cent. The waveform should be a true sine wave with not more than 5-per cent harmonic distortion.

The 9.96-kc subcarrier, which is frequency modulated at 30 cps, is produced by an electromechanical subcarrier generator in the form of a tone wheel. The subcarrier generator and goniometers are driven on a common shaft. Teeth, machined into the periphery of the wheel, vary the magnetic field during rotation, producing a voltage in a pickup coil.

The output from the pickup coil is somewhat more than 0.6 mw into a 600-ohm load. An accurate phase adjustment over a range of ± 10 degrees is provided to permit exact adjustment of the phase of the reference phase signal with respect to the variable phase signal. This adjustment is accomplished by moving the yoke containing the pickup coil. The wheel provides a reference-phase signal that can be securely locked in step with the 30-cps variable-phase signal.

Modulation Eliminator

The modulation eliminator was developed to remove the 9.96-kc subcarrier modulation and voice modulation from that portion of the omnirange transmitter output which is fed to the goniometer. Provision is made for proportioning the transmitter output between the carrier and sideband antennas.

Referring to Fig. 9(a), let us imagine a bridge circuit made of three equal lengths of coaxial transmission line and a fourth length which is 180 electrical degrees longer than the others. The transmitter supplies power to the bridge at the junction marked A. The other three terminals of the bridge are connected to impedances Z_0, Z_1, and Z_2, which are normally pure resistances. Z_0 represents the load impedance presented by the capacity goniometer. The characteristic impedance of each of the transmission lines of the bridge is also Z_0. When $Z_2=Z_1$, the bridge is balanced. A voltage E_s at point A sets up traveling waves in the branches ABC and ADC, which arrive at point C equal in magnitude and opposite

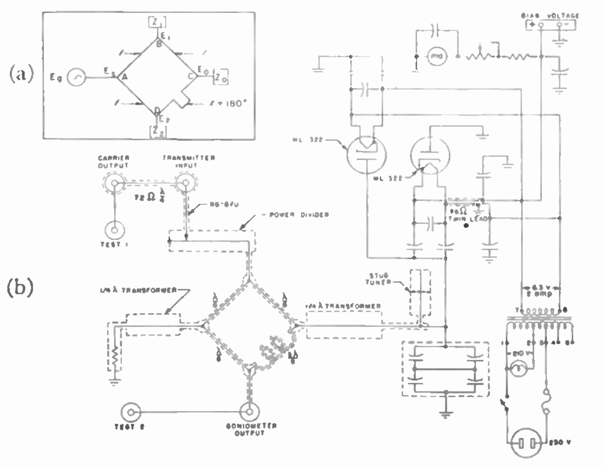


Fig. 9—Block diagram and schematic diagram of modulation eliminator.* 75Ω twin lead passes through stub tuner.

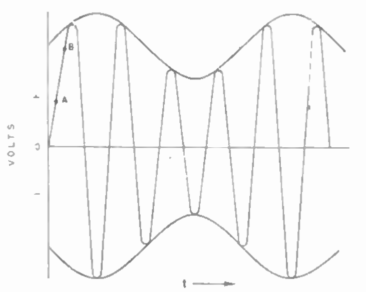
in phase. Thus, the voltage E_0 at point C is zero. When $Z_2 \neq Z_1$, the output voltage E_0 is not zero and power is delivered to the load. Fig. 9(b) shows a schematic diagram of the unit.

The oscillograms reproduced in Fig. 10 show the required performance of the modulation eliminator. All instantaneous voltages that exceed the modulation trough value are reduced to the trough value. This is accomplished by making both Z_1 and Z_2 resistive loads.

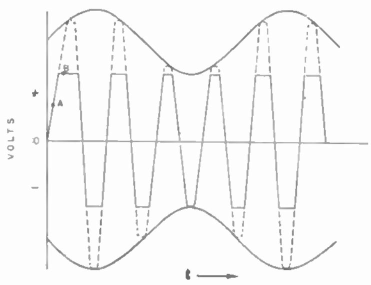
The value of Z_2 is a function of the modulation-eliminator input voltage. Point A, Fig. 10(a) corresponds to point A, Fig. 10(b); $Z_2 \neq Z_1$, therefore $E_0 \neq 0$. A short time later at point B, Fig. 12(a), $Z_2 = Z_1$ for the increment exceeding the trough value so that the instantaneous voltage output is that marked by point B, Fig. 10(b). Only that portion of the applied voltage which exceeds the trough value is cancelled at the output of the modulation eliminator.

A rear view of the modulation eliminator unit is shown in Fig. 11. Three of the four bridge arms are composed of solid dielectric lines, one-eighth wavelength long at 115 mc. The fourth arm of the bridge is also composed of solid dielectric line and is five-eighths wavelength long. The ML-322 diodes were especially designed for application in the modulation eliminator. The plate-cathode dc resistance is approximately 80 ohms in the forward direction and approaches infinity in the reverse direction. In order to obtain a low value of diode plate-cathode resistance for satisfactory operation of the circuit, the spacing between plate and cathode was held to about 0.010 inch. The plate-to-cathode capacity of the diodes is neutralized by a variable inductor, consisting of a short-circuited section of 70-ohm coaxial line.

The modulation eliminator is tested by observing the output of the unit on an oscilloscope, with the dc path



(a)



(b)

Fig. 10—(a) Voltage wave form at input to modulation eliminator. (b) Voltage wave form at output of modulation eliminator.

of the eliminator opened and then closed. When the dc path is open-circuited, the modulation eliminator is inactive so that the oscilloscope will display the modulation at the output of the transmitter. The dc path should then be closed to permit the unit to function and the envelope to be observed.

Typical power measurements of a modulation eliminator are

P_1 = power output of transmitter = 185 watts,
P_2 = power into eliminator = 58 watts,
P_3 = power output of eliminator = 13.1 watts,
P_4 = power into carrier antenna = 127 watts,
efficiency of modulation eliminator = P_3/P_2 = 22.6 per cent,
efficiency of system = $P_4 + P_3/P_1$ = 75.7 per cent.

These data were obtained with carrier only from the transmitter. The modulation eliminator current was 56 ma. Since the unit is set to limit to the trough value of the amplitude-modulated envelope, the theoretical

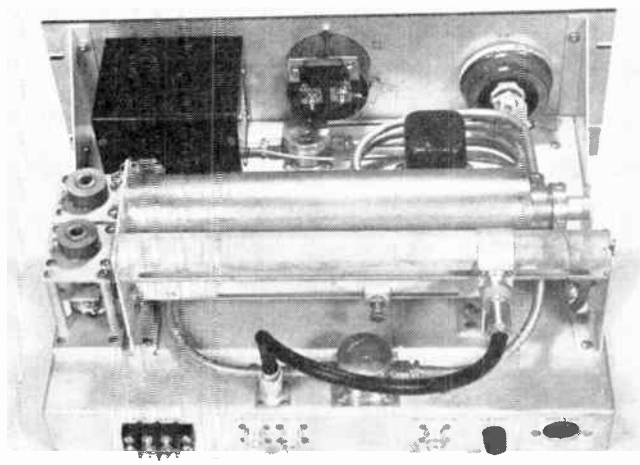


Fig. 11—Rear view of modulation eliminator.

maximum efficiency of the unit, as measured above with the 9.96-kc modulation off, is 67 per cent. This assumes that only the power represented by the voltage above the trough is lost.

Transmitter and Audio Equipment

The omnirange transmitter[13] is designed to deliver an output of 200 watts of carrier in the frequency range 108 to 127 mc, capable of being modulated 100 per cent over the range 60 to 12,000 cps. The input circuit of the audio section provides the simultaneous transmission of three modulations, including voice, a 9.96-kc subcarrier, and a 1,020-cps tone for the purpose of station identification. The voice and subcarrier levels are each set to produce 30-per cent modulation of the carrier, while the tone signal is adjusted for 10-per cent modulation.

Monitor

Proper operation of the omnirange requires that the 30-cps signal of the variable-phase channel and the 30-cps signal of the reference-phase channel be exactly in phase at magnetic north. The monitor provides an alarm for a shift in the relative phase of the two 30-cps signals and for a decrease in amplitude of either the variable-phase or reference-phase signals.

[13] United States Dept. of Commerce, CAA, "Installation Instructions for VHF Omnirange," Federal Airways Manual of Operations IV-B-2-3. For further equipment instructions, refer also to "Description and Theory of VHF Omniranges," Federal Airways Manual of Operations IV-B-1-3.

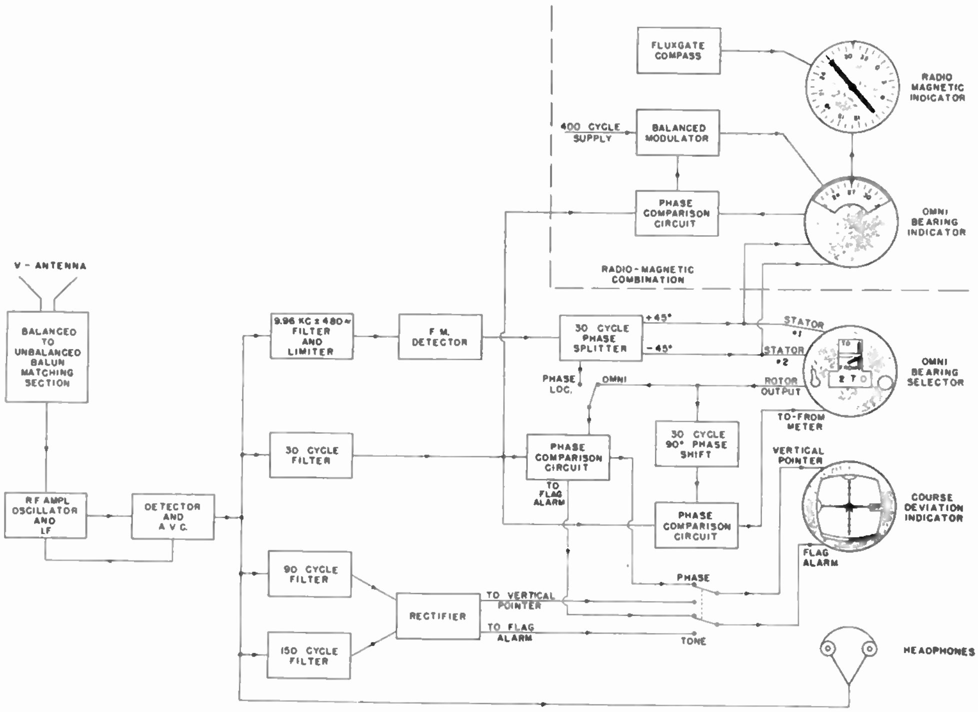


Fig. 12—Block diagram of vhf navigation receiver.

Station Identification

Identification of the omnirange is provided in two ways. One method utilizes an automatic voice-reproducing system for the benefit of aircraft pilots who are unfamiliar with the international Morse code. Station identification is transmitted approximately every 15 seconds by voice. The latter is recorded on a sound film which is moved continuously by a motor and a suitable arrangement of gears and pulleys so that the transmission at any desired interval between 10 and 20 seconds may be repeated.

A second method of station identification employs international Morse-code signals. A motor-driven keyer, with suitable cam and contact arrangement, transmits a series of dots and dashes in code to produce the identification letters assigned to the station.

Airborne Equipment

Fig. 12 is a block diagram of a typical navigation receiver which operates on the omnirange system and other navigational facilities, such as the two-course range, the 90–150-cps localizer, and the phase localizer.

General Performance Requirements

Some of the important requirements for the receiving equipment may be listed as follows:

1. The range courses and bearing will be shown in a visual manner.
2. It will be possible for the pilot to select rapidly any range course to or from the station.
3. The course-deviation indicator, when compared to a 360-degree indicator of similar size, will effectively amplify the course sensitivity in degrees by approximately five times.
4. An off-course condition will produce a direct voltage which can be used with automatic-pilot equipment and with a course-line computer.[14]
5. The receiving equipment will operate on localizers as well as on two-course and omnirange facilities.
6. The equipment will be capable of receiving voice communications from localizers, ranges, control towers, and communications stations.
7. It will be possible to provide an automatic indication of the magnetic bearing to the station.

Radio Frequency, Intermediate Frequency, and Detector

In general, the receivers are comprised of either a single, double, or triple superheterodyne rf section and an audio-converter section, consisting of filters and measuring circuits which connect to external navigational indicators. The frequency selection is accomplished by either manually tuning the rf-input circuits and simultaneously varying the frequency of a local oscillator, or by step tuning in conjunction with crystal control of the local oscillator. "Crystal-saver" circuits are used in step-tuned receivers to decrease the number of crystals required to provide 280 channels at 100-kc intervals over the frequency range of 108.1 to 136 mc. A delayed-action agc circuit is used to provide satisfactory operation of the measuring and indicating circuits. A high-level detector having excellent linearity is required to minimize cross-modulation components, including the unwanted components produced by propeller modulation. The detector output contains a combination of several signals. Suitable filters are necessary

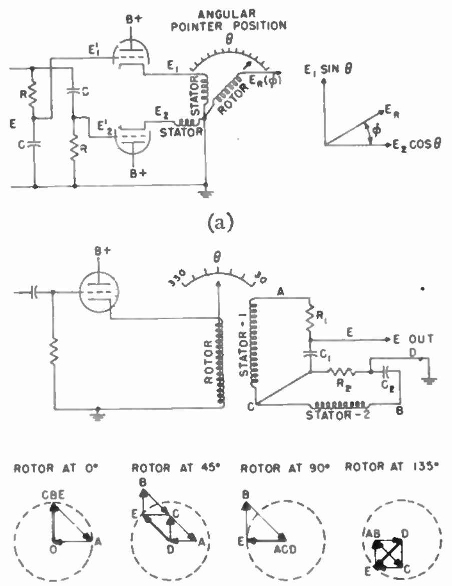
[14] F. J. Gross and H. A. Kay, "Initial Flight Tests and Theory of an Experimental Parallel Course Computer," CAA Technical Development Report No. 83; September, 1948.

to separate these signals. The most important is the 30-cps low-pass filter which attenuates the unwanted signal caused by propeller modulation.

Phase-Shift Methods

The phase angle to be measured by the phase shifter is the one between the 30-cps E_{REF} signal, which has the same phase at all directions from the range station, and the 30-cps E_{VAR} signal, which has a phase that is a function of the magnetic bearing of the aircraft from the station. The E_{VAR} signal is in phase with the E_{REF} signal at the magnetic north of the station. As the aircraft is flown clockwise around the station, the phase of the E_{VAR} signal lags behind that of the E_{REF} signal by an amount equal to the magnetic bearing of the aircraft from the station.

As shown in Fig. 13(a), a quadrature circuit ar-



(a)

(b)

Fig. 13—(a) Schematic of inductive phase shifter. (b) Schematic of improved inductive-type phase shifter.

rangement is used to provide the 90-degree phase difference required for the phase-shifter unit. If $R = 1/\omega c$ at 30 cps, the phase of E_1' will be advanced 45 degrees and the phase of E_2' will be retarded 45 degrees from the phase of E. The stators of the phase-shifter unit will have voltages E_1 and E_2 of equal amplitude, but a phase difference of 90 degrees. The magnitude of the resultant rotor output is proportional to $(E_1^2 \sin^2\theta + E_2^2 \cos^2\theta)^{1/2} = E_R$, where θ = rotor angle in degrees. If $E_1 = E_2$, then E_R is constant for any phase angle. The phase angle of E_R is ϕ and $\phi = \tan^{-1} E_1 \sin\theta / E_2 \cos\theta$. For equal voltages, when $E_1 = E_2$, $\phi = \theta$. The error introduced by unequal voltages E_1 and E_2 will be the difference of ϕ and θ in degrees. The error caused by a 5-per cent decrease in E_2 is 1.4 degrees.

Errors may also be caused by variation in power-supply frequency at the transmitter. The quadrature relationship of voltages E_1' and E_2' remains unchanged, but their amplitudes will vary with a change in frequency. At a frequency of 31.5 cps, the difference in amplitude between E_1' and E_2' is 5.0 per cent, which will cause an error of 1.4 degrees.

The error caused by incorrect phasing of E_1 or E_2 by 1 degree is 1 degree.

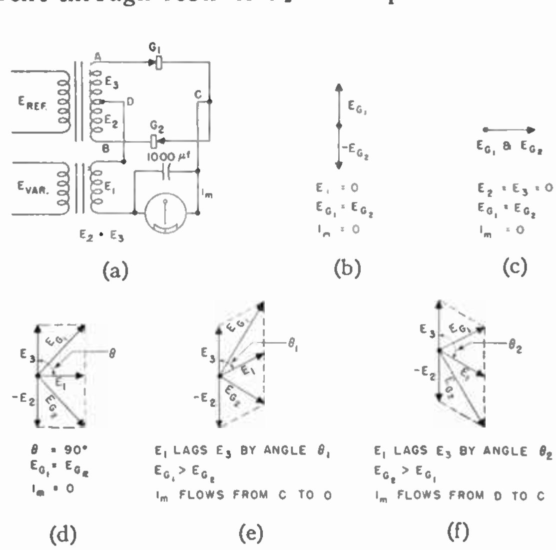
In order that errors produced by the inductive-type phase shifters be held to a negligible value, the units are usually constructed by high-permeability laminated core material, similar to MU-metal, with a skewed rotor winding connected to a high-impedance load. The number of turns on the two stators must be the same and the two magnetic fields must be in exact quadrature.

Fig. 13(b) shows an improved method for using the inductive-type phase-shifter unit. In this method a single tube is used to feed the phase shifter and an rc circuit is connected across each stator. The use of a single tube insures greater reliability. The vector diagrams show the operation of this circuit arrangement.

In general, the inductive-type phase shifter provides the simplest means of obtaining accurate phase variation, which is directly proportional to the angular position of the rotor and provides constant output for all phase angles. The resistance-type phase shifter does not provide constant output, and is difficult to construct so that accurate phase variation proportional to the angular pointer position may be obtained.

Phase-Comparison Methods

The E_{VAR} signal from the phase-shifter unit, commonly known as the "omnibearing selector," is mixed with the E_{REF} signal in a phase-comparison circuit in which a zero-center instrument indicates the phase condition. Fig. 14 shows one type of phase-comparison circuit which has been used in navigation receivers. The circuit operates a dc instrument, which is a desirable feature since adequate damping is easily obtained and operation into automatic-pilot control systems is possible. From an inspection of the circuit diagram, Fig. 14(a), it is evident that current through rectifier G_1 causes point C to be positive with respect to point D. Current through rectifier G_2 causes point C to be nega-



(a) (b) (c)

(d) (e) (f)

Fig. 14—Phase-comparison circuit and vector diagrams.

tive with respect to D. Other polarity conditions may be deduced as follows:

1. If the currents through rectifiers G_1 and G_2 are equal in magnitude, the net charge accumulated by the 1,000-μf capacitor, i.e., the sum of the currents through the indicator, is zero over each alternating-current cycle.

2. If the current through rectifier G_1 is greater than that through G_2, point C has a positive potential with respect to D. Indicator current I_m will flow from point C to D.

3. If the current through rectifier G_2 is greater than that through G_1, point D has a positive potential with respect to C, and indicator current I_m will flow from point D to C, causing the indicator pointer to deflect in a direction opposite to that in (2).

Vector diagrams b to f, inclusive, indicate phase relationships and indicator currents for various conditions. It has been shown that the resultant direct current through the indicator varies in amplitude and direction, depending on the phase relationship of the applied voltages. Since the instrument indicates average currents, the pointer deflection will be proportional to

$$KI = \left| \left[E_1^2 + E_2^2 + 2E_1E_2 \cos \theta\right]^{1/2} - \left[E_1^2 + E_2^2 - 2E_1E_2 \cos \theta\right]^{1/2} \right|,$$

where K is a constant. It follows that the current I changes sign as θ passes through 90 and 270 degrees, respectively. This action is desirable since it provides "sensing" to the instrument indication when used in the omnirange system. Sensing is defined here as the relative direction of motion of a course-deviation indicator needle, resulting from the departure of an aircraft in a definite direction from the desired flight path. The fact that there are two phase angles where the course-deviation indicator current is zero makes it necessary that both phase conditions be resolved and indicated. One method of doing this is shown in Fig. 15. The method used has a second phase-comparison circuit operating at a 90-degree phase displacement from the first phase-comparison circuit, together with an instrument to provide the indication desired. The sensing action of this indication is used in the omnirange system to show whether the numerical reading of the omnibearing selector for an on-course indication of the course-deviation indicator represents the bearing to or from the station. This instrument is commonly referred to as a "to-from indicator." In the circuit shown a small portion of the E_{VAR} and E_{REF} voltages is rectified and fed to the course-deviation indicator to operate the flag-alarm movement, which is now included in the latest type of indicators.

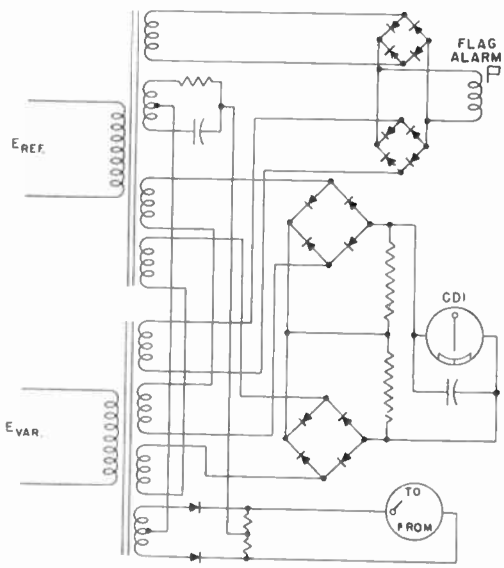


Fig. 15—Phase-comparison circuit with flag alarm and to-from indicator.

Omnibearing Indicators

To provide an automatic and continuous indication of the magnetic bearing to the omnirange station requires additional circuits and instrumentation. Fig. 16

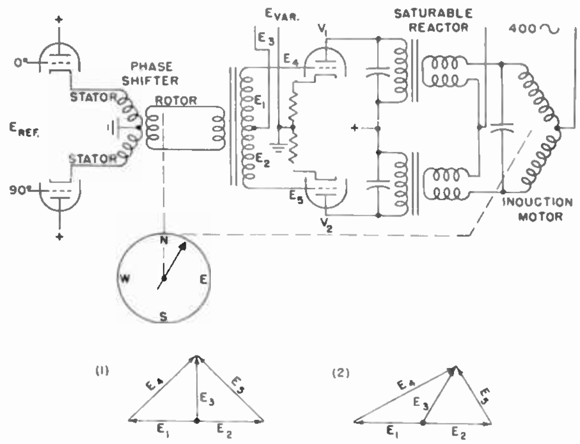


Fig. 16—Omnibearing indicator circuit.

shows a method of providing an automatic phase indicator. As shown in the diagram, the E_{REF} signal is split into quadrature components and fed to the stators of a phase-shifter unit. The rotor output is fed into a servo amplifier, where it is combined with the E_{VAR} signal in a phase-detector circuit. The phase-detector circuit consists of two tubes which are connected to two saturable reactors which, in turn, are connected to a balanced two-phase induction motor. When an unbalance exists, the motor rotates and, by means of a gear arrangement, turns the rotor of the phase-shifter unit to the point where a balanced condition exists. Vector diagram 1 shows the condition when the circuit is balanced and vector diagram 2 shows the condition when the circuit is unbalanced. Voltages E_4 and E_5 in diagram 2 are of different amplitude and, as a result, the currents through the saturable reactors are of different amplitudes. This creates an unbalance in the motor circuit, causing it to rotate and turn the rotor of the phase shifter to a position where a balanced condition again exists.

Radio-Magnetic Indicator

The navigation indicators described thus far provide omnirange bearing and course information independent of the aircraft heading. An instrument has been developed that is known as the "radio-magnetic indicator" (RMI) which, when used in combination with the omnirange and flux gate or gyrosyn systems, will automatically show both the heading and the magnetic bearing of the aircraft to the station. The instrument is, in effect, an omnibearing indicator, as previously described, together with a servo unit operating from the flux gate or gyrosyn system. The complete omnibearing indicator rotates to a position determined by the heading of the aircraft. The pointer of the instrument will indicate the station, and this action is analogous to that of the 360-degree indicators used with low-frequency automatic radio compasses. Fig. 17 shows the circuits of this instrument.

Omnirange Instrumentation

An example of omnirange instrumentation, using the instruments described, is illustrated in Fig. 18. In the example shown, the aircraft is south of the station and flying a course to the station and beyond. The omnibearing selector is set for a course of 350 degrees. When flying this course to the station, the sensing is such that, when the needle of the course-deviation indicator deflects from the zero center, it is necessary to fly the aircraft in the same direction in which the needle deflects so that the needle can return to the zero center; i.e., if the aircraft is flown to the left of the course, the needle will deflect to the right. To return to the course so that the needle will read zero, the aircraft must be flown toward the right.

For the 350-degree omnibearing course to the station, the magnetic compass in the aircraft reads 350 degrees also, neglecting instrument errors and the crab angle. As the aircraft passes directly over the station, the to-from and course-deviation indicators will show several rapid fluctuations. After passing the station and continuing on the same course, the course-deviation indicator will show on-course and have the same sensing as when approaching the station; but the to-from indicator will point to the word "from," indicating that the mag-

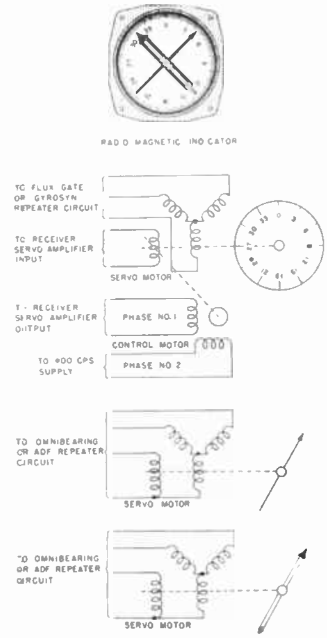


Fig. 17—Radio magnetic-indicator circuit diagram.

netic bearing of the aircraft from the station is 350 degrees. The magnetic compass will continue to read 350 degrees since no change of heading has been made.

If, after continuing on this course, the aircraft makes a turn of 180 degrees to fly the same course back to the station, the course-deviation indicator and to-from indicator will not change, except for small variations due to banking, and the like. However, the magnetic com-

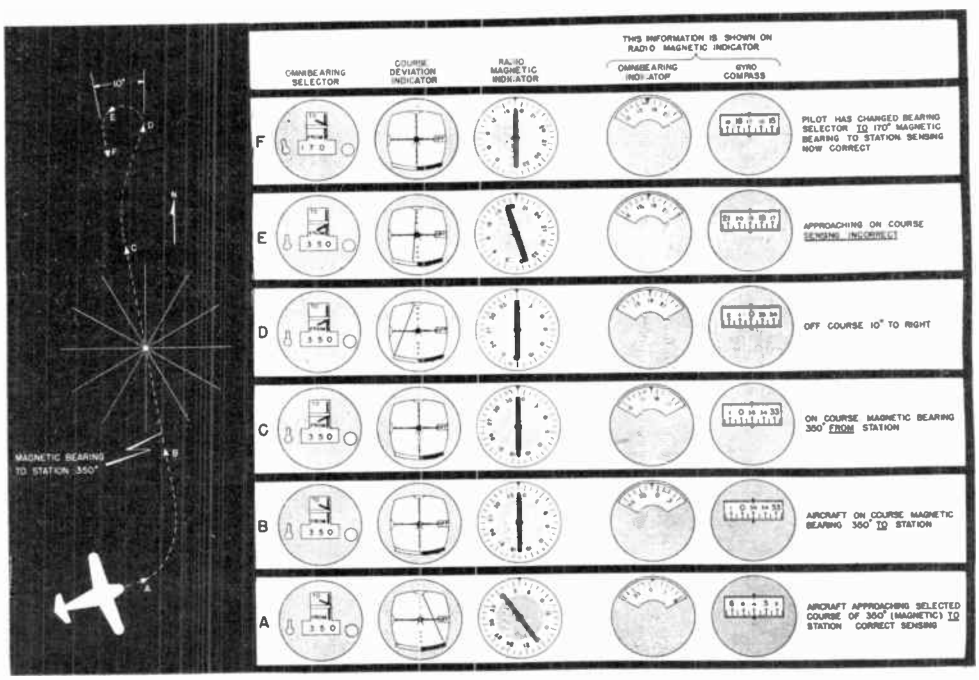


Fig. 18—Omnirange instrumentation combined with magnetic information.

pass will now read 170 degrees and the sensing will be reversed. Therefore, it is necessary to change the omnibearing selector to show a course of 170 degrees to the station. This will cause the to-from indicator to point to the word "to" and show the correct sensing of the course-deviation indicator. It should be pointed out that, under the correct conditions of flying omnirange courses, the omnibearing selector should read approximately the same as the magnetic compass.

Fig. 19 illustrates how the omnirange may be used

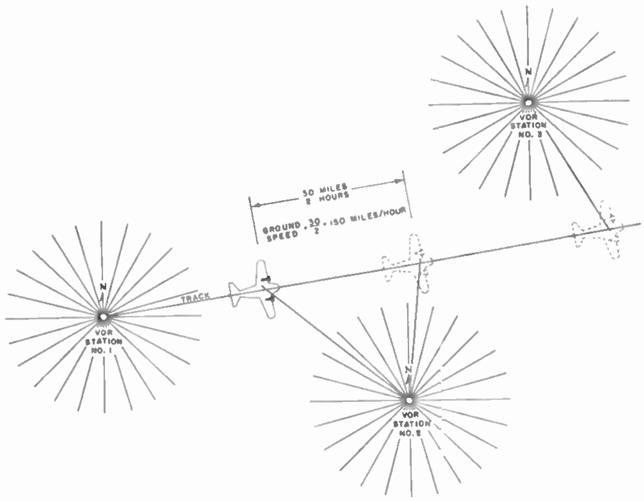


Fig. 19—Omnirange utilization for determining position and ground speed.

to provide position and ground-speed information for the pilot. The intersection of the two lines corresponding to the bearings observed from any two stations gives the position of the aircraft. A second similar observation made sometime later will provide information that can

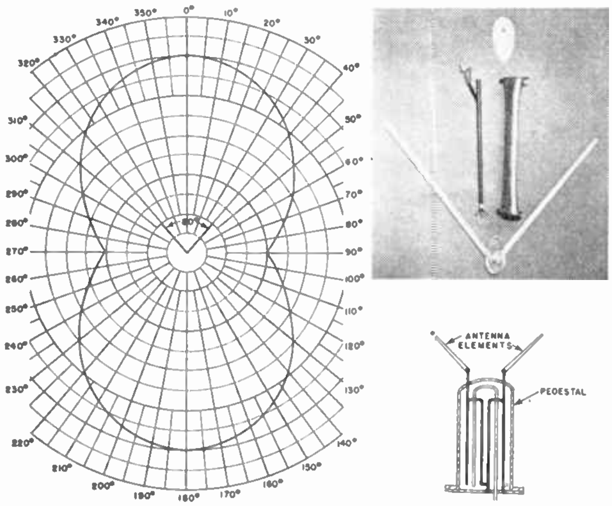


Fig. 20—Calculated field pattern in horizontal plane, balun arrangement and component parts.

be used to determine the actual ground speed of the aircraft.

Aircraft Antenna

Since the signal radiated from the omnirange station is horizontally polarized, it is necessary that the aircraft antenna be designed to receive horizontally polarized signals. One type of antenna which has been used extensively for reception in the frequency band of 108 to 122 mc is the V-type antenna, the parts of which are illustrated in Fig. 20. The diagram in this illustration shows the method used to connect a balanced antenna to a coaxial transmission cable. The impedance matching between the antenna, which has been determined to have a radiation resistance of approximately 33 ohms, and the standard 52-ohm RG 8/U cable, is accomplished by using a one-quarter wavelength balun section having a 42-ohms impedance. Also shown in this figure is a calculated field pattern for an antenna having one-quarter wavelength elements and an apex angle of 80 degrees.

The aircraft antenna is usually installed either on top of the fuselage behind the pilot's compartment or on the tail structure, as shown in Fig. 21. For this reason,

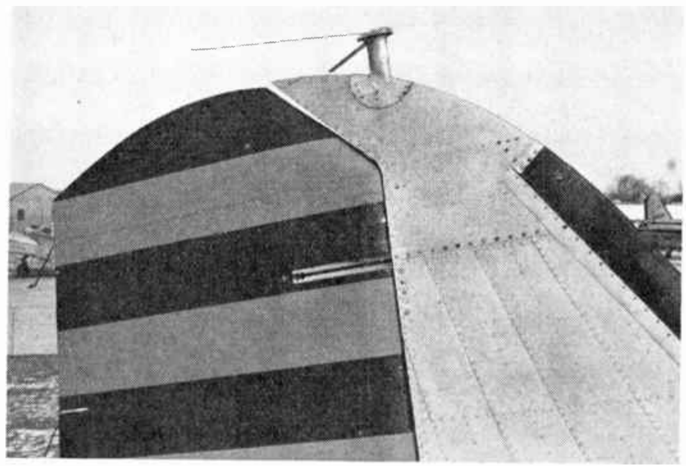


Fig. 21—Antenna mounted on vertical stabilizer.

it is necessary to make antenna measurements over a ground plane and also in free space. Standing-wave ratio measurements of the V-type antenna, between 108 and 122 mc for free-space conditions, showed a minimum of 1.3 and a maximum of 3.0. Similar measurements made over a ground plane showed a standing-wave ratio minimum of 1.4 and a maximum of 3.0.

Receivers and Their Characteristics

The first commercial omnirange-receiving equipment was developed in 1946 by the Aircraft Radio Corporation. The receiver is an eight-tube superheterodyne, having continuously variable tuning over the frequency range of 108 to 135 mc.

Fig. 22 shows the Type 51R-1 airline-type equipment developed by the Collins Radio Company. The receiver is a double superheterodyne which may be tuned from 108 to 135 mc in 100-kc steps by a remotely controlled mechanism, called an "autopositioner." Bearing information is provided by a complete set of instruments, including a frequency selector, radio-magnetic indicator, omnibearing indicator, course-deviation indicator with flag alarm, and an omnibearing selector with a to-from meter.

Late in 1947 a contract was entered into with the National Aeronautical Corporation to develop a low-cost, lightweight navigation receiver for private-flyer use. The receiver supplied on this contrast is a seven-tube superheterodyne, having continuously variable tuning over the frequency range of 108 to 122 mc. Bearing information is provided by a tapped-potentiometer-type bearing selector, a to-from indicator, and a single needle-type course-deviation indicator. The total weight of the unit, exclusive of cables, is 15 pounds, and provision is made for adding a vhf-transmitter unit to this equipment.

Omnirange Operational Characteristics

Siting of Ground Stations

The vhf omnirange requires reasonably good sites. Surfaces formed by objects, such as trees, buildings,

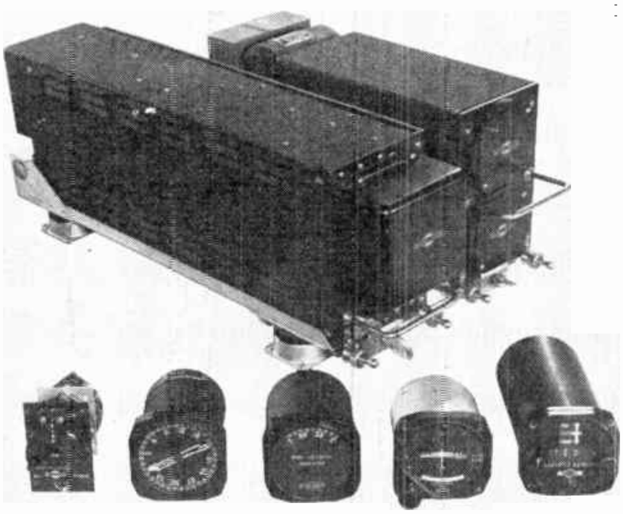


Fig. 22—Collins-type 51R-1 navigation receiver.

wires, hills, and the like, re-radiate or attenuate energy from the antenna system, causing course-deviation indicator fluctuations.[15] Course errors result if the fluctuations are very slow. A large number of installations offering a variety of siting conditions have been studied to determine siting requirements for satisfactory operation of the omnirange.

To briefly summarize siting requirements, it has been found that the site must be located on flat terrain or on top of a knoll that has uniform contours for a distance of 1,500 feet and is cleared of all obstacles, such as large buildings, woods, power lines, and the like, capable of serious re-radiation to a distance of 1,000 feet. In addition, all obstacles within 2,000 feet should subtend an angle in the vertical plane of less than 2.0 degrees, and the power line to the station should be installed underground for a distance of 750 feet.

Distance-Range and Course-Deviation Indicator Fluctuations

Flight tests have confirmed the fact that vhf-navigational facilities are limited to a usable distance range approximating line-of-sight. Certain rare cases have been reported where the distance range was found to be much greater than line-of-sight for a short time. However, this phenomenon has not been experienced at the experimental station.

Fig. 23 shows reproductions of flight recordings of the course-deviation indicator deflections, made when an airplane flew a straight ground track from an omnirange station and flew over the station at 10,000-foot altitude. The variation in course-deviation sensitivity[16]

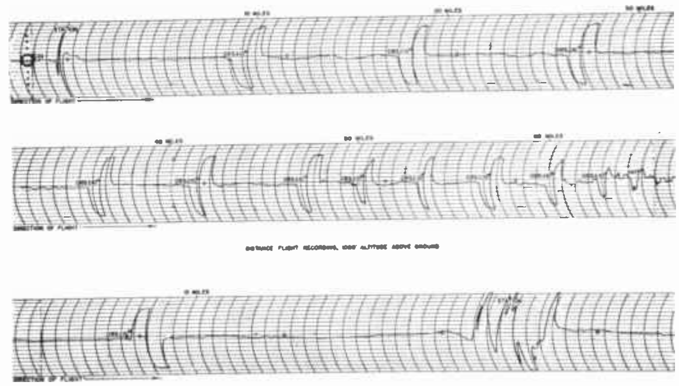


Fig. 23—Recordings of course-deviation indicator at two altitudes. Radial flight across VOR 10,000-foot altitude above ground. CDI-course-deviation indicator. OBS-omnibearing selector. Type-51R receiver, tail V-109 antenna, and C47-type aircraft.

is indicated by the deflections on the recording, for adjustments of the omnibearing selector, of ± 10 degrees. It will be observed that the course-deviation indicator fluctuations are very small in amplitude on this recording. Course fluctuations are caused mainly by poor site conditions at the range stations; consequently, some stations will show more course fluctuations than others. Although a 20-degree course-deviation sensitivity was used for these tests, flight experience has shown that a 30-degree course-deviation sensitivity is more satisfactory from the pilot's viewpoint. The site conditions at this range station are considered good. The course fluctuations shown on the recording made at 10,000 feet indicate a cone area directly over the station where the bearing information is not accurate. However, these violent fluctuations are such that they provide a positive-position indication of the station.

Course Stability

The prototype vhf omnirange, which has been operating for more than 3 years, has been monitored continuously. Faults have occurred at the station usually be-

[15] J. M. Lee, R. G. Pamler, and B. M. Lahr, "An Investigation to Determine the Characteristics of Horizontal and Vertical Polarization for Very High-Frequency Two-Course Visual Radio Ranges," CAA Technical Development Report No. 58; July, 1947.

[16] Course-deviation sensitivity is defined as the total number of degrees of adjustment of the omnibearing selector required to cause the pointer of the course-deviation indicator to move from full-scale left to full-scale right.

cause of a tube failure; this results, for example, in low power output with very high 9.96-kc modulation, at which time a course measurement by the monitor was not possible because of low signal levels. However, during normal operation the course monitor has indicated a maximum of 1-degree deviation from the correct reading. Fig. 24 shows a set of monitor readings which were taken over a period of 30 days. The monitor antenna is located on a line, which is magnetic north of the station and where the two figure-of-eight patterns are of equal magnitude, so that any change in the relative sizes of the patterns will be observed. Courses in other directions have been checked either by a theodolite flight calibration of the facility or by turning the antenna array through 360 degrees. Stability comparable to that indicated at the monitor was found.

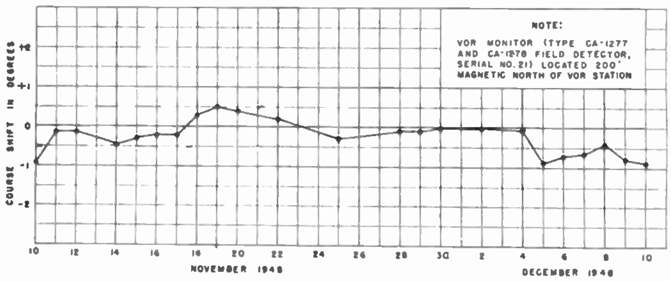


Fig. 24—Course shift as indicated by the VOR monitor versus time.

Flight Calibration

The flight calibration of an omnirange system has been described in a previously published report,[17] and only a brief description of the method used will be given here.

The method used consists of recording the voltage applied to the course-deviation indicator as the aircraft circles the station at a radius of 6 to 15 miles. The omnibearing selector is advanced in 10-degree steps to keep the course-deviation indicator on scale and to provide a recording whereby the indicated magnetic bearing from the range station may be obtained. The indicated bearing is compared with the magnetic bearing, as measured by a theodolite operated on the ground at the range station.

Polarization Errors

Tests have demonstrated that, while the omnirange antenna array radiates horizontally polarized energy, currents induced on the surface of the pedestals supporting the sideband loops radiate energy that is vertically polarized. This vertically polarized energy will produce omnibearing indications which are at quadrature with true bearing information, and errors caused by such radiation are termed "polarization errors."

If an aircraft equipped with a navigation receiver flies over a fixed ground-check point at a number of different headings, the indicated omnibearing may be found to vary with the heading. The aircraft, when over the ground-check point, has only one correct omnibearing, and it is independent of heading; consequently, the change in the omnibearing with heading is one form of polarization error, referred to as "push-pull error." The term comes from early navigational-aids work, where the course was observed to be pushed ahead of the aircraft, or pulled toward it when flying across the course. The aircraft receives the signal from the horizontally polarized wave primarily; however, as the heading is changed, the ratio of vertically to horizontally polarized pickup varies, producing different omnibearing indications when the aircraft is over a ground-check point. Polarization errors, in general, vary with the type of aircraft and with the location of the receiving antenna on the aircraft. Table I shows how the push-pull error may vary with heading.

TABLE I

PUSH-PULL POLARIZATION ERROR TESTS IN A DC-3 AIRCRAFT WITH A RECEIVING ANTENNA-LOCATED JUST BEHIND THE ASTRODOME (*Aircraft Located at 0-degree Azimuth*)

	Azimuth Selector Readings for the Following Headings							Maximum Over-all Error
Headings (Degrees)	0	90	135	180	225	270	315	
Pedestals Exposed	−1.5	0	−3.0	−2.5	+1.0	−3.0	−6.0	7.0
Uskon Cloth Around Pedestals	−1.0	0	−1.5	−1.6	−0.5	−1.0	−1.0	1.6

Many attempts have been made to reduce the polarization error. The most successful of these was the placement of Uskon cloth around the pedestals, as a group, and the cloth was insulated from the pedestals by wooden strips. This cloth has an rf resistance of about 377 ohms per square. Table I lists the indicated bearings observed when the vertically polarized-field effect was reduced by a factor of approximately four when Uskon cloth was used.

Another form of polarization error occurs when an aircraft is banking or turning. Under these conditions, the indicated omnibearing may change even though its direction from the station is unchanged. This type of polarization error is generally referred to as "attitude error." Flight tests made before and after the Uskon cloth was installed showed that its installation decreased the attitude errors. Table II lists the maximum errors observed in 360-degree turns, at a 30-degree bank, dur-

TABLE II

ATTITUDE ERROR AT VARIOUS DISTANCES WITH AND WITHOUT CLOTH INSTALLED

Distance From Station (Miles)	Maximum Error	
	With Cloth Installed (Degrees)	Without Cloth Installed (Degrees)
13.75	±1.5	±2.25
23	±1.75	±3.4
38	±2.1	±3.1

[17] T. S. Wonnell, "Flight Calibration of VHF Omnirange System," Technical Development Report No. 69; July, 1947.

ing a flight test in a DC-3 airplane. The tests were conducted at various distances from the station, and ARC-15 receiving equipment and the tail V antenna were used. Table III shows a comparison of the attitude errors noted when using the tail V antenna and the forward V antenna. These tests were also conducted in a DC-3 airplane, using ARC-15 receiving equipment. The Uskon cloth was in place during both tests.

TABLE III
COMPARISON OF ATTITUDE ERRORS WHEN USING TAIL V ANTENNA AND FORWARD V ANTENNA

Distance From Station (Miles)	Maximum Error	
	Tail V Antenna (Degrees)	Forward V Antenna (Degrees)
13.75	±1.75	±2.0
23	±1.86	±1.8
38	±1.75	±1.25

Propeller Modulation

In the early stage of development of the omnirange system it was observed that aircraft propellers caused the received signals to modulate. As a result of the modulation produced by the propellers, the course-deviation indicator will oscillate at an amplitude and rate depending on the amplitude and rate of the propeller modulation. Flight tests were conducted on omnirange systems using 30, 40, and 60 cps for the variable- and reference-phase signals. It was determined that a frequency of 30 cps would be the most desirable since the cruising propeller rpm of all known aircraft produces a propeller-modulation frequency higher than 30 cps. Simultaneous flight recordings of omnirange courses, using a 30-cps range system and a 60-cps range system, are shown in Fig. 25. The engine speed was adjusted to

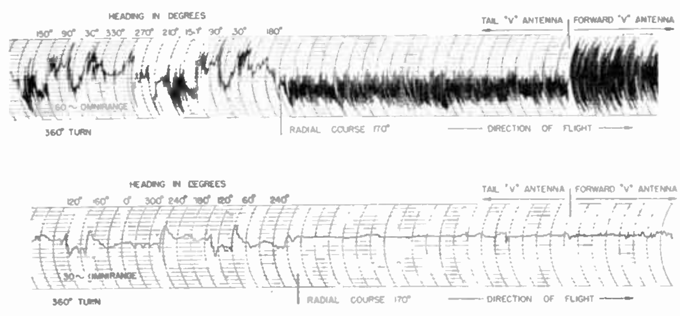


Fig. 25—Comparison of course-deviation indicator fluctuations for 30- and 60-cps omniranges. NC-182 (DC-3/c-47) 2,130 rpm both engines flight altitude 3,000 feet. Absolute distance from ranges 5 to 10 miles. Receiver 30∼omnireceiver Number 7. Limiter in 10-kc reference channel. Wein bridge in variable channel. 60∼omnireceiver number 6. Antennas: tail V coaxial type; forward V delta type. 60-omnitransmitter 125 mc. 30∼omnitransmitter 114.3 mc.

provide a propeller modulation of approximately 60 cps. It will be observed that the 30-cps range is free of variations caused by propeller modulation when the tail V antenna is used, and shows only a small variation when the forward V antenna, which is located near the propellers, is used. Recent improvements in navigation-receiver design provide further improvements in freedom from propeller-modulation effects.

Interference between Stations

Flight tests have been conducted to determine the interference characteristics existing between omnirange stations operating on the same carrier frequency and on adjacent frequency channels of 100-kc separation. ARC-15 navigation receivers were used for all tests.

To determine the interference area between two omnirange stations operating on the same frequency, those at Allentown, Pa. and Raleigh, N. C., separated by 383 miles, were both tuned to 117.5 mc. The area between the two stations was flown at various altitudes from 12,000 to 20,000 feet, and fixes were obtained by means of the MEW radar installation located at the Washington National Airport. The information obtained in the flight tests is shown in Fig. 26. The mini-

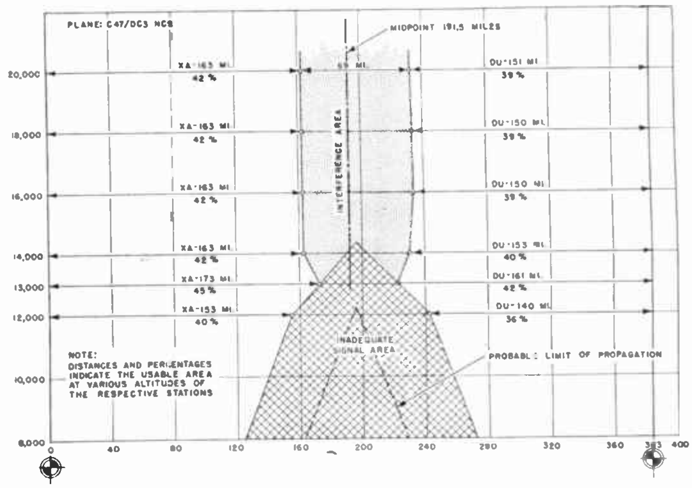


Fig. 26—Interference characteristics of two omnirange facilities operating on the same frequency. Distance in miles XA-117.5 mc. Allentown, Pa. VOR elev. 670-feet MSL. DU-117.5 mc. Raleigh, NC. VOR elev. 403-feet MSL.

mum interference altitude appears to be approximately 14,500 feet since below this level the limiting factor is inadequate signal strength. Above 14,500 feet, the interference area consists of a combination of severe audio heterodyning and variations of the course-deviation indicator. Summarizing, when two omnirange stations are operated on the same frequency, the useful distance range at high altitudes is approximately 40 per cent of the spacing between the stations.

The investigation of adjacent frequency-channel interference was made using omnirange stations located at Raleigh, N. C. and Spartanburg, S. C. The Raleigh station was operated on a frequency of 117.4 mc, and the Spartanburg station on a frequency of 117.5 mc. The geographical separation was 192 miles. The information obtained in these flight tests (see Fig. 27) indicates that from approximately 60 per cent at 6,000 feet to 75 per cent at 13,000 feet of the total distance is usable without interference. In this test it is important to note that the results obtained are dependent to a great extent on the selectivity characteristics of the receiving equipment.

The use of an airline-type receiver, such as the Collins 51R, would have provided vastly different results in the usable distance range.

Additional information concerning the co-channel interference was obtained by laboratory tests, and is shown in Fig. 28. For these tests, simulated omnirange signals from two signal generators were fed to an airline-type receiver, and the ratio of the rf signal levels was varied. The phase of the variable-phase signals was adjusted to cause the greatest interference on the indicators.

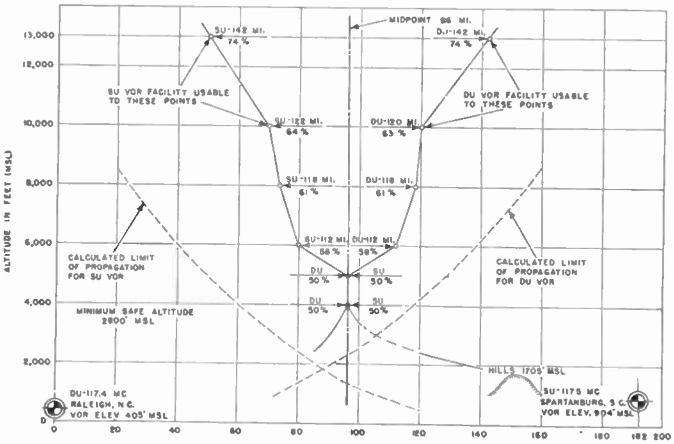


Fig. 27—Adjacent channel interference characteristics of two omnirange facilities. Distance in miles.

Power-Frequency Variation

In some remote sections of the country, where it may not be possible to obtain a standard commercial source of 60-cps power, it may be necessary to operate the station from local power or, perhaps, to utilize the emergency supply which is standard equipment at all CAA radio-range stations. In some cases these sources of power do not maintain a constant frequency of 60 cps, and deviations as much as ± 2 cps may be experienced. At those stations where the power-supply frequency variation exceeds ± 0.6 cps, a stabilized 60-cps tuning fork-type generator and amplifier will be installed to supply power to the goniometer and subcarrier equipment.

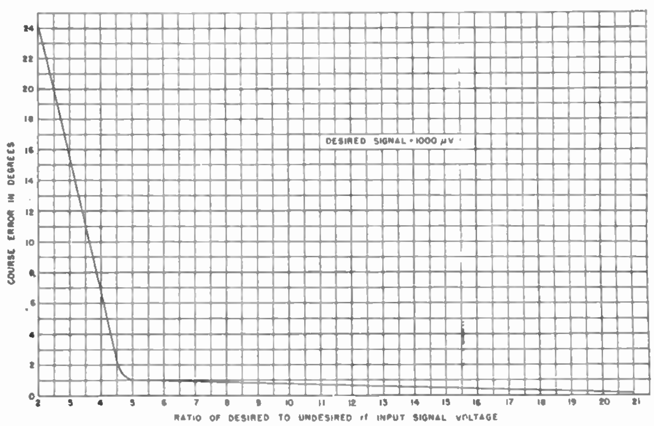


Fig. 28—Course error versus ratio of desired to undesired signal strength.

Conclusions

As a result of several years of study and development, in addition to 4 years of operation of omnirange facilities, the following general conclusions have been reached:

1. The theoretical aspects of the phase-comparison system for omnirange operation are sound.

2. Horizontal rather than vertical polarization provides steadier course indications.

3. A transmitter power rating of 200 watts is adequate for line-of-sight operation to a distance of approximately 200 miles at an altitude of 20,000 feet, 100 miles at 5,000 feet, 63 miles at 2,000 feet, and 45 miles at 1,000 feet.

4. An accuracy of approximately ± 1.5 degrees may be realized on radial courses where the sites are free of serious reflecting objects. The accuracy figure is for the complete omnirange system, receivers, plus ground equipment.

5. Sites located on flat terrain with all obstacles, such as large buildings, woods, power lines, and the like, cleared to 1,000 feet, and all obstacles within 2,000 feet subtending a vertical angle of less than 2 degrees have proven to be satisfactory. Nearby hills, higher than the site, reduce the distance range on radials passing over the hills.

6. Generation of sidebands by means of a rotating capacity goniometer was found more stable and reliable than methods using electronic sideband generators.

7. For unattended operation, a monitor must be included as part of the ground equipment. This provides the maintenance personnel with continuous knowledge as to the status of the facility and with a useful tool for corrective maintenance. The monitor will show up a fault as soon as it occurs, and will further insure that the pilot receives only correct information.

8. Either automatic or manually operated indicating devices may be used to provide the pilot with navigational information. The latter type, however, defines the range course more accurately.

9. Modulation factors of 30 per cent each are satisfactory for the reference and variable-phase signals, and the remaining 40 per cent is adequate for voice modulation and 1,020-cps tone identification.

10. Propeller modulation of received signals interferes with the omnirange courses under certain critical conditions of heading and propeller speed. A reduction in frequency from 60 to 30 cps, for the reference and variable-phase signals, has minimized this interference.

As a result of this development and operational experience, the vhf omnirange has been adopted as one of the basic elements of the Common-Civil-Military System, Transition Program, as recommended by the Radio Technical Commission for Aeronautics, Special Committee 31.

Information Theory and the Design of Radar Receivers*

PHILIP M. WOODWARD†

Summary—**The paper deals with the problem, frequently encountered in radar, of extracting simple numerical information from a noisy waveform. It is suggested that the only ideal way of doing this is to use the principle of inverse probability and convert the waveform into a probability distribution for the quantity sought. The method is applied to the problem of determining the time delay of a periodically modulated rf waveform in the presence of white Gaussian noise when the undelayed waveform without noise is exactly known. As a result, the matched predetection filter of Van Vleck and Middleton is automatically specified, and the theory of ideal detection is briefly indicated.**

I. Introduction

The object of this paper is to outline, in terms of a somewhat idealized example, a mathematical method by which a theoretically ideal radar receiver may always be specified in principle. It was for some time customary to regard signal-to-noise ratio as an all-important quantity in receiver design. Efforts were made to ensure that as high a ratio as ideally possible was obtained at the output. This seems now to be a mistaken philosophy, since signal-to-noise ratio does not measure information, and is something which can often be artificially enhanced by passing the waveform through a nonlinear device which does not alter the information content at all. The present method deals, not with signal-to-noise ratio, nor even with quantities of information, but with the information itself.

In radar, we have to answer such questions as whether a target is present or absent, what its range is, whether it is moving, and so on. If we attempt to design a receiver which would answer any or all of such questions exactly, we are attempting the impossible, because of the noise which must inevitably introduce false indications. But if we demand, on every occasion, an automatic assessment of the relative probabilities of all possible answers, we are being completely realistic and no receiving device can possibly do better. The present paper shows how this idea works out in one rather familiar problem—that of determining the time delay of a periodic waveform of known shape and amplitude. This amounts, in radar, to determining the range of a stationary target known to be present and giving an echo of known strength, and is obviously an artificial problem. But it suffices to illustrate the method, and is not altogether without practical interest. The quantity of range information latent in such a radar waveform has been evaluated elsewhere[1] without reference to any actual method of extracting it. Here the emphasis is on designing the ideal receiver rather than on evaluating its actual performance.

The whole of the present approach is based on the principle of inverse probability briefly summarized in the following section. There is nothing new in this principle, but in spite of the growing application of probability theory to such problems as arise in radar, as for instance by Kaplan and McFall,[2] there does not seem to have been a systematic attempt to apply inverse probability. The necessity for doing so becomes quite apparent, once the foundations of modern communication theory[3] have been studied. The present paper may perhaps help to "set the ball rolling," for the method can be applied to a very wide variety of radar problems.

II. The Principle of Inverse Probability and the Information Function

A "direct" probability describes the chance of an event happening on a given hypothesis, but if the event has actually happened and there are various hypotheses which would explain it, one is faced with a problem of inverse probability. Prior to the event, the various hypotheses may not have seemed equally probable, and such previous knowledge is expressed in terms of the *a priori* probabilities of the hypotheses. After the event, which will usually be an experimental observation specifically performed to test the hypotheses, their relative probabilities may become changed. The principle of inverse probability expresses the *a posteriori* probabilities in terms of the corresponding *a priori* ones, by utilizing the probabilities that the actual observation would have been obtained if each hypothesis in turn had been true. It is valid only when the hypotheses are mutually exclusive and exhaustive.

Let H_1H_2, and so on, denote the hypotheses,

$P(H_n)$	the *a priori* probability of H_n
$P(Ob \mid H_n)$	the probability of the observation if H_n were true
$P(H_n \mid Ob)$	the *a posteriori* probability of H_n after the observation is known.

The theorem[4] may then be written briefly in the form

* Decimal classification: R537.13. Original manuscript received by the Institute, June 23, 1950; revised manuscript received, March 9, 1951. The present paper is a revised version of the paper originally communicated to the Institute by G. G. Macfarlane, under the title "The decoding of radar information."

† Telecommunications Research Establishment, Ministry of Supply, T.R.E., Great Malvern, England.

[1] P. M. Woodward and I. L. Davies, "A theory of radar information," *Phil. Mag.*, ser. 7, vol. 41, p. 1001; October, 1950.
[2] S. M. Kaplan and R. W. McFall, "The statistical properties of noise applied to radar range performance," Proc. I.R.E., vol. 39, pp. 56–60; January, 1951.
[3] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. 27, pp. 379, 623; July and October, 1948.
[4] Harold Jeffreys, "Theory of Probability," chap. I; Oxford University Press; 1939.

$$P(H_n \mid Ob) \propto P(H_n)P(Ob \mid H_n). \tag{1}$$

In the present application, the hypotheses are all the possible delay times of a given periodic waveform, and the "observation" is simply the given waveform delayed an unknown amount, and with white Gaussian noise added to it. We shall call this the "received waveform," but it is not the output from a receiver because this would prejudge the whole question. It is the waveform as it enters the receiving system, including any noise which the receiver itself may subsequently introduce. The *a priori* probabilities of the hypotheses will form a continuous probability density distribution for the unknown delay time τ, and for simplicity this distribution will be taken uniform over an interval equal to one period of the waveform. In other words, we take all inherently unambiguous delay times to be equally likely *a priori*, though any other prior knowledge can subsequently be inserted in the theory.

The best that any receiver can do is to form the *a posteriori* distribution of probability for τ, from the received waveform. This distribution is the actual information sought, and is most conveniently handled in logarithmic form. We shall call its logarithm the "information function" and denote it by $Q(\tau)$. Equation (1) may now be written

$$Q(\tau) = \log P(Ob \mid \tau) + \text{constant}, \tag{2}$$

where $P(Ob|\tau)$ denotes the probability density for the received waveform on the hypothesis τ. The constant term is merely the logarithm of the normalizing factor for the *a posteriori* distribution. It serves no useful purpose and is in the future omitted.

III. Evaluation of the Information Function

Let us consider a particular occasion when the true value of the delay time τ happens to be τ_0, and write the received waveform in terms of real functions as

$$Y(t) = G(t - \tau_0) + I(t). \tag{3}$$

Here $G(t)$ is the rf waveform which would have been received in the absence of noise or of any time delay, and is assumed known *a priori*. It is also assumed that $G(t)$ is periodically modulated, and although it is convenient to use the language of pulses, the theory is, in fact, valid for any periodic modulation whatever, including frequency modulation. The function $I(t)$ represents added noise. The probability distribution for the magnitude of $I(t)$ at any particular time t is assumed Gaussian, but it is necessary here to generalize this concept. It can be shown, either by resorting to sampling-point analysis[5] or by means of a statistical mechanical argument, that the probability density for the whole waveform $I(t)$, in an appropriate number of dimensions, is proportional to

$$\exp\left[-\frac{1}{N_0}\int I(t)^2 dt\right], \tag{4}$$

[5] C. E. Shannon, "Communication in the presence of noise," Proc. I.R.E., vol. 37, p. 10; January, 1949.

where N_0 is the mean noise power per unit bandwidth.[1] The observer has access to $Y(t)$ but not to τ_0 directly, and must therefore try out all possible values of τ in turn. On the hypothesis τ, he can argue that the noise waveform alone would have to be $Y(t)-G(t-\tau)$, for which the probability density is proportional to

$$\exp\left\{-\frac{1}{N_0}\int [Y(t) - G(t-\tau)]^2 dt\right\}. \tag{5}$$

Consequently, by (2), we have

$$Q(\tau) = -\frac{1}{N_0}\int [Y(t) - G(t-\tau)]^2 dt. \tag{6}$$

The information function is thus proportional to the integrated square of the departure of the received waveform from a hypothetical noise-free waveform of lag τ. As the hypothetical τ is varied, the value which gives least-mean-square departure from the received waveform produces a maximum in the information function, and this, from the observer's point of view, is the most probable value of τ.

The limits of integration in (6) are chosen to correspond with whatever portion of the received waveform is being examined, and it is necessary to take this to be a whole number of repetition periods of the modulation. If the integrand be expanded into three terms, it will be found that the G^2 integral is independent of τ because of periodicity, and the Y^2 integral depends on τ_0 but not on τ, from (3). These two terms can consequently be omitted from $Q(\tau)$, since anything independent of τ may be absorbed into the normalization, which has already been omitted. We are left with

$$Q(\tau) = \frac{2}{N_0}\int Y(t)G(t-\tau)dt. \tag{7}$$

The integrand may be said to exist for all values of t, but τ is confined within certain fixed *a priori* limits, say between 0 and the repetition period R of $G(t)$. The domain of the integrand may be represented diagrammatically (Fig. 1) as a strip of indefinite length in the t direction, and of width R in the τ direction. If it is required to

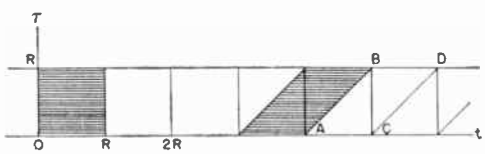


Fig. 1—Two methods of integrating range information.

form the information function $Q(\tau)$ to represent all the information contained in $Y(t)$ in the interval (O, R), then $Y(t)G(t-\tau)$ must be integrated with respect to t between the limits 0 and R for each value of τ as indicated by the shaded square. Further observation of the received waveform will entail further integrations in the obvious manner, and the limits of integration will always be represented by vertical lines on the diagram. The trouble with this process, from a practical point of view, is that all the accumulated information of any one interval becomes available instantaneously at the end of the interval, and is followed by a lull while fresh integration is being performed.

Fortunately, there is an alternative method which is more natural, though mathematically less straightforward. By carrying out the integration, not in successive squares, but in parallelograms such as the one indicated on the diagram, the information function becomes available from $\tau=0$ to R (A to B in the diagram) progressively in time. The moment B has been reached, a fresh trace can be developed from C to D. In this way, we have set up a saw-tooth relationship, a time base in fact, connecting τ and t.

This process of "progressive" integration does not, of course, correspond to fixed limits on the integral (7), but to a pair of limits advancing together in time. In fact, the value of the information trace at time $t=nR+\tau$, where n specifies the nth trace and τ is confined between 0 and R, is given by

$$Q_n(\tau) = \frac{2}{N_0}\int_{t-R}^{t} Y(t)G(t-\tau)dt. \tag{8}$$

It is mathematically inconvenient that the information function Q is the logarithm of an *a posteriori* probability distribution for τ only when the limits of integration are constant. In other words, each hypothesis τ should, strictly, be tested out on the same piece of the received waveform. Space does not permit a full discussion of this point, and it must suffice to remark that the progressive probability distribution

$$P_n(\tau) = e^{Q_n(\tau)}, \tag{9}$$

normalization omitted, behaves for all practical purposes as though it were a strict *a posteriori* distribution. In particular, when the information from successive periods of the received waveform is combined, either by summing the Q_n over n or by multiplying together the P_n, the resulting distribution differs from a true *a posteriori* distribution only because of end-effects, which become progressively less and less important.

IV. The Ideal Predetection Filter

The progressive information function given by (8) happens to have a very simple electronic interpretation. The form of the expression, being a linear superposition, will be recognized as that of the output from a linear filter. In fact, it is the output at time $t=nR+\tau$ from a filter whose input is the received waveform $Y(t)$, and whose impulsive response is given by

$$\delta(t) \to \begin{cases} \dfrac{2}{N_0}G(-t), & 0 < t < R \\ 0, & t < 0 \text{ and } t > R. \end{cases} \tag{10}$$

Such a filter, apart from the special scaling factor $2/N_0$, has been discussed by Van Vleck and Middleton,[6] who show that it is the unique linear filter which gives maximum peak signal-to-noise performance. (This property is, however, irrelevant to the present theory.) The filter has a frequency response which, in amplitude, has the same shape as the amplitude spectrum of one period of the input signal G, but which in phase is equal and opposite to that of G.

The output from the filter is, of course, a modulated radio-frequency waveform, and when passed through an exponential rectifying device in accordance with (9), it becomes the progressive *a posteriori* probability distribution for τ. It has already been pointed out that several traces of Q may be added together, before such rectification, if the information from several periods of the received waveform is to be combined. This is simply phase-coherent pulse-to-pulse summation, and must not be performed unless τ is completely independent of time, as has so far been assumed.

The underlying effect of this ideal filter is to cause the output signal peak to look exactly like a particularly large noise peak; all the pattern information, which originally distinguished signal from noise, has been extracted from the waveform and converted into amplitude discrimination. This may seem surprising in view of the fact that the signal and noise outputs have different power spectra. Or again, if the input signal is a square pulse, the noise output can be regarded as a multitude of overlapping square pulses, while the signal output will be a triangular pulse. However, the fact remains that a multitude of square pulses overlapping at microscopic intervals to form Gaussian noise provides a background against which the pattern of a single triangular pulse cannot be distinguished. If, indeed, pattern information could still be utilized in the filter output, one is led to a *reductio ad absurdum*, since it has been shown that the most probable value of τ is given by selecting, regardless of pattern, the largest amplitude in the filter output.

V. Discussion

The *a posteriori* probability distribution for the delay time τ has been shown to take the form of a modulated rf waveform, obtained at the output from a linear filter, distorted in amplitude or "rectified," by means of a device having an exponential characteristic. This naturally results in a function of τ whose envelope, if the signal is large enough, is peaked near the true value τ_G, but which contains under its envelope a multitude of fine peaks produced by the carrier. This fine structure represents a succession of probable and improbable values of τ resulting from comparison of the carrier phase in $Y(t)$ and $G(t)$. When this highly ambiguous knowledge of range is of no interest, it may be removed by smoothing or "detecting" the *a posteriori* distribution in such a way that areas over intervals of an rf cycle are preserved. In fact, when τ changes with time sufficiently rapidly to render the rf information out of date from one trace to the next, but not rapidly enough to affect the modulation appreciably, successive

[6] J. H. Van Vleck and D. Middleton, "A theoretical comparison of the visual, aural and meter reception of pulsed signals in the presence of noise," *Jour. Appl. Phys.*, vol. 17, p. 940; November, 1946.

traces may be combined only after removal of the rf. The detected *a posteriori* distributions must then be multiplied together, or alternatively their logarithms must be added. Mathematical readers will see immediately that, in this way, an ideal detection characteristic (of the form log I_0, where I_0 is the modified Bessel function) is uniquely specified by the theory when post-detection pulse-to-pulse summation is to be performed.

Without, for the present, developing any further the theory of removing any of the idealizations, it should be clear that any problem of extracting all the information from a noisy waveform, can, in principle, be solved uniquely and ideally by one universal method. One simply has to state the question, write down the *a posteriori* probability distribution for all possible answers to it, and interpret the resulting formula in terms of a physical device, on the principle that anything which can be computed mathematically, can also be computed electronically. No problem of waveform decoding then remains, for the *a posteriori* distribution *is* the required information; anything further is pure guesswork.

ACKNOWLEDGMENT

Acknowledgment is made to the Chief Scientist, British Ministry of Supply, for permission to publish this paper (British Crown copyright reserved). The paper is reproduced with the permission of the Controller, H.B.M. Stationery Office, England.

Notes on an Automatic Radio-Frequency Repeater System*

JAMES A. CRAIG†

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

***Summary*—The paper describes the basic principles involved in an operating system of rf repeaters, and discusses the planning and installation of this system in Cuba. It mentions the types of antennas used and briefly describes the physical layout of the equipment involved. Also included is a resumé of the difficulties that arose and how they were overcome.**

ONE OF THE principal radio-broadcasting chains indicated about 2 years ago their need for some form of network transmission system that would enable them to overcome the inadequacies of the available Cuban telephone lines. The service provided by these lines was undependable and, when available, was noisy and wholly unsuitable from a program-quality standpoint.

The plan involved the use of radio repeaters to relay programs originating in Santiago studios near the Eastern end of Cuba and to service a network of AM broadcasting stations extending some 500 miles west to Havana. At an average 50-mile distance between repeaters this would necessitate at least ten repetitions of the program material.

To use the conventional form of repeater would have been practically impossible since demodulation and remodulation at each repeater point would have introduced a prohibitive amount of noise and distortion due to nonlinearity in the detection and modulation systems. This factor alone usually dictates a maximum of 5 or 6 repeaters even for voice-communication circuits, and the requirement in this case was for program quality. The common form of repeater, when stripped of embellishments, remains essentially a radio receiver and a radio transmitter operating back-to-back.

To overcome the foregoing limitations a radio-frequency repeater system was developed. In this repeater, as the name implies, rf is used exclusively. Frequencies between 150 and 180 mc can be accommodated by this particular system although the principles are being readily applied to the other commonly used vhf and uhf bands. The actual frequency range used in the Cuban system is from 163 to 170.2 mc.

From the functional block diagram in Fig. 1 it is seen that a typical operating frequency of 165.0 mc has been chosen as the incoming carrier for the purpose of illustration. This frequency is fed from an antenna to a single stage rf amplifier. From this amplifier it is heterodyned with a locally generated signal originating in a "channel" oscillator. This oscillator is controlled by a quartz crystal operating as a harmonic oscillator at a frequency of approximately 45 mc. The crystal frequency is tripled and then amplified at 137.5 mc, as chosen in the block diagram, and is then fed to the first mixer stage. This crystal is the only one that requires changing when a different carrier frequency is to be used. The beat frequency between the incoming signal and the local oscillator is always 27.5 mc for any incoming carrier, and this frequency enters the first intermediate frequency amplifier consisting of two high-gain stages.

As the signal leaves the first IF amplifier, the user can elect to retransmit on a frequency which is either

* Decimal classification: R480. Original manuscript received by the Institute, March 22, 1951; revised manuscript received May 22, 1951. Presented, National Conference of the IRE Professional Group on Vehicular Communications, Detroit, Michigan, November 3, 1950.

† Link Radio Corp., 125 W. 17th St., New York 11, N. Y.

4.8 mc below or 5.2 mc above the incoming signal frequency. It will be noted that the signal can enter either one of two paths, each consisting of a second mixer and a second IF amplifier. These parallel circuits are so arranged that a selector switch energizes the cathode heaters of one branch or the other. High B+ voltage is continuously available to both branches, but only the one selected will draw B+ current.

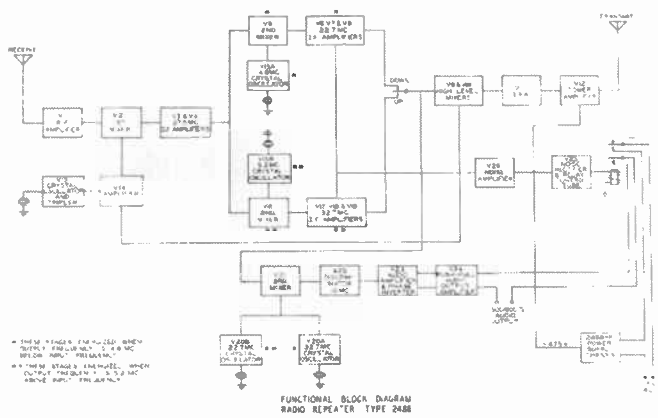


Fig. 1—Radio repeater, functional block diagram.

Assuming that retransmission is to occur on a frequency which is 4.8 mc below that of the incoming signal, the selector switch would be thrown to the "Down" (down in frequency) position. The upper branch on the block diagram would then be energized and the 27.5-mc signal would beat in the second mixer against a 4.8-mc signal locally generated by a crystal oscillator. This produces a 22.7-mc signal which is then amplified by the second intermediate amplifier. This intermediate amplifier consists of four stages. The first two stages function as straight amplifiers and the third and fourth stages act as cascaded limiters. Any incidental amplitude modulation of the IF signal is reduced in the third stage and fully removed in the fourth stage.

This purely frequency-modulated signal next enters a balanced high-level mixer where it again combines with the same 137.5-mc signal derived from the channel oscillator. This results in a 160.2-mc signal which is used to excite an intermediate-power amplifier stage which then, in turn, excites the final power amplifier having an output power of 50-watts nominal. The signal then passes to the outgoing antenna for transmission to the next repeater point at a frequency 4.8 mc below that of the incoming signal frequency, which was 165 mc.

Had the alternate second mixer and IF path been selected, by throwing the selector switch to the "UP" position, it can be seen that a local 5.2-mc signal would have converted the 27.5-mc signal from the first IF amplifier to a frequency of 32.7 mc. The signal would beat with the 137.5-mc channel oscillator frequency in the balanced high-level mixer to produce a final output frequency of 170.2 mc.

The purpose of shifting the transmission signal up or down by unequal amounts is to prevent interference between the different repeater stations in the system. Since the carrier frequency is raised by 5.2 mc and lowered by 4.8 mc at successive repeater points, no two repeater stations in such a chain will transmit at the same frequency.

As can also be noticed in the block diagram, an audio takeoff is provided for modulation of a member AM broadcast station of the chain or for monitoring purposes. This is accomplished by diverting a small portion of the 22.7- or 32.7-mc energy leaving the second IF amplifier, and combining it in a third mixer stage with a locally generated 32.7- or 22.7-mc signal, respectively. A second selector switch selects which of these crystal

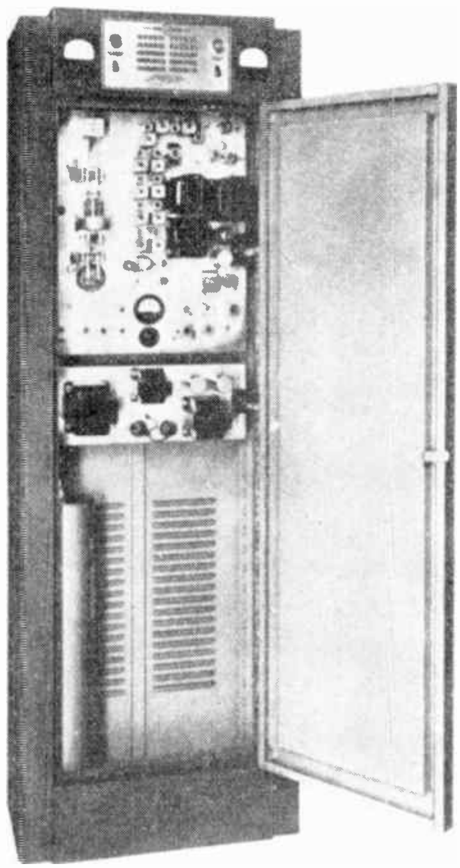


Fig. 2—Radio-repeater equipment, front view (Type 2486).

controlled oscillators is used, which depends, in turn, on the position of the first selector switch. In this manner, the output of the third mixer stage always produces a 10-mc signal which is fed to a balanced discriminator. The audio leaving the discriminator is then passed through a de-emphasis circuit and applied successively to an amplifier, a phase inverter, and a push-pull audio amplifier, which delivers audio at zero level across a termination impedance of 500 or 600 ohms.

The repeater is automatically turned on or off by a relay tube, which functions by virtue of noise present at the screen grid of the limiter tube in the second IF amplifier when an rf carrier signal is absent. The time constant of the limiter grid circuit is designed to be long compared to its IF frequency but sufficiently short to follow the rapidly fluctuating noise peaks. In the absence of a carrier this noise is amplified and then rectified and applied as negative bias to the grid of a tube having a relay winding as its plate load, thus causing the relay to be unoperated. In this condition the relay cuts off the high-voltage supply to the power-amplifier stage and opens the local audio output circuit. As soon as a carrier is received, the noise is reduced by the limiter, the relay operates, the repeater goes on the air and the local audio output circuit is closed. The bias on the noise rectifier diode can be adjusted to cause relay operation by an rf carrier of any level between 3 and 15 μv.

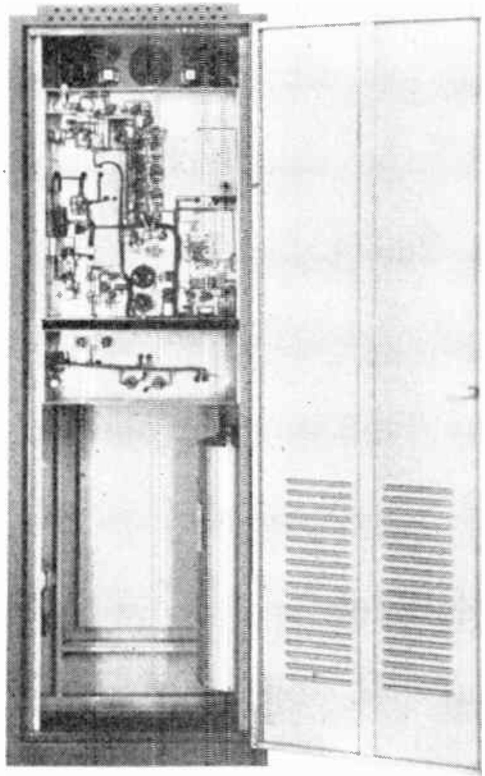


Fig. 3—Radio-repeater equipment, rear view (Type 2486).

Fig. 2 shows the front view of the repeater. The entire repeater is contained on the upper chassis, with the exception of its power supply which is contained on the lower small chassis. All necessary meters and switches for normal control of the equipment are accessible outside of the cabinet.

Fig. 3 shows the rear view of the same equipment. This cabinet is 72 inches high, providing sufficient room for the addition of a cavity resonator which is included to insure noninterference from signals outside of as well as within the system. These figures illustrate the accessibility of all parts of the equipment to provide easy servicing and maintenance.

In addition to the design of the equipment itself, other problems affected its utilization in the Cuban system. The map of the island of Cuba shows Santiago as the point of origin and Havana as the terminal (Fig. 4). It was imperative that, if at all possible, the repeater jumps should be so located that repeaters would occur at points where there were broadcast stations so that audio modulation could be immediately available to that station. A more economically important reason is that the AM radiators could be utilized as repeater towers for the repeater antennas. These AM station locations are at Guantanamo, Holguin, Tunas, Camaguey, Ciego de Avila, Santa Clara, Cienfuegoes, Colon, Matanzas, and Havana.

A complete study of the topography was made and profile curves prepared for each tentative jump. A minimum signal of 5 microvolts was decided upon as being necessary to insure proper operation of the system. Standard nomographs were employed to estimate propagation over each of the paths. These took into account the basic distance involved and the effect, if any, of intervening hills. These results were then adjusted to accommodate known factors involving antenna height, antenna gain, rf power being used, possible

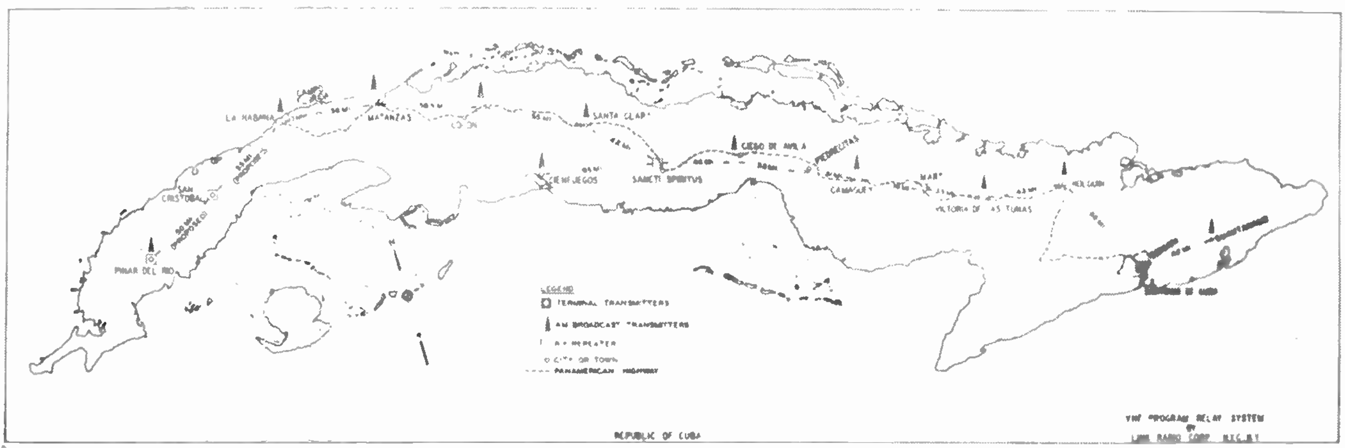


Fig. 4—Map of Cuban repeater installation.

6-db fades, and transmission-line losses. In the majority of cases these predictions worked out remarkably well in practice. However, tests showed that in two of the jumps, where the terrain was comparatively smooth, only insufficient signal was available. (These two jumps were from Tunas to Camaguey and from Camaguey to Ciego. As shown by the map, these two jumps were broken up into four jumps, with added repeaters at Marti and Piedrecitas.) The signal arriving at Marti from Tunas averages 21 μv with practically no recorded fades. The signal at Piedrecitas runs about 15 μv with 6-db fades occurring mostly at night.

A 250-watt terminal transmitter is located at Mt. Boniato about 4 miles north of Santiago, and the antenna was originally directed at Holguin. This terrain is quite rough, indicating the need for 250 watts of power on this jump. However, it was found that 50 watts would give ample signal into Holguin. This error, if it can be so called, was used to good advantage, however, because now the 250 watts of power at Boniato are being split into two antennas and directed to Holguin as originally planned and also to Guantanamo to the east, with approximately 100 watts used in each direction.

All antennas used in the repeater system are high-gain, broadband Yagi arrays, such as that shown in Fig. 5. The design of these antennas was originally prompted by the requirements of this repeater system although they are now being used successfully in many other installations. Actually, four types of Yagi arrays,

Fig. 5—Yagi antenna array used at terminal and repeater stations (Type 2488 A and B and 2489 A and B).

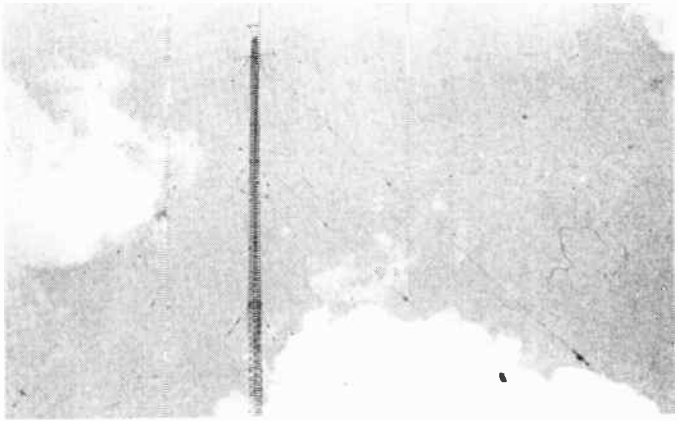


Fig. 6—Yagi antennas mounted at the top of the Camaguey AM radiator.

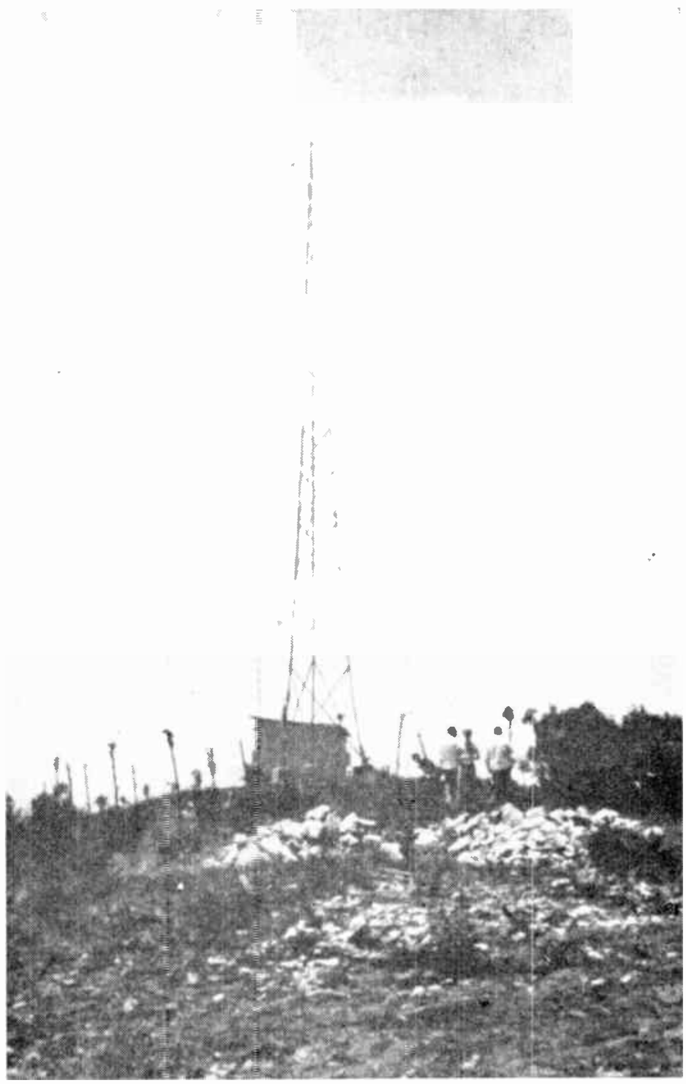


Fig. 7—Yagi antennas at top of tower. Mount Boniato terminal transmitter.

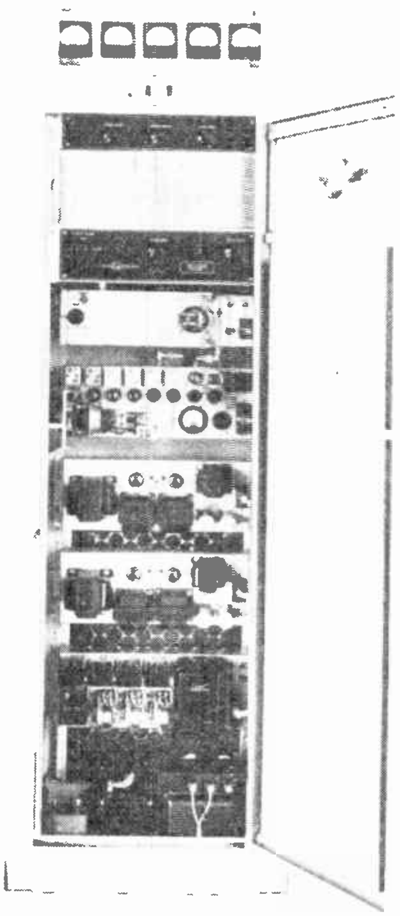


Fig. 8—Terminal-transmitter equipment, front view (Type 2485).

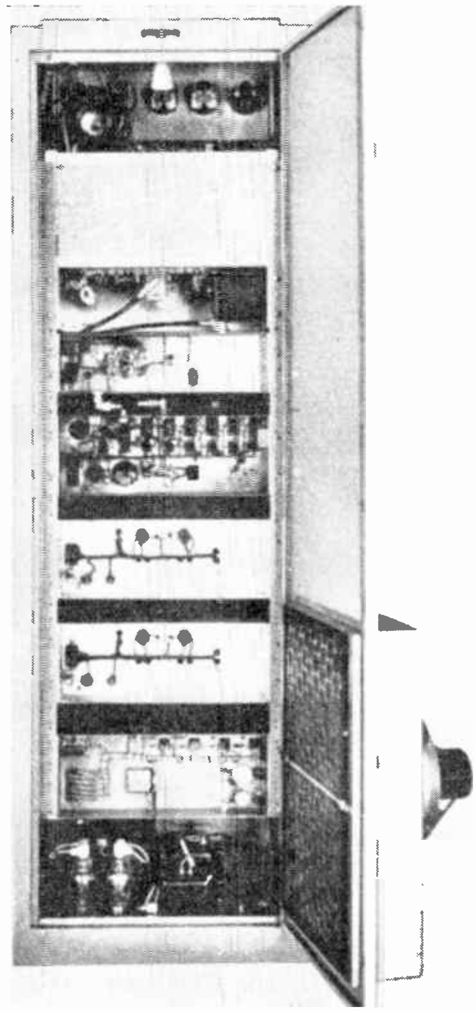


Fig. 9—Terminal-transmitter equipment, rear view (Type 2485).

each having a bandwidth of 5 mc are available for use between 152 and 172 mc. The characteristics of each type are almost identical, each having a front-to-back ratio of 22 db and a forward gain of 9 to 10 db over a half-wave dipole. The beam width of the half-power points is about 35 degrees. Each antenna has a nominal impedance of 50 ohms and each has a "bazooka" converter to adapt the balanced antenna to the unbalanced transmission line. Mechanically, the arrays are constructed of brass and are sufficiently strong to amply withstand wind velocities in excess of 100 miles per hour with severe ice-loading. Mounting facilities permit the use of the arrays with either horizontal or vertical polarization.

Figs. 6 and 7 show some of the typical tower installations. In each case, the Yagi arrays include those used for transmitting and for receiving. The tower shown in Fig. 6 is the Camaguey AM broadcast radiator, such as those used where available. In other cases, towers were erected at repeater sites where no AM broadcast stations existed. Fig. 7 shows one of these towers at the terminal transmitter location on Mt. Boniato. Another consideration in the location of the repeater sites was the availability of primary ac power. This power in Cuba leaves something to be desired in both voltage and frequency regulation. However, each repeater is equipped with a voltage regulator to accommodate wide voltage fluctuations and also with an automatic emergency power supply.

At the terminal transmitter on Mt. Boniato there is a wide-band FM receiver which is energized 24 hours a day. This receiver includes a carrier-operated relay which operates upon receipt of a signal from a 15-watt FM transmitter located on top of the Santiago studio building. When the carrier-operated relay operates, it automatically places the terminal transmitter on the air through suitable time-delay relays. As the transmitter comes on the air, it then energizes the Holguin repeater, which energizes the next repeater and so on to the other end of the system, automatically placing the entire repeater chain on the air.

It is interesting to note that the Santiago AM transmitter now receives its modulation from a receiver which picks up the signal from the terminal transmitter on Mt. Boniato. Since the distance involved is so short, sufficient radiation is received from the back-lobe of the Yagi arrays at the terminal transmitter to produce a usable signal at the AM transmitter site. Originally, the terminal transmitter was controled by an AM receiver located on Mt. Boniato and tuned to the Santiago AM transmitter. The AM transmitter was at that time connected to the studios by a leased line. However, since changing over to the present arrangement, even the the studio-to-transmitter leased line in Santiago has been dispensed with, which not only eliminated the line charges, but also definitely improved the quality of the audio arriving at the AM transmitter.

Fig. 8 shows a front view of the terminal transmitter. The power-amplifier power supply is located on the cabinet floor. Situated on the first chassis are the control, time-delay, and overload protective circuits. The next two chassis are the exciter and intermediate-amplifier power supplies, respectively. Then follows the exciter chassis, intermediate power-amplifier chassis, and finally the power-amplifier stage. Fig. 9 shows the rear view of the same cabinet, and emphasizes the complete accessibility of all parts for servicing. Note that the rear door includes a blower and Dustop filters. This blower automatically cuts in at ambient temperatures above 85 degrees F., and pressure ventilates the entire cabinet.

Each piece of equipment in the system has been designed so that it would operate over long periods of time without the need for frequent service and so that it would make servicing, when required, as simple and easy as possible.

After the system was originally installed, some trouble was found to originate with the antennas which accounted for about 5 per cent of the system outages. It was found that in the torrential tropical downpours, water would find its way into the antennas and cause detuning. Since then, however, the antennas have been modified and waterproofed differently, with the result that interruptions from this cause have disappeared.

Furthermore, each repeater is now equipped with an emergency power source which automatically cuts in upon the loss of primary power. Initially, more than 90 per cent of the outages in the system were from loss of primary power, and called for an emergency source that would instantaneously take over. This has been provided in the form of a 28-volt storage-battery bank which is maintained by a trickle charger. In the event of a power failure, a rotary converter, generating 110 volts at 400 cycles, supplies the equipment with power. 400-cycles were chosen to effect a reduction in converter size, and this imposed no limitations as far as the repeater equipment was concerned since it was designed to operate on any power frequency from 50 to 400 cycles.

Since making these two changes, the outages have been reduced to a negligible amount and the system has been performing very nicely. In fact, its performance has been so satisfactory and so vastly superior to that previously afforded by telephone lines that the system is to be extended west of Havana to Pinar del Rio. Furthermore, Cadena Oriental de Radio is currently building new studios in Havana, which, when completed, will create the need for a similar system operating eastward from Havana to Santiago and Guantanamo; these will be installed in the near future. This fact alone is probably one of the best indications of the suitability of this system of rf repetition for multi-hop repeater systems.

Direction Finder and Flow Meter for Centimeter Waves*

KIYOSHI MORITA†

Summary—**A new direction finder for cm waves has been developed which indicates the wave direction by the minimum response of a dipole placed at the focus of a small paraboloidal reflector. It works well irrespective of whether the polarization of the waves is horizontal or vertical.**

A new flow meter has also been developed which indicates directly the active power flow of the waves, in magnitude and direction. It will serve most effectively in the study of wave diffraction. With an ordinary mirror galvanometer the author has succeeded in measuring the energy flow of 1 μw per cm² at the wavelength of 10 cm.

I. Introduction

THE INCREASING NEEDS of microwave communication necessitates an accurate and complete study of wave nature. Although much work has been done on plane waves of definite polarization, not much has been contributed on the study of waves diffracted by some obstacles. In this case the polarization may be altered and the direction of the waves changed; the energy flow is not necessarily determined by the squared value of electric-field intensity.

Consequently, it may not be meaningless to develop a new direction finder which can tell wave direction accurately whether the plane of polarization is vertical or horizontal, and to develop an energy flow meter.

It is well known that when plane waves are reflected back by a metallic sheet placed at right angles to the direction of wave propagation there will be standing waves developed showing that some of the energy is reflected by the sheet. To know the amount of net energy absorbed in the sheet we have to measure the crest and bottom values of the electric-field intensity of the standing waves. This paper deals with the development of a new indicator which can show directly the net energy flow based on a single point measurement.

* Decimal classification: R501×R271. Original manuscript received by the Institute, July 5, 1950; revised manuscript received, March 9, 1951.

† Tokyo Institute of Technology, Ookayama, Meguroka, Tokyo, Japan.

II. Direction Finder

A. Principle of Direction Finding

Fig. 1 shows a paraboloidal reflector having a dipole antenna placed at the focus and in a direction coincident with the axis. The antenna is connected to a receiver amplifier by a coaxial transmission line. When plane waves come along the axis of Z, toward the paraboloid, the receiver will show no indication. But when the paraboloid is slightly changed in its bearing, the antenna

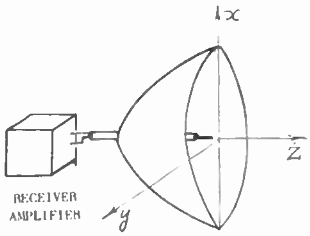


Fig. 1—Paraboloidal reflector and dipole.

will pick up part of the wave energy. The amount depends on the direction of electric vector in the coming waves and on the direction of steering. If the waves are vertically polarized, steering the paraboloid in a horizontal plane does not give rise to any indication in the receiver, whereas steering in a vertical plane does. Thus, we fail to determine the direction of coming waves in the horizontal plane when the waves are vertically polarized. Likewise, we will fail to determine the direction in the vertical plane when the waves are horizontally polarized.

In Fig. 2 a novel means to get rid of those defects is shown. It consists of a pair of quadrant shades made of metal sheet and placed at the face of the paraboloid. When the vertically polarized waves approach along

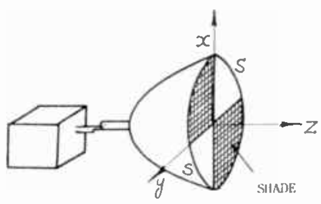


Fig. 2—Paraboloid with quadrant shades.

the Z axis, toward the paraboloid, a part of the energy is reflected back by the pair of quadrants and the rest enters into the paraboloid. The latter can be considered as consisting of two partial waves, one with its electric vector perpendicular to the arc of rim S and the other with its electric vector parallel to S. Of these two, the former partial waves only contribute to direction finding. They give an indication on the receiver when the paraboloid is steered.

Figs. 3(a) and 3(b) illustrate these circumstances. With an ordinary paraboloid the general tendency of the output of receiver versus steering angle is shown in

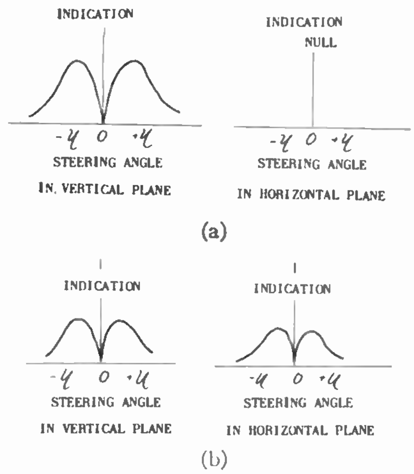


Fig. 3—Waves polarized in a vertical plane. (a) Characteristics of an ordinary paraboloid. (b) Characteristics of a quadrant paraboloid.

Fig. 3(a), while with the new quadrant paraboloid the characteristics are given in Fig. 3(b). Thus, in the latter case only we have a device which gives the direction of wave propagation in terms of the angles in vertical and horizontal planes.

B. Calculation of Directional Property

In calculating the directional property of the paraboloid with a dipole placed along the axis at its focus let us take an analytical figure of the paraboloid, as in Fig. 4. The focal point is indicated by F and the downcoming waves by SP. This contributes to the output of the dipole at F only if the electric vector of the waves has a component in the plane OPF. The contribution is calculated on the basis that it is inversely proportional to the distance r between F and P.

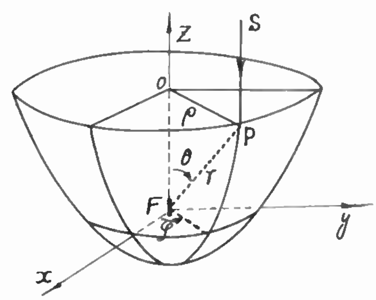


Fig. 4—Paraboloid.

Generally, the wave front is not necessarily perpendicular to the Z axis, and the waves which arrive at the opening of the paraboloid differ in their phases point by point; this must be taken into account in the calculation of the directional characteristics.

The calculated values are as follows:

$$\frac{E_p}{E_0} = j\left(\frac{2\pi f}{\lambda}\right)\cdot f_p \cdot \cos\eta;$$

$$f_p = \frac{1}{2}\int_0^{t_0}\left[\frac{4t}{4+t^2}\right]^2 \cdot J_1(k)dt \qquad (1$$

$$\frac{E_v}{E_0} = j\left(\frac{2\pi f}{\lambda}\right)\cdot f_v \cdot \cos\eta;$$

$$f_v = \frac{1}{\pi}\int_0^{t_0}\left(\frac{4t}{4+t^2}\right)^2\left[\frac{1-\cos k}{k}\right]\cdot dt,$$

where

E_0 = electric-field intensity incident on the paraboloid

E_p = induced electric-field intensity at the focus when the polarization is parallel with the incident plane (Fig. 5(a)).

E_v = induced electric-field intensity at the focus when the polarization is vertical to the incident plane (Fig. 5(b)).

f = focal distance

λ = wavelength

η = incident angle

$t_0 = a/f$

a = radius at the opening of the paraboloid

$t = \rho/f$

ρ = radius of the paraboloid at the arbitrary section

$k = (2\pi f/\lambda)t \cdot \sin\eta$

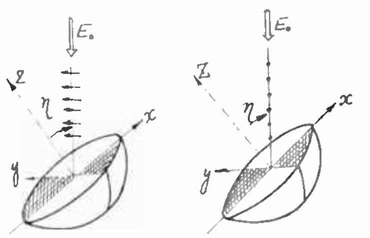


Fig. 5—Waves incident in the XZ plane.

$J_1(k)$ = Bessel function of the first order of modulus k. The integrals involved in E_p, and E_v are not easy to handle in more concrete form, and it is necessary to have their characteristics after numerical calculations.

The author calculated the directional characteristics for the case when

$$a = \rho_0 = 7.5 \text{ cm}$$
$$\lambda = 10 \text{ cm}$$
$$f = 2 \text{ cm}.$$

The results are shown in Fig. 6.

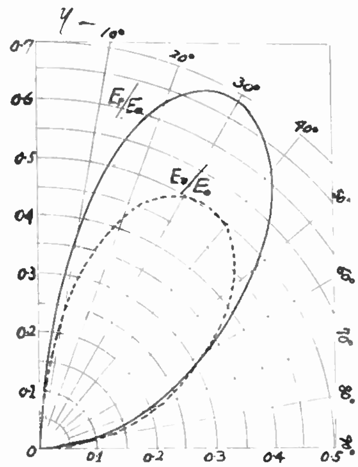


Fig. 6—E_p/E_0 and E_v/E_0 versus η for quadrant paraboloid of dimensions. $\rho_0 = 7.5$ cm and $f = 2$ cm working at $\lambda = 10$ cm.

C. Experimental Result

Experiments were carried on with a paraboloid shown in Fig. 7. It is shaped like a trophy cup and differs greatly from the ordinary paraboloid in that its focus is very deep near the bottom. Calculation shows that the trophy cup will give a much more sharp directivity than an ordinary paraboloid for a given diameter of the

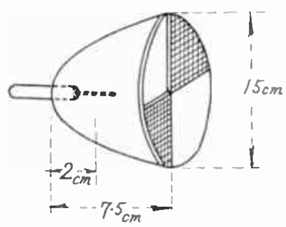
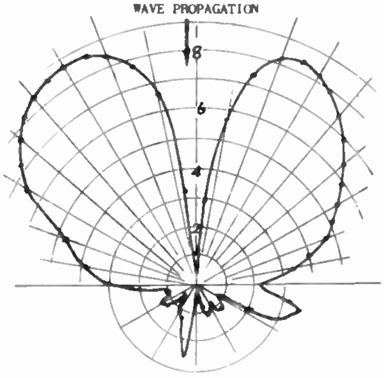


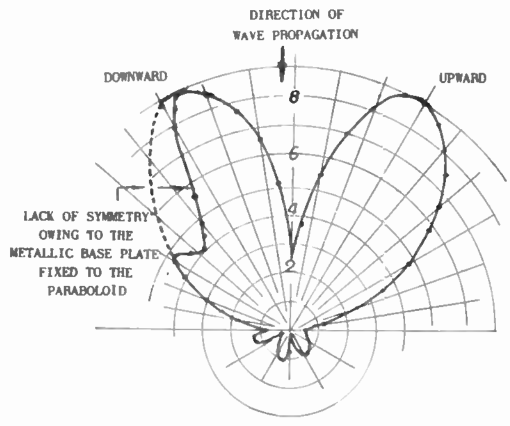
Fig. 7—Experimental paraboloidal reflector.

opening. In the experiment a 10-cm wavelength was used, and at the output terminals of the antenna a crystal detector is connected directly, which in turn gives a deflection on a galvanometer. The experimental results obtained on directional characteristics for horizontally polarized waves are given in Figs. 8(a) and 8(b).



(a)

Fig. 8(a)—Directional characteristics in a horizontal plane



(b)

Fig. 8(b)—Directional characteristics in a vertical plane.

Fig. 9 shows a series of contour lines with output figures as parameters when the bearing of the paraboloid is changed. In the figure the co-ordinates give the bearing in degrees when the electromagnetic field is horizontally polarized.

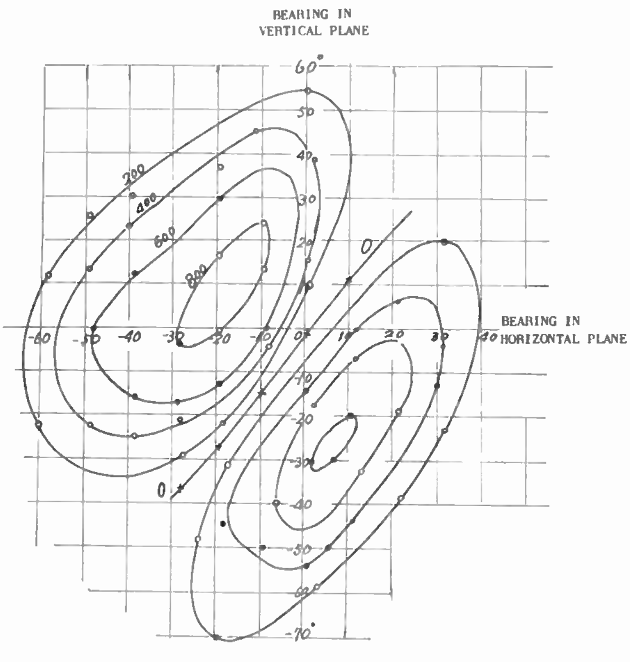


Fig. 9—Contour lines for a constant output. Figures denoted on each curve show the relative magnitude of output intensity.

We see that, instead of a single zero point, there is a line of zero intensity coinciding with the center line of the shading quadrant. Therefore, it is necessary to get another zero line crossing the above one to determine the true direction of the waves. This may be obtained by rotating the 90° quadrant around its periphery and again measuring the line of zero intensity. Thus, we can obtain two lines of zero intensity crossing each other, the cross point giving the direction of the waves under question, as shown in Fig. 10.

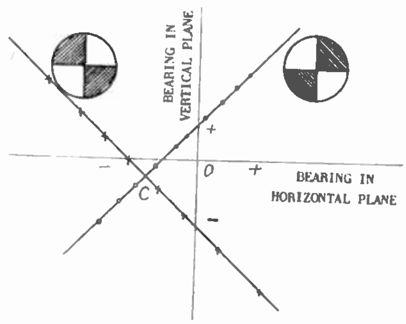


Fig. 10—Determination of direction of propagation by intersection of two lines of zero intensity.

D. *Note for Practical Application*

We know that waves are not necessarily polarized in a simple vertical or horizontal direction. To measure the direction of the wave for arbitrary polarization we have such an arrangement that the shading quadrant can be rotated around the periphery of the paraboloid and be set up at a suitable position. Thus in the case of waves polarized at 45° with the horizontal we have to turn the quadrant 45° more or less, as shown in Fig. 6, and examine the patterns.

In the above description the shading quadrant is made of plain metal sheet. It was found later that better results could be obtained in the directional pattern when a wet cardboard of a few millimeters thickness is placed on the outside surface of this metal shade. It absorbs rather than reflects the wave energy impinging on the shade, and consequently the pattern becomes simpler in shape, showing no subsidiary lobes nor back response.

III. Flow Meter

A. *Conception of Energy Flow*

The energy flow of an electromagnetic field can be indicated by the real part of Poynting vector, that is $(E \cdot H^*)$, where E is the electric and H^* the conjugate of the magnetic field vector. In the case of plane waves H is intimately connected to the electric field, and it may be said that $(E \cdot E^*)$ will give a measure of energy flow. This is the reason why we usually estimate the energy flow by a simple dipole. However, when there is any reflected wave, we will have standing waves in space and the output of a dipole will vary in wide range depending on its position. In such case a dipole never tells about energy flow; it is fundamentally a sort of field-intensity meter.

The author has devised a new flow meter which will give the magnitude of net energy flow, even when there are standing waves. By this apparatus, measurement at one point in space will give sufficient information about the density of energy flow and its direction at that point.

B. *Principles of the Apparatus*

Fig. 11(a) shows the principle of measurement. Here *abcd* indicate a small loop made of a copper wire and *ss* small metal plates placed perpendicular to the plane of the loop and capacitively coupled to the end straps *ac* and *bd*. Now assume a plane wave coming, with its electric vector, in the direction *ab* or *cd*. Then the current induced in the sides of the loop can be calculated with ease when we consider an equivalent circuit of the loop, as shown in Fig. 11(b). It has two generators supplying voltages of equal magnitude but of unequal phase in general. Z_c means the impedance of the loop for circulating current owing to the difference of these two voltages. Especially when the wave comes in a normal direction to the plane of the loop, no circulating current exists; but there still remains another current flowing in parallel through the sides of the loop, and its magnitude will be determined by the total impedance $\bar{Z}_p' = \bar{Z}_p + \bar{Z}_c/4$.

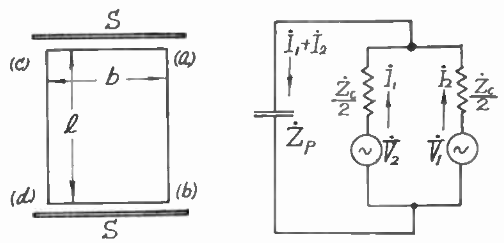


Fig. 11—Flow meter. (a) Simplified sketch. (b) Equivalent circuit.

When the dimensions of the loop are small in relation to the wavelength, Z_p' can be made high capacitive impedance. Now allowing $\bar{V}_1\bar{V}_2$ to be the induced voltage in each side of the loop and $\bar{I}_1\bar{I}_2$ the current through it, we have the following equation:

$$\begin{aligned} \bar{I}_1(\bar{Z}_c/2) + (\bar{I}_1 + \bar{I}_2)\bar{Z}_p &= \bar{V}_1 \\ \bar{I}_2(\bar{Z}_c/2) + (\bar{I}_1 + \bar{I}_2)\bar{Z}_p &= \bar{V}_2 \end{aligned} \tag{2}$$

and

$$\begin{aligned} \bar{I}_{1.2} &= \frac{1}{2\bar{Z}_p'}\left(\frac{\bar{V}_1 + \bar{V}_2}{2}\right) \pm \left(\frac{\bar{V}_1 - \bar{V}_2}{\bar{Z}_c}\right) \\ \bar{Z}_p' &= \bar{Z}_p + \frac{\bar{Z}_c}{4}. \end{aligned} \tag{3}$$

Now we want to choose the value of the impedances in such a way that I_2 becomes zero when the plane wave impinges upon the loop from right to left. Let $\bar{E}_0$ be the electric-field intensity at the center of the loop and l and b the length and span, respectively, of the loop. We then have

$$\begin{aligned} \bar{V}_{1.2} &= \bar{V}_0 e^{\pm j\theta/2}; & \bar{V}_0 &= l\bar{E}_0 \\ \theta &= 2\pi b/\lambda; & \lambda &= \text{wavelength.} \end{aligned} \tag{4}$$

Substituting these into (3) we obtain

$$\bar{I}_{1.2} = 2\bar{V}_0\left(\frac{\cos\theta/2}{4\bar{Z}_p'} \pm j\frac{\sin\theta/2}{\bar{Z}_c}\right). \tag{5}$$

The condition that $\bar{I}_2$ be zero will lead to

$$\bar{Z}_c = j4\bar{Z}_p \tan(\pi b/\lambda). \tag{6}$$

This forms the basic requirement of the flow meter, and under this condition we have

$$\begin{aligned} \bar{I}_1 &= j(4\bar{V}_0/\bar{Z}_c)\sin(\pi b/\lambda) \\ \bar{I}_2 &= 0. \end{aligned} \tag{7}$$

If we use thermojunction to measure current, the output G will be proportional to the squared value of the current. Thus we have

$$G_1 = 16\,k\cdot V_0^2\frac{1}{|\bar{Z}_0|^2}\sin^2\left(\frac{\pi b}{\lambda}\right) \tag{8}$$

$$\begin{aligned} G_2 &= 0 \\ G &= G_1 - G_2 \\ k &= \text{a certain coefficient of the galvanometer} \\ G &= \text{the energy flow of the plane wave.} \end{aligned}$$

We shall show further that if there be any reflected wave the difference of the junction output $G_1 - G_2$ gives the net energy flow. Assume the plane wave comes from right to left, that is, in the $-Z$ direction, and a part of its energy is reflected back the opposite way; the electric field then becomes

$$\bar{E} = E_i e^{j(2\pi/\lambda)z} - E_r e^{-j(2\pi/\lambda)z}, \tag{9}$$

where E_i means incident wave and E_r reflected wave. The induced voltage on each side of the loop becomes

$$\begin{aligned} \bar{V}_1 &= V_i e^{j(2\pi/\lambda)(z+b/2)} - V_r e^{-j(2\pi/\lambda)(z+b/2)} \\ \bar{V}_2 &= V_i e^{j(2\pi/\lambda)(z-b/2)} - V_r e^{-j(2\pi/\lambda)(z-b/2)} \end{aligned} \tag{10}$$

and the substitution of this equation into (3) and (6) will give

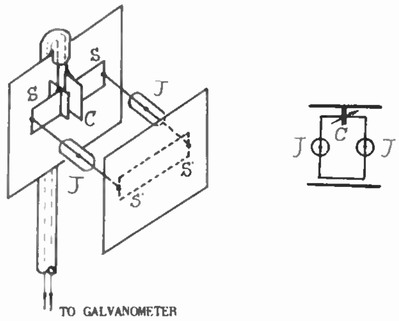
$$\bar{I}_1 = j4\frac{V_i}{\bar{Z}_c}e^{j(2\pi/\lambda)z}\sin\frac{\pi b}{\lambda} \tag{11}$$

$$\bar{I}_2 = j4\frac{V_r}{\bar{Z}_c}e^{-j(2\pi/\lambda)z}\sin\frac{\pi b}{\lambda}. \tag{12}$$

Therefore, $\bar{I}_1$ again indicates the incident wave and $\bar{I}_2$ the reflected wave, and the difference of the output of the thermojunctions will give $V_i^2 - V_r^2$, which means the net energy flow.

C. *Construction of the Apparatus*

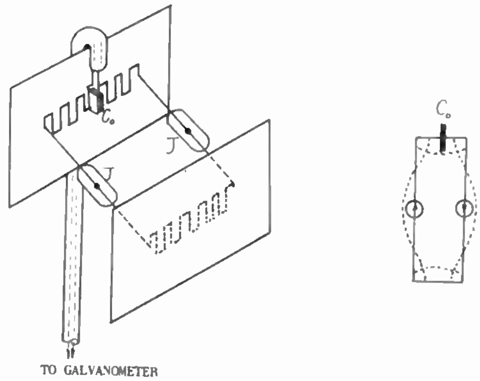
It is necessary to make the flow meter dimensions small in proportion to the wavelength under measurement. The author suggests two types of flow meter, as shown below: one for the decimeter wave region and another for the centimeter region. Figs. 12(a) and 12(b) represent a flow meter for decimeter waves and its circuit diagram. In Figs. 12(a) and 12(b) J is a thermoconverter sealed in an evacuated glass bulb and serves as a generator of thermo emf as well as electric heater; C is a variable capacitor for getting near resonance in a loop circuit; and ss are pieces of insulated copper plates forming a capacitor in free space. At the same time ss are coupled capacitively to the end of the loop, where they are to be likened, in ordinary sense, to some loading capacitor to an antenna. The size of ss, the gap between ss and the end straps, and the resistance of the junction wire should be so chosen that (6) will be satisfied. Whether the impedances are properly chosen or not can be checked by exposing the apparatus to a field of plane waves where the voltage-standing-wave ratio is already known and by observing the output from each junction separately. If the indication does not change when the apparatus is displaced a little, say $\frac{1}{4}$ wavelength along the line of propagation, the condition may be considered satisfied.



(a) (b)

Fig. 12—Flow meter for decimeter waves.

Furthermore, it is necessary to have such an arrangement that, in favor of the screening action of ss, emf will not be induced in the end straps of the loop even if the loop is tilted. This will benefit good directivity characteristics of the flow meter.



(a) (b)

Fig. 13—Flow meter for centimeter waves.

Fig. 13(a) is a flow meter for a centimeter-wave region. As the wavelength is too short, it is advisable to utilize harmonic excitation of the loop, as shown in Fig. 13(b) and to fold it up at the terminals so that small dimensions as a whole may be obtained. C_0 is a blocking condenser to avoid short-circuiting the thermo emf.

D. *Sensitivity*

We constructed a flow meter for a 15-cm wavelength with a thermojunction having 40 ohms resistance and generating 6 mv at the current of 10 ma, and a loop having dimensions 3×2.2 cm (i.e., $b=2.2$ cm). It can detect 1 μw per square cm per 10-mm deflection when a galvanometer of 8.3×10^{-8}amp/cm sensitivity is used as detector.

We utilized this flow meter to study some diffraction phenomena, in the case where the plane waves are distorted by a metallic sphere having a diameter 1.2 times that of the wavelength used, and fairly good results coinciding with calculations were obtained.

CORRECTION

W. W. von Wittern, co-author with H. E. von Gierke of the paper, "Condenser Microphone Sensitivity Measurement by Reactance Tube Null Method," which appeared on pages 633–635 of the June, 1951 issue of the PROCEEDINGS OF THE I.R.E., has requested that the editors publish the following correction:

The caption of Fig. 3 should read: Sensitivity and phase lag of two different microphones. Radial scale = sensitivity in db ref. 1 volt/dyne/cm². Angle of the radii = phase lag in degrees between pressure acting on the membrane and output voltage. Numbers along the curves = frequencies in cps.

Modified Locked-Oscillator Frequency Dividers*

PETER G. SULZER†, ASSOCIATE, IRE

Summary—A simple locked-oscillator frequency divider is described. The circuit, which consists of an *LC* oscillator modified to facilitate locking, is capable of operating over a 7-to-1 plate-supply-voltage range when dividing by a factor of 30.

Introduction

FREQUENCY DIVIDERS have seen wide application where precise time intervals or accurate audio frequencies are required. This results from the fact that it is possible, with present techniques, to obtain the highest oscillator frequency stability at frequencies of the order of 50 or 100 kc. Lower frequencies may be required for running clocks, for interpolation, or for many other purposes, and therefore frequency division is required.

Although several types of frequency dividers have been employed,[1–7] a need was found for a simple but reliable circuit capable of dividing by steps of 10 or greater in a single stage.

Improved Locked Oscillator

In developing the frequency divider to be described, an attempt was made to combine the frequency stability of the locked sinusoidal oscillator with locking ability of the multivibrator.

An analysis has shown[7] that if a voltage of frequency Nf_0 is injected into an oscillator of frequency f_0, synchronization may occur because of cross modulation between the injected voltage and harmonics of f_0. The modulation product can be considered as a synchronizing signal for locking with a one-to-one frequency ratio. It is well known that one-to-one locking is comparatively easy in any nonlinear oscillator.[8,9] Thus it might appear that the synchronizing ability of a sinusoidal oscillator as a frequency divider might be improved by increasing its harmonic content.

A regenerative sinusoidal oscillator may be regarded as a positive-feedback amplifier with a filter in the feedback loop. This filter—normally a tuned circuit or a tuned transformer—permits little loop gain at the harmonic frequencies. By-passing the filter with an all-pass network will increase the gain at harmonic frequencies; however, if this process is carried too far, the tuned circuit will lose control of the operating frequency, and relaxation oscillations will be produced.

As an example, consider the cathode-coupled oscillator[10–12] shown in Fig. 1(a). With proper design, a sine

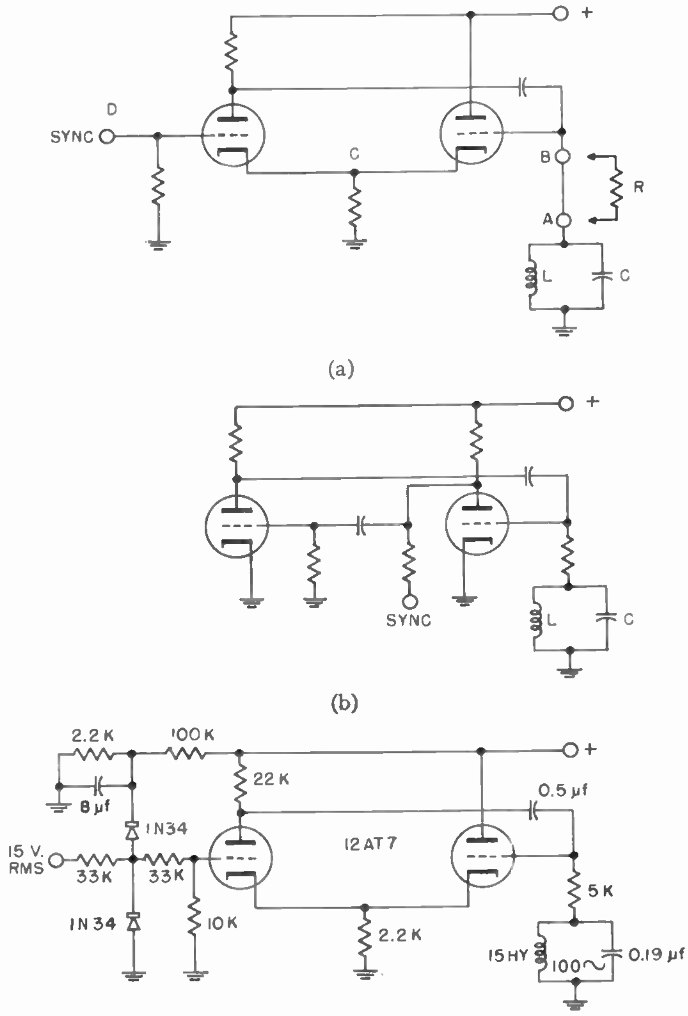


Fig. 1—Locked-oscillator frequency dividers.

* Decimal classification: R213.2×R357.3. Original manuscript received by the Institute, November 22, 1950; revised manuscript received, March 13, 1951.

† National Bureau of Standards, Washington, D.C.

[1] R. L. Fortescue, "Quasi-stable frequency dividing circuits," *Jour. IEE.*, vol. 84, pp. 693–698; June, 1939.

[2] E. Normann, "The inductance-capacitance oscillator as a frequency divider," Proc. I.R.E., vol. 24, pp. 799–803; October, 1946.

[3] C. R. Schmidt, "Frequency division with phase-shift oscillators," *Electronics*, vol. 23, pp. 111–113; June, 1950.

[4] F. C. Williams and T. Kilburn, "Time discriminators, automatic strobes, and pulse-recurrence frequency selectors," Pt. II, IEE Convention; March, 1946.

[5] R. L. Miller, "Fractional frequency generators utilizing regenerative modulation," Proc. I.R.E., vol. 27, pp. 446–457; July, 1939.

[6] L. M. Hull and J. K. Clapp, "A convenient method for referring secondary frequency standards to a standard time interval," Proc. I.R.E. vol. 17, pp. 252–271; February, 1929.

[7] Post Office Engineering Department Radio Report No. 1475, "An investigation of the theory and performance of triode multivibrator frequency division chains with capacitive inter-stage coupling," London, Eng.; 1945.

[8] R. Adler, "A study of locking phenomena in oscillators," Proc. I.R.E., vol. 36, pp. 351–357; June, 1946.

[9] J. Croszkowski, "The interdependence of frequency variation and harmonic content, and the problem of constant-frequency oscillators," Proc. I.R.E., vol. 21, pp. 958–981; July, 1933.

[10] M. G. Crosby, "Two terminal oscillator," *Electronics*, vol. 19, pp. 136–137; May, 1946.

[11] G. C. Sziklai and A. C. Schroeder, "Cathode-coupled wide-band amplifiers," Proc. I.R.E., vol. 33, pp. 701–709; October, 1945.

[12] P. G. Sulzer, "Phase and amplitude stability of the cathode-coupled oscillator," Proc. I.R.E., vol. 38, pp. 540–542; May, 1950.

wave is produced at points A and B, while a half-sine wave is produced at C. Hence harmonics are present, and frequency division by factors of two or three is possible if sufficient synchronizing voltage is inserted at point D. It is found, however, that there is little tendency for stable operation at larger frequency ratios, indicating, from the above, a possible need for greater high-order harmonic content.

A simple way of increasing the loop gain at harmonic frequencies is to insert a resistance R at points A and B. As the value of R is increased, a sine wave continues to be produced at A, but rapid transitions are obtained in the wave form at B, indicating the presence of the desired high-order harmonics. With a moderate value of R, the tuned circuit LC retains control of the oscillator frequency. As R is increased, however, a critical point is reached at which relaxation oscillations are obtained. The oscillograms of Fig. 2, taken from the circuit of Fig. 1(c), show this process.

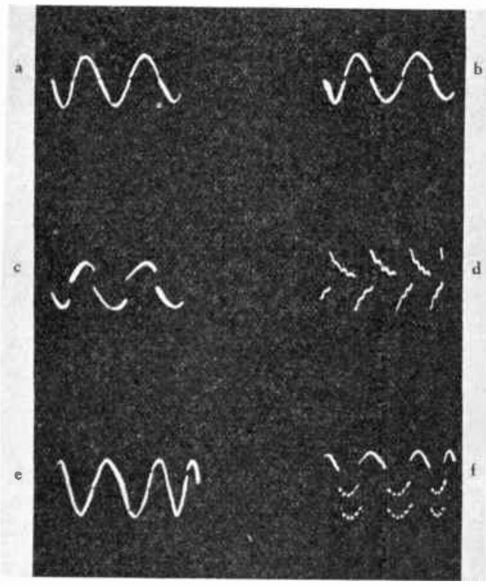


Fig. 2—Oscillograms showing the voltage at point B for the following conditions: (a) $R=0$; (b) $R=3{,}000$ ohms; (c) $R=10{,}000$ ohms; (d) $R=11{,}000$ ohms (relaxation oscillations—sweep frequency was decreased by factor of 10); (e) $R=0$, with 5 volts peak-to-peak applied to point D; (f) $R=10{,}000$ ohms, same synchronizing voltage.

With an intermediate value of R, excellent locking is obtained and stable frequency division by large integers is possible. At the same time, division by rational numbers is observed. Thus an output at 60 cycles can be obtained with an input of 100 cycles, or 50 cycles can be obtained from 60 cycles. The optimum value of R for this or any other condition is best determined by experiment. It appears to be a function of the frequency ratio, the impedance presented by the negative-resistance circuit, and the resonant impedance and Q of the tuned circuit. In any case, the setting of R is a compromise; harmonic content sufficient for locking is required, but excessive harmonic content may decrease the frequency stability of the oscillator to such an extent that synchronization is lost.

A second feedback oscillator[13] that is capable of being modified for frequency division is shown in Fig. 1(b). It is possible that the transitron[14] and screen-coupled[15] oscillators might also be useful in this application. The use of single-tube, transformer-coupled oscillators has not been considered, although a wide-band transformer might permit the use of this type of circuit.

Circuit Performance

Although it may be difficult to establish a criterion for judging the reliability of the circuit, measurement of the plate-supply-voltage range over which locking occurs at a desired frequency ratio appears to be a reasonable choice. Thus the circuit of Fig. 1(c), when dividing by 10, will lock as the plate supply is varied from 26 to 350 volts. This compares favorably with the range of 180 to 350 volts obtained with a well-designed astable multivibrator operating at the same frequency ratio.[16]

One requirement of the modified locked oscillator is that the synchronizing voltage be made proportional to

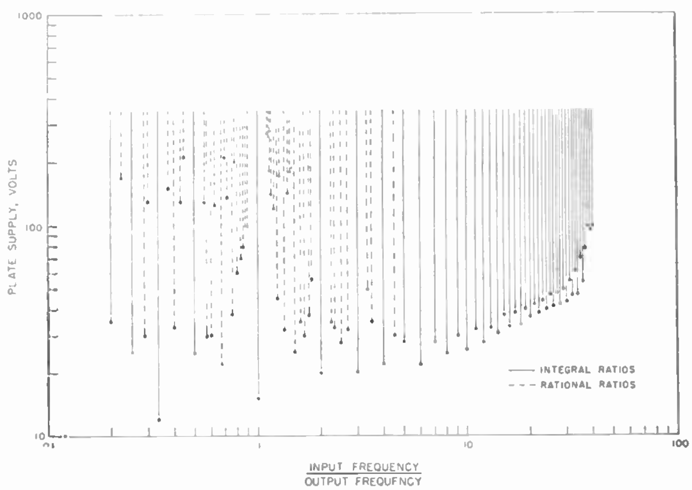


Fig. 3—Plate-supply-voltage range over which locking was obtained versus frequency ratio. Solid lines indicate integral ratios; dashed lines indicate rational frequency ratios.

the plate-supply voltage. This is accomplished in Fig. 1(c) by means of a biased clipper. Such clippers are not required between stages when several dividers are connected in a chain.

Fig. 3 shows the performance of the circuit of Fig. 1(c) at many different frequency ratios. Here the vertical-line length indicates the plate-voltage range over which locking was obtained at the frequency ratio shown by the horizontal scale. It will be noted that synchronization was obtained at frequency ratios from

[13] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., p. 509; 1943.

[14] C. Brunetti, "The transitron oscillator," Proc. I.R.E., vol. 27, pp. 88–94; February, 1939.

[15] P. G. Sulzer, "Applications of screen-grid-supply impedance in pentodes," *Communications*, vol. 36, pp. 10–12; August, 1948.

[16] B. Chance, "Waveforms," McGraw-Hill Book Co., Inc., New York, N. Y., p. 577; 1949.

1/5 to 40. If a plate-supply voltage range of 7-to-1 is taken as indicative of reasonably good stability, frequency ratios up to 30 are possible.

It is worth mentioning that the circuit under consideration divided by 10 with a wide range of type 12AT7 tubes. In addition it functioned with other tube types, including the 12AU7, 12AX7, 6SN7GT, and 6SL7GT, without readjustment. Although long-time tests have not been completed, this is believed to be an indication that a long effective tube life will be obtained, barring accidents such as open heaters or short circuits.

Design and Adjustment

In designing locked oscillators for use as frequency dividers it appears to be necessary to employ empirical methods; however, the following suggestions may be found helpful:

1. The negative-resistance circuit employed should be designed to present the lowest possible value of negative resistance[17] consistent with the available plate-supply voltage and desired plate dissipation. This will aid in obtaining a rapid transition, which will produce high harmonic content.

2. It has been found by experiment that the optimum tuned-circuit Q is approximately equal to the frequency ratio employed, when the circuit is used as a frequency divider. A lower value of Q will produce poor frequency stability which may cause a loss of synchronization, while a much higher Q will make adjustment difficult, and will also result in a poor phase lock and restricted operating bandwidth.

The tuned circuit must contain linear elements. Molybdenum-permalloy inductors have been found very satisfactory. When very high inductances are required, nickel-iron alloy cores can be used if a sufficiently large gap is provided to stabilize the value of the inductance under conditions of changing flux density. If this precaution is not observed, variations in oscillation amplitude due to supply-voltage or tube changes may cause sufficient detuning to effect loss of synchronization.

In selecting the value of the inductance it is satisfactory to take product henrys × cycles (operating frequency) equal to 1,000. Deviations by factors of 1/3 to 3 from this rule will permit satisfactory results providing a sufficiently high Q can be obtained.

3. In aligning the circuit, a small synchronizing voltage is applied, and the resonant circuit is set for locking at the desired frequency ratio by means of a decade capacitor. The synchronizing voltage is then increased until the oscillator stops functioning. The optimum synchronizing voltage is approximately one half this value. If an excessive amount of synchronization is applied, oscillation will cease at low plate-supply voltage or with low-transconductance tubes.

Divider-Chain Example

The circuit of Fig. 4 is used to divide from 100 kc to 1 kc to obtain driving voltage for a 1-kc clock. Sine-wave and pulse outputs are provided for timing purposes.

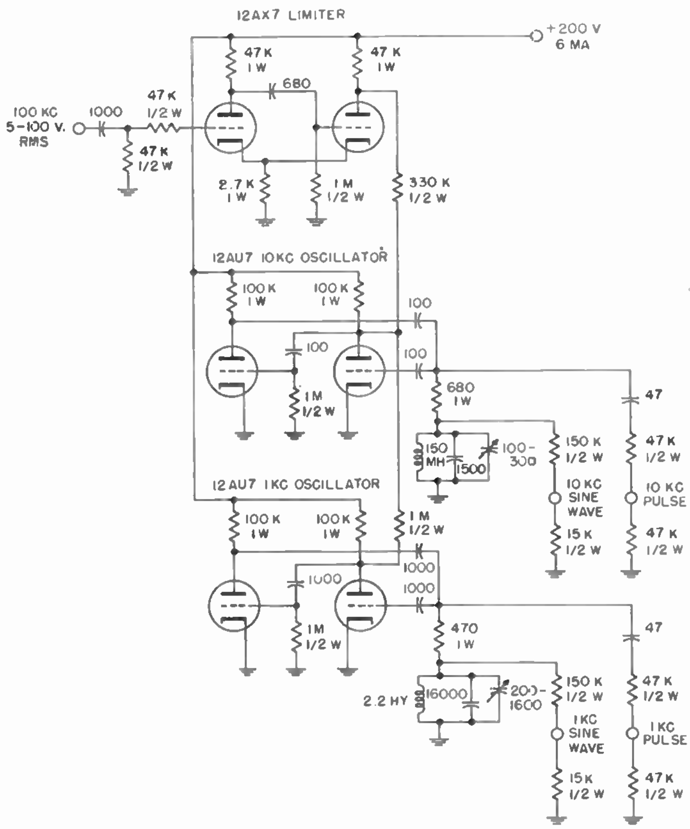


Fig. 4—Divider chain producing 10-kc and 1-kc sine waves and pulses from a 100-kc input. All capacitances in $\mu\mu f$. $K=10^3$. $M=10^6$.

The divider chain will operate properly as the plate supply is varied from 30 to 350 volts; at the same time, the heater supply can be varied from 2.8 to 12 volts.

The approach used in this paper has involved the modification of a sinusoidal oscillator; however, the same result can be obtained by considering the process as one of stabilizing a multivibrator by means of a tuned circuit. Builder[18] has described a successful gas-tube divider using resonant stabilization, although performance data were not given.

[17] P. G. Sulzer, "Cathode-coupled negative-resistance circuit," Proc. I.R.E., vol. 36, pp. 1034–1039; August, 1948.

[18] Geoffrey Builder, "A stabilized frequency divider," Proc. I.R.E., vol. 29, pp. 177–181; April, 1941.

The Source of Long-Distance Backscatter*

WILLIAM G. ABEL†, AND LEONARD C. EDWARDS†, ASSOCIATE, IRE

Summary—This paper describes a method of determining the source of the long-distance backscatter and the results obtained using this method at frequencies of 9, 12, 16, and 22 mc.

Introduction

IN THE PAST there has been conflicting opinion among research workers in ionospheric physics as to whether the backscatter has its origin in the *E* region of the ionosphere or at the surface of the earth. Eckersley[1] concluded that all backscatter is from the *E* region. Benner,[2] and Hartsfield, Ostrow, and Silberstein[3] reported that long-distance backscatter is made up of both *E* region and ground scatter. Peterson[4] concluded that, in most cases, the backscatter is from the ground. It is the purpose of this paper to present a method for determining the source of long-distance backscatter, and to report the results of an investigation, using this method at frequencies of approximately 9, 12, 16, and 22 mc.

Description of Backscatter

Scattering of electromagnetic radiation occurs when it encounters irregularities along the path of propagation which are large compared to the wavelength. For frequencies above several mc such irregularities are found both at the surface of the earth and in the *E* region of the ionosphere.

If electromagnetic waves are transmitted at a frequency above the critical frequency of the *F* region for vertical incidence, the skip-distance ray will follow a path similar to *ACE* in Fig. 1. If the waves are scattered after reflection from the *F* region, either in the *E* region at $\overline{DD'}$ or on the ground at $\overline{EE'}$, some of the energy may be returned as the long-distance backscatter along path $\overline{DD'}C'CA$ or path $\overline{EE'}C'CA$.

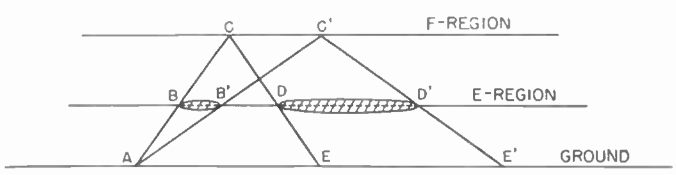


Fig. 1—Sources of backscatter.

* Decimal classification: R113.242.2. Original manuscript received by the Institute, November 20, 1950; revised manuscript received, March 16, 1951.

† Raytheon Manufacturing Company, Waltham, Mass.

[1] T. L. Eckersley, "Analysis of the effect of scattering in radio transmission," Jour. *IEE* (London), vol. 86, pp. 548–563; June, 1940.

[2] A. H. Benner, "Predicting maximum usable frequency from long-distance scatter," Proc. I.R.E., vol. 37, pp. 44–47; January, 1949.

[3] W. L. Hartsfield, S. M. Ostrow, and R. Silberstein, "Backscatter observations by the central radio propagation laboratory—August, 1947 to March, 1948," *Jour. Res. Nat. Bur. Stand.*, vol. 44, pp. 199–214; February, 1950.

[4] A. M. Peterson, "The interpretation of long scatter echo patterns," *Proc. Conf. Ionospheric Res.*, vol. 1, pp. 1F–11F; June, 1949.

Long-distance backscatter received at the transmitter and presented on an oscilloscope with a linear horizontal time sweep produces a pattern similar to that in Fig. 2. The transmitted pulse appears as a single pip

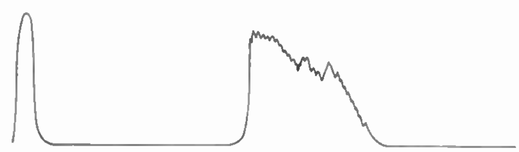


Fig. 2—A-scope presentation of backscatter pattern.

at the left of the trace; the backscatter pattern, as a succession of pips of slowly varying amplitude near the center.

Method of Determining Backscatter Source

For frequencies greater than about 1.5 times the vertical-incidence critical frequency of the *F* region, the leading edge of the backscatter pattern is made up of scatter produced by the skip-distance ray either in the *E* region at *D* or on the ground at *E* in Fig. 1. Delay time to the leading edge of the pattern is then a measure of propagation time between the transmitter and the near edge of the backscatter source, along either *ACDCA* or *ACECA*. If a beacon transponder located at *E* were triggered by the skip-distance ray, delay time to the beacon response would be a measure of propagation time between the transmitter and the ground along the path *ACECA*.

Since the position of the beacon response, with reference to the leading edge of the backscatter pattern, can be measured accurately when displayed with suitable time markers on a linear oscilloscope trace, it is possible to determine whether the leading edge of the pattern is produced by the *E* region or the ground scatter. If the delay time to the beacon response and that to the backscatter pattern are equal, it is evident that the leading edge of the pattern is produced by the ground scatter at *E*. If, however, the delay time to the beacon response exceeds that to the backscatter pattern by the propagation time between the *E* region and ground along the path *DED*, it is equally evident that the leading edge of the pattern is produced by the *E* region scatter at *D*.

If the leading edge of the backscatter pattern is found to be produced by the ground scatter, the remainder of the pattern can also be expected to be produced by the ground scatter. If the leading edge is produced by the *E* region scatter, then the remainder of the pattern can be expected to be produced by both the *E* region and the ground scatter. The reason for this is clear from the geometry of the propagation path when we remember that scatter is produced when electromagnetic radiation encounters irregularities which are large compared to the wavelength. The vertical angle of arrival of waves arriving in $\overline{DD'}$ (Fig. 1) decreases from D to D'. The effective size of the irregularities parallel to the wave front will also decrease with the decreasing angle so that, at the low angles of arrival, there will be less tendency for scatter production in the E region than at the higher angles. If the skip-distance ray does not produce the E-region scatter but penetrates this region to produce scatter at the ground, there is no reason to believe that waves arriving in the E region at lower angles will produce the E-region scatter. In this case the entire backscatter pattern is produced by the ground scatter. However, if the skip-distance ray does produce scatter in the E region, some of the energy will be scattered in a forward direction and form scatter at the ground. Waves arriving in the E region at lower angles than those of the skip-distance ray, and producing the E-region scatter, will also penetrate to the ground to produce ground scatter. Under these circumstances, the leading edge of the backscatter pattern is produced by the E-region scatter, the remainder of the pattern by a combination of the E region and the ground scatter.

If a beacon transponder is located beyond the minimum skip distance for the operating frequency, the distance to the transponder will become the skip distance twice during the day: in the morning as the skip distance decreases, and in the evening as it increases. Thus, two records per day are obtained when the delay time to the leading edge of the backscatter pattern can be compared with that to the beacon response. In this investigation, four transponders located at distances of approximately 700, 1,000, 1,500, and 2,000 miles from the transmitter, were used.

Equipment

The equipment of the transmitting station consisted of a pulse transmitter, a Hammarlund Super-Pro receiver modified for pulse reception, and timing and recording equipment. The transmitter was operated at approximately 20-kw peak-power output, using 100-microsecond pulses at a repetition rate of 20 pulses per second. The output from the receiver was displayed in two different ways on two separate oscilloscopes, one an A-scope presentation with time markers for monitoring and photographing and the other with an intensity-modulated time base for continuous photographic recording. Three-element Yagi antennas were used at frequencies of 9, 12, and 16 mc, and a five-element Yagi antenna at 22 mc.

Each beacon transponder comprised a pulse receiver and a pulse transmitter operated at a peak-power output of 1 kw. A delay of 1 millisecond was inserted in the beacon between the arrival of the first interrogating pulse and the beacon response. The antennas used with the beacons were vertical whips for the 9- and 12-mc frequencies and horizontal half-wave dipoles at a half-wave above ground for the 16- and 22-mc frequencies.

Analysis of Data

From the continuous, intensity-modulated film records, measurements were obtained of the difference in delay time between the leading edge of the backscatter pattern and the beacon response for all times when a transponder was at the skip distance. These times were readily identified on the film by the converging of the high- and low-angle rays of the beacon response prior to fadeout, and the diverging of these rays following the first reception of the beacon response. A typical film record of the backscatter pattern and the beacon response when a transponder is at the skip distance is shown in Fig. 3. In this case, the distance from the transmitter to the transponder is approximately 1,000 miles. As the high- and low-angle rays converged, the

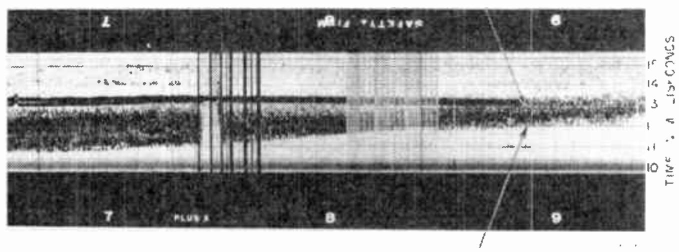


Fig. 3—Intensity-modulated film record of backscatter pattern and beacon response.

delay time to the leading edge of the backscatter pattern increased. At the moment when the beacon response faded out because of the skip distance increasing beyond the distance to the transponder, it was approximately 1 millisecond ahead of the beacon response. Since one millisecond of delay was inserted at the transponder between the triggering signal and the beacon response, the delay time to the beacon response is actually the same as that to the leading edge of the backscatter pattern. Therefore, the backscatter in this case is from the ground, the leading edge of the pattern being produced by scatter at the skip distance, and the remainder of the pattern by scatter at greater distances.

Fig. 4 (see page 1540) shows the various differences in delay time measured between the leading edge of the backscatter pattern and the beacon response when a transponder was at the skip distance, and the number of times each measurement was obtained during operation on all frequencies. The large number of times when the difference in delay time was approximately equal to the inserted 1-millisecond beacon delay clearly indicates that in the majority of cases backscatter was from the ground.

To determine whether there were any instances of E-region scatter, in all cases where the delay time between the leading edge of the backscatter pattern and the beacon response was greater than 1 millisecond, the measured delay time was compared with the delay time required for E-region scatter, i.e., propagation time between E region and ground plus 1 millisecond of delay.

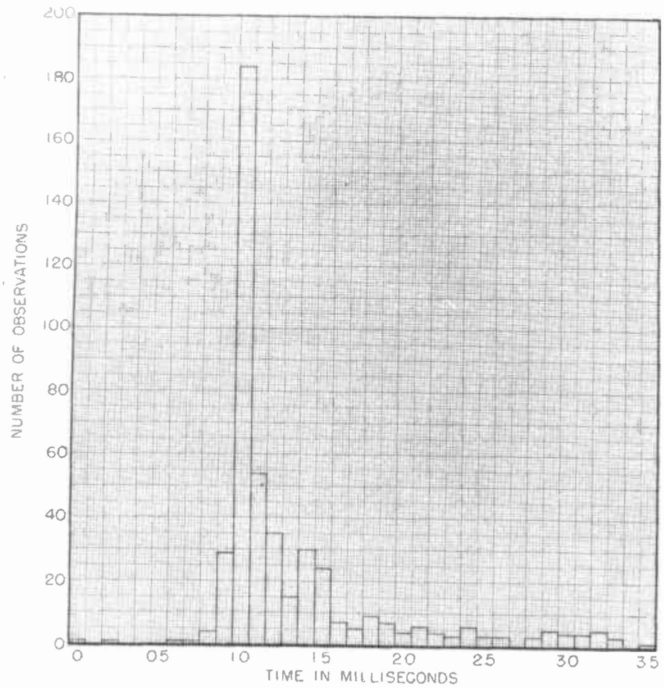


Fig. 4—Delay time between leading edge of backscatter pattern and beacon response averaged for 9, 12, 16, and 22 mc.

The differences between the measured delay time between the leading edge of the backscatter pattern and the beacon response, and the time difference required for *E*-region scatter and the number of times each difference was obtained, are shown in Fig. 5. The few cases where the differences were approximately zero shows that they are probably random cases and not cases of *E*-region scatter.

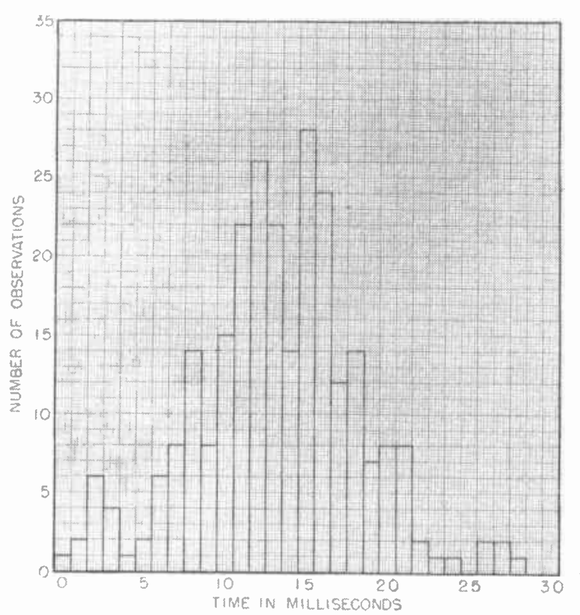


Fig. 5—Difference between measured delay time and delay time required for *E*-region backscatter averaged for 9, 12, 16, and 22 mc.

Since delay times between the leading edge of the backscatter pattern and the beacon response greater than 1 millisecond could not be attributed to *E*-region scatter, the reason for these delay time was further investigated. In most such cases the backscatter pattern on the intensity-modulated film was found to have a somewhat ragged leading edge. This ragged edge corresponds to slowly varying, low-amplitude pips in the pattern displayed on a linear oscilloscope trace. It was also noted that with the antenna pointed at the transponders, and a pattern of low-amplitude pips at the the leading edge followed by an amplitude peak at some distance, this amplitude peak could be moved to the leading edge of the pattern, with no change in delay time to the leading edge, merely by rotating the antenna slightly to the south. This is illustrated in Figs. 6a and 6b. Fig. 6a is an A-scope picture taken with the antenna pointing toward the transponder sites, and Fig. 6b one with the antenna directed approximately

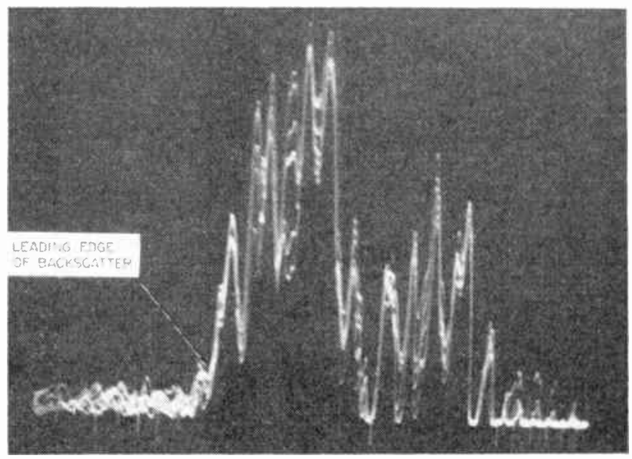


Fig. 6(a)—A-scope presentation of backscatter on 22 mc at azimuth of transponders.

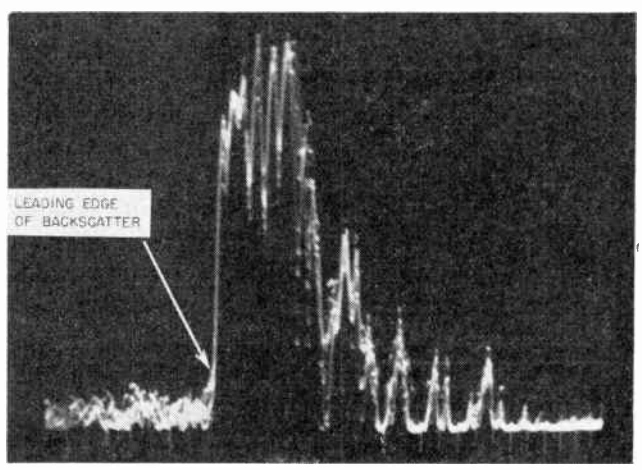


Fig. 6(b)—A-scope presentation of backscatter on 22 mc 30 degrees south of azimuth of transponders.

30 degrees to the south of the first position. The weak leading edge of the pattern in Fig. 6a is from the same source as the strong leading edge in Fig. 6b. It is weaker in the first picture because the backscatter producing this portion of the pattern is arriving at an angle of approximately 30 degrees to the principal axis of the antenna. The arrival of the backscatter from azimuths off the principal axis of the antenna ahead of that along the principal axis masks the true leading edge of the backscatter pattern, particularly on intensity-modulated film. This low-amplitude backscatter at the leading edge of the pattern explains many of the cases shown in Fig. 4, where the delay time between the leading edge of the backscatter pattern and the beacon response was greater than 1 millisecond.

Conclusions

The amplitude peak, either at the leading edge of the backscatter pattern or slightly behind it, is produced by ground scatter at the skip distance measured along the principal axis of the antenna; that portion of the pattern following this peak, by ground scatter beyond the skip distance. Delay time to this peak is therefore a true measure of slant range to the skip distance.

That portion of the backscatter pattern ahead of this amplitude peak is produced by backscatter from azimuths off the principal axis of the antenna, and at shorter distances, arriving ahead of backscatter along the principal axis.

Acknowledgment

This paper is based on work performed under Contract AF28(099)-182 with Watson Laboratories, AMC, USAF, and under the supervision of J. T. deBettencourt of the Raytheon Manufacturing Company.

Space-Charge and Ion-Trapping Effects in Tetrodes*

KARL G. HERNQVIST†

Summary—**Under the assumption of absolute vacuum in a tetrode, the space-charge depression in the screen-grid anode region will at sufficiently high current prevent secondary electrons released at the anode from reaching the grid, thus making the anode-current, anode-voltage characteristic similar to that of a pentode. Under pressure conditions present in ordinary high-vacuum tubes, positive ions formed by collisions between electrons and residual gas molecules will be trapped in the potential trough between the screen grid and the anode. Provided there exists no transverse ion-sweeping fields from beam-forming plates (as in the beam-power tetrode) or from charged end insulators, ion trapping proceeds automatically until the electron charge is neutralized. The effects of ion-trapping are discussed and a pulse method is described for measuring the conditions with and without presence of ions and for studying the ion-buildup process. Plasma ion oscillations have been observed.**

I. Introduction

An ion trap is defined as such a region in an electron flow where the space-charge field of the electrons, together with the primary electrostatic field, acts as a potential trough for positive ions. If the electron velocity is above the ionization potential, such a trough will be filled with ions formed by collisions between electrons and residual gas molecules. This trapping process will proceed automatically until the electron charge is neutralized, the equilibrium condition being determined by such factors as rate of formation of ions and ion temperature. Even at pressures used in ordinary high-vacuum tubes the number of residual gas molecules will be sufficient to supply ions for electron flow of high current density. The trapping of ions in electron beams, observed through its effect on the beam spread, has been studied by Field, Spangenberg and Helm,[1] and by Linder and Hernqvist.[2] However, it seems that the effect has not been studied for other types of electron flow.

The space-charge effect in the grid-anode region of vacuum tubes has been studied both theoretically and experimentally by several workers. Here will only be mentioned the well-known papers by Plato, Kleen, and Rothe,[3] Fay, Samuel, and Shockley,[4] and Salzberg and Haeff.[5] Conditions for ion trapping sometimes occur in such a grid-anode region. However, for reasons discussed below, this effect appears not to have been observed.

* Decimal classification: R138.1. Original manuscript received by the Institute, October 27, 1950; revised manuscript received, March 20, 1951.

† The Research Institute of National Defense, Stockholm, Sweden.

[1] L. M. Field, K. Spangenberg, and R. Helm, "Control of electron beam dispersion at high vacuum by ions," *Elec. Commun.*, vol. 24, pp. 108–121; March, 1947.

[2] E. G. Linder and K. G. Hernqvist, "Space-charge effects in electron beams and their reduction by positive ion trapping," *Jour. Appl. Phys.*, vol. 21, pp. 1088–1097; November, 1950.

[3] G. Plato, W. Kleen, and H. Rothe, "Die Raumladegleichung für Elektronen mit Anfangsgeschwindigkeit," *Zeit. für Phys.*, vol. 101, pp. 509–520, 1936; and Nos. 11, 12, pp. 711–723, 1937.

[4] C. E. Fay, A. L. Samuel, and W. Shockley "On the theory of space charge between parallel plane electrodes," *Bell Syst. Tech. Jour.*, vol. 17, pp. 49–79; January, 1938.

[5] B. Salzberg and A. V. Haeff, "Effects of space charge in the grid anode region of vacuum tubes," *RCA Rev.*, vol. 2, pp. 336–374; January, 1938.

II. Effects of Space Charge and Secondary Emission in the Grid-Anode Region of Tetrodes

In this section the conditions will be discussed under the assumption of absolute vacuum. A brief summary of some of the results obtained in the above-mentioned papers[3-5] will be given.

Consider a parallel stream of electrons being injected into a plane-parallel grid-anode space, the electrodes at first being assumed to yield no secondary electrons. Fig. 1 is a plot of the corresponding potential distribution. Here curve 1 represents the electric field without presence of electrons. Due to the space charge of the electrons, the potential will be decreased as shown by curve 2. Increasing the injected current will further decrease the potential, and for sufficiently high current a potential minimum V_m will be formed, as at 3. However, this depression will not affect the current to the anode; only the transit time of the electrons when passing from G to A will be increased. But when

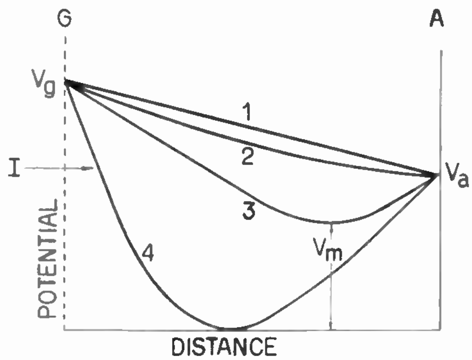


Fig. 1—Potential distributions in a plane-parallel grid-anode space as the injected current is varied.

the injected current is further increased, there will be a certain current at which there is an abrupt decrease in the anode current. This corresponds to a sudden drop to zero in the value of the potential minimum. A so-called virtual cathode has been formed. Some electrons will be reflected to the grid, causing a current partition and an anode-current versus injected-current characteristic with the well-known hysteresis effect. The anode current just before formation of the virtual cathode is the maximum current that can be transmitted through the grid-anode region. It is determined by

$$(I_a)_{\max} = 2.33 \times 10^{-6} \times \frac{(V_g^{1/2} + V_a^{1/2})^3}{a^2} \times S \text{ amperes,} \qquad (1)$$

where V_g and V_a are the grid and anode voltages, respectively, a is the distance between grid and anode, and S is the area of the electron flow.

Now consider the practical case when the anode A is a secondary-electron-emitting surface. Fig. 2[6] is an example of a current-energy distribution of the secondary electrons emitted by such a surface. Curves for

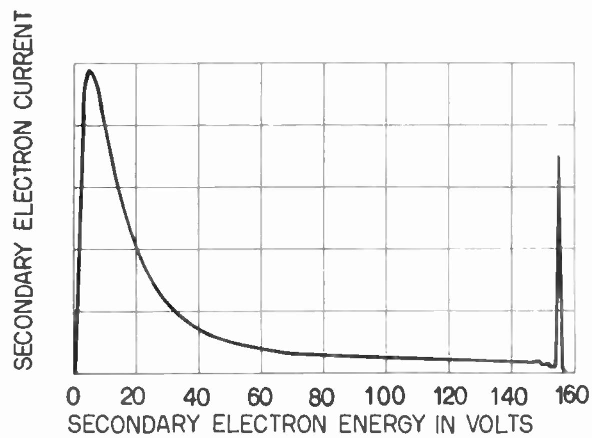


Fig. 2—Current-energy distribution of the secondary electrons emitted by gold at a primary electron energy of 155 volts.

other primary electron velocities and other materials are similar in shape, having a maximum at about 2 to 6 volts, practically independent of the primary electron velocity. If the anode voltage is equal to or less than the grid voltage, secondary electrons from the anode will enter the grid-anode region and their space charge will influence the potential distribution curve of the primary electrons shown in Fig. 1. To determine quantitatively the resulting potential distribution curve would be almost impossible. If there is a depression in the resulting potential distribution, only a part of the secondary electrons whose velocity distribution is determined by a relation similar to that of Fig. 2 will overcome the retarding potential. Introducing the following symbols,

I = current injected into the grid-anode space
I_a'' = secondary electron yield of the anode
$(I_a'')_{\text{refl}}$ = the part of I_a'' that will not overcome the retarding space-charge potential at the anode
I_a = resulting anode current
$\delta = I_a''/I$ = ratio of secondary- to primary-electron currents of the anode,

one obtains the current measured in the anode circuit

$$I_a = I - I_a'' + (I_a'')_{\text{refl}} \qquad (2)$$

or

$$\frac{I_a}{I} = 1 - \delta + \frac{(I_a'')_{\text{refl}}}{I}. \qquad (3)$$

Here $(I_a'')_{\text{refl}}$ will be determined by the depth $(V_a - V_m)$ of the potential depression. For small currents there will be no space-charge potential depression; most of the secondaries will reach the grid and I_a/I will approach the value $1-\delta$. If the injected current is increased, more and more of the secondaries will be reflected back to the anode, and, finally, for high currents none of the

[6] E. Rudberg, "Inelastic scattering of electrons from solids," *Phys. Rev.*, vol. 50, pp. 138–150; July 15, 1936.

secondary electrons will overcome the potential depression, thus making $I_a/I=1$. For still higher currents I_a will reach its maximum value, and finally a virtual cathode will be formed as discussed above. Due to the space charge of the secondaries, the value of $(I_a)_{max}$ will probably be somewhat decreased, compared to the value set by (1).

The results of the above discussion can be transferred to the anode current-anode voltage characteristics of a tetrode as shown in Fig. 3. Here, it is assumed that the conditions for formation of a virtual cathode, determined by (1), are not reached anywhere, and that there is no secondary-emission current from the screen grid. It is seen that for small currents the curves show the well-known dip due to the secondary-emission yield of the anode. However, for reasons discussed above, the dip disappears at high currents and the characteristics will be similar to that of a pentode.

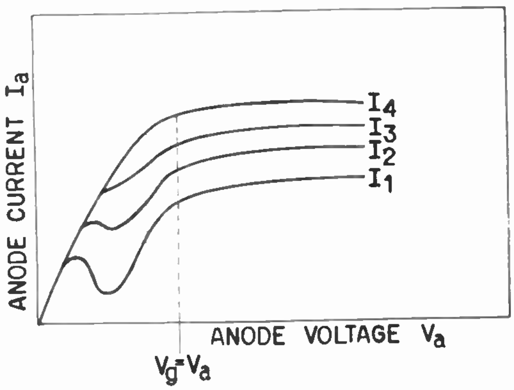


Fig. 3—Anode current-anode voltage characteristics of a tetrode under the assumption of absolute vacuum. Parameter is the current I injected into the screen-grid anode region. $I_1<I_2<I_3<I_4$.

III. Ion Trapping and Its Effects on Space Charge and Secondary-Emission Currents

The discussions of this section will be applied to the practical conditions in an ordinary high-vacuum tube with the presence of residual gas molecules.

Assuming the electron flow discussed in the previous chapter to be infinitely wide, or that there exists no ion-sweeping fields transversal to the direction of the electron flow, conditions for ion trapping are present when the electron velocity is above the ionization potential. The ideas and results obtained in the investigation of ion-trapping effects in electron beams[1,2] are, in general, applicable, even to the type of flow discussed here. In the beam type of flow, positive ions formed by collisions between electrons and residual gas molecules will be attracted to the beam axis, due to the depression in the *transversal* potential distribution of the beam. The same process will occur for conditions such as those present in curves 3 and 4 in Fig. 1, the trapping here being due to the depression in the potential distribution *along* the electron flow. The equilibrium conditions, however, will be somewhat different. Also, the behavior is not so well adaptable to numerical calculations as in the beam case.

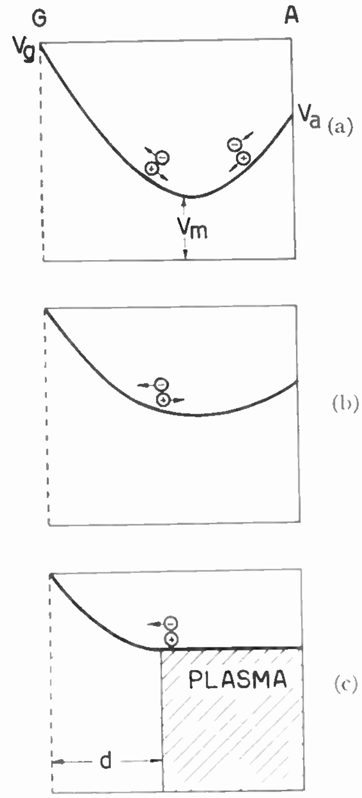


Fig. 4—Potential distributions in a grid-anode space with various degrees of space-charge neutralization.

For a better understanding of the behavior, different steps in the ion-trapping process have been indicated in Fig. 4 for the case discussed in Fig. 1. In Fig. 4(a) the potential distribution is shown for the unneutralized case when the electron flow is just injected. As shown by the arrows, positive ions will be attracted to the potential minimum. Low-velocity electrons formed in the ionization process will be pushed away from the depression. After some time part of the space charge will be neutralized by the trapped ions, and the potential will be increased as shown in Fig. 4(b). Finally, all the space charge in the trough will be neutralized and equilibrium conditions as in Fig. 4 (c) will be reached. Here a plasma region has been formed close to the electrode with lowest potential. The electrode with higher potential is surrounded by a sheath of unneutralized electron flow, the sheath-thickness d being determined by the Poisson equation and the boundary conditions. The boundary conditions are that the potentials V_1 and V_2 at the electrodes, and the electric field strength be zero at the transition to the plasma region. For an injected current density J amperes per

[7] E. G. Linder, "Sheath formation in ion-neutralized electron beams," *Phys. Rev.*, vol. 80, pp. 99–100; October 1, 1950.

square cm, one obtains[3,7] if $V_1 > V_2$

$$d = 1.53 \times 10^{-3} \times \sqrt{\frac{1}{J}}\sqrt{\sqrt{V_1} - \sqrt{V_2}}\,(\sqrt{V_1} + 2\sqrt{V_2})\ \text{cm}. \qquad (4)$$

The ion buildup time, i.e., the time required for the ion trap to be completely filled, has been calculated by Linder and Hernqvist.[2] If the electron velocity is determined by the voltage V, one obtains the ion buildup time measured in microseconds as

$$\tau = \frac{0.0169}{pP(V)\sqrt{V}}, \qquad (5)$$

where p is the pressure in mm of mercury and $P(V)$ is the ionization probability in number of ions formed per cm per electron at a pressure of 1 mm of mercury. Here V is assumed to be constant during the ion buildup period, and the formula does not account for the fact that the potential is increased when space-charge neutralization takes place (e.g., from V_m to V_a for the sheet at the potential minimum in Fig. 4). However, in the practical case the errors caused by other effects will be much larger.

The practical effects of the ion trapping will be:

(1) The transit time of the electrons when passing from G to A will be decreased.

(2) The formation of a virtual cathode will be prevented.

(3) The secondary-emission currents will be influenced.

Transit-time effects will have significance under high-frequency operation, but will not be discussed further here.

Due to ion trapping, ***a virtual cathode can exist only under certain conditions*** although the limit set by (1) is surpassed. Either of the following conditions must be fulfilled: (a) the electron velocity is below ionization potential; or (b) there exists an ion-sweeping transversal field or certain types of pulsed and ac conditions.

Rothe, Plato, and Kleen[3] proved the existence of a virtual cathode under condition (a). In a beam-power tetrode a virtual cathode can be formed because of the ion-sweeping transversal field from the beam-forming plates. Salzberg and Haeff[5] verified experimentally in a specially built tube the existence of a virtual cathode. However, since no detailed information of the structure is given, it can only be supposed that transverse ion-sweeping fields from charged insulators have existed. If none of the above conditions is present, there will be no such limit for the current that can be passed through an interelectrode space as set by (1).

In the practical case when the anode is emitting secondary electrons, as discussed in connection with Fig. 3, the ion trapping will have a noticeable effect on the current distribution. After space-charge neutralization, as in Fig. 4(c), most of the secondary electrons from the anode will reach the grid. If the value of the anode current under space-charge neutralized conditions is $(I_a)_{SCN}$, the relation (3) will become

$$\frac{(I_a)_{SCN}}{I} = 1 - \delta. \qquad (6)$$

Also, all of the anode current-anode voltage characteristics will show the typical dip, independent of the injected current.

From what is said above, the ways of indicating ion trapping are very limited although such effects may be quite common in vacuum tubes. Studying these effects in connection with the formation of a virtual cathode will be rather difficult because of the excessive heating of the grids at the high current density required. Studies of the effects on secondary-emission currents, described below, are limited to voltages where transfer of secondary electrons from one electrode to another is actually secured.

IV. Experimental Verification

A. Method of Measurements

The tube used in the tests (Fig. 5) is a commercial high-power transmitting tetrode, QB 2.5/250 made by Philips.[8] It is an all-glass tube with powder glass base,

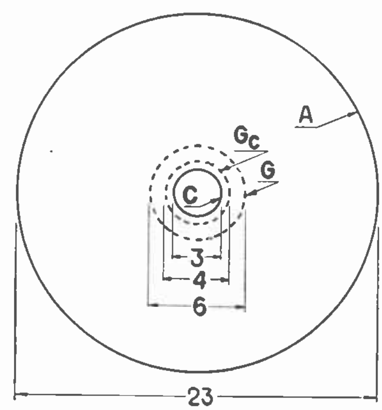


Fig. 5—Drawing showing the approximate dimensions in mm of the tetrode QB 2.5/250.

zirconium-coated carbon anode, molybdenum terminals, and a double helical thoriated tungsten filament. The screen-grid and the anode are supported directly by the terminals and no end insulators exist between these electrodes. Due to the treatment of the anode, the secondary emission has been brought to a minimum. On the other hand, the relatively long screen-grid anode distance, approximately 8.5 mm, has made the space-charge effects very pronounced, and therefore, this tube has been especially suitable for the measurements made here. The anode length is about 30 mm and the emitting cathode length about 25 mm.

In the tests, a pulse technique similar to the one

[8] The tube QB 2.5/250 was later followed by a somewhat modified type QB 3/300 (equivalent to the American-type 4-125A). A tube of this type was tested and showed the same effects as the QB 2.5/250. Since the function of this tube for anode voltages less than the screen-grid voltage is of little importance for the normal operation of the tube, no significant characteristics are given by the manufacturer in this voltage region. The tubes tested were perfectly satisfactory new ones.

used by Linder and Hernqvist[2] was applied. Fig. 6 is a circuit diagram showing the measurement arrangements. The pulser in the control-grid circuit produces a negative pulse of variable width and repetition rate,

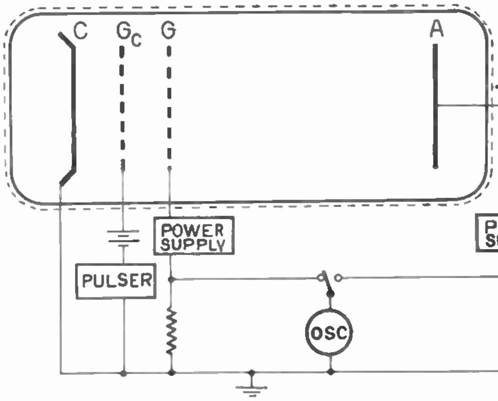


Fig. 6—Schematic diagram of tube and circuit.

and of a sufficiently high amplitude to cut off the cathode current. The oscilloscope could be connected to observe the screen-grid current or the anode current. A typical oscillogram is sketched in Fig. 7. Here control-grid voltage, screen-grid current, and anode current are plotted against time. From a to b are shown the equilibrium conditions with constant currents. At b a negative pulse is applied to the control grid of an amplitude high enough to cut off the currents and having a width great enough to allow trapped ions to be removed from the interelectrode spaces. At e the pulse is terminated, and the grid voltage as well as the cathode-emission current return to their initial values. The anode and screen-grid currents, however, do not immediately return to their equilibrium values. The anode current rises to a higher value, the screen-grid current to a lower. The reason for this is that the positive ions were removed during the application of the pulse and that there is no space-charge neutralization, and therefore, a depression is formed in the potential distribution as shown by curve 3 in Fig. 1. Consequently, some of the secondary electrons from the anode turn back and do not reach the screen grid. Thus the current at e represents the primary anode current minus the part of the secondary-emission current that can overcome the potential depression. If the depression is deep enough, all of the secondaries are reflected and the anode current at e is the primary current. However, in the interval from e to f, ions are again formed and trapped. At f the ion trap is completely filled, corresponding to a potential distribution shown in Fig. 4(c), and most of the secondaries from the anode will reach the screen-grid. Because of the current partition between screen grid and anode, the current to the former behaves as shown in Fig. 7(c). Fig. 8 shows an oscillogram of such a pulse, taken at $V_a = 200$, $V_g = 250$ volts, and a cathode-emission current $(I_a + I_g)$ $= 115$ ma. The pulse length T_1 is 50 microseconds and the interval between pulses $T_2 = 2{,}000$ microseconds.

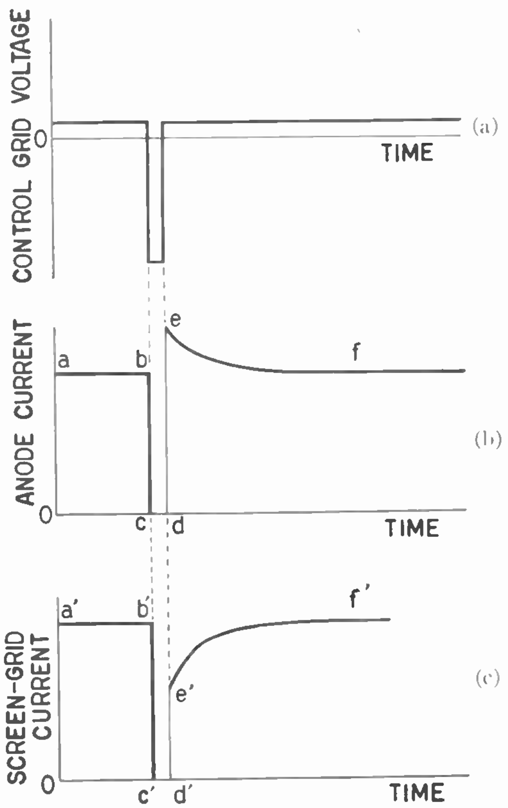


Fig. 7—Typical pulse shapes showing control-grid voltage (a), anode current (b), and screen-grid current (c).

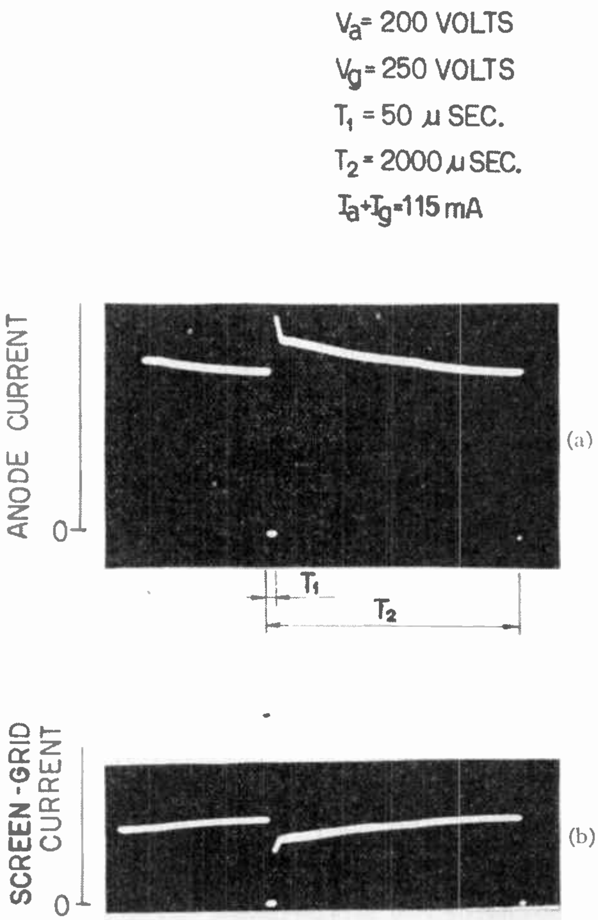


Fig. 8—Oscillogram of anode current (a) and screen-grid current (b).

B. Effects of Anode Voltage and Current Variation

Anode current-anode voltage characteristics of a tetrode without space-charge neutralization as discussed in Fig. 3, can be determined directly from oscillograms such as that of Figs. 7 and 8 (see page 1545). The ratio of the anode current without and with space-charge neutralization $I_a/(I_a)_{SCN}$ is the ratio of the lengths ed/bc. The numerical value of $(I_a)_{SCN}$ may be measured with a meter, while the pulser is switched off. The actual measurements were done on screen-grid oscillograms, such as that of Fig. 8(b), due to the better accuracy obtained here, the secondary-emission current from the anode being a bigger part of the screen-grid current. Since (I_a+I_g) was monitored and observed not to change its value after the pulse, the current represented by length $b'c'-e'd'$ equals that represented by $ed-bc$.

A series of oscillograms of the screen-grid current for different anode voltages are shown in Fig. 9. Here the grid voltages and the cathode-emission current (I_a+I_g) are kept substantially constant. For small values of the anode voltage, the secondary-emission current is small. When the anode voltage is increased, more and more secondaries are released from the anode, as indicated by the point e' moving downward. For anode voltages higher than the screen-grid voltage, again very few secondaries reach the screen-grid due to the retarding action of the primary electrostatic field. From such a series of oscillograms the corresponding anode current-anode voltage characteristics have been plotted in Fig. 10. Here is also shown the the value of $(I_a'')_{refl}$ which is the difference $I_a-(I_a)_{SCN}$. Obviously, the potential depression has not been high enough to prevent all secondaries from reaching the screen-grid, as indicated by the remaining incurved part of the I_a curve.

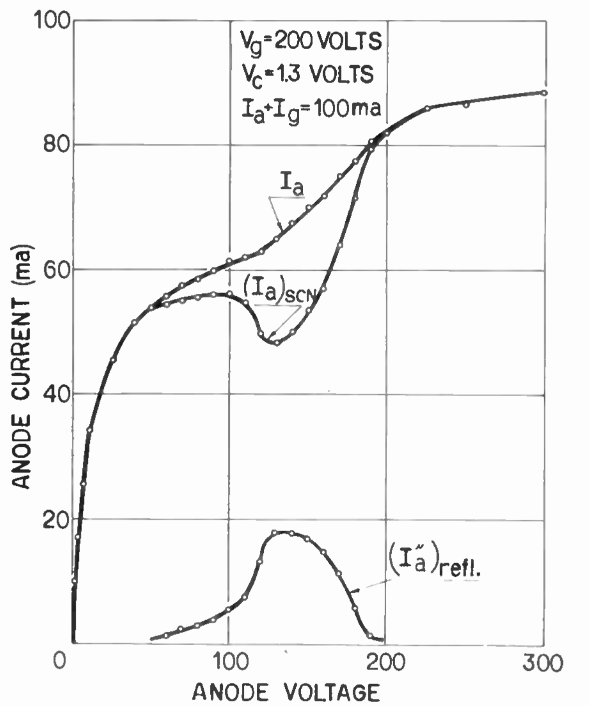


Fig. 10—Anode current-anode voltage characteristics with and without space-charge neutralization.

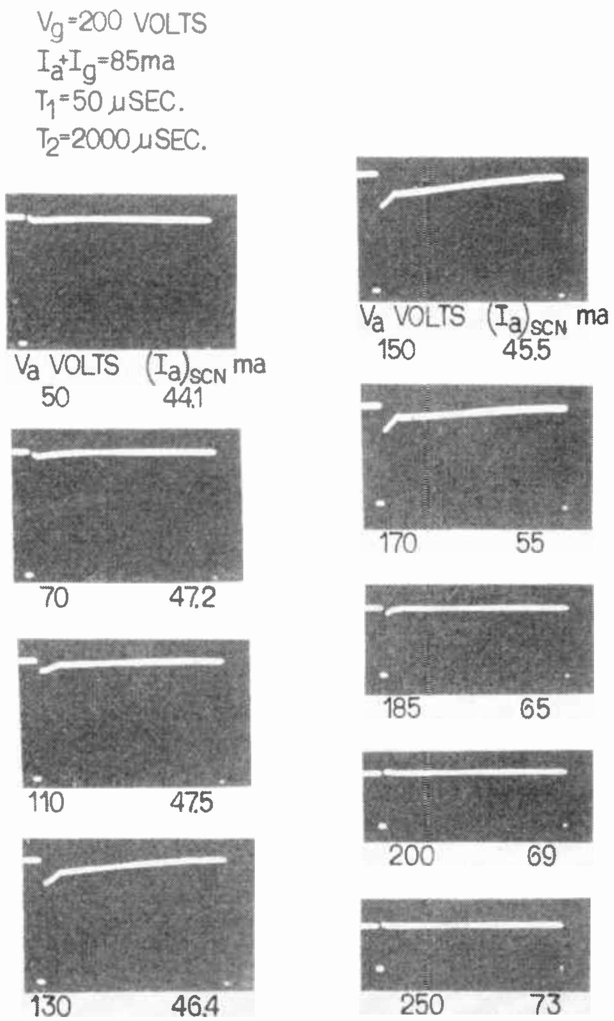


Fig. 9—Oscillograms of the screen-grid current showing the effect of varying anode voltage.

To show the effect of the injected current a number of anode current-anode voltage characteristics with the control-grid voltage V_c as parameter have been plotted in Fig. 11 (see page 1547). Here the filament current is kept constant and the injected current is varied by adjusting V_c. Along each of these characteristics (I_a+I_g) is substantially constant. It is seen how the dip in the I_a curve disappears at high injected currents.

C. Effect of Pressure Variation. Discussion of Pulse Shape.

Plasma Ion Oscillations: The tests described so far have been made on a sealed-off tube. Measurements of the ion current indicate that the data have been taken in a pressure region of approximately 10^{-6} to 5×10^{-6} mm of mercury.

The tube was put on a pumping system providing controlled adjustment of the pressure. A series of oscillograms of the screen-grid current for a range of pressures from 5.3×10^{-6} to 8×10^{-5} mm of mercury are shown in Fig. 12. The decrease in buildup time with increasing pressure, as required by the theoretical formula (6), is of an easily apparent magnitude in the

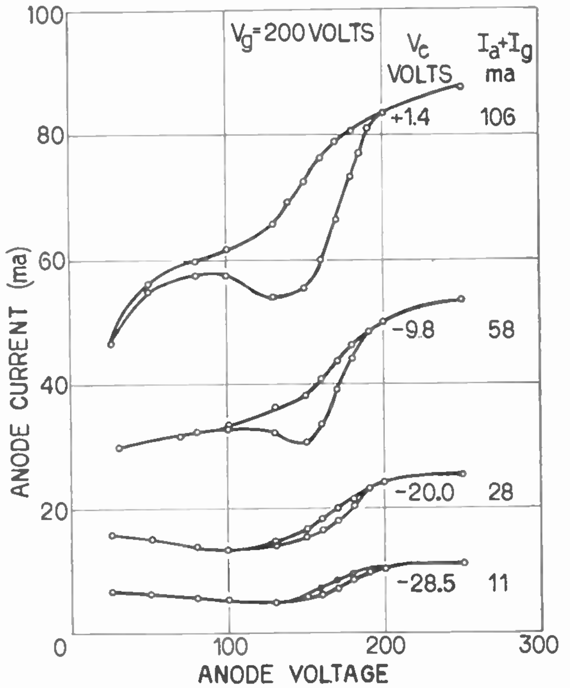


Fig. 11—Anode current-anode voltage characteristics with control-grid voltage as parameter.

photographs. As indicated in Fig. 12, there is a small decrease in the anode current $(I_a)_{SCN}$ under space-charge neutralized conditions when the pressure is increased. This effect may be explained in the following way. Due to the thermal velocities of the ions, there will be a tendency for the ions to escape from the trap. This loss of ions will have to be compensated by the formation of new ions. At low pressures, the rate of formation of ions is not as high as the rate of loss of ions would be for the field-free case in Fig. 4(c), and therefore there will have to be a remaining small potential depression ΔV_m between the screen grid and the anode to satisfy equilibrium conditions. When the pressure is increased, the rate of formation of ions also increases, and consequently ΔV_m decreases. At a sufficiently high pressure the rate of formation of ions will be equal to the rate of loss of ions under field-free conditions, and ΔV_m vanishes, as in Fig. 4(c). Thus the influence of ΔV_m on the low-velocity secondary electrons will account for the decrease in anode current with increasing pressure. These equilibrium conditions have been treated by Linder and Hernqvist.[2]

It is seen from the oscillograms, for instance that of Fig. 8(b), that the ion buildup takes place in two different steps. At first, the screen-grid current rises quite rapidly and with stability. Thereafter follows a discontinuous decrease in the rate of rise of the current accompanied by heavy irregular oscillations (barely visible on the photographs). The nature of these instabilities is not known for certain, but they are most likely caused by the low-velocity secondary electrons at the maximum in the current-energy distribution (Fig. 2). These electrons, which are the ones transmitted through the anode-grid region in the later part of the ion buildup process, represent a considerable space charge. The fact that these electrons mostly have velocities below the ionization potential of the residual gas molecules and do not contribute to the ion formation further increases the ion buildup time.

The unstable oscillations discussed above change to quite stable oscillations when the ion buildup is completed, and the ion density is approximately equal to the electron density. These oscillations can be synchronized with an oscilloscope or measured with a receiver. Oscillations were detected in the frequency region 0.5 to 1.5 mc. Comparison with the theory[9] for plasma ion oscillations indicated that the observed frequencies were of the right order of magnitude. Plasma ion oscillations in electron streams containing ions have been observed by other workers.[2,10]

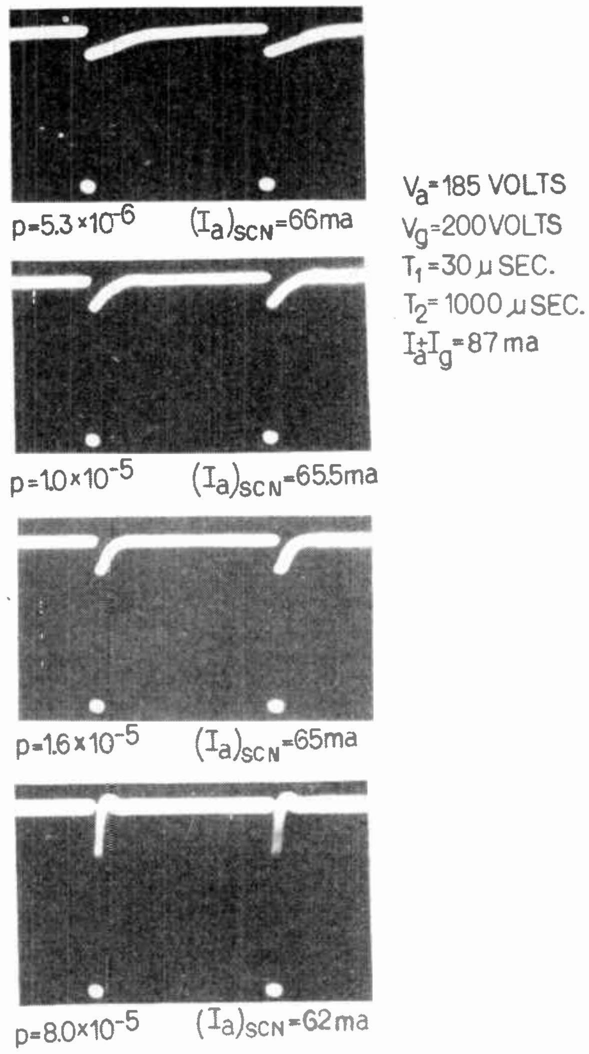


Fig. 12—Oscillograms of the screen-grid current showing the effect of pressure variation (p in mm of mercury).

Acknowledgment

The author wishes to thank Herbert Steyskal for helpful discussions during progress of this work.

[9] I. Langmuir and K. T. Compton, "Electrical discharges in gases," *Rev. Mod. Phys.*, vol. 2, p. 239; 1930.
[10] J. R. Pierce, "Possible fluctuations in electron streams due to ions," *Jour. Appl. Phys.*, vol. 19, pp. 231–236; March, 1948.

Space-Charge Effects in Reflex Klystrons*

M. CHODOROW†, SENIOR MEMBER, IRE, AND V. B. WESTBURG‡, ASSOCIATE, IRE

Summary—Space charge in reflex tubes has an effect which causes considerable departure from the existing reflex theory. To a first order, it modifies the electronic admittance by a bunching effectiveness parameter, designated F, which, in general, takes some value between one and three.

Several approximate methods have been used to estimate this F factor and the results are presented graphically for comparison with experimental measurements on the 723A/B reflex tube, showing good first-order agreement.

I. Introduction

THE SIMPLE THEORY of the reflex klystron has been given by various authors.[1–4] Although the form of the theory differs somewhat among the various presentations, essentially they all give the same result. One useful form of the theory represents the effect of the electron beam as an equivalent admittance across the resonator gap. This is called the "electronic admittance." It is a nonlinear function of the radio-frequency voltage across the gap, and a steady state of oscillation occurs when the radio-frequency voltage and frequency are such that the electronic admittance is the negative of the passive resonator admittance. This electronic admittance depends on the detailed trajectories of the electrons in the reflector region, and, in the conventional treatment, it has been assumed that the field in the reflector region is constant as between plane-parallel electrodes. This neglects the effects of finite reflector size, reflector curvature, and space charge. In this paper some calculations and measurements are discussed which deal with the space-charge effects and which lead to some semiquantitative values for the magnitude of these effects. In the calculations, the approximation of plane-parallel electrodes in the reflector region is retained, but, in comparing with the experimental results, a correction is made for curvature of the reflector. Neglecting the finite cross section of the electron beam should not be too serious if the beam diameter is large compared to the reflector spacing. This condition holds for many existing reflex tubes.

The electronic admittance in general is defined as the ratio of the effective radio-frequency current in the beam to the radio-frequency gap voltage. In the simple theory assuming constant reflector field this turns out to have the magnitude

$$|Y_e| = 2\mu \frac{I_0 J_1(X)}{V_1} \tag{1}$$

where X is the bunching parameter defined as

$$X = \pi N \frac{\mu V_1}{V_0}. \tag{2}$$

I_0 and V_0 are beam current and beam voltage; V_1 is radio-frequency gap voltage; N is the number of cycles transit time in the reflector region; μ is the beam modulation coefficient; and J_1 is the first-order Bessel function. It is assumed here that all the bunching takes place in the reflector region so that N represents the total transit time between successive passages through the gap. If there is any field-free drift region preceding the reflector region, such as a grid of finite thickness, N must be replaced by the difference between reflector transit angle and field-free transit angle.

Although the admittance as defined in (1) is customarily used in discussing reflex tubes, it is possible to derive a more general equation for the electronic admittance which at least formally contains the effect of a nonconstant field.[5] In this more general case, the argument X of the Bessel function is replaced by FX, where

$$F = \left(\frac{2V_0}{\theta}\right)\frac{d\theta}{dV}$$

($\theta = 2\pi N$ is transit angle in radians) can be called the "drift effectiveness factor." As defined, F represents the effect of nonconstant fields only for small radio-frequency voltages, but it is probable that for usual operating conditions in practical tubes it does remain a valid concept. Although a formal definition of F is possible, a quantitative evaluation for any specific geometrical configuration of the reflector or set of space-charge conditions is not simple.

In succeeding sections some approximate methods of calculating the F factor due to space charge are described, and also some measurements on existing reflex tubes. Although the calculations are necessarily crude

* Decimal classification: R355.912.3×R138.1. Original manuscript received by the Institute, August 1, 1950; revised manuscript received, March 8, 1951. Presented, 1950 IRE National Convention, New York, N. Y.; March 9, 1950.

The research reported in this paper was made possible through support extended the Electronics Research Laboratory, Stanford University, jointly by the Navy Department (Office of Naval Research) and the U. S. Army Signal Corps under ONR Contract N6-onr-251 Consolidated Task No. 7.

This work was initiated under the Sylvania Electric Products fellowship in electrical engineering at Stanford University for the year 1947–1948.

† Stanford University, Stanford, Calif.

‡ Stanford Research Institute, Stanford, Calif.

[1] J. R. Pierce and W. G. Shepherd, "Reflex-oscillators," *Bell Sys. Tech. Jour.*, vol. 26, pp. 460–681; July, 1947. See page 481.

[2] E. L. Ginzton and A. E. Harrison, "Reflex-klystron oscillators," Proc. I.R.E., vol. 34, pp. 97–113; March, 1946.

[3] J. C. Slater, "Microwave Electronics," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[4] D. R. Hamilton, J. K. Knipp, and J. B. H. Kuper, "Klystrons and Microwave Triodes," McGraw-Hill Book Co. Inc., New York, N. Y.; 1948.

[5] A. H. W. Beck, "Velocity-Modulated Thermionic Tubes," The Macmillan Co., New York, N. Y., p. 120; 1948. Beck uses R, where $R = F/2$.

because of the inherent multistream character of the electron flow in the reflector region, the agreement among the various methods of calculation and the measurements is quite good. The value of F is found to range between one and three for operating conditions quite typical of those for existing tubes. These large F values indicate that space charge quite drastically modifies the behavior of reflex tubes and must be considered seriously in the design and performance analysis of such tubes.

II. The Static Solution

Before considering the problem of velocity-modulated electrons in the reflector region, it is useful to consider the trajectories, field distribution, and transit time for the static case, i.e., with no modulation.

Neglecting space charge, the field in the reflector region is constant and the time trajectories are simple parabolas as shown in Fig. 1. When space charge is present the trajectories have the same general shape, but space-charge components must be added to the field and potential in the reflector region. The resultant field and potential are shown in the same figure. It may be noted that the derivative of the field has a singularity at the turn-around plane for the electrons.

For the derivation of the trajectories, considering space charge, one has Gauss' law and the current identity

$$\partial E'/\partial x = \rho/K \tag{3}$$

and

$$\rho = -2J dt/dx \tag{4}$$

where E' is the field, ρ is the charge density, J is the current density, and K is the permittivity of free space. The factor 2 must be added to take account of the current due to both entering and returning electrons.

What is needed to solve the equation of motion is the field E', as seen by the moving electron. One can write for the *total* derivative of the field, i.e., the rate of change as seen by the moving electron,

$$dE'/dt = \partial E'/\partial t + (dx/dt)(\partial E'/\partial x)$$

but since $\partial E'/\partial t = 0$, one has simply

$$\partial E'/\partial x = (dE'/dt)(dt/dx).$$

Substituting this relation in (3) together with the expression for ρ, one obtains

$$dE'/dt = -2J/K$$

which upon integration becomes

$$E' = -2Jt/K + C. \tag{5}$$

Each electron being subject to this field, the equation of motion is simply

$$\ddot{x} = -eE'/m. \tag{6}$$

For this static case, even though the velocity at each point is double-valued, a simple solution is possible because of the symmetric nature of the trajectories, such that electrons traveling in opposite directions have velocities of equal absolute magnitude.

The acceleration of the electron, from (5) and (6), becomes

$$\ddot{x} = -(e/m)(-2Jt/K + C). \tag{7}$$

This is one form of Llewellyn's basic relation,[6] used to treat similar space-charge problems.

Integrating twice and evaluating the constants in terms of the boundary conditions, namely, the field $V_r/(d-a)$ at the distance of greatest penetration of the electrons a, (7) becomes

$$x = -(e/m)[-Jt^3/3K + V_r t^2/2(d-a)] + a. \tag{8}$$

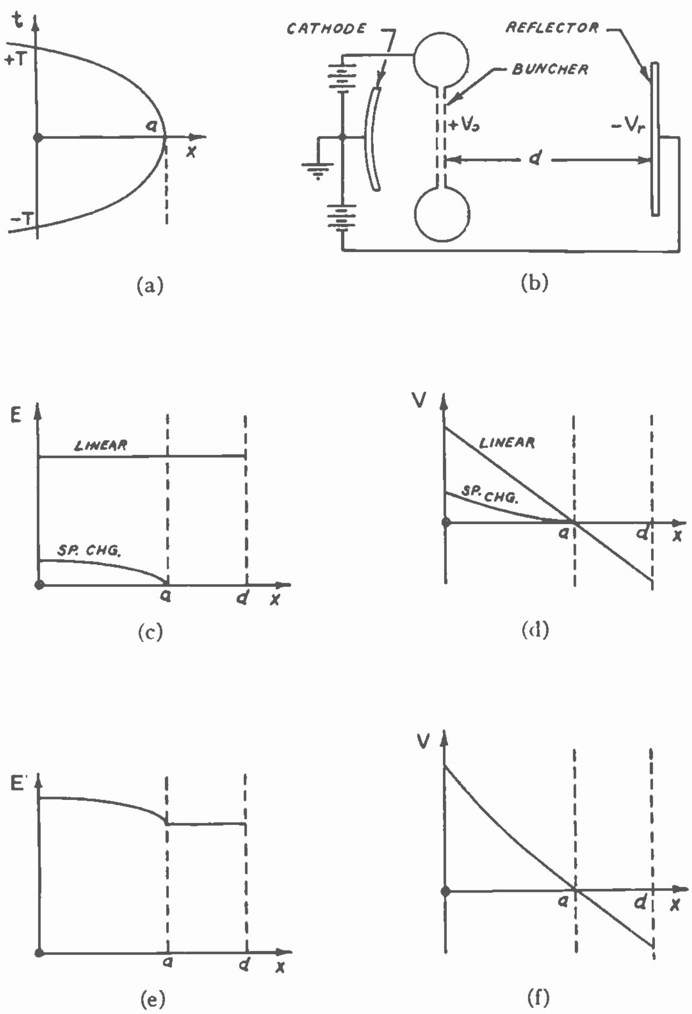


Fig. 1—Field and potential as a function of distance in the reflector region. (a) Electron trajectory. (b) Reflex tube. (c) Linear and space-charge components of field. (d) Linear and space-charge components of potential. (e) Resultant field in reflector region. (f) Resultant potential in reflector region.

[6] F. B. Llewellyn, "Electron-Inertia Effects," Cambridge University Press, London; 1941.

In terms of the initial velocity v, another boundary condition is utilized, yielding the second of the two equations which may be solved simultaneously for the half-transit time T,

$$v = (e/m)[JT^2/K + V_rT/(d - a)] \tag{9}$$

where

$$v = \sqrt{2eV_0/m}.$$

Rewriting (8) for the corresponding conditions, which are $x=0$ and $t=-T$

$$0 = -(e/m)[JT^3/3K + V_rT^2/2(d-a)] + a. \tag{10}$$

In order to proceed with the simultaneous solution it is desirable to reduce the number of variables in (9) and (10). By application of dimensional analysis[7] it may be shown that the half-transit time, when properly normalized, is expressible as a function of just two parameters, a normalized current density and reflector voltage. Normalized quantities will be underlined hereafter and are defined as follows:

$$\underline{V} = V_r/V_0 \qquad \underline{J} = Jd^2/V_0vK$$
$$\underline{x} = x/d \qquad \underline{a} = a/d$$
$$\underline{t} = tv/d \qquad \underline{T} = Tv/d$$
$$\underline{E} = Ed/V_0, \quad = \underline{V}/(1 - \underline{a}) \qquad \dot{\underline{x}} = \dot{x}/v.$$

Here, for convenience, we summarize the notation previously introduced. V_r and V_0 are the reflector and accelerating voltages. v is the entering velocity, x is the displacement, and a is the penetration distance of an electron. t is the time corresponding to the displacement x, and T is half the transit time. d is the distance between the entrance plane and the reflector electrode and E is the field at the turn-around plane, i.e., $V_r/(d-a)$. K is the permittivity of free space.

Using these normalized parameters, (9) and (10) assume in the dimensionless form

$$\underline{J}\underline{T}^2/2 + \underline{V}\underline{T}/2(1 - \underline{a}) - 1 = 0 \tag{11}$$

$$\underline{J}\underline{T}^3/3 + \underline{V}\underline{T}^2/2(1 - \underline{a}) - 2\underline{a} = 0. \tag{12}$$

The solution of these equations for the half-transit time $\underline{T}$ is given in graphical form in Fig. 2.

One of the tacit assumptions has been that the electrons at any plane have the same absolute velocity; in other words, the thermal spread of velocities has been neglected. This assumption is justified for the magnitude of accelerating voltage which is used in klystron work, and the more general development of Fay, Samuel, and Shockley[8] has proceeded on the same basis. In their work, the aim has been to find the potential as a function of distance and the results are less suited for the purpose at hand. In this paper, the electron trajectory itself is of greater interest and it becomes possible to express both the potential, as seen by the electron, and distance as rather simple explicit functions of time. In normalized form these functions are

$$\underline{x} = \underline{J}\underline{t}^3/6 - \underline{V}\underline{t}^2/4(1 - \underline{a}) + \underline{a} \tag{13}$$

and

$$\underline{V}' = \dot{\underline{x}}^2 = [\underline{J}\underline{t}^2/2 - \underline{V}\underline{t}/2(1 - \underline{a})]^2 \tag{14}$$

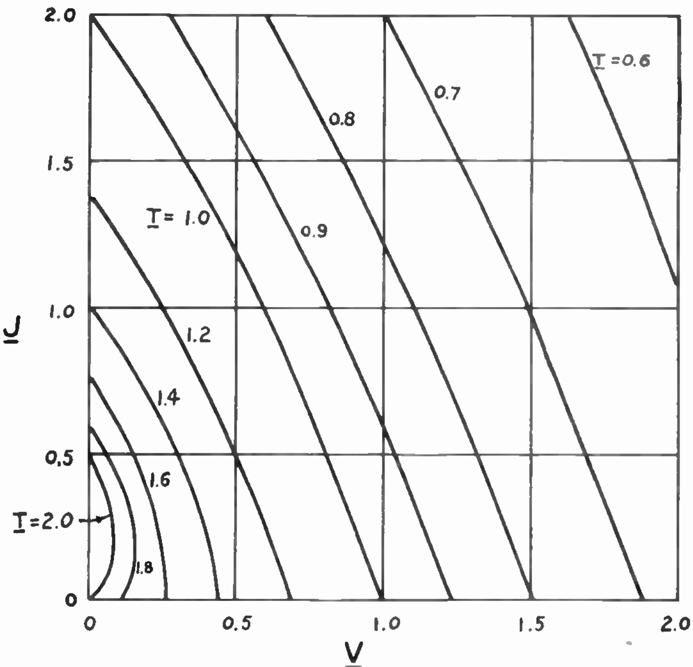


Fig. 2—The static solution for the half-transit time.

III. General Analysis for Nonstatic Case

In order to evaluate the F factor exactly, it is necessary to deal with time-varying fields in the reflector region. This is the result of the fluctuating nature of the space charge in the reflector region under conditions of modulation and the fact that the electron transit time is of the order of several cycles.

There is no general and exact analysis for the fields and trajectories under conditions where the various electrons at a given point and time may have several different velocities, positive and negative. The problem is made difficult because the electron trajectories depend on the fields directly through Newton's second law, while the fields in turn are determined by the charge distribution (and hence the trajectories) through Gauss' law.

The general solution to the problem can be expressed in the form of two equations for the geometry and boundary conditions which have been set forth. To do this, it is necessary to recall that the beam can have two or several components at a given point, corresponding

[7] V. Westburg, "The Electronic Admittance of the Reflex Klystron in the Presence of Space Charge," Ph.D. Thesis, Stanford University, Stanford, Calif.; 1949.

[8] C. E. Fay, A. L. Samuel, and W. Shockley, "On the theory of space charge between parallel plane electrodes," *Bell Sys. Tech. Jour.*, vol. 17, pp. 49–79; January, 1938.

to electrons traveling with different velocities. Associated with each component there is a component charge density and current density. Let these quantities for the nth component (or nth sheet, considering a contour surface of the functions plotted over the $x-t$ plane) be designated J_n and ρ_n, such that

$$\rho(x, t) = \sum_n \rho_n(x, t) \quad \text{and} \quad J(x, t) = \sum_n J_n(x, t).$$

In terms of these quantities the first equation, based on Newton's law and Gauss' law, is

$$\partial(J_n/\rho_n)/\partial t + (J_n/\rho_n)\partial(J_n/\rho_n)/\partial x$$
$$= (e/m)\left\{(V_0 + V_r)/d + (1/Kd)\int_0^d \left[\int_x^d \rho(X, t)dX\right]dx - (1/K)\int_x^d \rho(X, t)dX\right\}. \tag{15}$$

The second equation, the charge conservation principle, is

$$\partial J_n/\partial x + \partial \rho_n/\partial t = 0. \tag{16}$$

The solution of (15) and (16) simultaneously for $J(x, t)$ and $\rho(x, t)$ as n-sheeted functions in the $x-t$ plane constitutes the general solution of this space-charge problem.

Not only is it difficult to carry out the exact solution for the trajectories, but many obvious approximate solutions for the F factor also fail.

Two methods of approximation have been used by Sollfrey.[4,9] Although these have certain limitations for vanishing signals, it is possible to compute the F factor for assumed values of Δ, the modulation coefficient, defined in Section V. His results are tabulated below:

$JT^2\backslash\Delta$	0.025	0.05	0.10
0.34		1.38	
0.70	3.12	2.54	1.70
1.07	4.06	2.76	

Several such approximate methods, including the one of the following section, ignore the fact that the F factor is a function of the transit time relative to the modulation period. This is probably a serious omission, however, since the time variation of the fields near the turn-around plane is great and the electrons pass through this region slowly.

A method of field perturbation, discussed in detail in Section V, is an attempt to include this factor. The method gives a certain insight into the way in which transit-time variations take place from point to point in the orbit, and, of various approximate methods, it provides the best agreement with experimental results.

[9] W. Sollfrey, "Space Charge Effects in Reflex Klystrons," Sperry Gyroscope Co., Great Neck, L. I., N. Y., Report No. 5221-1084, p. 3.

IV. The F Factor from the Static Solution

The exact static solution of Section II may be extended to cover the case where the entering velocity of the electrons is modified by an incremental amount of velocity modulation. The F factor is proportional to the first derivative of transit time with respect to the entering velocity, subject to the restriction that the fields in the reflector region are sensibly constant during the time of transit of an electron, corresponding to modulation at a very low frequency.

To find the first derivative, the variable $\underline{a}$ is eliminated between (11) and (12), at the same time including the entering velocity $\underline{v}$ (although it has a normalized value of unity when there is no modulation)

$$-\frac{\underline{J}^2\underline{T}^5}{6} + \frac{4\underline{J}\underline{T}^3\underline{v}}{3} - 2\underline{J}\underline{T}^2 - 2\underline{V}\underline{T} - 2\underline{v}^2\underline{T} + 4\underline{v} = 0.$$

The variable $\underline{V}$ appears in this equation, but it is expressible as

$$\underline{V} = (1 + V_r/V_0) - \underline{v}^2$$

since the square of the entering velocity represents the difference in potential between the entrance and turn-around planes.

Differentiating with respect to $\underline{v}$ and simplifying, the F factor is found to be

$$F = \frac{1}{\underline{T}}\left(\frac{d\underline{T}}{d\underline{v}}\right) = \frac{2\underline{J}\underline{T}^3 + 6}{\underline{J}^2\underline{T}^5 - 4\underline{J}\underline{T}^3 + 3\underline{J}\underline{T}^2 + 6}. \tag{17}$$

When $\underline{J} = 0$, the charge-free case, this expression reduces to unity properly enough.

A plot of the F factor from the static solution is given in Fig. 3.

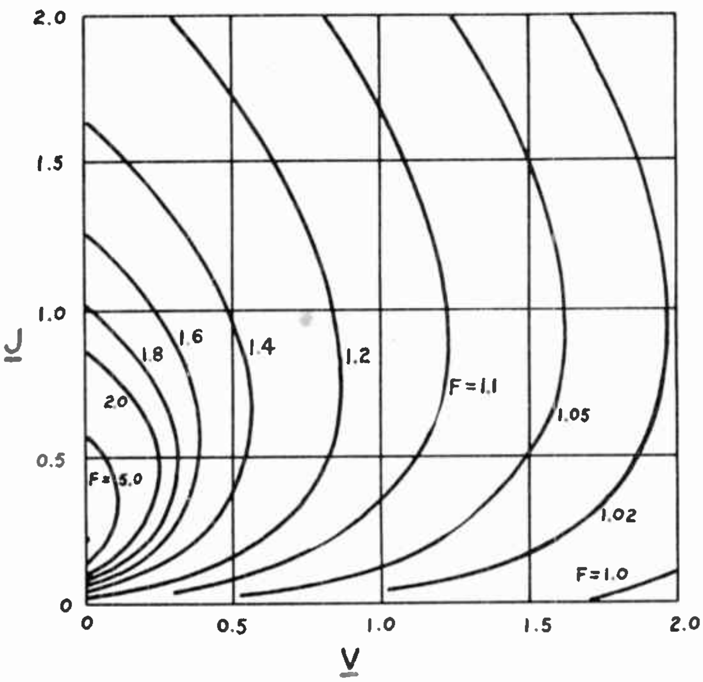


Fig. 3—The F factor from the static solution.

The restriction on the validity of the F factor from this method is simply that $T \ll T_0$, where T is the half-transit time and T_0 is the modulation period. In other words, this F factor is only valid when the time variations are very slow.

V. The F Factor by a Method of Field Perturbation

An approximate value for the F factor can be derived by a method in which one makes use of a time-dependent perturbation of the field in the reflector region. In this method, assuming that the electron trajectory, to a first approximation, remains unchanged from the space-charge-free case, one calculates the effect of a time-varying field on the electron motion. This is entirely analogous to the method of calculating gap transit effects in triodes and klystrons, where one calculates the effect of the gap field on the motion of the electron, assuming the electron has a velocity corresponding to the dc potential.[4,6]

Otherwise stated, one computes the trajectories of velocity-modulated electrons in a constant (space-charge-free) field. From these, a time-dependent space-charge contribution to the field is calculated and then the effect of this extra field as a perturbation is determined, using the original trajectory and the modified field in the transit time integral

$$T = 2 \int_{\underline{v}_1}^{\underline{v}_2} [1/\underline{E}'(\dot{x})] d\dot{\underline{x}} \tag{18}$$

where $\underline{v}_2$ and $\underline{v}_1$ are the entering and leaving velocities, respectively.

$\underline{T}$ from (18) does not correspond to the transit time of an electron free to move at will after being introduced into the field $\underline{E}'(x, t)$, but rather it is an integration along a certain "synthetic trajectory." The property of this synthetic trajectory is that an electron "sees" a time-modulated field in this hypothetical orbit similar to that experienced by an electron in the true orbit. This perturbing field $\underline{E}'(x, t)$ has the desirable property of approaching the actual time-varying field which exists in the reflector region in the limit of no space charge. The accuracy of the results will depend on how much the perturbing field affects the original unperturbed trajectory. As applied here, the accuracy of the method is undoubtedly less than in the original applications by Llewellyn *et al.*[6] Nevertheless, the displacement of an electron with time is the second integral of the field determining its motion; hence, relatively large field fluctuations, and corresponding accelerations, are not reflected by large changes in the trajectory.

To determine $\underline{E}'(\underline{x}, \underline{t})$ it is first necessary to know the charge distribution assuming the parabolic trajectories of the Webster theory for a given modulation level. While an analytic treatment is possible, the Appelgate diagram on $\underline{x}-\underline{t}$ co-ordinates affords an expedient way of determining the charge distribution. From such a diagram, the field $\underline{E}'$ can be determined at any point $(\underline{x}, \underline{t})$ as the sum of the original constant field plus a time-dependent space-charge contribution. The velocity of a given electron at a point $(\underline{x}, \underline{t})$ on its trajectory is easily found from the Webster theory, and with the knowledge of $\underline{E}'$ and $\dot{\underline{x}}$ at all points on a given trajectory the transit time can be computed from (18).

Fig. 4 shows the variation of $\underline{E}'$ versus $\dot{\underline{x}}$ along the trajectory for the electrons with the greatest and least entering velocities on $\underline{E}'-\dot{\underline{x}}$ co-ordinates. The curve for a given electron when $\Delta = 0$ is also shown. Δ, the modulation coefficient, is defined as $\underline{v}_2 = (1+\Delta)\underline{v}$, where $\underline{v}_2$ and $\underline{v}$ are the entering and unmodulated velocities of the electron, respectively. The area on these co-ordinates has the dimensions of transit time; hence the F factor is simply the difference in area between the curves for the fast and slow electrons, divided by the difference in entering velocity 2Δ.

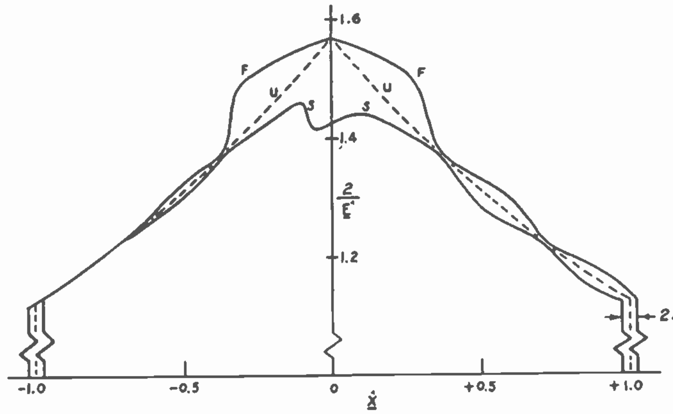


Fig. 4—Graphical integration for transit time by the method of field perturbation for a typical case.

$J = 0.22$
$\overline{V} = 0.5$
$J\overline{T}^2 = 0.37$
$\overline{N} = 2.77$
$\Delta = 0.025$
$N^2\Delta = 0.2$
F = fast electron
S = slow electron
U = unmodulated electron.

Curves for the fast and slow electrons in Fig. 4 give a clear conception of the manner in which the electron experiences field variations at the signal frequency during its transit through the reflector region. In particular, much of the difference in transit time is seen to accrue near the turn-around plane $\dot{\underline{x}} = 0$, and the curves are by no means symmetrical about this ordinate.

Fig. 5 gives the values of the F factor for various degrees of modulation and transit time in cycles. The abscissa $\underline{JT}^2$ is a single parameter which includes the effects of both $\underline{J}$ and $\underline{V}$ and which defines a certain relative static field shape in the reflector region. The use of this parameter as a simplification appeared possible in applying the method of field perturbation, and as an approximation seemed to be adequately justified.

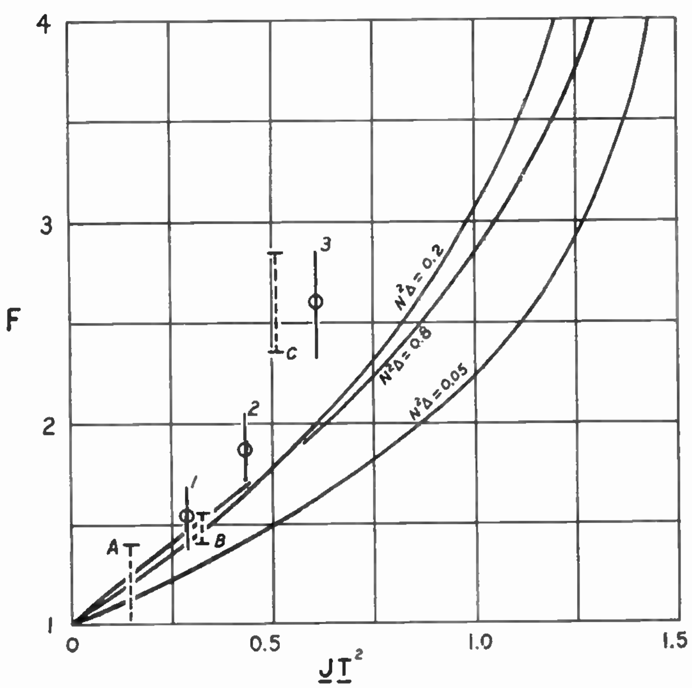


Fig. 5—The F factor from the method of field perturbation and measured values.

Measurement	Tube	N	$N^2\Delta$	$\underline{JT}^2$	F
1	723A/B	4.25	0.523	0.29	1.54
2	723A/B	5.25	0.560	0.44	1.87
3	723A/B	6.25	0.505	0.61	2.59
A	SRC-2	2.43	0.63	0.15	1.07
B	SRC-2	3.43	0.78	0.33	1.47
C	SRC-2	4.43	0.59	0.53	2.61

The curves in Fig. 5 are labeled with respective values of $N^2\Delta$. It may be shown from the parabolic shape of the trajectories that for constant values of the parameter $N^2\Delta$ the family of trajectories on the Appelgate diagram has very nearly the same appearance near the turn-around plane.[7] Thus the parameter $N^2\Delta$ may be thought of as defining the relative positions of the trajectories on the Appelgate diagram near the turn-around plane. Since it seems reasonable to suppose that the F factor is determined by this factor, computations were made for two different transit-time modes having the same value of $N^2\Delta$ and such indeed was found to be the case. The curves of Fig. 5 were computed for the following three values of the parameter $N^2\Delta$,

$$N^2\Delta = 0.05 \qquad \Delta = 0.025 \qquad N = 1.385$$

$$N^2\Delta = 0.20 \qquad \Delta = 0.025 \qquad N = 2.77$$

$$N^2\Delta = 0.80 \qquad \Delta = 0.10 \qquad N = 2.77$$

By way of examination, the method of field perturbation may be used to compute the F factor for a case in which the exact value is known. Such a case, corresponding to the conditions of validity for the F factor from the static solution, is when the electron transit time is a small fraction of a modulation period. For representative parameters, $\underline{J} = 2/9$ and $\underline{V} = 1/2$; these values of the F factor are 1.25 by the method of field perturbation and 1.29 from the static solution.

VI. Experimental Determination of the F Factor

It is possible, by a suitable combination of measurements and use of the theory of a reflex oscillator, to determine approximately the F factor of a reflex tube operating in any mode. These measurements were made on several tubes and the results are in good agreement with the theoretical calculations. To describe this method, a number of relations are required from the theory of a reflex oscillator. These have been derived elsewhere[1–4,10] and are listed here without proof.

First, the electronic admittance at the center of a mode can be written as

$$Y_e = -\mu^2 \frac{\theta F I_0}{V_0} \frac{J_1(FX)}{FX} = -2G_e \frac{J_1(FX)}{FX} \tag{19}$$

where G_e, as defined here, is known as the small signal electronic conductance. μ, the beam modulation coefficient, is defined as

$$\mu = \frac{\sin \phi/2}{\phi/2} \tag{20}$$

where ϕ is the transit angle between the grids. The condition determining the steady-state oscillation is that the electronic admittance be the negative of the circuit admittance. At the center of the mode, the admittances are pure conductances and the condition for oscillation can be written as

$$2G_e \frac{J_1(FX)}{FX} = \omega C\left(\frac{1}{Q_0'} + \frac{G}{Q_e}\right). \tag{21}$$

The cavity admittance on the right consists of two parts—the first due to the internal losses (i.e., cavity plus beam loading losses), and the second due to the external load, coupled to the cavity by means of some transmission line. Thus, in terms of the loaded Q, Q_L, given by

$$\frac{1}{Q_L} = \frac{1}{Q_0'} + \frac{G}{Q_e},$$

$1/Q_0'$ represents the internal losses and G/Q_e the load losses. Here G is the external load conductance normalized to the characteristic admittance of the transmission line and measured at a suitable plane in the line. Q_e is a factor representing the coupling between cavity and line and is more precisely defined below where the method of measuring is discussed. ωC is a geometric factor equal to the ratio of cavity Q to cavity impedance. The power output is equal to the product of load conductance as seen at the gap, by the mean-square voltage, i.e.,

[10] M. Chodorow, "Theory of Reflex Oscillators," part 1, Sperry Gyroscope Co., Great Neck, L.I., N.Y., Report No. 5221–1079; 1947.

$$P = V_1^2 \frac{\omega CG}{2Q_c} = \frac{V_1^2}{2}\left[2G_e \frac{J_1(FX)}{FX} - \frac{\omega C}{Q_0'}\right] \quad (22)$$

which can be written as

$$P = \frac{P_0}{F\theta}\left[2(FX)J_1(FX) - \frac{(FX)^2\omega C}{G_e Q_0'}\right] \quad (23)$$

where $P_0 = I_0 V_0$.

To find maximum power possible in a given mode, one differentiates with respect to the bunching parameter FX, and setting the derivative equal to zero, obtains

$$J_0(FX) = \frac{\omega C}{G_e Q_0'} \quad (24)$$

as the condition determining FX for optimum power. By means of a recurrence formula for the Bessel functions and (21), one obtains a related equation for the optimum load

$$J_2(FX) = \frac{\omega CG}{G_e Q_c}. \quad (25)$$

Simply stated, if one adjusts the load for optimum power, then (24) and (25) are simultaneously satisfied or, taking their ratio,

$$\frac{J_2(FX)}{J_0(FX)} = \frac{GQ_0'}{Q_c}. \quad (26)$$

One more equation is required, that for the electrical tuning, i.e., rate of frequency change with reflector voltage. This can be written as

$$\frac{\Delta f}{\Delta V_r} = -\frac{f_0}{2}\left(\frac{\Delta\theta}{\Delta V_r}\right)\left(\frac{1}{Q_0'} + \frac{G}{Q_c}\right) \quad (27)$$

where f_0 is the resonant frequency of the cavity and $\Delta\theta/\Delta V_r$ is the rate change of transit angle with reflector voltage. Equation (27) is valid for changes in θ such that $\tan(\Delta\theta) = \Delta\theta$.

Using these formulas then, in particular (22), (26), and (27), and some measured quantities, one can determine the F factor. By measuring G the optimum load, Q_c, and Q_0' one can get FX for optimum power from (26); from the optimum power and ωC one gets V_1 from (22). By determining μ and θ and using these values of FX and V_1, F is given by

$$F = \frac{2V_0(FX)}{V_1 \mu\theta}. \quad (28)$$

The various measurements involved will be briefly described; most of them are described in greater detail in the literature. To determine θ, a mode plot of reflector voltage versus accelerating voltage for constant frequency of oscillation can be made. In particular, it can be shown from dimensional analysis that for constant ratio of V_r/V_0, $\theta\sqrt{V_0}$ will be constant. By finding a number of modes it is possible to assign values of θ unambiguously.

μ (20) can be found from the gap spacing and the acceleration voltage. Q_c is obtained by a cold measurement. Q_c represents the coupling between the tube resonator and the transmission line which couples the external load to the tube. It can be shown that the input admittance Y, looking into the cavity from a particular plane in the transmission line, is given by

$$\frac{Y}{Y_0} = \frac{Q_c}{Q_0}(1 + 2j\delta Q_0) \quad (29)$$

where Y_0 is the characteristic admittance of the transmission line, δ is the fractional deviation from the resonant frequency of the cavity, and Q_0 is the cold Q of the cavity. The particular plane for which (29) applies is that at which there is a voltage minimum in the line if a measurement is made at the resonant frequency of the cavity with the cavity detuned. This plane is called the "detuned short." By measuring Y as a function of frequency, Q_c can be determined. The value of G, the normalized load conductance for optimum power, which appears in (26), is also to be measured at the same "detuned short" plane. Q_0 in (29) is *not* identical with Q_0' in general since it contains no beam loading effects. To find Q_0' it is necessary to measure the modulation sensitivity, $\Delta f/\Delta V_r$. By use of (27) it is possible to get Q_0'. This measurement is made with G having the value for optimum power. $d\theta/dV_r$ is found from the same mode plot measurements used in getting θ, and finally ωC is determined with sufficiently good accuracy assuming a field distribution in the cavity and calculating the conductance-Q ratio. It has been shown that this ratio, and indeed the separate factors in it, are quite insensitive to the assumed field.[11]

VII. Discussion of Experimental Results and Conclusions

The experimentally determined values of the F factor for the 4.75, 5.75, and 6.75 modes in the 723A/B tube operating at a wavelength of 3.2 cm are plotted in Fig. 5. Each of these points represents the average of the values in a given mode for the three tubes used in the measurements. The vertical bar through each of these points indicates the 95-per cent confidence limits on the measurements. The vertical dashed bars represent the spread of measurements on a quite different reflex tube—the SRC-2 at a wavelength of 6 cm.

The experimental F factor, F, includes both the effects of space charge and of the nonconstant reflector field because of the reflector geometry itself. Since the latter effect is not appreciable, it was assumed that the two effects were independent and could be superimposed.

[11] M. Ettenberg, "Calculations of Q and Shunt Impedance of Reentrant Cavities," Sperry Gyroscope Co., Great Neck, L.I., N.Y., Report No. 5221-1076; January 3, 1947.

Measurements of the static transit time as a function of reflector voltage for the 723A/B tube indicate that the field in the reflector region is nearly that between two concentric spheres having a ratio of radii of 1.15. Curves calculated by Zitelli[12] for this case were used to find the F factor due to the nonconstant reflector field alone, of the order of 1.2 to 1.3.

A correction has also been applied to account for the field-free bunching which takes place between the resonator grids. The experimental F factor in the reflector region alone F_r may be computed from the expression

$$F_r = (F\theta + \phi)/(\theta - \phi)$$

where F is the measured F factor, θ is the total transit angle, and ϕ is the transit angle between resonator grids. With the inclusion of such corrections, the measured values in Fig. 5 express the effects of space charge alone in the reflector region.

It is likely that the difference between the experimental and theoretical values of the F factor is not experimental error but inherent in the approximation of the method of field perturbation itself. Notwithstanding, the first-order agreement between the measurements and theory is token in itself of the usefulness of the method of field perturbation.

The plotted points in Fig. 5 have required an estimate of the beam current density $\underline{J}$. This was done by assuming a uniform current density across the entire grid aperture. Depending on the efficiency of focussing, however, the effective current density could be somewhat greater if the beam diameter were less than that of the aperture. This effect, if present, would reduce the difference between the experimental points and the theoretical curves.

Certain general observations can be made concerning the F factor, as seen from Figs. 2, 3, and 5. First, for constant $\underline{J}$, the F factor increases monotonically with N. Second, for cases of general interest, the F factor in a given mode (for constant N) increases with $\underline{J}$, the current density, or with the beam perveance for a constant reflector distance d. Thus, space charge in reflex tubes results in an F factor greater than unity, and the effect becomes increasingly more important as the current density is increased and as the reflector voltage is decreased.

The effects of this F factor on general reflex-tube behavior are the following: Essentially, the effective value of the bunching angle in the reflector region is no longer N, but FN, and in considering all the properties of a reflex tube depending on the bunching angle this distinction between the two must be considered. The only place in the reflex-tube theory where the true transit angle is relevant is in determining the phase of the bunched current. In all properties relating to amplitude, efficiency, power output, and so on, FN is the relevant parameter and must be used in calculating these quantities.

An important consequence of the space charge causing an F factor greater than unit relates to the design of low-frequency reflex tubes. From what has been said previously, it is apparent that, for a constant current density, as the transit time increases the F factor also increases. Quantitatively, if N is kept constant and the frequency varied, the parameter $\underline{J}T^2$ varies as J/f^2, and F varies somewhat more rapidly than $\underline{J}T^2$. Therefore, FN increases and the conversion efficiency and the power output eventually decrease to a negligible value as the frequency is lowered. This can be somewhat circumvented by going to smaller N; but there is a smallest value of N permitted, for phase reasons, i.e., 3/4 cycle, and beyond this no decrease is possible by simple means.

An interesting consequence concerning the maximum conversion efficiency in a given mode, i.e., the power delivered to the cavity plus load divided by the dc input power, is indicated by the calculation of F. Since F is greater than unity, FN is greater than N, and therefore the maximum conversion efficiency, being equal to $0.398/FN$, is always reduced by space charge. The maximum power in a given mode, however, is equal to the product of the power input and the maximum conversion efficiency for the mode

$$P = 0.398 V_0 I_0/FN. \tag{30}$$

By referring to Fig. 5 it is seen that the F factor is a function of the beam current I_0 in such a way that the ratio I_0/F first increases, reaches a maximum, and then decreases as the current density is increased. This means that, because of the effects of space charge, there is a certain optimum value of beam current (with given values of V_0 and N) for which maximum power can be obtained.

While the maximum efficiency is decreased by the presence of space charge, the zero-signal admittance is increased in direct proportion to the F factor, with the result that signal power can be obtained at a proportionately lower impedance level.

Acknowledgment

The authors are grateful to Karl Spangenberg for his helpful criticisms and comments on this work. They also wish to thank J. Sadler of Sperry Gyroscope Company for making the data on SRC-2 available to them.

[12] L. T. Zitelli, "Repeller Field Shape as a Factor in the Design of Broadband Reflex Klystron Oscillators," Contract N6-onr-251, Consolidated Task 7, Technical Report No. 3; May 25, 1948.

A Circuit for Generating Polynomials and Finding Their Zeros*

F. W. BUBB, JR.†, ASSOCIATE, IRE

***Summary*—This paper describes an instrument for generating any polynomial and for finding the roots of any polynomial equation to any desired degree of accuracy.**

I. Introduction

THE PURPOSE of this paper is to describe a device for generating the polynomial

$$\sum_{m=0}^{n} a_m z^m = a_0 + a_1 z + a_2 z^2 + \cdots + a_n z^n = f(z) \quad (1)$$

and for solving the equation

$$\sum_{m=0}^{n} a_m z^m = 0, \quad (2)$$

where $z = x + jy$ and $a_m = b_m + jc_m$.

The basic problems in the synthesis of an instrument for generating polynomials are the analogical representation of its individual terms and the addition of these terms. Since z is complex in general, a representation of a complex number is necessary, while to obtain z^2, a representation of the product of two complex numbers must be found. Presuming this can be done, z^3 may be found as the product of z^2 and z, z^4 as z^3 multiplied by z, and so on for all the powers of z, while the multiplication of the various powers of z by their coefficients may be similarly represented. Finally, the terms of (1) must be added to obtain $f(z)$.

The usual shorthand notation used by the electrical engineer to represent the steady-state solutions of the integrodifferential equations describing an electrical network suggests that an electrical circuit might be the basis for representing the operations with complex numbers if sinusoidal steady-state currents and voltages exist in the network. Circuits for multiplying and for adding complex numbers are explained in section II.

It is pointed out here that these circuits are capable of analogically performing the required arithmetic operations only for $|x|$, $|y|$, $|b_m|$, and $|c_m| \leqq 1$; although these restrictions in no way cripple the usefulness of the instrument, for in solving (2) $|b_m|$ and $|c_m|$ may be made $\leqq 1$ by dividing (2) by the largest real or imaginary coefficient part. Furthermore, to find roots of (2) with real and/or imaginary part >1, z may be replaced by $1/w$ to get

$$a_0 w^n + a_1 w^{n-1} + \cdots + a_n = 0, \quad (3)$$

whose roots have parts $\leqq 1$ corresponding to those roots of (2) which have parts $\geqq 1$. Thus (3) satisfies the requirements of the solver and enables the solution of (2).

Presuming that the coefficients of (1) have parts $\leqq 1$, $f(z)$ is suited to the circuits later described for $|x|$ and $|y| \leqq 1$, while to find $f(z)$ for $|x|$ and/or $|y| > 1$, (1) may be divided by $|z|^n$, where $|z|^n > 1$, to get

$$\phi(z) = \frac{f(z)}{|z|^n} = \frac{a_0}{|z|^n} + \frac{a_1}{|z|^{n-1}}(z/|z|) + \frac{a_2}{|z|^{n-2}}(z/|z|)^2 + \cdots + a_n(z/|z|)^n, \quad (4)$$

or letting $z' = z/|z|$, $a_{m'} = a_m/|z|^{n-m}$,

$$\phi(z) = \sum_{m=0}^{n} a_{m'}(z')^m, \quad (5)$$

where the parts of z' have moduli $\leqq 1$. Thus $\phi(z)$ satisfies the conditions required by the instrument and can be generated by it, while $f(z)$ is found from $\phi(z)$ by simple multiplication. If the largest coefficient part is >1, $f(z)$ or $\phi(z)$ divided by this largest coefficient part can be generated by the instrument, and again $f(z)$ is found by simple multiplication.

It is also possible to "contract" the variable by letting $z = kz''$, $k > 1$, so that

$$f(kz'') = a_0 + (a_1 k)z'' + (a_2 k^2)z''^2 + \cdots + (a_n k^n)z''^n, \quad (6)$$

which satisfies the condition $|x''| \leqq 1$, $|y''| \leqq 1$, while division by the largest coefficient part, if it is greater than unity, enables the generation of $f(z)$.

The origin of the variable z may also be shifted to obtain various results; this is covered in section IV C.

II. Circuits for Performing the Basic Operations

It has been shown[1] that the input and output voltages of the feedback amplifier shown in block form in Fig. 1

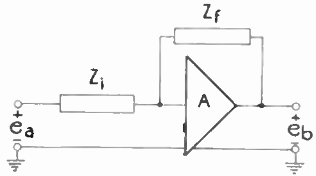


Fig. 1—A multiplying device.

* Decimal classification: 621.375.2. Original manuscript received by the Institute, August 22, 1950; revised manuscript received, March 5, 1951.

† St. Louis University, St. Louis, Mo.

[1] J. R. Ragazzini, R. H. Randall, and F. A. Russell, "Analysis of problems in dynamics by electronic circuits," Proc. I.R.E., vol. 35, pp. 444–452; May, 1947.

are related by

$$e_b = -(Z_f/Z_i)e_a, \tag{7}$$

provided the open-loop gain of the amplifier A is very large and the phase shift from the input to the output is approximately 180 degrees with the feedback loop open.

It has been shown also that the output voltage of Fig. 2 is related to the input voltages by

$$-e_t = (R_f/R_0)e_0 + (R_f/R_1)e_1 + \cdots + (R_f/R_n)e_n, \tag{8}$$

where the amplifier must meet the conditions above.

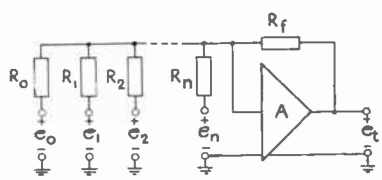


Fig. 2—Adding circuit.

It will now be shown that the input and output voltages of the circuit of Fig. 3 are related by

$$e_b = ze_a = (\pm|x| \pm j|y|)e_a, \tag{9}$$

where $|x|$ $(|y|)$ is the fraction of the resistance of the potentiometer P_x (P_y) between brush and ground. It is assumed that the resistances R_3 are large enough so that

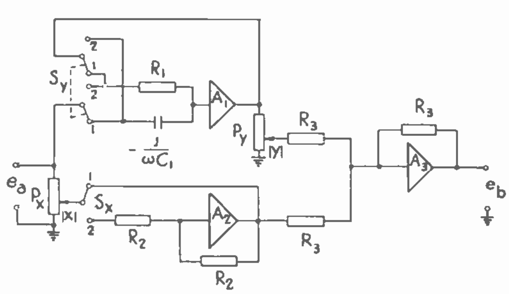


Fig. 3—A circuit for multiplying by any complex number.

they do not load the potentiometers P_x and P_y, and that the amplifiers A_1, A_2, and A_3 have large gain and approximately 180-degree phase shift with their feedback loops open. The components R_1 and C_1 are such that $R_1\omega C_1 = 1$, where ω is the angular frequency of e_a. The switch S_x is a single-pole double-throw type, while S_y is a double-pole double-throw switch.

Suppose S_x and S_y are both in the positions marked "1." By (7) the output voltage of A_1 is $e_1 = -jR_1\omega C_1 e_a = -je_a$, in view of the restrictions on the components. A portion of e_1, $|y|e_1$ is applied to the input of A_3 through a resistance R_3, while the voltage $|x|e_a$ at the brush of P_x is applied to the input of A_3 through another resistance R_3. By (8), the output voltage of A_3 is then

$$e_b = (-|x| + j|y|)e_a. \tag{10}$$

If S_x is in position 1, while S_y is in position 2, the feedback and input impedances of A_1 are interchanged. The output of A_1 is then $j|y|e_a/R_1\omega C_1 = j|y|e_a$, so that the output voltage of A_3 is

$$e_b = (-|x| - j|y|)e_a. \tag{11}$$

As yet, no mention has been made of the purpose of switch S_x, which enables either the application of $|x|e_a$ directly to the input of A_3 through R_3, or the application of $|x|e_a$ to the input of A_2, and thence to R_3. The amplifier A_2 has equal input and feedback resistances (R_2): Its function is to change the sign of the voltage applied to its input. Thus, if S_x is in position 2, the output of A_3 is either

$$e_b = (|x| + j|y|)e_a \tag{12}$$

or

$$e_b = (|x| - j|y|)e_a, \tag{13}$$

depending upon the position of S_y.

Each of equations (10)–(13) is a special case of (9), hence the circuit of Fig. 3 will multiply its input voltage by any type of complex number whose parts have a magnitude of unity or less. This circuit is used to raise z to its various powers.

As a device for multiplying the various powers of z by their coefficients, the circuit of Fig. 4 is used. This

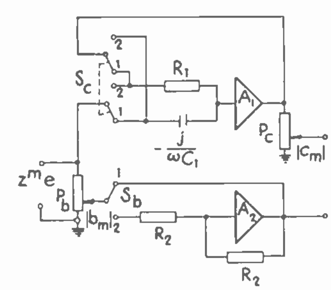


Fig. 4—A coefficient unit.

circuit is identical to part of Fig. 3; the amplifier A_3 and the resistances associated with it are omitted. Here $|b_m|$ is the fraction of the resistance of P_b between its brush and ground, while $|c_m|$ is the fraction of P_c between brush and ground. The input voltage to this unit is $z^m e$, so that the voltage at the output of A_2, or at the brush of P_b depending upon the position of S_b, is $\pm|b_m|z^m e$. The voltage at the brush of P_c is $\pm|c_m|z^m e$, where, as in the unit of Fig. 3, $R_1\omega C_1 = 1$ and the signs are accounted for by the position of S_c.

III. A Circuit for Solving Polynomials with Complex Coefficients

The new circuit for solving polynomials is shown in block form in Fig. 5. Each of the cascaded blocks across the top of the diagram is meant to contain the circuit of Fig. 3; they are called "involution units" because they perform this operation upon the variable z. All of the P_x shafts are mechanically ganged so that the $|x|$'s in all units are the same. The P_y shafts are also ganged. To insure the same signs for the x's and y's in all units, the switches S_x are mechanically connected, as are the S_y's.

An alternating voltage e, of angular frequency ω, is applied to the first involution unit. By (9), the output voltage of this first unit is ze, and this voltage is applied to the input of the second involution unit so that its output voltage is z^2e, and so on. The outputs of the cascaded involution units are then successively ze, z^2e, z^3e, $\cdots$, z^ne.

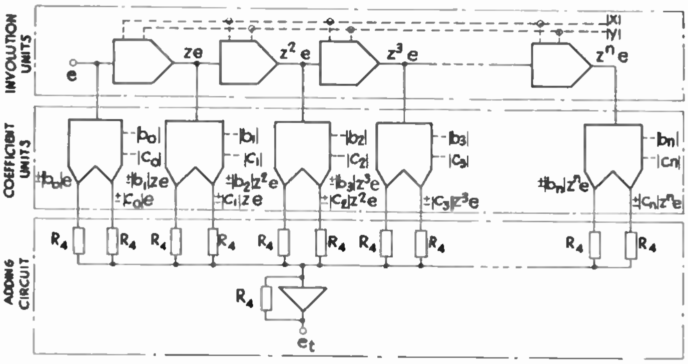


Fig. 5—Block diagram of an instrument for solving polynomials with complex coefficients.

Each of the coefficient units (the blocks connected to the output of each involution unit and going to the adding circuit) is the circuit of Fig. 4. The outputs of these latter blocks are evident from the voltages marked on the circuit. The adding circuit is the same as Fig. 2; the components marked R_4 are equal resistances.

By (8), the output voltage of the adding amplifier is

$$\begin{aligned} -e_t &= (\pm|b_0| \pm j|c_0|)e + (\pm|b_1| \pm j|c_1|)ze + \cdots \\ &\quad + (\pm|b_n| \pm j|c_n|)z^ne \\ &= \sum_{m=0}^{n} (\pm|b_m| \pm j|c_m|)z^me \end{aligned} \tag{14}$$

or

$$\begin{aligned} -\frac{e^t}{e} &= \sum_{m=0}^{n} (\pm|b_m| \pm j|c_m|)z^m \\ &= \sum_{m=0}^{n} (b_m + jc_m)z^m = \sum_{m=0}^{n} a_mz^m, \end{aligned} \tag{15}$$

which is the polynomial (1), provided each $|b_m|$ and $|c_m|$ is properly adjusted (by adjusting the potentiometers P_b and P_c) and the switches S_b and S_c are appropriately set for each a_m.

Assuming the coefficients are properly set into the solver, $f(z)$ is found by comparing e_t to e for magnitude and phase. To change the variable, $|x|$ and $|y|$ are changed by adjusting the P_x potentiometers and the P_y potentiometers, while the positions of all of the S_x's and all of the S_y's determine the form of z. In order to generate $f(z)$ for $|x|$ and/or $|y|>1$, $\phi(z)$ or $f(kz'')$ mentioned in section I may be used. To generate $f(z)$ in the vicinity of a given z, the origin may be shifted to this value of z as shown in section IV C; this enables the accurate determination of $f(z)$ in a given vicinity.

To find a root of the polynomial equation, $|x|$ and $|y|$ are independently varied[2] until $e_t=0$; the values of $|x|$ and $|y|$ resulting in $e_t=0$ and the positions of S_x and S_y determine the root. This process is repeated until all roots whose parts are less than unity are found, when (3) is set into the instrument enabling the location of all roots with $|x|$ and/or $|y|>1$. By the successive approximation process shown in section IV C, a root may be found to any desired degree of accuracy.

IV. Operating Notes

A. An Example of the Preparation of a Polynomial for the Instrument

First, consider the solution of polynomial equations, and as an example, use the equation

$$400{,}000 + 24{,}000z + 220z^2 + z^3 = 0. \tag{16}$$

Although a cubic is used as the example, it should be remembered that the instrument is for solving polynomials that do not have literal solutions. It is also pointed out here that most of the calculations are of slide-rule accuracy only.

To set (16) into the instrument, first divide the equation by 400,000, the largest coefficient. This gives

$$1 + 0.06z + 0.00055z^2 + 0.0000025z^3 = 0. \tag{17}$$

Now it would be difficult to set the number 0.00055 into the solver with any degree of accuracy, even using a vernier attached to the P_b potentiometer, and it would be virtually impossible to set in the number 0.0000025. It appears that this polynomial must be "prepared" for the instrument.

Let $z=20z'$. Inserting this in (17), gives

$$1 + 1.2z' + 0.22z'^2 + 0.02z'^3 = 0 \tag{18}$$

or

$$0.8333 + z' + 0.1833z'^2 + 0.0167z'^3 = 0. \tag{19}$$

The coefficients of (19) could be set into the solver with much more accuracy than those of (17). Thus, by "contracting" the variable so that the new variable is $z'=z/20$, the "contracted" polynomial may be set into the instrument with a higher degree of accuracy.

It is also possible to shift the origin of z so that the coefficients are more nearly equal.

B. The Root-Finding Process

To show that a change in $|x|$ or $|y|$ resulting in a decrease in e_t is a step toward finding a root, write the polynomial as

$$f(z) = a_n(z-z_1)(z-z_2)(z-z_3)\cdots(z-z_n) = -e_t/e, \tag{20}$$

where $z_m(m=1, 2, 3, \cdots, n)$ are the zeros of (20).

[2] It is shown in section IV B that an adjustment of either $|x|$ or $|y|$, resulting in a decrease of e_t is a step toward finding a root.

Suppose that the polynomial has a zero $z_1=x_1+jy_1$ as shown in Fig. 6.

If the solver is originally set to $z_a=x_a+jy_a$, then $f(z_a)$ will have some value not zero. Suppose x is varied, while $y=y_a$ is not changed. As x is increased, $|f(z)|$ will tend to decrease because $|z-z_1|$ decreases. However, the minimum value of $|f(z)|$ may not occur at $z_c=x_1+jy_a$, although $|z_c-z_1|=y_1-y_a$ is certainly the smallest absolute value of the factor $z-z_1$, if only x is varied. It is probable that some value of $x=x_b$, other than x_1, will result in a minimum $|f(z)|$ because the product of the factors $|z_c-z_m|$ is enough larger than the product of the factors $|z_b-z_m|$, $m=2, 3, 4, \cdots, n$, so that $|f(z_b)|<|f(z_c)|$. Thus, if only x is varied until the magnitude of e_t reaches a minimum, then the value of $x=x_b$ will be near x_1, the real part of the root z_1.

Now suppose that only y is varied, while $x=x_b$ is not changed. When $|f(z)|$ reaches its minimum value, the value of z may be some value such as $z_d=x_b+jy_d$, where $y_d\neq y_1$. At this time x only may be varied again, then y, and so on. It is evident that each variation of x or y which reduces $|f(z)|$ (or e_t) will result in values of x and y closer to x_1 and y_1, respectively. When $|f(z)|=0$, then $z=z_1=x_1+jy_1$ and a root has been found.

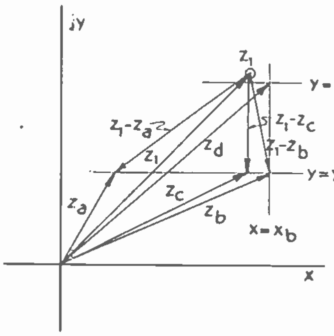


Fig. 6—The root-finding process.

C. Shifting the Origin to Obtain Higher Accuracy

Consider, as an example, the cubic equation with complex coefficients

$$f(z)=(b_0+jc_0)+(b_1+jc_1)z+(b_2+jc_2)z^2+(b_3+jc_3)z^3 \quad (21)$$

and replace the variable z by $z'+r+js$. This in effect shifts the origin by an amount $r+js$ as shown in Fig. 7. The new equation in z' is

$$\begin{aligned} f(z'+r+js)=\ &[(b_0+b_1r-c_1s+b_2r^2-b_2s^2-2c_2rs+b_3r^3 \\ &-3b_3rs^2-3c_3r^2s+c_3s^3)+j(c_0+b_1s+c_1r \\ &+2b_2rs+c_2r^2-c_2s^2+3b_3r^2s-b_3s^3)] \\ &+[(b_1+2b_2r-2c_2s+3b_3r^2-3b_3s^2-6c_3rs) \\ &+j(c_1+2b_2s+2c_2r+6b_3rs+3c_3r^2-3c_3s^2)]z' \\ &+[(b_2+3b_3r-3c_3s)+j(c_2+3b_3s+3c_3r)]z'^2 \\ &+[b_3+jc_3]z'^3=\psi(z'). \end{aligned} \quad (22)$$

Now suppose that a zero of a polynomial has been found by means of the instrument of Fig. 5, but upon inserting the zero into the polynomial, it does not reduce to zero exactly, indicating that the zero is slightly in error. If the zero which the instrument has found is $r+js$, while the actual zero is $z_n=x_n+jy_n$, different from $r+js$ by $\delta+j\epsilon$, then one of the zeros of (22) is $\delta+j\epsilon$. This is evident from Fig. 7, where $z_n=(r+\delta)+j(s+\epsilon)$, and the fact that $f(z_n)=0$. Thus, $f(z_n)=\psi(\delta+j\epsilon)=0$.

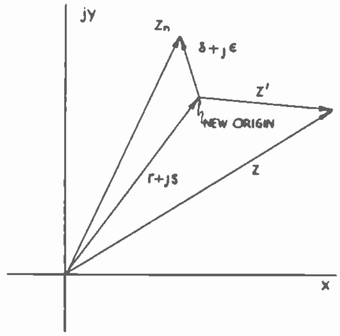


Fig. 7—Shifting the origin.

By means of (22), z_n may be found more exactly than $r+js$ by inserting r and s into (22), setting (22) into the solver, and finding $\delta+j\epsilon$. Then z_n (to a better degree of accuracy) $=(r+\delta)+j(s+\epsilon)$. Of course, δ and ϵ will be subject to errors, but the "shifting" process as well as the "stretching" process may be repeated just as many times as are necessary to get any desired degree of accuracy for the zeros.

It should be understood that the specific case of a cubic has been discussed here, but the same shifting process to obtain accurate roots can be applied to any degree polynomial. Suppose one had twelfth-degree polynomials to solve as a daily process. One would set up an equation similar to (22) for the twelfth-degree polynomial; and while one engaged in finding roots within the accuracy of the instrument, the usual type of arithmetic calculators could be used by someone else to set up the next equation to be solved for the error. Thus, this instrument could be used to a great time advantage over the usual numerical calculations.

It should also be mentioned that the use of (22), or of its higher degree counterpart, will enable the accurate generation of $f(z)$ in the vicinity of the point $r+js$. It is presumed that a large number of values in the vicinity are desired, warranting the arithmetic involved.

D. Discussion of Distortion

In order that any of the equations derived with regard to the complex multiplying devices be valid, no distortion can occur anywhere in the circuit with the exception of the adding amplifier when the instrument is being used to find roots. For finding roots it is recommended than an oscilloscope be used to measure e_t, and switches be arranged so that the oscilloscope may be connected to either the output of A_1 in the first involution unit, or the output of the last involution unit in the instrument to check for distortion. In cases where

$|z|<1$, the largest voltage in the involution circuit is usually at the output of A_1 in the first involution unit. In cases where $|z|>1$, the largest voltage is at the output of the last involution unit. Of course, if any coefficient a_m is such that $|a_m|>1$, then the output of the corresponding coefficient unit may be larger than either of the voltages mentioned above. As an added precaution, the applied voltage e, should be kept small, at least until a null ($e_t=0$) is approached. Then e may be increased so that e_t increases and is easier to measure. A check should be made for distortion when e is increased; but as the settings of x and y will be close to the root, the operator will now know at which point the largest voltage will appear and need check only that one point.

In generating a polynomial, distortion is not permitted at all; and since the largest voltage will usually be e_t, it should be monitored for distortion by a reliable means.

V. The Values of the Circuit Constants

This discussion presupposes an amplifier which meets the requirements mentioned in section II;[3] the amplifier must be capable of delivering 2.5 watts at about 50 volts rms.

It has already been stated that $R_1\omega C_1=1$ (see Figs. 3 and 4). Letting the operating frequency be 60 cps and C_1 be 0.003 microfarads, then $R_1=883{,}000$ ohms (slide-rule accuracy). These are merely nominal values for C_1 and R_1; the necessary relation is $R_1\omega C_1=1$. Thus, if C_1 is not exactly 0.003 microfarads, R_1 may be adjusted to make the gain of A_1 exactly $\pm j$ with the feedback loop closed. It should be noted that the magnitude of the impedance of C_1 or of R_1 is 883,000 ohms; this high an impedance will have practically no loading effect upon the amplifier preceding it.

Now consider the potentiometers P_x, P_y, P_b, and P_c. Each of these is at the output of an amplifier, except those at the input to the solver where a voltage e exists whether or not this is the applied emf. The resistance of these potentiometers cannot be so small that they will overload the amplifiers. It seems reasonable now to assume that a value of 1,000 ohms will be satisfactory in this respect. Each of these potentiometers should be helically wound with, say, 10 turns, and the shaft of each potentiometer should be equipped with a dial so that 0.01 of one turn may be read directly from the dial. Of course, all of the P_x's must be mechanically connected together as indicated by the dashed lines in Fig. 5. The P_y's are also mechanically connected so that all of them may be varied simultaneously. For the coefficients a_m to be accurately set into the solver, it is only necessary that the potentiometers P_b and P_c be linear, while for the value of z in each unit to be identical with the others, it is necessary that the potentiometers P_x and P_y be linear and that the mechanical connections be accurate—it is not necessary that the total resistance of each potentiometer be exactly that of the others. Fortunately, helical potentiometers with good linearity (0.05 per cent) are commercially available.

Having chosen the potentiometer resistance equal approximately to 1,000 ohms, R_2, R_3, and R_4 must be chosen large enough so that the voltages at the brushes are linear with brush position within reasonable accuracy. If a particular resistance is 100 times as large as the resistance of the potentiometer to which it is connected, the departure from linearity of the brush to ground voltage will not exceed 0.25 per cent. If a resistance 1,000 times as large as the potentiometer's is used, the largest departure from linearity is about 0.025 per cent. It appears that a resistance from 100,000 ohms to 1 megohm would be suitable for R_2, R_3, or R_4, depending upon the accuracy requirements; one megohm, resulting in negligible error, is recommended.

It is not necessary that the resistances R_2 in the sign changers A_2 be exactly 1 megohm, but it is essential that for a given sign changer the two R_2's be equal to each other within, say, 0.05 per cent. Thus, for each sign changer the resistances designated R_2 must be trimmed to the same value close to 1 megohm. Similarly, the three resistances R_3 in a given involution unit must be trimmed, while the resistances R_4 in the adding circuit all must be trimmed to the same value close to 1 megohm.

VI. Simplifications in the Instrument Resulting in a Circuit for Solving Polynomials with Real Coefficients Only

It will be noted that the polynomials which arise in engineering problems usually have real coefficients; it is an origin shift to a complex z which gives rise to complex coefficients. Thus, in the event that extreme accuracy is not required of the polynomial solver, the use of the origin shift may be eliminated and the instrument much simplified accordingly. It is at once apparent that the A_1 amplifiers and associated components in the coefficient units (Fig. 4) may be discarded. Furthermore, the A_2 amplifiers in the involution units (Fig. 3) and the S_y switches may be eliminated. This is not so apparent, and a few explanatory remarks are necessary.

By reducing the involution units as mentioned above, the real part of z is always negative, and the imaginary part may be either plus or minus. Choose the connections to the A_1 amplifiers of the involution units so that $z=-|x|+j|y|$. Now, since a polynomial with real coefficients has zeros in conjugate complex pairs, while $f(\bar{z})=\overline{f(z)}$, where $\bar{z}$ is the conjugate of z and $\overline{f(z)}$ is the conjugate of $f(z)$, it is apparent that it is not necessary to change the sign of the imaginary part of z either to find the zeros of the polynomial or to generate it, pro-

[3] A high-gain alternating-current amplifier with approximately 180-degree phase shift from the input to the output at the operating frequency is required. To discourage any tendency to oscillate at a frequency other than the operating frequency, a band-pass filter at the input, or output, or both, should be used. Furthermore, to eliminate "hum" (since the instrument will operate at 60 cps) the filaments of all tubes should be supplied by a direct-current source.

vided one accounts for the sign of the imaginary part of either the zero or the polynomial appropriately.

In order to change the sign of the real part of z, it is noted that if z were $|x| - j|y|$ (instead of $-|x| + j|y|$), the signs of the odd powers of z would be changed, but those of the even powers would remain the same, By changing the sign of every odd power of z by shifting the burden of the sign to the coefficient, the output of the adding circuit is as if z were $|x| - j|y|$, which enables one to deal with plus values of x. The fact that the sign of the imaginary part is minus is unimportant, but of course the remarks above on conjugates must be regarded.

VII. Conclusion

This paper describes an instrument for generating polynomials and for finding the roots, real or complex, of any polynomial, of any degree, with either real or complex coefficients, and to any degree of accuracy. It is also shown that this instrument may be built with commercially available parts.

The Measurement of Current Distributions along Coupled Antennas and Folded Dipoles*

T. MORITA†, ASSOCIATE, IRE, AND C. E. FAFLICK†, STUDENT, IRE

Summary—**The theoretical analysis of coupled antennas is facilitated by the use of symmetrical and antisymmetrical currents. This paper presents experimental results verifying this physical concept of current on closely coupled antennas and folded dipoles.**

I. Introduction

The theoretical analysis of coupled antennas is carried out advantageously by representing the current in each element as a superposition of symmetrical (codirectional) and antisymmetrical (oppositely directed) components of current.[1] If the spacing between the elements satisfies the near-zone conditions $\beta_0 b = 2\pi 1\lambda \ll 1$, the symmetrical components maintain the far-zone field and may therefore be called the radiating or antenna currents; while the antisymmetrical components which contribute insignificantly to the far-zone field are essentially nonradiating or transmission-line currents. It is the purpose of this paper to verify experimentally this physical concept of currents in coupled antennas and folded dipoles.

II. Measuring Apparatus and Techniques

Fig. 1 is a block diagram of the measuring setup. The basic structure on which all current-distribution measurements were made consists of a coaxial line, the shield of which ends at a conducting ground plane; the extension of the inner conductor over the ground plane forms the antenna.[2] Protruding from a slot in the center conductor is a small shielded loop which can be moved along the entire length of the antenna and into the coaxial line to measure both the current distribution on the antenna and line and the standing-wave ratio on the line. In this way the impedance of the antenna is obtained. The length of the antenna may be increased by extending the inner conductor a greater distance over the plane by a rack-and-pinion movement. The second element of the coupled antenna is formed by a quarter-inch rod which is parallel to the slotted antenna and composed of threaded sections that may be adjusted in length to any integral number of centimeters. This element may be connected to or insulated from the ground plane; in the latter case the antenna is connected to a type-N connector.

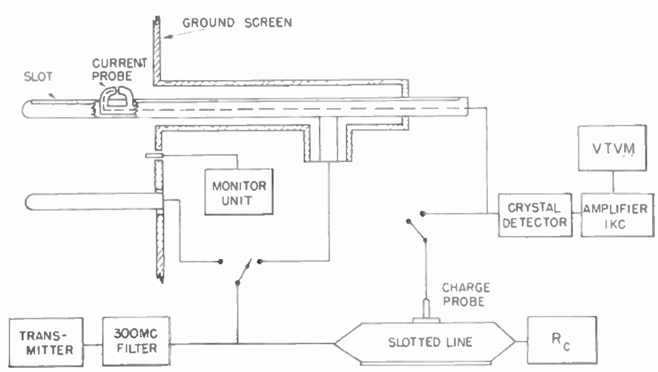


Fig. 1—Block diagram for measuring setup.

The problem is to use the same slotted antenna to measure the current distribution on both the driven and parasitic elements. By symmetry, the feed points may be interchanged so that the slotted antenna can be used as the parasitic antenna. A small charge probe, located symmetrically between the two elements, is used both as an amplitude and a phase-reference monitor so that when the two elements are interchanged,

* Decimal classification: R129×R142. Original manuscript received by the Institute, September 12, 1950; revised manuscript received, March 13, 1951.

The research reported in this paper was made possible through support extended Cruft Laboratory, Harvard University, by the Naval Department (ONR), the Signal Corps, the United States Army, and the United States Air Force, under ONR Contract N5ori-76, T. O. 1.

† Cruft Laboratory, Harvard University, Cambridge, Mass.

[1] King, Mimno, and Wing, "Transmission Lines, Antennas, and Waveguides," McGraw-Hill Book Co., New York, N. Y.; pp. 224–226; 1945.

[2] T. Morita, "Current distributions on transmitting and receiving antennas," Proc. I.R.E., vol. 38, pp. 898–904; August, 1950.

the relative amplitudes and phases of the currents on each element may be determined.

Relative phase is measured by adding the signal from the test probe to a signal obtained from a traveling probe on a slotted line terminated in its characteristic impedance. All measurements were made at 300 mc, modulated by a 1-kc signal. A 1N21B crystal and a tuned amplifier were used in the detector circuit and the voltage was read on a Ballantine vacuum-tube voltmeter.

To measure the current distribution on a folded dipole, a short-circuiting bar was placed at the ends of the coupled antennas described above. Measurements were made for coupled antennas and folded dipoles of lengths $h=\lambda/4$ and $\lambda/2$ with spacing between elements of $b=0.04\lambda$.

III. Resolution of Measured Distribution into Symmetrical and Antisymmetrical Currents

By application of the superposition theorem it is always possible to solve the problem of two coupled

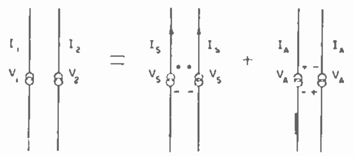


Fig. 2—Resolution of measured distribution into symmetrical and antisymmetrical currents.

antennas, center-driven by voltages V_1 and V_2, by obtaining the solutions for symmetrically and antisymmetrically driven pairs. From Fig. 2 it follows that

$$V_1 = \tfrac{1}{2}(V_S + V_A)$$
$$V_2 = \tfrac{1}{2}(V_S - V_A),$$

or

$$V_S = V_1 + V_2$$
$$V_A = V_1 - V_2.$$

Similarly,

$$I_1 = \tfrac{1}{2}(I_S + I_A)$$
$$I_2 = \tfrac{1}{2}(I_S - I_A),$$

or

$$I_S = I_1 + I_2 \tag{1}$$
$$I_A = I_1 - I_2,$$

and

$$Z_1 = \frac{V_1}{I_1}. \tag{2}$$

For the particular case under consideration, the second element is parasitic. Thus,

$$V_2 = 0,$$

or

$$V_S = V_A,$$

and

$$V_1 = \tfrac{1}{2}V_S,$$

so that

$$Z_1 = \frac{2V_S}{I_S + I_A} = \frac{2Z_A Z_S}{Z_A + Z_S},$$

with

$$Z_S = \frac{V_S}{I_S} \tag{3}$$

and

$$Z_A = \frac{V_A}{I_A}.$$

In measuring, the impedance of the antenna with its coupled element, as well as the relative amplitudes and relative phases of the currents in the elements, has been obtained. It is desirable to secure the absolute amplitudes and absolute phases of the measured distributions. If V_1 is arbitrarily chosen to be 1 volt, with a phase angle of 0 degrees, it follows from (2) that

$$|I| \underline{|-\theta^0} = \frac{1 \underline{|0^\circ}}{|Z_1| \underline{|\theta^0}}.$$

Therefore, the reciprocal of Z_1 gives the amplitude of current in amperes per volt, while the negative of the phase angle of the impedance gives the phase angle of the current at the gap. This establishes the absolute value of amplitude in amps per volt, and the phase of the current at any point.

With such a procedure it may be assumed that it is possible to define a unique voltage at the gap. From the point of view of the wave picture, the electromagnetic field at the gap can be represented in terms of the dominant TEM mode in the coaxial line, together with nonpropagating higher-order modes introduced at the gap, in order to satisfy the boundary condition at the discontinuity. The assumption of a voltage of 1 $\underline{|0^\circ}$ volt at the gap from measurements on the line is equivalent to neglecting these higher-order modes and representing the driving voltage only in terms of the dominant TEM mode. Theoretically, this means that the voltage assigned to the gap is actually the voltage defined back in the line, one wavelength from the gap, where only the dominant mode is present. By application of (1), the symmetrical and antisymmetrical components of current may be obtained. The symmetrical and antisymmetrical impedances may be derived from (3). Moreover, the mutual and self-impedances Z_{12} and Z_{11} are given by the following:

$$Z_{12} = Z_{21} = \tfrac{1}{2}(Z_S - Z_A)$$

and

$$Z_{11} = \tfrac{1}{2}(Z_S + Z_A).$$

IV. Results

A. Coupled Antennas

In Figs. 3 and 5 are shown the measured distributions of the magnitudes and phases of the currents on both the driven and parasitic elements of two coupled antennas, for which $\beta_0 h = \pi/2$ and π. By following the procedure developed in the foregoing section, these currents are decomposed into the symmetrical and antisymmetrical components of Figs. 4 and 6. The antisymmetrical currents may be identified with the equal and opposite currents of a two-wire transmission line. Since the symmetrical currents are equal and in phase in both elements, they may be identified with the radiating or antenna currents.

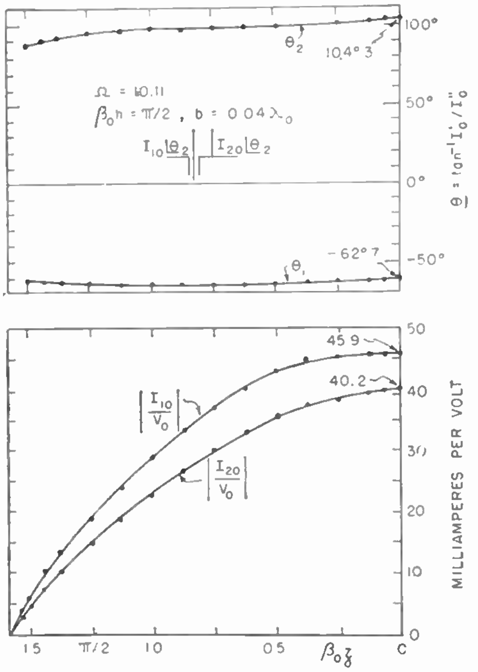


Fig. 3—Measured current distribution on coupled antenna, $h=\lambda/4$, $b=0.04\lambda$.

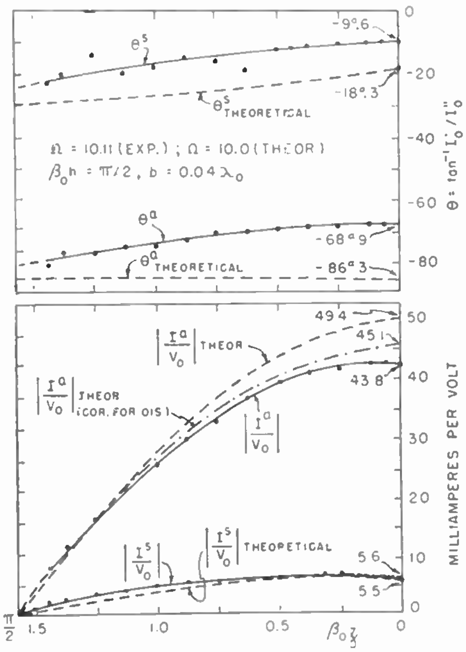


Fig. 4—Symmetrical and antisymmetrical components for coupled antenna, $h=\lambda/4$, $b=0.04\lambda$.

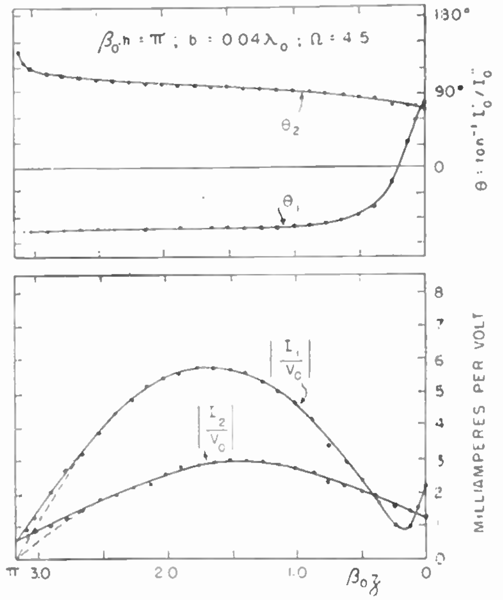


Fig. 5—Measured current on coupled antenna, $h=\lambda/2$, $b=0.04\lambda$.

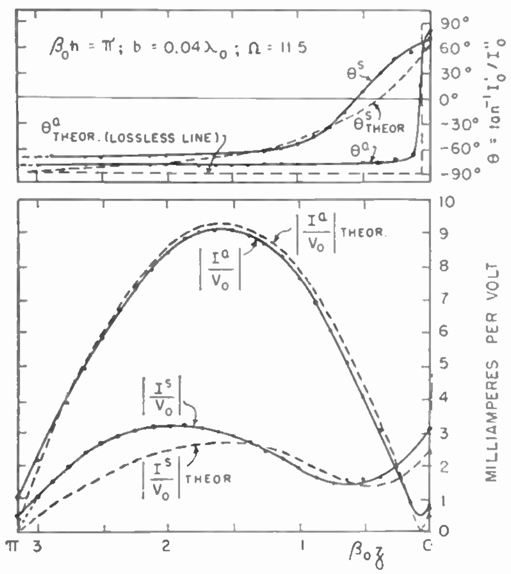


Fig. 6—Symmetrical and antisymmetrical currents for coupled antennas, $h=\lambda/2$, $b=0.04\lambda$.

It is interesting to compare the large amplitude of the antisymmetrical current with that of the symmetrical current. Since antisymmetrical components do not radiate, it is obvious that the antisymmetrical component has the effect of lowering the efficiency of the antenna if the transmission line has ohmic losses. Note that

TABLE I

$\beta_0 h=\frac{\pi}{2}$	*Experimental* $\Omega=10.11=2\log\frac{2h}{a}$	*Theoretical No Ohmic Dissipation* $\Omega=10$	*Theoretical with Added Series Resistance of 3.55Ω to Compensate for Ohmic Dissipation*
Z_1	$10.0+j19.2=21.8\ \angle 62.7^\circ$	$3.2+j19.0=19.3\ \angle 80.5^\circ$	$9.3+j17.9=20.2\ \angle 62.5^\circ$
$Z_{11}=Z_{22}$	$45.8+j12.8=47.6\ \angle 15.6^\circ$	$42+j19.3=46.3\ \angle 24.6^\circ$	$45.6+j19.3=49.5\ \angle 22.9^\circ$
Z_{12}	$41.6+j\ 1.9=41.7\ \angle 2.6^\circ$	$41.4+j\ 9.2=42.3\ \angle 12.4^\circ$	$41.4+j\ 9.15=42.3\ \angle 12.4^\circ$
Z_S	$87.4+j14.7=88.6\ \angle 9.4^\circ$	$83.4+j28.4=88.1\ \angle 18.8^\circ$	$86.95+j28.4=91.4\ \angle 18.1^\circ$
Z_A	$4.2+j10.9=11.7\ \angle 68.9^\circ$	$0.65+j10.1=15.5\ \angle 86.3^\circ$	$4.2+j10.1=11.0\ \angle 67.4^\circ$
$\beta_0 h=\pi$	$\Omega=11.3$	$\Omega=10$	
Z_1	$122-j454$	$170-j352$	
$Z_{11}=Z_{22}$	$127-j782$	$52.5-j987$	
Z_{12}	$-43+j494$	$52.5+j798$	
Z_S	$84.5-j289$	$105-j190$	
Z_A	$170-j1275$	$0-j1785$	

for $\beta_0 h=\pi$, the distances between minimum amplitudes are smaller for the symmetrical current distribution than for the antisymmetrical distribution. Theoretical curves[3,4] are shown for the symmetrical currents computed from the second-order solution to the integral equation; the antisymmetrical currents are compared with the sinusoidal distribution characteristic of a lossless transmission line.[3] In the theoretical formulation it is assumed that the antenna is driven by a difference of scalar potential applied by a ring generator, while the experimental model is driven by a coaxial line over an image plane. Since these methods of driving are equivalent only in the physically unavailable limit of a coaxial line with a dielectric of vanishingly small radius and a comparably small gap, only qualitative agreement is to be expected.

It is interesting to note that for $\beta h=\pi/2$ much better agreement with the experimental value is obtained if 3.55 Ω is added in series with the theoretical antisymmetrical resistance in Table I (above), to correct for the ohmic dissipation.

B. Folded Dipole

In Fig. 7 are shown the measured curves of resistance and reactance as a function of the length of the folded dipole over an image plane. The region of broad-band behavior for this structure is between points of antiresonance. For studying the behavior of the folded dipole in this range, a value of $\beta_0 h=\pi/2$ was selected. The measured distribution curve is shown in Fig. 8, and the symmetrical and antisymmetrical components in Fig. 9. The symmetrical components has a maximum at the gap while the antisymmetrical current has a maximum at the end of the antenna. The antisymmetrical current is the same as that of a short-circuited transmission line of length $l=h+b/2$, where b is the spacing between the two conductors.

A clearer picture of this concept of symmetrical and antisymmetrical currents may be obtained from a study of the phase diagram in Fig. 10. Here the voltage of the gap is taken as $1\ \angle 0^\circ$ volts. Since the antenna is slightly

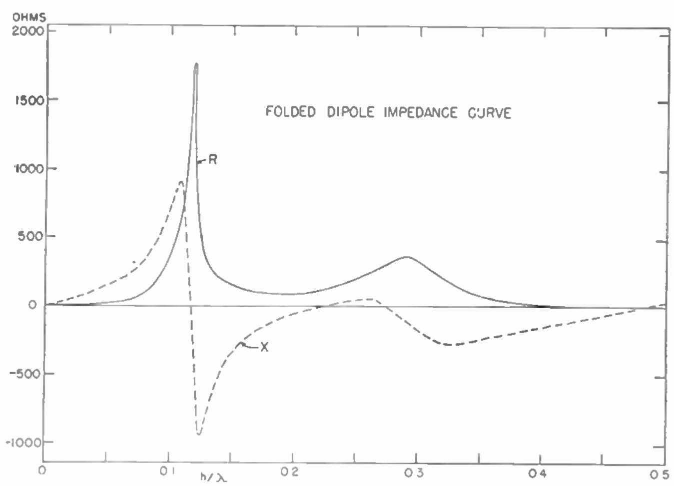


Fig. 7—Measured impedance for folded dipole, $b=0.04\lambda$.

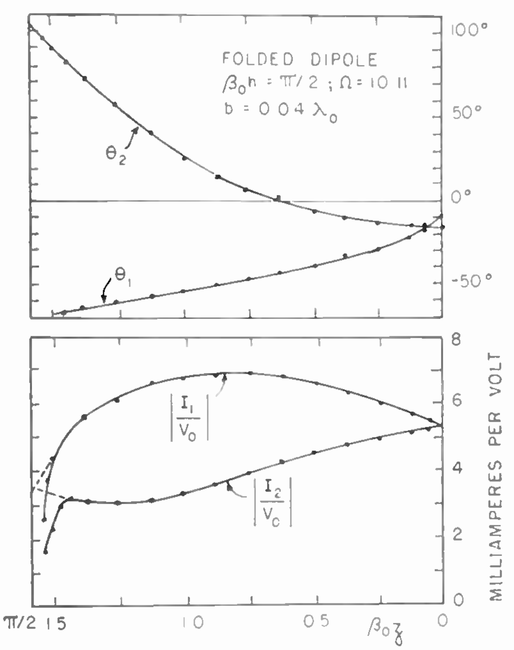


Fig. 8—Measured current on folded dipole, $h=\lambda/4$, $b=0.04\lambda$.

[3] C. T. Tai, "The theory of coupled antennas and its applications," PROC. I.R.E., vol. 36, pp. 487–497; April, 1948.
[4] R. W. P. King, "Self- and Mutual Impedances of Parallel Identical Antennas," Cruft Laboratory Technical Report No. 118, Harvard University, Cambridge, Mass.; November, 15, 1950.

longer than the resonant length of a dipole, the impedance for the symmetrical component is slightly inductive and small. Accordingly, the symmetrical current at the gap is large and lags behind the voltage by

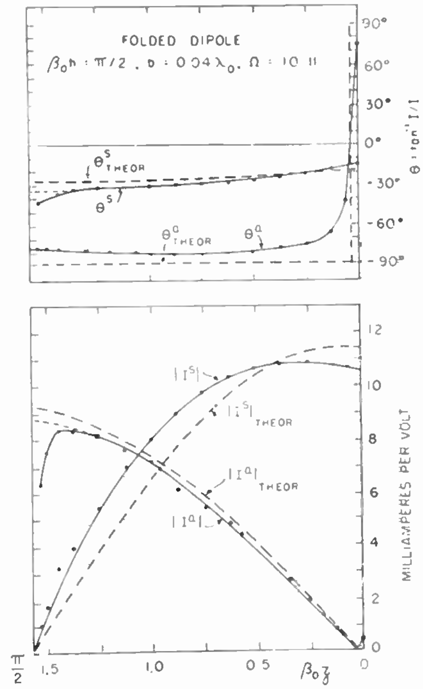


Fig. 9—Symmetrical and antisymmetrical components for folded dipole, $h=\lambda/4$.

a small angle. As we proceed along the antenna, the current amplitude decreases, falling to zero at the end, while the phase lags by a small amount as the end is approached, causing the phasor to rotate clockwise.

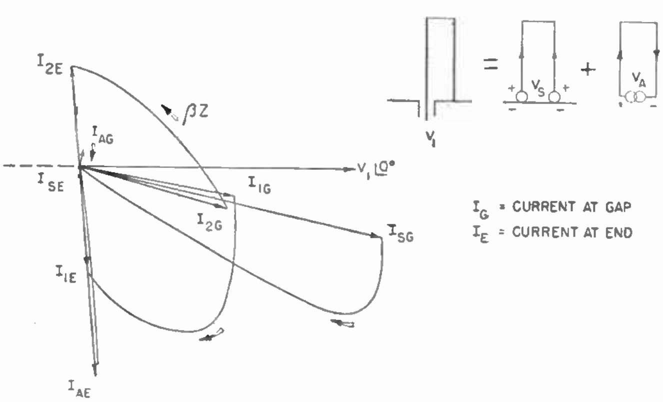


Fig. 10—Vector diagram for current distribution on folded dipole, $h=\lambda/4$.

Since the equivalent short-circuited transmission line is slightly longer than a quarter wavelength, the antisymmetrical impedance is high and capacitive. This means that the antisymmetrical current is small and leads the applied voltage by nearly 90 degrees at the gap. At the quarter wavelength from the end, phase reversal takes place, with the antisymmetrical current going to a minimum. Beyond this point the current increases sinusoidally, reaching a maximum at the short-circuited termination. The vector sum and difference of the symmetrical and antisymmetrical currents give the currents on the driven and parasitic elements, respectively. For this case the following impedances are obtained:

Experimental

$$Z_1 = 188 + j36.4$$
$$Z_{11} = 513 - j116.9$$
$$Z_{12} = -467 + j128.1$$
$$Z_S = 92.2 + j22.4$$
$$Z_A = 1{,}980 - j450.$$

In Fig. 11 we see the measured current distribution along a folded dipole, with $\beta_0 h=\pi$. Here the antisymmetrical current distribution is seen to be very much greater than the symmetrical current; hence, no reasonably accurate decomposition into symmetrical and antisymmetrical currents is possible.

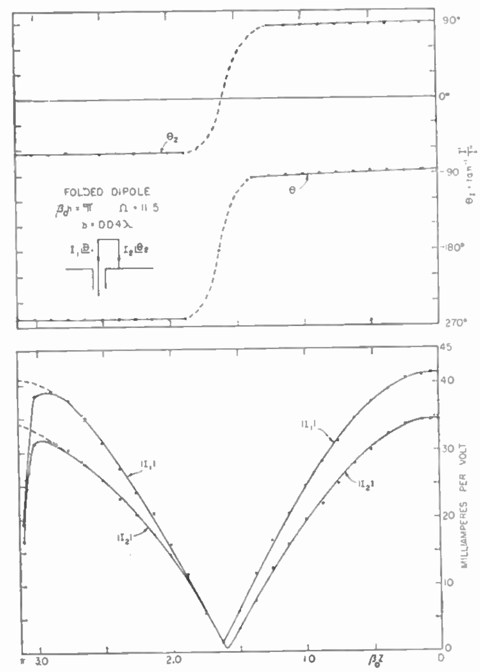


Fig. 11—Measured current on folded dipole, $h=\lambda/2$, $b=0.04\lambda$.

V. Conclusion

A method has been developed for measuring and interpreting distributions of current along both folded and parasitic dipoles. The results of the measurements have been compared with the theory, and qualitative agreement has been obtained. This is another illustration of the use of the concept of symmetrical and antisymmetrical components in physical problems.

Acknowledgments

The authors are indebted to R. W. P. King of the Cruft Laboratory, Harvard University, for his guidance in the work, and to Mrs. R. Dressler, who did part of the computing.

A High-Speed K-Band Switch*

MAURICE W. LONG†, ASSOCIATE, IRE

Summary—**A high-speed rotary three-way switch for 1.25 centimeter waves is described. An application for rapid scanning and performance characteristics are given.**

I. Introduction

The switch discussed here was designed to permit rapid scanning of a sector by means of a multiple-dish rotary antenna. For this application, each output arm shown in Fig. 1 is connected to a primary feed by an open choke-flange coupling. A stationary shorting barrier which has a window is between these open junctions. As the rotor spins with the antenna assembly, power is successively delivered to each feed while the others are short circuited. This switching arrangement reduces the required rotating speed in rapid-scan systems, thereby minimizing mechanical difficulties inherent in high-speed systems, and decreases the dead time which is present in a single antenna system.

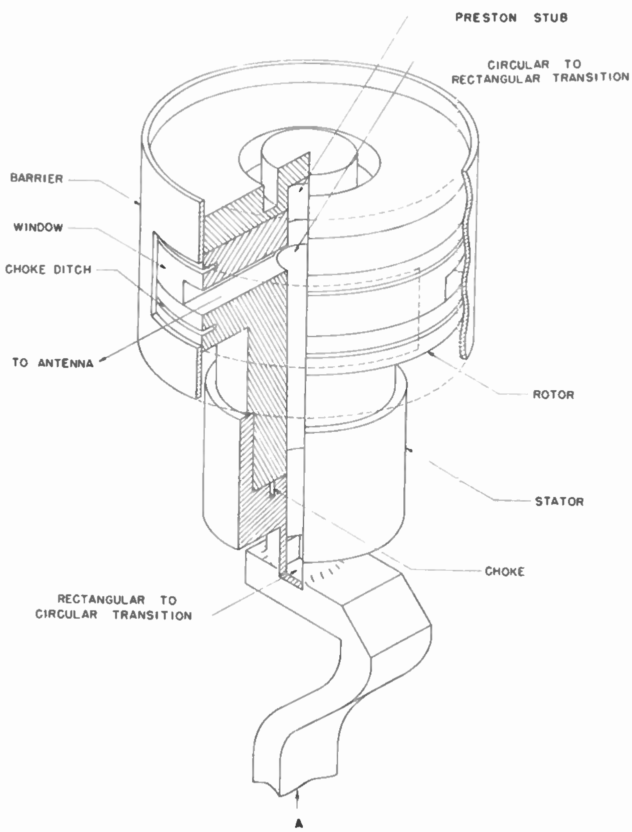


Fig. 1—Detailed drawing of three-way rotary switch.

The switch is based on a TM_{01}-mode rotary joint.[1,2] As shown in Fig. 1, TE_{10}-mode microwave energy is introduced at A and is propagated through a standard TM_{01} transition or stator into the circular waveguide section. Since the TM_{01} mode is symmetrical about the axis of the circular waveguide, the energy incident upon the output transition or stator is not a function of its position in azimuth. During the on time, a 1.25 VSWR (voltage standing-wave ratio) is not exceeded over a two per cent band. The maximum VSWR during switching is slightly less than 2. The output power variation with rotation is less than 0.05 db at any frequency within the 2 per cent band. The insertion loss is approximately 0.50 decibel.

II. Description of Switch

The rotor diameter was chosen so as to facilitate the use of broad-band matching techniques by means of inductive irises placed in the output waveguides. This method of matching to the loads separately insures that equal power be delivered to each antenna. The on time of the switch is slightly greater than twice the dead time. By sacrificing bandwidth, the switching angle may be reduced by increasing the rotor diameter by an integral number of wavelengths. Increasing the number of output waveguides will further decrease the bandwidth. An extension of the round waveguide below the rectangular waveguide, terminated by a short circuit, similar to that used in the Preston TM_{01} transition,[3] was employed to suppress the dominant TE_{11} mode within the rotor. The aperture is cut to slightly overlap two adjacent waveguides in order to reduce reflections during switching. A rectangular window has been used for simplicity; however, slanting the short sides of the window may reduce the switching VSWR. If the relative motion of the rotor and barrier window is considered, it may be seen that the barrier is a variable inductive iris during the time a window edge moves across the face of an output waveguide.

Since the field in the stator and rotor are relatively free from asymmetry, it follows that large TE_{11}-mode

* Decimal classification: R310. Original manuscript received by the Institute, December 18, 1950; revised manuscript received, May 24, 1951.
The work was conducted under The United States Navy Bureau of Ordnance, Contract No. 10020.
† Research Department, Georgia Institute of Technology, Atlanta, Ga.

[1] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Co., New York, N. Y., vol. 9, p. 385; 1948.
[2] See pp. 538–539 of footnote reference 1.
[3] See pp. 388–389 of footnote reference 1.

reflection coefficients are present at the transitions. Thus, cavity action is present when the round-shaft length is an integral number of half wavelengths in the TE_{11} mode. As the number of wavelengths in the round waveguide is increased, the bandwidth between resonances is decreased. Therefore, the circular waveguide is kept short to avoid resonances within the operating band. The shaft length was adjusted so as to prevent resonances and to result in impedance cancellations which minimize the over-all VSWR. In Fig. 2, the VSWR's for the stator, rotor, and complete switch are plotted versus free space wavelength.

The barrier is spaced several thousandths of an inch from the rotor face. Choke ditches are milled in the rotor face above and below the output apertures to provide a good shorting characteristic at the barrier, thereby reducing energy leakage and erratic performance. The output waveguides are then spaced a few thousandths of an inch from the barrier face. The switch is made of annealed brass and silver plated to mprove the conductivity.

ACKNOWLEDGMENTS

The author is indebted to J. E. Boyd, J. S. Hollis, R. E. Honer, and T. E. Roberts for inspiration and assistance.

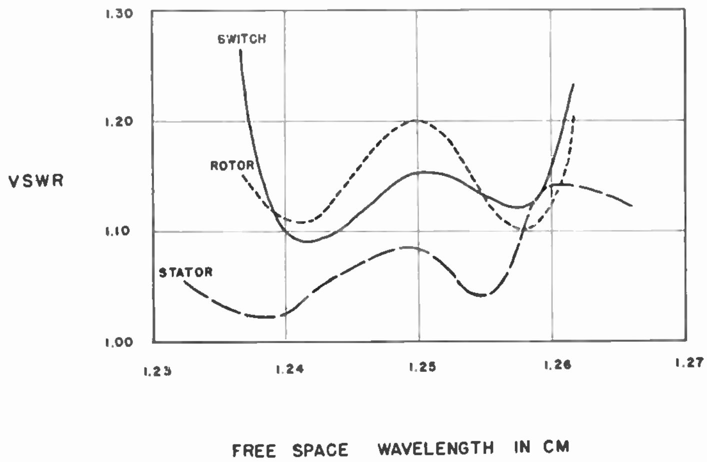


Fig. 2—VSWR versus free-space wavelength for stator, rotor, and complete switch.

I.R.E. Electroacoustics Standards: Genesis of the Glossary of Acoustical Definitions

In the May, 1951 issue of the PROCEEDINGS OF THE I.R.E., the "Standards on Electroacoustics: Definitions of Terms, 1951" appeared. Questions of credit for work done have arisen; the IRE is happy to thank the persons named below, as well as many others who contributed to the formulation of the standard and have not been mentioned. An explanation is in order:

At the end of the war the IRE Committee on Electroacoustics undertook the revision of the published IRE Standards on Electroacoustics of 1938. By the end of 1947 a "final draft" was distributed for criticism and by April, 1948 the proposed standard had been presented to the IRE Standards Committee, revised, and was ready for final action by the IRE Standards Committee. The IRE was thus quite close in 1948 to having new standard definitions of terms on electroacoustics.

In April, 1948 the American Standards Association subcommittee Z24A (sponsored by the Acoustical Society of America), having undertaken the revision of the ASA Acoustical Terminology of 1942, asked IRE to join with it. IRE agreed, and the processing of the new electroacoustics standard within IRE was stopped. Mr. Eginhard Dietze, Chairman of IRE Electroacoustics Committee, was made co-chairman of Z24A.

In 1950 the final form of a proposed standard, which was the direct responsibility of "the combined IRE Electroacoustics Committee and ASA Subcommittee Z24A on Acoustical Terminology," was adopted. For substantial reasons the IRE Standards Committee asked IRE representatives on ASA Z24 (not including Mr. Dietze, who died recently) when voting for adoption to request (but not require) an editorial qualification concerning a category of terms. (This is mentioned because these terms do not appear in the published IRE standard.)

The proposed standard from the combined group was presented (late 1950) to the IRE Standards Committee by the IRE Electroacoustics Committee for adoption and publication as an IRE standard. This is the course which would have been pursued if the Electroacoustics Committee's work up to 1948 had continued to follow IRE procedure, and had the great advantage that it would allow publication in the PROCEEDINGS so that all interested IRE members might benefit. The IRE Standards Committee requested extensive changes (these changes consisted of the omission of a group of terms; the items retained in the standard agree in wording with those of the combined IRE-ASA group report), and the result is the standard published in the May, 1951 issue of PROCEEDINGS.

The upshot of this situation may be partially summarized thus:

(1) The published IRE standard is not the ASA proposed standard, but agrees with the latter to the extent to which the former goes.

(2) Almost all the terms appearing in the IRE standard were included in the draft being circulated in 1947 by the IRE Electroacoustics Committee, but many definitions have undergone revision since then.

(3) IRE would not normally adopt or reproduce an ASA standard as its own, but in the present case the work was a joint effort of IRE on the one hand and ASA on the other.

(4) In the period between April, 1948 and Fall, 1950 many persons as well as the IRE Electroacoustics Committee worked on the standard, and to them acknowledgment and thanks are due. In particular, Dr. Leo L. Beranek, who became Chairman of ASA Sectional Committee Z24 some time after the merger of the work took place, and Mr. C. F. Wiebusch, who was co-chairman with Mr. Dietze of Subcommittee Z24A, should be singled out.

IRE members will have noticed that, preceding each standard published in the PROCEEDINGS, there are listed the names of members of the Standards Committee and of members of the originating IRE committee. The main purpose of this is to indicate *responsibility*, and only incidentally to give *credit*. However, in the present case it may be noted that of the combined group 17 were members of the IRE Electroacoustics Committee and were listed with the standard; the remaining 17 are:

Laurence Batchelder	R. F. Norris
V. L. Chrisler	R. V. Parsons
Floyd A. Firestone	H. I. Reiskind
C. W. Hewlett	M. S. Richardson
J. W. Horton	A. M. Small
T. A. Kvaas	S. P. Thompson
James J. Lamb	H. M. Trent
J. C. R. Licklider	C. F. Wiebusch
R. W. Young	

Correction

The following errors in the IRE Standards on Electroacoustics appeared on page 529 of the May, 1951 issue of the PROCEEDINGS:

line 9: 10^2 should read 10^5
line 25: 0.000045 should read 0.0000045
Table VI, last column: 1523.8 should read 1532.8

Contributors to Proceedings of the I.R.E.

William G. Abel was born in Louisville, Ky., on March 3, 1915. He received the B.S. degree in physics from Catholic University (Washington, D. C.) in 1939, and the M.S. degree in physics from Fordham University in 1947. Since 1947 he has been with the Raytheon Manufacturing Company, where he is at present engaged in research involving communications and propagation in the high-frequency band.

William G. Abel

❖

Sterling R. Anderson (A'41) was born in Toledo, Ohio, on December 6, 1916. He received the degrees of B.S. in electrical engineering and B.S. in engineering mathematics from the University of Michigan in 1940. Employed by the consulting radio engineer, Paul F. Godley, at Upper Montclair, N. J., from 1940 to 1942, he designed and adjusted directive antenna arrays for broadcasting stations.

Sterling Anderson

Since 1942, Mr. Anderson has been working on antennas and associated equipment for radio aids for air navigation at the United States' Civil Aeronautics Administration Technical Development and Evaluation Center, Indianapolis, Ind. At the present time, he is serving as chief of the antenna and propagation branch of the center.

❖

Sanborn C. Brown was born on January 19, 1913, at Beirut, Lebanon. He received the A.B. degree from Dartmouth College in 1935, and the M.A. from the same institution in 1937. He was an assistant in the Department of Physics at Dartmouth from 1935 to 1937. He held a fellowship from that college at Massachusetts Institute of Technology from 1937 to 1938, and was a teaching fellow from 1938 to 1941 and instructor from 1941 to 1944 in the Physics Department. He received his Ph.D. from M.I.T. in 1944, and since then has been a member of its physics faculty, now holding the rank of associate professor. At present he is in charge of the microwave gas-discharge work in the Research Laboratory of Electronics of M.I.T., and is Chairman of the Division of Electron Physics of the American Physical Society.

Sanborn C. Brown

❖

F. W. Bubb, Jr. (A'44) was born on July 12, 1921, at St. Louis, Mo. He received the B.S. degree in electrical engineering with honors in 1943, and the M.S. degree in applied mechanics in 1949, both from Washington University. From 1943 to to 1948 he was successively instructor at the Massachusetts Institute of Technology Radar School in Boston; chief product engineer at the Navy division of McQuay-Norris Manufacturing Company, St. Louis, Mo., a VT fuze assembly plant; senior design engineer at the Naval ordnance division of Eastman Kodak Company, Rochester, N. Y.; and research engineer engaged in work on automatic control devices at Washington University, St. Louis, Mo. During 1948, Mr. Bubb invented and has applied for patent on an electronic instrument for measuring and controlling fluid phases and extremely low humidities. In 1948, he joined the staff of St. Louis University as an assistant professor of engineering, and was recently appointed associate profession of engineering.

F. W. Bubb, Jr.

Mr. Bubb is a member of Tau Beta Pi, Sigma Xi, and Pi Mu Epsilon, and is a registered professional engineer in the fields of applied mechanics and electrical engineering.

❖

Marvin Chodorow (A'43–SM'47) was born on July 16, 1913, in Buffalo, N. Y. He received the B.A. degree in physics from the University of Buffalo in 1934, and the Ph.D. degree from the Massachusetts Institute of Technology in 1939. During 1940, he was a research associate at Pennsylvania State College. Dr. Chodorow was an instructor of physics at the College of the City of New York from 1941 to 1943, when he became associated with the Sperry Gyroscope Company as a senior project engineer. He remained at Sperry until 1947, when he joined the physics department of Stanford University, where he is now associate professor.

Marvin Chodorow

Dr. Chodorow is a member of the American Physical Society, and of Sigma Xi.

❖

Ralph I. Cole (A'29–SM'46) was born in St. Louis, Mo., on August 17, 1905. Shortly after receiving the B.S. degree from Washington University, St. Louis, Mo., in 1927, Mr. Cole entered Government Service as a radio engineer for the Signal Corps, engaged in research and development at their Fort Monmouth Laboratory. Projects assigned to his cognizance included direction finding and tank radio, and it was under his technical direction that the first modern tank radio was developed, using such components as Radio Receiver BC312. He is the holder of several patents concerning improvements in radio direction finding systems and special radio components.

R. I. Cole

In 1936 Mr. Cole received the M.S. degree in physics from Rutgers University. He assumed charge of Research and Development of Radio Direction Finding and Fighter Control Equipments intended for Signal Corps and Air Force use in 1940. With the advent of hostilities, Mr. Cole was directly commissioned a Major in the Army of the United States and assigned to the Signal Corps Laboratory as Officer in Charge of the Radio Direction Finding Branch of the Eatontown Signal Laboratory. He was promoted to Lieutenant Colonel, Signal Corps, in July, 1944. Transferring to the Air Forces on February 1, 1945, he was placed in charge of all engineering activity concerning the Development of Ground Electronics Equipment for the Air Forces at Watson Laboratories.

Mr. Cole returned to civilian life in June, 1947, after having received the Army Commendation Ribbon and the Legion of Merit for his efforts in the research and development field. He is still active in Air Force Reserve matters and holds the rank of Colonel. Mr. Cole is now Technical Director, Electronics Development Division, Rome Air Development Center, Rome, New York.

❖

James A. Craig was born in 1907, at Hackensack, N. J. After graduation from high school, he was employed by the Bell

Contributors to Proceedings of the I.R.E.

Telephone Laboratories for seven and a half years. During this time he attended Cooper Union Night School of Engineering, and graduated in 1932 with the degree of B.S. in electrical engineering.

James A. Craig

From 1932 until 1936 Mr. Craig was engaged in aviation as a pilot and mechanic. After this period he did teletype-maintenance work at Hearst Radio, Inc. He was then transferred to Finch Telecommunications Laboratories, where he was engaged in development work on picture transmission and facsimile.

In 1938 Mr. Craig joined the Link Radio Corp., where he is currently employed as systems engineer.

❖

Leonard C. Edwards (A'50) was born on December 27, 1925, at Lowell, Mass. He received the B.S. degree in electrical engineering from the University of New Hampshire in 1949. Since that time he has been with the Raytheon Manufacturing Company, where his work has been concerned with communication, propagation, and pulse technique in the high-frequency band. He is a member of Tau Beta Pi and Phi Kappa Phi.

L. C. Edwards

❖

Carl E. Faflick was born in Cleveland, Ohio, on March 10, 1922. He attended Oberlin College from 1940 to 1943 and from 1946 to 1947, and in 1943 received the B.A. degree. He served in the Signal Corps, United States Army, from 1942 to 1946, and during 1944 and 1945 took the Harvard - Massachusetts Institute of Technology Army Electronics course. In 1947 he was an instructor in physics at New Mexico State College.

Carl E. Faflick

In 1949 Mr. Faflick received an S.M. degree from Harvard University, where he has been a teaching fellow and research assistant. At present he is in Europe as a Sheldon Traveling Fellow. He is a member of Sigma Xi.

Harry H. Goode was born in New York, N. Y., on July 1, 1909. He received the B.S. degree at New York University in 1931, the B.Ch.E degree from Cooper Union in 1940, and the M.A. degree in mathematics from Columbia University in 1945. He served as assistant to the Deputy Commissioner of the New York City Health Department from 1935 to 1938, as assistant director for statistics on the Commission for Study of Crippled Children, New York City, from 1938 to 1940, and as statistician-in-charge of the Central Statistical Division at the New York City Department of Health from 1940 to 1942.

Harry H. Goode

During World War II Mr. Goode was a research associate at Tufts College, where he worked chiefly in probability applied to war problems. In 1946 he joined the staff of the Office of Naval Research at the Special Devices Center, and, as research section head, computer consultant, and special projects branch head, did work in computer research, simulation, training, aircraft instrumentation and control design, antisubmarine warfare, and weapon system design.

Early in 1950 Mr. Goode became the head of the Systems Analysis and Simulation Department of the Willow Run Research Center at the University of Michigan. At present he is chief project engineer for the Center, whose work is concerned with research and development on complex systems.

Mr. Goode is a member of the American Association for the Advancement of Science, the Institute of Mathematical Statisticians, the American Mathematical Society, the Association for Computing Machinery, and Mu Alpha Omicron.

❖

Karl G. Hernqvist was born in Boras, Sweden, on September 19, 1922. He was graduated from the Royal Institute of Technology, in Stockholm, Sweden, in 1945. Since August, 1946, Mr. Hernqvist has been employed by the Research Institute of National Defense in Stockholm, except for the period from October, 1948, to November, 1949, when he was associated with the RCA Laboratories, Inc., in Princeton, N. J. At RCA he was a trainee under the auspices of the American-Scandinavian Foundation.

K. G. Hernqvist

In addition to his work at the Research Institute of National Defense, Mr. Hernqvist is pursuing graduate studies at the Royal Institute of Technology.

❖

Henry C. Hurley (A'40) was born in Star, N. C., on November 22, 1903. He received the B.S. degree in electrical engineering from North Carolina State in 1927.

Henry C. Hurley

Following graduation, Mr. Hurley was employed in the radio engineering department of the General Electric Company, Schenectady, N. Y. From 1930 to 1938, he was employed by the RCA-Victor Company at Camden, N. J., where he worked on special apparatus and test-equipment development and design. Since 1938, he has been engaged in the development of radio navigational aids at the United States' Civil Aeronautics Administration Technical Development and Evaluation Center, Indianapolis, Ind.

❖

Hugh F. Keary (A'46) was born in Ireland in 1899. He attended radio technical school in 1920, and received the United States' first-class commercial operator's license. As a marine radio operator, he spent five years aboard United States' merchant vessels.

Hugh F. Keary

From 1925 to 1929, Mr. Keary served as junior engineer at RCA radio central stations at Rocky Point and Riverhead, Long Island, N. Y. He resigned from this position to join the laboratory of the Wired Radio, Inc., Ampene, N. J., where he assisted in the development and testing of broadcast systems for transmission over power-line networks.

From 1932 to 1934, Mr. Keary was employed by the DeForest Radio Company and the Hygrade Sylvania Company as radio-test engineer in the development and testing of radio-communication equipment, and from 1935 to 1937, he was development and test engineer in the manufacture of air navigational aids and communication equipment for the Radio Receptor Company, N. Y.

In 1937 Mr. Keary joined the staff of the Bendix Radio Corporation, Washington, D. C., and served as development and test engineer for radio-range systems and com-

Contributors to Proceedings of the I.R.E.

munication equipment. He resigned in 1943 to assist in the design and development of vhf navigational aids systems for the United States' Civil Aeronautics Administration, Technical Development and Evaluation Center, Indianapolis, Ind.

❖

Maurice W. Long (S'47–A'51) was born on April 20, 1925, in Madisonville, Ky. He received the B.E.E. degree from the Georgia Institute of Technology in 1946, and the M.S.E.E. degree from the University of Kentucky in 1948. The next two years were spent in graduate studies at the University of Kentucky and Columbia University.

MAURICE W. LONG

Mr. Long was employed for part of 1946, 1947, and the summer of 1948, as a research assistant at the Georgia Institute of Technology. From 1947 to 1949 he was an instructor of electrical engineering at the University of Kentucky. At present, he is a research engineer in the Electronics Research Laboratories, at the Georgia Institute of Technology.

Mr. Long is a member of the American Society of Engineering Education, Sigma Pi Sigma, and an associate member of Sigma Xi.

❖

Kiyoshi Morita was born in Tokyo, Japan, on March 18, 1901. He was graduated in 1921 from Tokyo Higher Technical School, which was chartered later as the Tokyo Institute of Technology, and won the Tejima honor prize the same year. In 1933 he received a doctorate in electrical engineering from the Tokyo Imperial University. In 1941 he was appointed professor at the Tokyo Institute of Technology, where he was primarily concerned with the development of measuring apparatus for dielectric loss, paraboloidal reflectors, triple-line feeders, and the like.

KIYOSHI MORITA

Since then he has given lectures on high frequency and electronic engineering.

Dr. Morita won an honor prize for his contribution on uhf techniques from the Nippon Radio Kyokai in 1946. The same year he became chief of the Special Committee for UHF Measurement, sponsored by the Ministry of Education, and was active in the development of the wattmeter for uhf. Since 1949 he has headed the Committee for the Development of Vacuum Tubes for Microwaves.

In April, 1950, he was sent to the United States by the Galioa Fund. From that time until June, 1950, he visited Cornell University, Brooklyn Polytechnic Institute, the University of Michigan, and other universities, from which he studied the American engineering educational system. He returned to Japan and contributed much to the improvement of teaching in the universities there. He also did actual work, in collaboration with many American professors sent to Japan, in the engineering section of the Institute for Educational Leadership.

❖

For a photograph and biography of TETSU MORITA, see page 1463 of the November, 1951, issue of the PROCEEDINGS OF THE I.R.E.

❖

Peter G. Sulzer (A'46) was born in Media, Pa., on August 3, 1922. He attended Drexel Institute of Technology in Philadelphia from 1940 to 1943. During that time he also spent one year with the Westinghouse Radio Division in Baltimore, Md.

PETER G. SULZER

Mr. Sulzer was in the United States Army Signal Corps from 1943 to 1946, engaged for the most part in ionospheric work. He received the B.S. degree in 1947, and the M.S. degree in 1949, both in electrical engineering, from the Pennsylvania State College, where he had been engaged in designing ionosphere equipment. Since September, 1949, Mr. Sulzer has been employed at the Central Radio Propagation Laboratory of the National Bureau of Standards in Washington, where he is concerned with ionospheric instrumentation.

❖

Vernon B. Westburg (S'46–A'49) was born in Chicago, Ill., on October 9, 1920. He received the degree of B.S. in electrical engineering from Purdue University in 1941, and the M.S. degree in 1942.

V. B. WESTBURG

From 1942 to 1946 Dr. Westburg was employed by Zenith Radio Corporation, where he worked on radar indicator circuits and antennas for automotive radio. From 1946 to 1948 he was the recipient of the Sylvania Electric Products fellowship at Stanford University, and in 1950 he received the Ph.D. degree in electrical engineering. At present he is a research engineer in the vacuum-tube laboratory of Stanford Research Institute.

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

❖

Philip Woodward was born at Gnosall, England, on September 6, 1919. He received the B.A. degree in mathematics from Oxford University in 1942, and since 1941 has been employed by the Telecommunications Research Establishment in England, where he is now a Principal Scientific Officer. During the war, he was trained in research under professor H. G. Booker of Cornell University, who was then in charge of the theoretical work at the Telecommunications Research Establishment.

PHILIP WOODWARD

Mr. Woodward's work was devoted to research on the theory of antennas, radio propagation, and turbulent diffusion in the atmosphere. His mathematical interests now concern the theory of probability, the Fourier transformation, and the theory of information.

Correspondence

More on Telepathic Communication*

Hollmann's[1] correspondence brings to light the following: Loefgren of The Royal Institute of Technology, Stockholm, Sweden carried out extensive pioneer work in the radio sciences, which also extended to brain radiations. As his assistant, I was in correspondence, during 1936 and 1937, with Cazzamalli, and received various papers and reports with circuit diagrams and photographs of Cazzamalli's equipment, as well as oscillograms of brain waves. The emotional excitation of the patient via "radio transmission" resulted in jerks in the curve of the oscillogram, more or less as the corresponding case would be, using conventional encephalographic methods. From the material presented no definite conclusion could be drawn at that particular time as to whether or not true brain-wave radiation was at hand, particularly since Cazzamalli's equipment was rather nonprofessional. There was no uncertainty in Cazzamalli's statements, however, that he actually received communications at a distance from active human brains without the use of wires.

Much more interesting, perhaps, is an earlier contact I had in this field as the guest of von Ardenne of Berlin, Germany, a well-known radio scientist. In his residence, converted into a huge and well-equipped four-story electronics laboratory, one room was used for brain-wave tests. Here, in 1929, von Ardenne demonstrated various short-wave receivers suitable for brain-wave reception, and he claimed successful results. These receivers, superheterodyne and super-regenerative types, went down to 5 or 10 meters, and were equipped with the Ardenne-developed "Loewe" tube following the detector. After confronting von Ardenne again in 1937 with the Cazzamalli results, correspondence developed in which von Ardenne claimed to have received brain-wave vhf communication on special receivers during previous years. It may be, therefore, that Cazzamalli made the discovery of the radiation phenomena in medical science around 1923, while von Ardenne was the originator of modern receiving and measurement methods of electromagnetic brain waves around 1927

Just as research work in Stockholm had begun to check Cazzamalli's and von Ardenne's results, both of these scientists declared that no further correspondence could be carried on since this field of research had been made secret by Mussolini and Hitler.

As the author was connected, in 1947, with the United States Air Force, as the Chief Communications Laboratory, Cambridge Field Station, he strove to have the USAF make funds available for triple-continued research investigations. These would include (1) modified encephalographic recording of the patients' conditions, and (2) microwave and infrared reception of the radiation; these technical approaches being associated with such techniques as may develop from (3) a rigorous study of the extra-sensory perception results obtained by Duke University. The latter's pioneer work in esp provided the missing link (not just the most exciting controversy of modern scientific thought) and also grounds on which *the radio scientist enters the field of human psychology* to unite with the psychologist in research work. My requests for funds in 1947 were not approved by the USAF because of the urgency of other research work.

No doubt this is the right time for serious research work along the lines indicated above. Such work is bound to yield not only startling, but also highly useful results. Thanks to Shannon and others a firmly grounded information theory applicable to this field, is available, as was indicated by Mr. Bibbero[2] in the original correspondence on these pages.

HARRY STOCKMAN
Electronics Consultant
Waltham, Mass.

* Received by the Institute, July 25, 1951.

[1] H. E. Hollmann, "Telepathic communication," PROC. I.R.E., vol. 39, p. 841; July, 1951.

[2] R. J. Bibbero, "Telepathic communication," PROC. I.R.E., vol. 39, pp. 290–291; March, 1951.

Checking Analogue Computer Solutions*

An important problem faced by the users of analogue computers is the checking of solutions. Although the digital solution of sample cases is likely to be both tedious and expensive in coding and checking time, particularly for complicated problems, nevertheless the desirability of such parallel solution was several times mentioned at the recent IRE National Convention.

In the course of testing our analogue computer, we devised a direct check on analogue solutions which, so far, has obviated the need for digital check solutions. The method will catch mistakes in scale factors and machine setup, if the error they cause is somewhat greater than the general level of precision expected from the machine solution.

Essentially, the check consists of substituting the machine solution back into the original set of equations. In the case of a set of differential equations, the procedure is to record all the variables and their various derivatives. The machine values are converted back into mathematical quantities by applying the scale factors. If, on substitution, one or more equations do not yield identities, either a scale factor error exists in those equations or the proper terms are not being summed by the solving amplifier. As the final logical step, it is necessary to check, by numerical integration, whether each derivative is the integral of the next lower derivative.

E. LAKATOS
Bell Telephone Laboratories, Inc.
Murray Hill, N. J.

* Received by the Institute, March 29, 1951.

Mu-Beta Effect Calculator*

An additional application of some value has been found for the calculator described by Felker.[1] The stability margin, described by Duerdoth,[2] is a measure of the closeness of approach of the $\mu\beta$ locus to the critical point $1|\underline{0}$ The advantage which the stability margin offers is ease of measurement: It is, as Duerdoth shows, simply the ratio of gain with feedback to gain without feedback, and it provides the most convenient production test of stability.

The stability margin is defined as

$$\sigma = 20 \log_{10} | \mu_f/\mu |,$$

where

$$\mu_f = \mu/(1 - \mu\beta),$$

and μ, β have the usual meanings. After some rearrangement, we obtain

$$\sigma = 20 \log_{10} \frac{-1/\mu\beta}{1 - 1/\mu\beta}.$$

This equation may be compared with Felker's equation (3), which is equivalent to

$$\gamma_{db} = \frac{-\mu\beta}{1 - \mu\beta}.$$

It is thus clear that the calculator can be used equally to determine σ and γ. The stability margin is obtained for any particular $\mu\beta$ by finding the mu-beta effect corresponding to $-|\mu\beta|$ db instead of $|\mu\beta|$ db.

As an example we may consider the servoamplifier of Felker's paper. The stability margin obtained in the "normal gain" and "12 db up" conditions is tabulated in the following:

normal gain

ω	$-\lvert\mu\beta\rvert$ db	θ	σ db
1.5	−17	37°	−16
2	−13	44°	−11.6
4	−5	56°	−3.4
10	4	57°	+1.5
20	15	23°	+1.6

gain 12 db up

ω	$-\lvert\mu\beta\rvert$ db	θ	σ db
1.5	−29	37°	−29
2	−25	44°	−25
4	−17	56°	−16.5
10	−8	57°	−6.6
20	+3	23°	+7.1

Duerdoth recommends that a production limit of +6 db should be used. It is seen at once that with gain up 12 db this amplifier does not fall within limits.

It has been found that this direct use of the calculator for stability margin offers worth-while advantages in designing amplifiers in terms of the final test criterion.

H. JEFFERSON
Telefonaktiebolaget
L. M. Ericsson
Stockholm, Sweden

* Received by the Institute, May 14, 1951.

[1] J. H. Felker, "Calculator and chart for feedback," PROC. I.R.E., vol. 37, p. 1204; October, 1949.

[2] W. T. Duerdoth, "Some Considerations in the Design of Negative-Feedback Amplifiers," *Proc. IEE* (London), vol. 97, Pt. III, p. 138; May, 1950.

Correspondence

Preferred Numbers*

In the February, 1951 issue of the PROCEEDINGS OF THE I.R.E. there appeared a guest editorial on preferred numbers by Mr. A. F. Van Dyck. This was followed by a second guest editorial on the same subject by Mr. Virgil Graham. The writer of the following communication, Mr. John B. Moore, presents in the following an analytical commentary on the entire selection and utilization of preferred numbers. The information in his letter should be of interest to the members of the Institute and to those committees of the Institute concerned with the use of preferred numbers.—*The Editor.*

During roughly the past year there has been published a number of articles, editorials, and pamphlets dealing with the question of using so-called "preferred numbers" for sizes or values of items to be carried in stock. The aim is laudable. There seem, however, to be two basic questions that need to be answered in any specific case:

1. Is a geometric, rather than an arithmetic, series suited to the practical requirements of the particular case?
2. What should be the basic step ratio?

Question 1 is the basic question, in any particular case, and is the one with which this present discussion will deal primarily. Before proceeding to it, however, let us briefly review some of the geometric or logarithmic series and bases that have been referred to in other published discussion.

Step Ratios

It will suffice to comment briefly, as follows, regarding the bases of certain existing series:

ASA	$\sqrt[20]{10}$	$= 1.122$
RTMA	$\sqrt[24]{10}$	$= 1.100$
Brown and Sharpe (AWG) wire gauge	$\sqrt[6]{2.005}$	$= 1.123$
Decibel	$20 \times \log_{10} 1.122$	

The ASA plan adheres strictly to the decimal system, whereas the RTMA plan uses a decimal base and a duodecimal (12) root. The latter evidently is advocated chiefly because it results in a series that fits nicely into a pattern of ± 20- or ± 10- or ± 5-per cent tolerances—an established pattern in certain lines—and results in no rejects. In view of other practical considerations, it is a question if these particular tolerance values should be considered so sacred. These could be increased slightly to fit the ASA-sponsored series of "preferred numbers." There then would be the not inconsiderable advantage that only one set of so-called "preferred numbers"—odd and confusing at best—would have to be remembered or looked up.

The Brown and Sharpe (AWG) wire gauge uses the basic ratio $\sqrt[6]{2}$ for the smaller and medium sizes, then departs from the geometric concept and follows an essentially arithmetic and decimal system for the larger sizes, expressed in circular mils rather than in gauge numbers.

* Received by the Institute, June 5, 1951.

The decibel corresponds very closely to the loss in one mile of so-called "standard cable," which latter was the practical unit of telephone transmission loss or gain before adoption of the bel and decibel nomenclature.

Geometric or Logarithmic versus Arithmetic

The really basic question, in any particular case, is whether the practical requirements of actual use will be best met by an arithmetic or a geometric series of values. This warrants serious consideration.

Of the four geometric or logarithmic series mentioned above, the Brown and Sharpe wire gauge (AWG) and the decibel have been used and generally accepted for a good many years. The reasons for this are that first, they provide the required geometric series having practical ratios for the specific applications concerned, and that, secondly, in their practical everyday use, both of them are essentially simple arithmetic series.

That last statement warrants a bit of explanation. In using the Brown and Sharpe (AWG) wire gauge, the practical man almost invariably specifies by gauge number. He does not state exact diameter or cross section carried to three or more significant figures. He does not need to. The gauge numbers give him a simple and consecutive arithmetic series by which he can specify and obtain exact sizes that follow a geometric but unmemorizable series of values. So, with an absolute minimum of mental strain, he gets exactly what he wants.

The decibel provides a logarithmic ratio or geometric series, of practical size steps, and yet requires the practical user to mentally manipulate only a simple arithmetic system of values having the decimal arrangement to which he is accustomed by training and by use in his daily life.

Consider our monetary system. As an application of the decimal system, it is based on the fact that humans found it convenient to add and subtract by using the ten fingers on their two hands. Read that sentence carefully. It brings out the fact that our natural method of computation is addition or subtraction, and the units or bases employed therein are 1, 2, 5, and 10. Except for the 25 cent piece and the $2.50 gold coin, our monetary system fits that pattern exactly. That is the basis of the average person's mathematical processes.

To be of practical usefulness, away from the drawing board or the machine shop, sizes and values must be such that they can be referred to in terms of numbers closely associated with, or simply related to, the basic numbers 1, 2, 5, and 10. This requires that geometric or logarithmic series, involving odd and unmemorizable values, should be dealt with in some manner such as illustrated by the decibel system or the Brown and Sharpe Wire Gauge. These both provide the practical man with an easily remembered and easily used arithmetic system.

Conclusions

Geometric or logarithmic series would seem to be best adapted to such cases as: (1) dimensions of wire or sheet-gauge sizes, (2) nominal dimensions of unfinished stock that must be machined to finished dimensions, (3) sizes of items that are not used in additive combination, and (4) fields in which the ratio rather than the absolute value is of primary importance.

Arithmetic series, of either the consecutive-numbers type or the decimal (1, 2, 5, and 10) type, would seem to be generally preferable in most cases to those listed in the preceding paragraph.

Resistors, such as those used in the circuits of electrical and electronic equipment, should be rated and sold according to an arithmetic decimal series of values. If, however, a geometric series is to be used—for certain technical reasons—industry should at least be consistent and use a basic ratio which corresponds with that of the decibel system. It just so happens that this also agrees with the ASA proposal.

Recommendation

If the electronics industry wants a geometric or logarithmic series of values, for resistors, it would seem most consistent to rate such resistors in dbr—decibels above and below one ohm. As contrasted with any new series of so-called "preferred numbers," the dbr series of values would appear to offer many advantages.

Addendum

In certain published discussions regarding so-called "preferred numbers," there seems to have been considerable stress placed on the practical utility of the Brown and Sharpe wire gauge. Some elaboration of this point appears warranted.

The Brown and Sharpe wire gauge employs the basic ratio $\sqrt[6]{2.005}$. This base—2—is one of the natural preferred numbers, if I may call it that, in the 1, 2, 5, and 10 decimal system used daily by practically every individual, regardless of the extent of his or her formal education.

In view of the foregoing, it would seem to be in line with other discussions, and also with the requirements of practical use, to incorporate the basic 2-to-1 ratio in any system of so-called "preferred numbers" Fortunately, the well-established db system does just this. Reference to the following tabulation will bring out this point.

db (Ratio)	*Exact Values*	*Rounded Values*
0	10	10
2	12.59	12.5
4	15.85	16
6	19.95	20
8	25.12	25
10	31.62	32
12	39.81	40
14	50.12	50
16	63.10	63 or 64
18	79.43	79 or 80
20	100	100

Note that the 2-to-1 relationship prevails throughout the list of rounded values.

J. B. Moore
66 Broad St.
New York, N. Y.

Institute News and Radio Notes

IRE OFFICIALS ATTEND WESTERN CONVENTION

Discussing Western IRE Convention agenda are (from left to right): A. R. Ogilvie, Chairman of the IRE San Francisco Section; L. J. Black of Berkeley, Convention Chairman; George Bailey of New York, National IRE Executive Secretary; R. A. Isberg of San Francisco, Chairman of the Convention's first Technical Session; and A. V. Eastman of Seattle, Director of the IRE 7th Region.

1951 WESTERN IRE CONVENTION HUGE SUCCESS

The 1951 Western IRE Convention and 7th Annual Electronic Exhibit, held in the Civic Auditorium, in San Francisco, on August 22, 23, and 24, proved to be one of the most successful of all Western Conventions. The attendance at the exhibit exceeded 8,000, with over 1,100 registrants at the IRE Convention. The excellent quality of the exhibit coupled with the well-planned technical sessions provided the attendants with a complete view of the activities in the electronic field. The technical program featured a new theme wherein each session was arranged to cover a specific topic of a current interest in the electronic field. The chairman of each session was a leading authority on the subject under discussion. This well-balanced technical program gave a maximum of information for those in attendance. The climax of the technical sessions came Friday evening with the presentation of a paper by D. E. Foster on, "System Aspects of Color Television," with approximately 500 persons attending.

The Convention was also extremely successful for the wives of the IRE members. The highlight of the women's activities was a fashion show staged by Nelly Gaffney in the Peacock Court of the Hotel Mark Hopkins, with Paul Speegle as mirth-provoking commentator. The ladies were presented with leis that had been flown in from Hawaii especially for this occasion.

Plans are now underway for the 1952 Western IRE Convention and 8th Annual Pacific Electronic Exhibit to be held in the Municipal Auditorium, Long Beach, Calif., August 27, 28, and 29, 1952. Arrangements should be made now to insure that all IRE members, who wish to attend the Convention, may do so. Plans and information of the 1952 Convention will be carried in the future issues of the PROCEEDINGS.

GROUP ON TRANSISTORS ORGANIZED

Formation of a group on transistors, to assist its committee on electronics to establish sound policies for the development and functional application of transistors by the Armed Services, has been announced by the Research and Development Board of the Department of Defense.

Included in the group are six representatives from university and industrial laboratories: J. W. McRae (A'37–F'47), Bell Telephone Laboratories, chairman; E. F. Carter (A'23–F'36), Sylvania Electric Products; E. W. Engstrom, (A'25–M'38–F'40), RCA Laboratories; I. A. Getting (SM'46), Raytheon Manufacturing Company; A. G. Hill (SM'46–F'50), Massachusetts Institute of Technology; and G. F. Metcalf (A'34–M'38–F'42), General Electric Company.

The military service members are: Colonel C. J. King (A'30–VA'39–SM'50), Office, Chief Signal Office; Colonel G. F. Moynahan (S'49–A'50), Office of the Assistant Chief of Staff, Department of the Army; J. M. Bridges (SM'47), Bureau of Ordnance; C. L. Stec (S'35–VA'39–SM'49), Bureau of Ships; H. V. Noble (A'34–M'35–SM'43), Wright Air Development Center, Dayton; and E. W. Samson (M'47–SM'51), the Air Force Cambridge Research Center.

The first meeting of the group was held on October 11, 1951, New York, N. Y. At this time the group discussed its functions, responsibilities, and methods of operation.

NBS MEETING ON STORAGE

A Symposium on Williams Electrostatic Storage will be held at the National Bureau of Standards, Washington, D. C., on December 13 and 14, 1951. Those interested in this type of storage who wish to participate should contact the conference chairman, J. H. Wright, Electronics Division, National Bureau of Standards, Washington 25, D. C.

Calendar of COMING EVENTS

IAS-ION-IRE-RTCA Conference on Air Traffic Control, Astor Hotel, New York, N. Y., January 30.

1952 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6.

Radio Component Show, London, England, April 7–9.

Radio and Television Show, Manchester, England, April 23–May, 3.

IEE Television Convention, London, England, April 28–May 3.

IRE-AIEE-RTMA Symposium on Progress in Quality Electronic Components, Washington, D. C., May 5–7.

IRE National Conference on Airborne Electronics, Hotel Biltmore, Dayton, Ohio, May 12–14.

4th Southwestern IRE Conference and Radio Engineering Show, Rice Hotel, Houston, Tex., May 16–17.

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 27–29.

NEW ITALIAN PUBLICATION

Announcement has been made of a new Italian publication, *Marconi*, published by Marconi Societa Industriale, Via Hermada 2, Genova-Sestri, Italy. It is a quarterly review of telecommunications and electronics techniques. Technical articles are published in Italian; summaries in Italian, English, and German.

IRE EMPORIUM SECTION SEMINAR HAS LARGEST ATTENDANCE

The 12th Annual Summer Seminar of the Emporium Section of the IRE, held August 17 and 18, 1951, enjoyed the largest attendance in its history. Average session attendance was over 200, with nearly 100 out of town guests. A total of 175 were entertained at the picnic held Saturday afternoon.

The technical section of the program was divided into two main parts consisting of a papers' session Friday evening, and a Reliable Tube Symposium Saturday morning. Speakers and members of the Symposium panel were all outstanding men in the electronics industry who had made noteworthy contributions to the reliability programs now so important to the Armed Forces. Information on the program schedule can be found on page 971, in the August, 1951, issue of the PROCEEDINGS.

1952 IRE Convention Set for March 3–6

March 3–6, have been selected as the dates for the largest and most important radio engineering meeting and show in 1952—the IRE National Convention. The Waldorf-Astoria Hotel and Grand Central Palace in New York City will again be the scene for the 4-day program of technical sessions, exhibits, and social activities.

This Convention takes on special significance for Institute members in that 1952 marks the 40th anniversary of the founding of IRE. Many of the Convention activities will be keyed to this important milestone and to the Convention slogan, "Forty years sets the pace," making it an event of particular interest to the membership.

The opening meeting of the Convention will be the Annual Meeting of the Institute which, in honor of the anniversary, will feature an interesting presentation of 40 years of IRE by Dr. Alfred N. Goldsmith and John V. L. Hogan, co-founders of IRE, entitled, "The IRE: From Acorn to Oak." This meeting will be held at 10:30 A.M., Monday, March 3, in the Jade Room of the Waldorf-Astoria Hotel. The Radio Engineering Show, enlarged to include four floors of exhibits, will again occupy the Grand Central Palace, two blocks south of the Waldorf on Lexington Avenue. Sessions of the extensive technical program will be held in both locations and the nearby Belmont-Plaza Hotel.

The social activities will start on the first evening of the convention, March 3, with a "get-together" cocktail party, which this year will be held in a new and more spacious location, the Grand Ballroom of the Waldorf-Astoria. On the following day the world-famous Starlight Roof of the Waldorf will be the scene for the traditional President's Luncheon in honor of Dr. Donald B. Sinclair, the IRE President for 1952. Tables will be set aside for members of the Professional Groups in order that they may become better acquainted with engineers in their own field of interest. Charles E. Wilson, Director of Defense Mobilization, will address the Annual Banquet on Wednesday evening, March 5, in the Grand Ballroom. The recipients of the 1952 IRE Awards will be introduced at that time. For the wives of members and visitors an entertaining 4-day program of tours, plays, and fashion shows is being prepared. Members are urged to buy their tickets now. Ticket prices are as follows: Banquet, \$12.00; Luncheon, \$5.75; Cocktail Party, \$3.80.

Full details of the program, including 100-word summaries of technical papers, will appear in the February issue of the PROCEEDINGS.

Technical Committee Notes

The **Standards Committee,** under the Chairmanship of Ernst Weber, held a meeting on October 11, 1951. Acting-Chairman Ernst Weber, in the absence of Chairman A. G. Jensen, reviewed certain items that had been discussed at the meeting of the Administrative Committee on October 10. At this meeting, Chairman A. G. Jensen appointed a Subcommittee to consider the need of a **Technical Committee on Pulse Systems.**

A Meeting was held by the **Committee on Modulation Systems** on September 18, Chairman W. G. Tuller presiding. This Committee has found it necessary to revise its scope of activity and is anticipating a change in its title. Recommendations of these changes will be submitted to the Standards Committee for their approval.

The Executive Committee approved the change in the title of the IRE Technical **Committee on Mobile Communications** to the IRE Technical **Committee on Mobile Communications Systems.** This Committee has broadened its scope of interests. The revised scope is as follows: Responsibility for mobile communication system matters such as general engineering studies, frequency allocations, and the consideration of new developments or techniques which may have a major effect on the engineering of mobile communication systems; To act as consultant with RTMW, RTCA, FCC, and other bodies as may desire its assistance on engineering matters of mobile communication systems.

Under the Chairmanship of H. E. Roys, the **Sound Recording and Reproducing Committee** held a meeting on October 13. The Subcommittee on Mechanical Recording, of this Committee, convened in the morning of October 13, joining the main Committee in the afternoon. Lincoln Thompson of the Subcommittee reported on their activities. Preparations are underway to outline tasks on the calibration of frequency records, and plans are being made to write material on the amplitude method of measurement.

H. R. Terhune has been appointed to serve as alternate to A. F. Pomeroy, who is the IRE Representative on the ASA Sectional Committee Y-1 Abbreviations.

Professional Group Notes

Delegates from all of the **IRE Professional Groups** attended the organizational meeting on November 9, of the Technical Program Committee, which is formulating plans for the **IRE National Convention** on March 3–6, 1952. The Committee Chairman, W. H. Doherty of Bell Telephone Laboratories, Murray Hill, N. J., will assist the Professional Groups in arranging the Symposia to be sponsored by the various Groups at the Convention.

The **IRE Professional Groups on Airborne Electronics, Audio, Vehicular Communications, and Antennas and Propagation** are now levving assessments on their members for publication expenses. Arrangements are being made by the **IRE Professional Groups on Instrumentation and Information Theory** to adopt the assessment payment for their respective Groups.

Plans to hold a membership drive are being made by the **IRE Professional Group on Airborne Electronics** during its participation in the IAS-IRE-RTCA-ION Annual Convention on Electronics and Aviation, January 28 to February 1, 1952, at the Hotel Astor, New York, N. Y. This Group's meeting will take place on January 30, 1952.

The recently formed **IRE Professional Group on Information Theory,** headed by Acting Chairman Nathan Marchand, is now in the process of drawing up its Constitution and is also making plans for a membership drive. The Group is tentatively planning a symposium to be held in January, 1952, and expects to sponsor a session at the IRE National Convention in March, 1952.

The IRE Professional Group on Nuclear Science held a very successful symposium on December 3 and 4, at the Brookhaven National Laboratory, Upton, L. I., in cooperation with the Atomic Energy Commission. Full details on the papers presented, the banquet, and the panel discussion highlighting the Conference may be found on page 1343, of the October, 1951, issue of the PROCEEDINGS.

RDB Announces New Program

The Research and Development Board, Department of Defense, has received the recommendations of a group created by the Board in December, 1950, to look into the question of reliability of military electronic equipment and to recommend measures which will insure more reliable performance with a minimum of maintenance. Based upon these recommendations, RDB has established a clearing house for reliable information which will be responsible for the collection and dissemination of such information to laboratories concerned with government work. A. F. Murray, Radio and Television Consultant, Washington, D. C., who headed the previous study, and M. Barry Carlton of the RDB secretariat have been assigned the task of monitoring future research and development programs in this field. Mr. Carlton's title is that of, RDB, Coordinator of Reliability.

The Secretary of Defense and the Chairman of RDB have issued statements to the Military Department emphasizing the importance of this problem.

According to RDB officials concerned, improvement in reliability must include all the following links in the chain leading finally to the application of the equipment in the field: military characteristics, experimental models, specifications, manufacturing control procedures, service tests, final inspection, packaging and shipping, storage, installation, operational use, and maintenance. Consequently, the combined efforts of the Munitions Board, Joint Chiefs of Staff, the three Services, and the Research and Development Board are required. These organizations are represented in the group.

The Military Reliability Program can be effective only if it has the backing and cooperation of industry. It is vital that the precepts of reliability are disseminated to and carried out by the electronics industry upon which the Department of Defense depends for research, development, and high quality control during reliability production.

IRE People

Robert M. Page (SM'45–F'47), superintendent of Radio Division III at the Naval Research Laboratory, has returned recently from a three-month survey of electronics developments in Germany. He was employed as technical consultant to the Military Security Board, the Office of the High Command for Germany. His mission, sponsored by the State Department, was to establish closer working relations with German scientists, thus aiding mutual interchange of scientific information. Dr. Page visited 58 university and commercial laboratories while in Germany.

Robert M. Page

Dr. Page was born on June 2, 1903, in St. Paul, Minn. He received the B.Sc. degree in 1927 from the Hamline University and the M.A. degree from the George Washington University in 1932. In 1943 he received an honorary Doctor of Science degree from the Hamline University. He is the holder of several awards including the United States Navy Civilian Service Award, The Citation O.S.R.D., and the President of the United States Certificate of Merit.

In 1947 Dr. Page received the IRE Fellow Award for, "recognition of his pioneering achievements in solving some of the early problems of basic importance to radar." He has been a member of the IRE Board of Editors since 1946

❖

Robert W. Annis (S'47–A'51), **William C. Coombs** (S'33–A'35–VA'39), and **Charles A. Culver** (M'20–SM'43), have recently been appointed to the physics department staff of the Southwest Research Institute, San Antonio, Tex. The appointments were announced by **Paul Erlandson** (SM'50), the Institute Physics chairman.

Mr. Annis a senior research physicist from Champaign, Ill., is a specialist in electrical engineering, radio, and radio direction finding. He holds a Master of Science degree from the University of Illinois.

Mr. Coombs, supervisor of the measurements laboratory, is an authority in the fields of radio direction-finder design for intelligence intercept-, navigation-, and military-countermeasures application, radio development for production, and electronics development in physical research.

Dr. Culver, senior physicist, whose studies include courses at the University of Cambridge, Eng., specializes in the fields of electroacoustics, electrical and physical measurements, and optical, x-ray, and mass spectroscopy. Dr. Culver is the author of several technical books.

William A. Edson (M'41–SM'43) has been named Director of the school of electrical engineering at the Georgia Institute of Technology, it was recently announced.

William A. Edson

Dr. Edson, a native of Burchard, Neb., was born October 30, 1912. He studied electrical engineering at the University of Kansas and received the B.S. and M.S. degrees in 1934 and 1935, respectively. He then attended Harvard University on a fellowship, receiving the degree of Dr. Sc. in communication engineering in 1937. Following this, he joined the Bell Telephone Laboratories as a member of the technical staff but resigned that position to become assistant professor of electrical engineering at the Illinois Institute of Technology. He later returned to the Bell Laboratories to engage in research for the Army, working on the development of the microwave echo box, a device for testing and maintaining radar systems.

Dr. Edson accepted the position of professor of physics at the Georgia Institute of Technology in 1945, transferring to professor of electrical engineering in 1946, the position he held until the present appointment. He holds several patents in connection with oscillators and resonators, and has contributed numerous articles to technical periodicals in his field. He is a licensed engineer in Georgia and a consultant to the United States Bureau of Standards.

Dr. Edson is active in various professional and educational organizations including the American Physical Society, the American Association for Advancement of Science, the American Society for Engineering Education, and the National Society for Professional Engineers. He has served as Chairman of the IRE Committee on Standards of Piezoelectricity, and Chairman of the Subpanel on Frequency Research and Development Board. He is the past Chairman of the IRE Atlanta Section.

❖

Elmer W. Engstrom (A'25–M'38–F'40) has been named as Vice-President in charge of the RCA Laboratories Division, according to an announcement by Brigidier General David Sarnoff. Dr. Engstrom has been Vice-President in charge of research of the RCA Laboratories Division since 1945, and prior to this he served 2 years as Director, supervising research and engineering which resulted in wartime advances in radar, television, radio, and other electronic developments. He had previously served for 13 years in various RCA research positions.

Elmer W. Engstrom

Dr. Engstrom was born on August 25, 1901, in Minneapolis, Minn., and received the B.S. degree in electrical engineering from the University of Minneapolis in 1923. He then joined the staff of the General Electric Company, where he worked on transmitters, receivers, and sound motion pictures.

Dr. Engstrom has been a member of numerous IRE committees, including the Board of Editors, Awards, and Nominations. He served on the IRE Board of Directors in 1949. In June, 1949, he received an honorary degree of Doctor of Science from the New York University for his contributions as a research engineer.

❖

Allen F. Pomeroy (A'42–SM'43), a member of the technical staff of the Bell Telephone Laboratories at Murray Hill, N. J., met with experts from other countries at Montreux, Switzerland early in November to discuss graphical symbols for use on drawings. The President of the United States National Committee of the International Electrotechnical Committee appointed Mr. Pomeroy as the United States member of the Committee of Experts of Technical Committee 3 (Electrical Graphical Symbols).

Allen F. Pomeroy

While in Europe he visited organizations manufacturing and using microwave communication equipment in England and Switzerland.

Mr. Pomeroy has been with the Bell Telephone Laboratories since his graduation from Brown University in 1929. Since 1941 he has been engaged primarily in the design and testing of microwave components such as are used in the New York to San Francisco radio relay system now utilized for telephone service and to bring television programs from the Pacific coast. This system was developed at the Murray Hill Laboratory of Bell Telephone.

Mr. Pomeroy served as chairman to the IRE Symbols Committee in 1949–1951, and is currently the Institute Representative on the Drawing and Symbols Correlating Committee of the American Standards Association. He is also Co-chairman of the ASA Sectional Committee on Graphical Symbols, and was recently appointed the industry liaison member of the Standard Drawing Practice Committee of the Munitions Board Standards Agency in Washington, D. C.

His Institute activities have included work on the Annual Review, Circuits, Papers Review, and Standards Committees.

IRE People

Trevor H. Clark (A'32–SM'45) has been appointed as director of the division of Military Research and Development and special consultant to the physics department at the Southwest Research Institute, San Antonio, Tex. He was formerly with the Federal Telecommunication Laboratories, Nutley, N. J.

TREVOR H. CLARK

Mr. Clark is the holder of many patents on electronic devices. His special fields include vacuum tubes, communication equipment, microwave relay systems, radio-direction finders, electronic countermeasures equipment, research administration, and manufacturing methods. He was one of four scientists selected to establish the United States Laboratories of the International Telephone and Telegraph Company, and has held a number of technical and administrative positions in the foreign and domestic research development and manufacturing subsidiaries of that organization. In 1948 he was named assistant to the president of the Federal Telephone and Radio Corporation of Clifton, N. J. More recently he has been active in introducing new developments of the Federal Telecommunication Laboratories, particularly microwave links and television broadcasting. Mr. Clark was awarded the United States Navy Certificate of Commendation for his work in direction finders during World War II.

A native of Kansas, Mr. Clark received the B.A. degree from the Friends University in 1930, and the M.S. degree from the University of Michigan in 1933. He has been a member of various IRE Committees and served on the IRE Board of Directors in 1950. He is a member of the American Institute of Physics, the Accoustical Society of America, and the Armed Forces Communication Association.

❖

Maxwell K. Goldstein (A'30–SM'46) has been appointed as vice president and technical director of the Balco Research Laboratories, Newark, N. J. He was formerly associated with the Air Navigation Development Board.

M. K. GOLDSTEIN

Dr. Goldstein received his doctorate degree from the Johns Hopkins University in 1933, and subsequently was engaged in the development of automatic remote indicating systems for antiaircraft gun control apparatus and radio navigation aids. In 1939 he joined the Naval Research Laboratory and for a number of years was in charge of its radio direction-finder work; he later became the head of its navigation section.

In 1947 Dr. Goldstein was awarded the Distinguished Civilian Service Award by the Secretary of the Navy, "for distinguished contributions to the Naval Service in developing high-frequency direction finding as a vital weapon for combating the German submarine menace during the crucial Battle of the Atlantic."

Dr. Goldstein is a member of Sigma Xi.

❖

Charles B. Jolliffe (M'25–F'30), who has served as Executive Vice-President in charge of RCA Laboratories Division since December, 1945, has been elected to the position of Vice-President and Technical Director of RCA. In his new position, he will be responsible for the development of long-range plans for RCA and generally will supervise the execution of such plans by the division and subsidiary companies. In addition, he will co-ordinate broad engineering policies and will direct the representation of RCA in technical matters before the various public and governmental bodies.

CHARLES B. JOLLIFFE

Dr. Jolliffe, a native of Mannington, W. Va., was graduated from West Virginia University with a B.Sc. degree in 1915 and the M.Sc. degree in 1920. He was awarded a Ph.D. degree in 1922, at Cornell University, and in 1942 he received an honorary LL.D. degree from West Virginia University. He then served as a physicist in the Bureau of Standards' radio section and later became chief engineer of the Federal Radio Commission.

He first joined RCA as engineer in charge of its frequency bureau in 1935 and, subsequently, he was named chief engineer of the RCA Laboratories and assistant to the president of RCA Victor in 1941 and 1942, respectively. In September, 1942, he became vice president and chief engineer of the RCA Victor Division.

Dr. Jolliffe has been a member of many IRE Committees and has also served on the IRE Board of Directors in 1936–1937, and again in 1944.

❖

LaVerne R. Philpott (SM'48) has been appointed as chief of the Electronics Division of the Balco Research Laboratories. He was formerly with the Air Navigation Board.

A native of Nebraska, Dr. Philpott received his B.S. degree from the College of Idaho in 1925, and later worked as a research engineer at the Westinghouse Research Laboratories from 1928 to 1934. He then went to the United States Naval Research Laboratories where he remained until 1947.

In 1945, Dr. Philpott was awarded an honorary D.Sc. degree from the College of Idaho for his work in radar, and in 1947 he was awarded the Certificate of Merit by the President of the United States for his contribution to the design of the first United States Naval radar.

Dr. Philpott is also the holder of the Distinguished Civilian Service Award.

❖

F. W. Tatum (A'43–SM'47), professor of electrical engineering at the Southern Methodist University, Dallas, Tex., has been named Chairman of the Department of Electrical Engineering at S.M.U.

F. W. TATUM

Professor Tatum, a native of Texas, received his technical education at Columbia University, where he graduated with the B.S. and M.S. degrees in electrical engineering, in 1935, and 1946

From 1935 to 1947 he was associated with the American District Telegraph Company and held the position of engineering supervisor at the time of his resignation to join the faculty at S.M.U.

Professor Tatum is a member of the American Institute of Electrical Engineers, the American Society for Engineering Education, and the Texas Society of Professional Engineers.

❖

Robert Hertzberg (A'46), long active in the technical magazine field as writer, editor, and publisher, has established a consulting publications and public relations service for radio manufacturing and merchandising firms and their associated advertising agencies, in Jackson Heights, N. Y. Their service includes counsel on the preparation of military manuals and instruction books.

ROBERT HERTZBERG

Mr. Hertzberg worked as a radio feature writer for New York newspapers during the early days of broadcasting and later became managing editor of *Radio News*, advertising manager of the Wholesale Radio Service Company, Incorporated, and, most recently, executive editor of ***Mechanix Illustrated***.

Mr. Hertzberg was born on February 26, 1905, in New York, N. Y., and studied at the College of the City of New York, and Columbia University. He was commissioned in the Signal Corps Reserve for contributing to the development of the Army Amateur Radio System in the middle "20's," and was called to active duty in 1940. He served for more than five years in the United States and Europe, and after VJ Day was released to inactive status with the grade of colonel.

Mr. Hertzberg is a member of the Armed Forces Communications Association, the Quarter Century Wireless Association, and the New York Technical Section of the Photograph Society of America.

James Harold Weiner (M'50) has been promoted to the grade of Colonel, according to information received from the Air Defense Command Headquarters, Ent Air Force Base, Colorado Spring, Colo. He is now serving as Assistant Director of Communications and Electronics at ADC Headquarters.

Before entering the service in 1942, Colonel Weiner attended the Lowell Institute of Cambridge, in 1932–1934, studied at the Boston University until 1936, and was employed as a communications engineer for the New England Telephone and Telegraph Company.

Colonel Weiner has been awarded the Legion of Merit, the Army Commendation Ribbon, the American Defense Service Medal, and the American Campaign Medal. He is a native of Winthrop, Mass.

❖

Raymond S. Timm (A'47–M'50), formerly of the Programs Research Division, Office of Naval Research, has been named as electronics engineer at the Balco Research Laboratory, Newark, N. J.

A native of Michigan, Mr. Timm received his B.S. degree in electrical engineering in 1942, at the Lawrence Institute of Technology, and later did graduate work at the University of Maryland.

Books

Short Wave Wireless Communication by A. W. Ladner and C. R. Stoner

Published (1951) by John Wiley & Sons Inc., 440 Fourth Ave., New York 16, N. Y. 703 pages+8-page appendix+6-page index+xvi pages. 417 figures. 5½×8½. $8.00.

Mr. Ladner was formerly principal of the Marconi School of Wireless Communication and superintendent of instruction, Marconi's Wireless Telegraph Co., Ltd. Mr. Stoner, now a reader in electrical engineering at the University of London, was also formerly associated with Marconi's Wireless Telegraph Co., Ltd.

This is the fifth edition of a book which was first published in 1932. The present edition differs from the fourth by the deletion of three chapters, and substitution of three new chapters, which are of greater current interest. Brief summaries of the chapters of the text follow.

The introduction outlines the requirements of communications systems and includes some historical data. This is followed by a chapter on sound and vision signals, one of the new chapters, which discusses the physiological and mathematical characteristics of sound and vision signals. A statement on page 37 that, "there is no immediate prospect of the transmission of color pictures," might be open to question.

This is followed by a chapter on modulation, with consideration of noncarrier, carrier, suppressed carrier, phase, frequency, and various pulse methods. This is a rather unusual approach, since ordinarily, the subject of modulation follows a discussion of methods of producing waves. It is perhaps unfortunate that the authors retain the term "ether" in their discussion. Two chapters on the propagation of short and ultra-short waves follow, which are excellent summaries of the essential features of propagation phenomena, the ionosphere, and related phenomena. A chapter is then devoted to the general aspects of radio-frequency lines, with reference to impedance matching with stubs, tapered lines, and quarter-wave lines. The second of the new chapters gives a brief account in the field of waveguides, and this is followed by two chapters on antennas and antenna arrays, which begin with the concept of an antenna as a nonterminated line. This covers various phases of antenna design, including arrays.

The next five chapters, which are largely of a circuit nature, include tuned power amplifiers; oscillators, the frequencies of which are determined largely by circuit elements external to the tube, e.g., tank circuits, crystals, and RC networks; electron oscillators, which depend more directly on the electron beam, e.g., magnetron, klystron, Barkhausen-Kurtz, among others; modulation, amplitude, frequency, and pulse, and reception, with considerations of noise, selectivity, sensitivity, and the general characteristics of receivers for various types of modulated signals. The discussions of these sections are mainly theoretical, with frequent reference to schematic diagrams.

Following is a chapter dealing with commercial receivers which gives the specialized practices used in special types, and another chapter on radio telephone circuits discusses specialized commercial equipment, although it also outlines the requirements for single-sideband, suppressed-carrier, and other systems.

The third new chapter is on radio telegraph circuits, and the book then closes with a chapter on some typical commercial equipment. In general, the material dealing with principles is clearly expressed. The material which deals with detailed commercial practice seems to be somewhat out of place in a work intended as a text.

This book will be found useful by practicing engineers, and would serve as a useful reference book for students in their radio and communication courses. Its use as a text in such courses at the college level is seriously questioned.

Samuel Seely
Department of Electrical Engineering
Syracuse University
Syracuse, N. Y.

Electromagnetic Waves by F. W. C. White

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 103 pages+2-page index+2-page appendix+viii pages. 25 figures. 4¼×6½. $1.25.

The fourth edition of a monograph published originally in 1933, this book surveys briefly the fundamentals of electromagnetic theory, Maxwell's field equations, Lorentz's equations of motion, the theory of dispersion, and the propagation of radio waves.

It is written in strictly classical style similar to that of Jean's text, although much abbreviated, and has been revised only slightly to cover the results of studies carried out by Sir Charles Darwin into the influence of electromagnetic waves on the charged particles of materials through which they pass.

The text employs Gaussian units throughout. Vector analysis is used very sparingly. Very little work appearing in the literature since 1933 has been included.

The book may serve quite well as a survey text for readers interested in obtaining a birdseye view of problems and methods of attack in this field. The specialist, however, will not find it particularly helpful.

Lloyd T. DeVore
General Electric Co.
Syracuse, N. Y.

Theory and Design of Television Receivers by Sid Deutsch

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 521 pages+14-page index+xix pages. 376 figures. 6×9¼. $6.50.

Mr. Deutsch is an instructor in advanced television at the Madison Institute in Newark, N. J.

This book, the first of its type, attempts to meet the need of the practising television-design engineer and of the senior or graduate electrical engineer for a coverage of television-receiver design which includes supporting theoretical background

Books

For the most part, the author has done a very good job. Chapter 6, "Detector and Automatic Gain Control Circuits," Chapter 7, "Video Amplifiers and D-C restorers," and Chapter 9, "Synchronizing Section," are excellent in their completeness and clarity of exposition. Chapter 10, "Electrostatic Deflection and Focusing," Chapter 11, "Magnetic Deflection and Focusing," Chapter 12, "Power Supplies," Chapter 13, "Automatic Frequency Control," Chapter 15, "Direct- and Projection-Viewing Systems," are adequate in their treatment and coverage. The foregoing chapters are supported by Chapter 1, "Television Standards," and Chapter 16, "Television Receiver Servicing." The latter includes an excellent pictorial treatment of the effects of various receiver faults.

The author is weakest in his discussion of the rf and IF portions of the television receiver (Chapters 2, 3, 4, 5, 8, and 14). Some specific criticisms follow:

Chapter 2, page 57, is misleading in its discussion of transit-time versus lead-inductance loading. In general, for the two types discussed, cathode-lead inductance has a larger effect on input resistance than transit time and the limitation on the use of the 6AC7 is not loading of the previous stage but self-resonance of the input circuit which occurs at a considerably lower frequency.

In Chapter 3, under the discussion of IF gain on page 80, stability and maximum gain are not considered; this is a serious omission. In the same chapter, on pages 90 and 91, where series-tuned coupling circuits are described, the conclusion that the response curve has approximately the same shape as a single-tuned circuit is incorrect in that the effect of the shunt capacity on the inductance is omitted, an effect which produces a response curve with an "S" shape with a minimum at a higher frequency.

The discussion of antennas in Chapter 4 seems to this reader quite inadequate and cursory. Here on pages 150 and 160, the importance of maintaining a balance of each wire to the shield in a shielded pair is neglected and a rather loose statement is made concerning compensations for the unbalance of a coaxial line by rotating the antenna. On pages 119 to 123, there is a discussion of various antenna-input circuits. An unsupported statement is made that the width of the response curve at the image frequency will be approximately four times the IF bandwidth. This certainly does not hold for the multiple resonance of the semi-distributed circuits. The circuit diagrams all show blocking condensers in the antenna lead-in and do not show the requisite shunt-leakage path required for static protection. Also, the gain figures given for the examples are misleading and unduly high in that the maximum stable gain that can be obtained at 200 megacycles is substantially less. On page 126, under the discussion of "Local Oscillator Circuits" the statement is made that "The correct mode of oscillation is assured through the use of the Colpitts type of circuits." This is a bit of wishful thinking. In the discussion of the oscillator, no mention is made of the importance of high-C circuits in the minimizing of frequency drift, and this is another notable omission.

In Chapter 5, under "Video I-F Section," the entire discussion is of a 21-megacycle IF, and there is no mention of the RMA standard of 41.25-megacycle IF which has been standard for more than a year, and has been under discussion for the past three years. Another serious omission in the choice of the intermediate frequency is the lack of consideration of the second-order effects which have been so troublesome with present-day television receivers, particularly those employing 20-megacycle IF.

Likewise Chapter 8 on "The Sound Section," and Chapter 14, "Intercarrier Sound Reception," contain no mention of the problems introduced by a 41.25-megacycle IF. The discriminator circuit shown on page 271 is believed to be a Conrad circuit, and is incorrectly attributed to Travis. In the discussion of intercarrier-sound reception, the author seems unfairly prejudiced against the system, and fails to point out the improvement in reception which can be obtained with intercarrier-type receivers by tuning the picture signal to the top of the IF curve.

The author must be criticized, also, in a more general way, for the scarcity of reference material, and for his failure to indicate outside support for many of the statements made.

In summation, the author has attempted to cover a large field, and, on the whole, has done so in an excellent fashion.

JOHN D. REID
American Radio and Television, Inc.
North Little Rock, Ark.

Electronics by Jacob Millman and Samuel Seely

Published (1951) by McGraw-Hill Book Co., 330 W. 42 st. New York 18, N. Y. 559 pages+16-page index+22-page appendix+ix pages. 392 figures. 6×9¼. $7.25.

Mr. Millman is associate professor of electrical engineering at The City College of New York, and Mr. Seely is Chairman of the Department of Electrical Engineering at Syracuse University, Syracuse, N. Y.

This book, a second edition, is an improved and modernized version of an outstanding book which in its original form very successfully correlated physical electronics and engineering electronics.

Starting with the fundamental processes of physics, the authors squarely face the problems of electronics in sufficient detail to give the reader an excellent background. With this foundation, they describe the electronic behavior of many practical devices.

By concentrating primarily on the electronic aspects of these devices, they have covered a large field remarkably well. The problem of motion of charged particles in the electric and magnetic fields is carried on through applications to practical devices, such as the cathode-ray tube, the klystron, the betatron, and the FM cyclotron. Electron phenomena in metals is traced from the Fermi-Dirac-Sommerfeld function to contact potential, barrier layer rectifiers, and electron emission. The several mechanisms of electron emission, field emission, photoelectric emission, thermionic emission, and secondary emission are discussed, and applications are given to every important kind of commercial tube. Fundamental processes in gases are used to explain the behavior of practical gas tubes of all sizes, including the ignitron and excitron.

Although this book is primarily intended as a textbook to set the ground work for specialized courses in communications, electron-tubes, industrial electronics, etc., it would be a welcome addition to the reference shelf of most practicing engineers. Familiarity with the calculus and with elementary differential equations is assumed. The presentation is from the engineering viewpoint and presupposes a working knowledge of the basic principles of alternating-current theory, including elementary Fourier analysis.

This reviewer is favorably impressed by this book. By their clear presentation and agreeable style, the authors have produced an easily readable book which successfully bridges the seldom filled gap between fundamental theory and engineering applications.

R. R. LAW
RCA Laboratories Division
Princeton, N. J.

Principles of Electrical Engineering by W. H. Timbie and Vannevar Bush

Published (1951) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 609 pages+8-page index+8-page appendix+ix pages. 417 figures. 5¾×8¾. $6.50.

W. H. Timbie is Professor Emeritus, Electrical Engineering and Industrial Practice, Massachusetts Institute of Technology, Cambridge, Mass. Vannevar Bush is president of the Carnegie Institution, Washington, D. C.

This is the fourth edition of a text book for college student use which first appeared twenty-nine years ago. While the first edition treated all fundamental concepts from the viewpoint of the power engineer, the new edition approaches these fundamental concepts from the various viewpoints of the electrical engineers in many specialized fields of power, communication, or electronics. The problems included in each chapter likewise reflect this treatment. They are presented in such a way that they will appear up to date many years from now, even though the fields continue to advance at a rapid rate. In only a few instances does the text or the problems refer to values or equipment which has been superseded.

The new edition sets forth modern theories; advanced methods for analyzing and solving electric and magnetic problems and their relation to engineering circuit analysis. The information is presented in a manner which is easy to read and understand, and the exertion needed by a student to cover any assignment is thereby eased.

Books

It is interesting to note various warnings given as to probable errors due to improper use of terms, as in the case of the decibel. An attempt is made to avoid incomplete or colloquial engineering expressions even though some, such as "ohm per mil-foot," are in fairly common use. Yet it is noticed that no reference was made to "potentiometer," which commonly is used in the shortened form "pot," to indicate a variable resistor used as a voltage divider.

The Index is presumably adequate for its purpose, although this reviewer would prefer some expansion of it. The Appendix carries the usual tables for conductor sizes, current-carrying capacities, temperature constants, etc., as well as conversion factors. One wonders why the Appendix does not include some brief data as to time, place, and occupation of early pioneers such as Maxwell, Weber, Ampere, Volta, Gauss, and others whose names denote some electrical quantity, law, or characteristic. In some of the rewritten or new material in the text, an occasional attempt is made in this direction, such as: "It was discovered by Oerstad in 1820 . . . ," or, "Euler, writing in 1761 . . . ," but this does not occur frequently enough, and hence this might well be included in the Appendix.

Generally speaking, this book would not be purchased by the radio engineer although some of the older engineers, in some infrequent excursions outside of their specialized fields, may find helpful information, therein, not contained in the usual handbook because of its brevity.

ALOIS W. GRAF
135 S. La Salle Street
Chicago 3, Ill.

Elements of Television Systems by George E. Anner

Published (1951) by Prentice-Hall Inc., Publishers, 70 Fifth Ave., New York 11, N. Y. 771 pages+14-page index+11-page appendix+18-page problems +xii pages. 5½×8½. $10.35.

The author is an assistant professor of electrical engineering at the New York University, New York 53, N. Y.

A very complete coverage of the wide range of subjects related to television is effected in this book. The material presented includes not only rather detailed information on television systems and equipments as they exist today, but also contains some features which might properly be considered as belonging to the history of the development of television. Several outmoded procedures and methods of operations are described in addition to those which are standard at present. To the reader who is looking for information on present systems, these descriptions of outmoded systems seem to have a masking effect on the information sought. To one interested in the progress mankind has made in the last few years toward the realization of the long-dreamed-of near-instantaneous transmission of pictures, the historical approach makes interesting reading.

The book is well organized and most of the subjects are clearly presented. It would appear that the writer had in mind both readers who are looking for word explanations of physical principles, and those who understand more quickly these principles through the language of mathematics. The tendency is to lean somewhat heavily toward the mathematical treatment, although the physics of the processes are very adequately presented.

The method of presentation indicates rather clearly that the book has been written primarily for physicists and engineers rather than those who are interested in the very elementary aspects of the subject.

The first chapters of the book apparently have for their objective a complete description of the manner in which picture material is converted into electrical signals which can be used to reconstruct the original picture. The theory of scanning is quite adequately treated and references are given to authoritative sources of material. Camera tubes are included for detailed explanations.

The following chapters discuss the commercial telecasting system. Synchronizing pulses are treated at length, and the relations between transmitter and receiver are rather completely explained. Throughout the book the parts played by the FCC and co-operating industries are described.

The final chapter relates to color television. The three major contending systems, CBS, CTI, and RCA, are described. This chapter might be considered less complete than the sections devoted to the black-and-white systems, and perhaps less accurate in regard to three-color mixing processes, but the fundamental principles involved in each color-television system are adequately presented.

Overall, the book is an abundant source of rather detailed information on the television system as it has developed in the United States up to the present. The reader, on finishing the book, is apt to be more firmly convinced than ever that television is the most complicated form of communication yet to be commercialized.

JOHN M. BARSTOW
Bell Telephone Laboratories, Inc.
463 West Street
New York 14, N. Y.

Basic Electron Tubes by Donovan V. Geppert

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 325 pages+4-page index+2-page appendix. 257 figures. 6×9¼. $5.00.

Mr. Geppert is affiliated with the General Electric Co., Syracuse, N. Y.

This book has been written as an undergraduate text to introduce students to the basic principles of commercially available electron tubes. Microwave tubes have not been included since it is expected that the student will not have covered field theory at the time this subject is given. The reviewer feels, however, that some mention of high-frequency limitations on low-frequency triodes and tetrodes would be in order.

The chapter headings are: (1) High-vacuum and gas phototubes; (2) High-vacuum thermionic diodes; (3) High-vacuum triodes; (4) Tetrodes and pentrodes; (5) Beam power tetrodes; (6) Cathode ray tubes; (7) Glow-discharge tubes; (8) Thermionic gas diodes; (9) Thyratrons; (10) Mercury pool arc rectifiers; and (11) Ignitrons. The author introduces each of the types of tubes with illustrations pertaining to the class of the tubes. The circuit diagram for obtaining static characteristics follows, together with typical sets of static curves. A qualitative theory accounting for the observed characteristics is then followed by the mathematical equation defining the static curves. The simpler equations are derived. This approach of introducing the subject first with explanation later is novel; however, the author's handling of the subjects justifies his use of this method.

Readers may find the descriptive theory somewhat scanty, but the author has had to compress the subject into so little space that such criticism is inevitable. The book has very few technical errors, although some of the discussion on thoriated tungsten and oxide cathodes is out of date. For example, the maximum plate voltage at which thoriated tungsten filament tubes are used is no longer 4 or 5 kv. Also, the upper frequency limit on triodes is now at least 4,000 mc.

The figures are adequate and well done. Potential distributions, as obtained on a rubber sheet, are included to aid in explaining electron motion. A number of problems are presented in each chapter, and typical problems are solved in the text using both cgs and mks units.

Although the book is primarily an undergraduate text, it would be useful for the advanced engineer who wished to review the subject.

H. D. DOOLITTLE
Machlett Laboratories Inc.
Springdale, Connecticut

New Publications

Tables of the Bessel Functions of the First Kind of Orders Seventy-Nine Through One Hundred and Thirty Five, by the Staff of the Computation Laboratory of Harvard University, published by the Harvard University Press, 1951, $8.00. The tables in this volume were computed by the Automatic Sequence Controlled Calculator. This is the calculator, new in principle, which was presented in 1944 to Harvard University by the International Business Machines Corporation for operation by the Computation Laboratory.

These tables, carried to 10 decimal places, are valuable timesavers in the design departments of electrical and radio industries, in government, industrial, and business laboratories, and to scientists and mathematicians.

Mathematical Methods in Electrical Engineering by Myril B. Reed and Georgia B. Reed

Published (1951) by Harper & Bros., 49 E. 33 St., New York 16, N. Y. 332 pages+6-page index+viii pages. 91 figures. $6\frac{1}{4}\times9\frac{1}{4}$. $5.00.

Myril B. Reed is a Professor at the University of Illinois, Urbana, Ill.

The stated purpose of this book is to emphasize the manipulative aspects of mathematical disciplines applicable to engineering problems. The authors proceed with this by presenting a very large number of completely worked-out numerical examples. In most cases these examples do not give further insight into the methods, but rather present the routine operational aspects of a given mathematical technique. The book gives a standard treatment of the elementary phases of topics in algebra and analysis. The algebraic topics treated are: the algebra of complex numbers, linear equations and matrix algebra, partial and continued fractions, the Graeffe method for determining polynomial roots, and vector analysis. Other subjects treated are linear differential equations with an introduction to Laplace transforms, the partial differential equations of transmission lines, Bessel functions, and complex function theory. In every case only well-known preliminary aspects are examined and the treatment of the topics never deviates from the purely manipulative. Thus a reader will learn how to compute the roots of a polynomial by consulting the 21 pages of numerical examples devoted to this topic, but it is doubtful whether he will have any insight into the qualitative location of polynomial roots in the complex plane. Again the reader is given 13 pages of illustrative computation for solving linear algebraic equations, but basic considerations such as linear dependence are not presented.

The book may find some application as an undergraduate text since it has a large number of problems for class assignment; it adds little, however, to an already crowded field and cannot be heartily recommended over its numerous competitors.

HERBERT J. CARLIN
Microwave Research Institute,
Polytechnic Institute of Brooklyn
Brooklyn, N. Y.

Fundamentals of Electrical Engineering by Fred H. Pumphrey

Published (1951) by Prentice-Hall, Inc., 70 Fifth Ave., New York 11, N. Y. 662 pages+6-page index +xii pages. 460 figures. $5\frac{1}{4}\times8\frac{1}{4}$.

Fred H. Pumphrey is Head Professor, Electrical Engineering, Univ. of Florida, Gainesville, Fla.

Professor Pumphrey's new book is a considerably enlarged version of his earlier work entitled "Electrical Engineering," intended as a text for students of engineering who are specializing in fields other than electrical engineering. The book is purely descriptive, with much attention given to the practical aspects of the design, construction, and use of various electrical devices. The concise and very readable style, the numerous diagrams and photographs, and the carefully worked out examples combine to produce an understandable presentation of a subject which many engineering students consider most difficult.

The discussion of fundamental concepts takes up the first four chapters of the book. Here the author introduces the concepts of current, voltage, charge, resistance, and so forth; he also discusses Kirchoff's laws and their application to the analysis of simple dc circuits. The author's treatment of these topics, although very superficial by the standards of the electrical engineer, is probably adequate for an average student of engineering.

The next 11 chapters deal principally with ac circuit analysis and machinery. The treatment of machinery is quite good, but in this reviewer's opinion, the same cannot be said of the treatment of ac-circuit analysis. In particular, the author might well have avoided introducing the difficult concept of "vector operator."

The following six chapters cover electron tubes and their applications. Included here are descriptions of various types of vacuum tubes, amplifier and oscillator circuits, cathode followers, switching circuits, multivibrators, and various electronic instruments. Magnetic amplifiers are briefly discussed in Chapter 18.

The remainder of the book is given over to a survey of the principal applications of electricity in industry. The last chapter contains a brief discussion of typical communication systems and their components.

All in all, the author has succeeded rather well in the difficult task of presenting the fundamentals of electrical engineering in a manner that is comprehensible to the non-electrical student of engineering.

L. A. ZADEH
Columbia University
New York 27, N. Y.

Introduction to Servomechanisms by A. Porter

Published (1951) by John Wiley & Sons, Inc. 440 Fourth Ave., New York 16, N. Y. 151 pages+2-page index+vi pages. 70 figures. $6\frac{1}{2}\times4\frac{1}{4}$. $1.75.

An excellent little book, in which the maximum amount of information is crammed into a minimum of space. British and American sources are quoted and very well blended into the text. This is, however, a "First Course in Servoanalysis," and not by any means an "Introduction to Servomechanisms," a much-needed book which still remains to be written.

P. LE CORBEILLER
Harvard University
Cambridge, Mass.

Ferromagnetism by Richard M. Borzoth

Published (1951) by D. Van Nostrand Co., Inc., 250 Fifth Ave., New York, N. Y. 861 pages+20-page index+12-page appendix+71-page bibliography+xvii pages. 775 figures. 6×9. $17.50.

In this book the author has made the first attempt to summarize the knowledge relating to ferromagnetism since Becker and Doring's "Ferromagnetisms" (Springer, 1939); he has succeeded admirably, and his book will serve as a standard reference for those interested in the theory, development, or utilization of ferromagnetic materials.

The book is divided into two main parts: In the first, after a brief introduction to magnetic concepts, the author takes up in detail the properties of high-permeability materials starting with iron and its alloys, and proceeding through nickel, cobalt, manganese, and their alloys. Permanent magnet materials are considered next, and the details of manufacture and processing of commercially important alloys are given special attention. The author has striven to make the book of value to readers of widely different backgrounds; for example, he begins the chapter on the magnetic properties of pure iron with a brief review of the chemistry of an iron blast furnace, and finishes it with a discussion of the diffusion rates of impurities in iron. The treatment of such extraneous topics is consistently brief and pertinent, and serves to enhance the value of the book as a reference.

The second main part of the book presents a lucid account of magnetic phenomena and theories. Here also the author has taken into consideration the varying backgrounds of the readers. The emphasis is on the physical principles involved rather than on detailed mathematical analysis. This part of the book is concluded with a brief discussion of magnetic measurements.

Among the attractive features of the book are the large number of well-chosen illustrations (almost one per page), and the chronological bibliography of some 1,700 references.

B. H. ALEXANDER
Sylvania Electric Products Inc.
P.O. Box 6,
Bayside, L. I., N. Y.

New Publications

A new and readily understandable **Notebook on UHF Television and UHF-VHF Tunners,** generously illustrated with basic and comprehensive circuit schematics, block diagrams, and waveforms, has been announced by the Paul H. Wendel Publishing Co., Inc., of Indianapolis, Ind.

Prepared by Edward M. Noll, author of "Television for Radiomen," and "Color Television Notebook," this new publication describes the characteristics of vhf tuners with respect to gain, sensitivity, bandwidth, signal-to-noise ratio, various types of interference, and alignment. The vhf tuner section also details typical commercial tuners including RCA, Sarkes Tarzian, Standard Coil, Zenith, Stromberg-Carlson permeability tuner, Hallicrafter printed-circuit tuner, Philco tapered-line tuners, and General Instrument.

The section treating uhf television describes the nature of propagation, influence of the weather, and some of the results obtained by RCA with their experimental transmitter at Bridgeport, Conn., operating in the 529 to 535-megacycle region. Polycasting, uhf converter methods, uhf antennas, uhf converter circuits, and experimental RCA uhf tuners are also described.

The notebook contains a bibliography, a table of proposed channels for vhf-uhf television indicating frequency ranges, and a comprehensive tabulation of proposed allocations of vhf-uhf channels by cities and states. It measures $8\frac{1}{2}\times11$ inches, and is bound to lie flat when opened for study or reference. Copies may be obtained by sending $1.00 to the Paul H. Wendel Publishing Co., Inc., P.O. Box 1321, Indianapolis, 6, Ind.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London, S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

ACOUSTICS AND AUDIO FREQUENCIES

016:534 2608

References to Contemporary Papers on Acoustics—R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 484–489; July, 1951.) Continuation of 2309 of November.

534.213.4:534.373 2609

Boundary Layer Attenuation of Higher Order Modes in Rectangular and Circular Tubes—R. E. Beatty, Jr. (*Jour. Acous. Soc. Amer.*, vol. 23, p. 481; July, 1951.) Correction to paper noted in 1045 of June.

534.231 2610

On the Acoustical Radiation of an Emitter Vibrating Freely or in a Wall of Finite Dimensions—J. Pachner. (*Jour. Acous. Soc. Amer.*, vol. 23, p. 481; July, 1951.) Correction to paper noted in 1816 of September.

534.24 2611

Sound Scattering by Solid Cylinders and Spheres—J. J. Faran, Jr. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 405–418; July, 1951.) The theory of the scattering of plane sound waves by isotropic circular cylinders and spheres is extended to take into account shear waves, which can exist as well as compression waves within solid scattering bodies.

534.321.9:534.232 2612

Generation of Ultrasonic Oscillations by means of Volume Magnetostriction—H. H. Rust. (*Z. angew. Phys.*, vol. 3, pp. 9–14; January, 1951.) Description of an experimental generator. An excitation coil is enclosed in a vessel fitted at one end with a sound-transmitting plate and filled with a suspension of carbonyl/iron powder in insulating oil. The circuit of a pulse generator of this type is shown. Qualitative tests have proved satisfactory.

534.321.9:537.228.2 2613

Magnetostrictive Ultrasonic Apparatus—H. Thiede. (*Funk u. Ton*, vol. 5, pp. 32–42; January, 1951.) Illustrated description of the construction and applications of ultrasonic transducers for the frequency range 15 to 200 kc. The electroacoustic efficiency, resonance, and attenuation factors of different forms of oscillator are discussed, and the advantages of nickel as the magnetostrictive element are reviewed. The circuit of a tube-operated drive unit providing a modulated hf excitation voltage is shown.

534.373 2614

The Absorption of Sound in Fog—Y. Rocard. (*Rev. Sci.* (Paris), vol. 89, pp. 42–43; January to March, 1951.) Measured values of the absorption of sound in fog are very different from theoretical values, particularly at high frequencies. It is tentatively suggested that the discrepancy may be due to resonant vibration of drops of moisture excited by the sound wave.

534.373:534.213.4 2615

Wall Viscosity and Heat Conduction Losses in Rigid Tubes—R. F. Lambert. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 480–481; July, 1951.) Expressions are derived for the attenuation constant and Q of rectangular and round acoustic tubes for various modes of excitation.

534.64 2616

Principle of and Apparatus for Measurement of Sound Transmission through Partitions—M. Kobrynski and A. Neyron. (*Ann. Télécommun.*, vol. 6, pp. 34–42; February, 1951.) Description of a cro installation designed to record automatically the transmission coefficient of panels 1 m square, over the frequency range 50 to 10,000 cps. The test panel is fixed between two microphones at one end of a sealed echo-free chamber, and a uniform sound pressure is applied to the external face of the panel. Voltage proportional to the logarithm of the instantaneous frequency is applied to the timebase plates of the cro, while voltage proportional to the decibel difference at the two microphones, obtained by use of logarithmic-sensitivity amplifiers, is applied to the deflecting plates.

534.75 2617

The Frequency Selectivity of the Ear as determined by Masking Experiments—S. de Walden. (*Jour. Acous. Soc. Amer.*, vol. 23, p. 481; July, 1951.) Comment on 2692 of 1950.

534.75:621.39.001.11 2618

Information and the Human Ear—H. Jacobson. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 463–471; July, 1951.) Calculations of the capacity of the human ear to handle information are made by computing the number of sound patterns per second that can be discriminated, and applying the Shannon information theory. A maximum of 10^4 bits per second is found. This is compared with corresponding values for other sound channels, and with the observed rate of perception of information from speech and music. A capacity of 5×10^4 bits per second is necessary for high fidelity. The brain is able to use <1 per cent of the information transmitted by the ear.

534.78 2619

The Influence of Interaural Phase on Masked Thresholds. I. The Role of Interaural Time-Deviation—F. A. Webster. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 452–462; July, 1951.) The masking of a tone or narrow-band noise, the signal, by wide-band noise, is discussed with particular reference to the interaural phase differences for signal and noise, respectively. Audibility is best when the interaural phase difference for the signal is 180°, and that for the masking noise is zero. Experimental results are discussed in terms of the interaural phase- and time-deviations produced by the superposition of the signal on the noise.

534.78 2620

A Phoneme Detector—C. P. Smith. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 446–451; July, 1951.) Description of a device in which an incoming speech signal is compared with a set of stored energy-distribution reference patterns, and the best fit among the reference patterns is selected. Contiguous band filters covering the range 100 cps to 7 kc are used to dissect the incoming signal, and special arrangements are incorporated to provide improved selectivity and discrimination against noise.

534.851 2621

Intermodulation Distortion in Gramophone Pickups—S. Kelly. (*Wireless World*, vol. 57, pp. 256–259; July, 1951.) A special test record is used having negligible intermodulation components. The frequencies are 60 cps and 2,000 cps on one side, and 400 cps and 4,000 cps on the other. The output from a pickup playing this record is observed either on an intermodulation test set, or visually on a cro.

534.861:621.396.61 2622

Justification for raising the Standard of Broadcast Transmissions—M. Beurtheret. (*Onde élect.*, vol. 31, pp. 184–187; April, 1951.) Improvements in the quality of broadcast transmissions can improve reception for a large number of listeners. Conventionally accepted criteria of quality do not take adequate account of the worst faults, viz., frequency distortion due to intermodulation, and transient distortion. Use of a high degree of negative feedback in the transmitter is recommended to improve quality. Perfect linearity of the af amplitude/frequency characteristic of the transmitter is essential only in the range 50 to 2,500 cps; the production of high-quality broadcast transmitters is simplified if it is not attempted to extend the linear range unnecessarily.

621.395.61+534.321.9 2623

On the Possibility of a True Mechanical Transformer—Y. Rocard. (*Rev. Sci.* (Paris), vol. 88, pp. 236–238; October to December, 1950.) A rod, the cross section of which varies exponentially, has the properties of an impedance transformer for longitudinal, transverse, and torsional vibrations. Practical applications of the device in transducers, etc., are described.

621.395.616:534.612.4 2624

Condenser Microphone Sensitivity Measurement by Reactance Tube Null Method—H. E. von Gierke and W. W. von Wittern. (Proc. I.R.E., vol. 39, pp. 633–635; June, 1951.) The capacitance variations produced by energizing voltages in the frequency range 20 cps to 200 kc, are balanced against corresponding capacitance changes produced in a calibrated reactance-tube circuit. The capacitance variations are arranged to modulate the amplitude of a 3-mc carrier. Balance is observed by means of a cro to which the demodulated signal is applied.

621.395.623 2625

The Transmission Factor and Distortion Factor of an Electrodynamic Sound Receiver—G. Haar. (*Funk u. Ton*, vol. 5, pp. 17–26; January, 1951.) Description of measurement circuits, and procedure for determination of the characteristics, of a headphone designed for acoustic tests in the range 50 to 5,000 cps. Its advantages over a loudspeaker are outlined. Its low inherent distortion factor makes it especially useful for investigation of nonlinear distortion in an af system.

621.395.623.7 2626

Loudspeaker Diaphragm Control—J. Moir. (*Wireless World*, vol. 57, pp. 252–255; July, 1951.) Loudspeakers of current designs are critically damped by an amplifier having an output impedance >10 to 20 per cent of the dc voice-coil resistance. The use of loud speakers in small rooms prevents any advantge being obtained from over-critical damping. The use of an exponential horn is recommended to increase damping of the cone oscillation at high frequencies, and to flatten the voice-coil impedance curve.

621.395.623.7:534.13 2627

Growth of Subharmonic Oscillations—W.J.Cunningham. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 418–422; July, 1951.) The conditions necessary for the excitation in a resonant system of oscillations at half the driving frequency are discussed. An experiment is described, using a suitably energized lc tuned circuit. The mode of production of subharmonic vibrations in a direct-radiator loudspeaker is also discussed, and experimental data are given.

621.395.623.7:534.231 2628

Sound-Field Regulator—P. Riety. (*Ann. Télécommun.*, vol. 6, pp. 43–48; February, 1951.) Description of a system providing a sound field of constant strength over the range 20 cps to 15 kc. The input to the loudspeaker amplifier is fed through a regulating circuit to which is applied a control voltage corresponding to the emf developed by a "pilot" microphone in the sound field of the loudspeaker, so that the input to the latter is inversely proportional to its acoustic efficiency. The time lag of the regulation is discussed. Amplification range is >30 db.

621.395.625.3 2629

A Professional Magnetic-Recording System for Use with 35-, 17½- and 16-mm Films—G. R. Crane, J. G. Frayne, and E. W. Templin. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 295–309; discussion, p. 309; March, 1951.) Description of portable equipment for producing a high-quality sound track for motion pictures.

621.395.625.3 2630

A.C. Magnetic Erase Heads—M. Rettinger. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 407–410; April, 1951.) Various ring-shaped heads for erasing magnetic records are compared. Measurements were made of the degree of erasure obtained, and the temperature rise of the head as the 70-kc erase current was varied. Best results were obtained with two heads in cascade.

621.395.625.3 2631

A German Magnetic Sound Recording System in Motion Pictures—M. Ulner. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 411–422; April, 1951.) For economy, magnetic tape (6.5 mm) is used for all original sound records in German film studios, film being used only in re-recording on the final negative. A description is given of the magnetic recorder and play-back amplifier, and of the method of synchronization with the picture.

621.396.645.029.3 2632

Extended Class-A Audio—Sterling. (*See* 2680.)

ANTENNAS AND TRANSMISSION LINES

621.315.212 2633

The Behaviour of Curved Uniform Lines at High Frequencies—H. H. Meinke. (*Arch. elekt. Übertragung*, vol. 5, pp. 106–112; March, 1951.) The case of a line bent into an arc of a circle is considered. An equation is derived, giving the change of characteristic impedance introduced for a line of arbitrary cross section. A formula giving a very close approximation is derived for the coaxial line, and some other cross sections are discussed. The effective length of the line is given by the circumference of a circle whose radius is equal to the mean radius of curvature as particularly defined. For the sake of simplicity, the discussion is restricted to cases in which axial field components may be neglected.

621.315.212:621.3.09 2634

The Effect of Connectors on the Properties of a Cable—M. Cotte. (*Câbles & Trans.* (Paris), vol. 5, pp. 84–87; January, 1951.) The effect on the propagation characteristics of discontinuities introduced by connectors between lengths of cable, is discussed. The levels of parasitic echo signals so introduced are calculated for transients and for steady-state conditions. Phase distortion due to differences in group velocities is estimated. To keep these distortion effects within given limits, the product of the reflection coefficient between a cable length and a connector, multiplied by the length of the connector, must be smaller than a predetermined value. In practice, with typical cables and connectors, it is found that echo-effect distortion is more important than phase distortion. The requirements for good transmission are met by correctly designing the connectors.

621.315.212:621.392.43 2635

Coaxial-Stub Filter—J. A. Craig. (*Electronics*, vol. 24, pp. 132, 134; June, 1951.) Describes the use of correcting stubs for coaxial transmission lines.

621.392.09 2636

Single-Conductor Surface-Wave Transmission Lines—G. Goubau. (Proc. I.R.E., vol. 39, pp. 619–624; June, 1951.) General information and design data relating particularly to the dielectric-coated conductor. See also 281 of March and 812 of April.

621.392.21 2637

Note on the Variations of Phase Velocity in Continuously-Wound Delay Lines at High Frequencies—I. A. D. Lewis. (*Proc. IEE*, vol. 98, pp. 312–314; July, 1951.) "The effect of the self-capacitance of the winding on the velocity and on the characteristic impedance is here determined by a simple method. A direct treatment is given of the decrease in the inductance per unit length which occurs when the wavelength along the coil is not very large compared with the winding diameter. The two effects have opposite influences on the velocity, and a fair degree of cancellation can be attained if a certain relationship between the parameters is satisfied."

621.392.26† 2638

On the Representation of the Electric and Magnetic Fields produced by Currents and Discontinuities in Wave Guides—N. Marcuvitz and J. Schwinger. (*Jour. Appl. Phys.*, vol. 22, pp. 806–819; June, 1951.) The calculation of the fields, produced by prescribed and induced currents at a discontinuity, is treated by representing the fields in terms of a complete set of vector modes characteristic of the possible transverse field distributions in the waveguide cross section. This representation transforms the field problem into one-dimensional modal problems of conventional transmission-line form. The eigenvalue problem of finding the characteristic modes is discussed in detail for the case of a uniform guide with perfectly conducting walls. A typical modal analysis and synthesis are presented for infinite and semi-infinite waveguides of arbitrary cross section.

621.392.26†:621.39.09 2639

The Significance of the Transmission-Line Equations and Characteristic Impedance for Waves of Arbitrary Type in Cylindrical Lines—F. Borgnis. (*Arch. elekt. Übertragung*, vol. 5, pp. 181–189; April, 1951.) The familiar transmission-line equations expressing the spatial distribution of current and voltage along a Lecher line, apply also to the propagation of an arbitrary electric wave in a cylindrical line (e.g., a waveguide) provided it is possible to excite a single electric or magnetic mode in the line. The special interpretation to be attached to the concepts "current," "voltage," and "characteristic impedance" in this case, is discussed.

621.392.26†:621.39.09 2640

The Velocity of Propagation of H_{0n} Waves in a Multilayer Waveguide—N. N. Malov. (*Zh. Tech. Fiz.*, vol. 20, pp. 1509–1510; December, 1950.) A discussion of the possibility of using waveguides filled with two or three layers of dielectrics as phase changers, with comments on a paper by Fox (1255 of 1948).

621.392.26†:621.396.67 2641

Application of Array Theory to Nonresonant Slotted Waveguides—J. B. Tricaud. (*Onde élect.*, vol. 31, pp. 122–132 and 188–200; April, 1951.) The directive properties of halfwave dipole arrays, fed by a waveguide, are compared with those of a slotted guide. The slotted guide is discussed in detail as a particular class of linear array. Schelkunoff's space-factor theory is developed to give the amplitude-distribution function. Approximate sinusoidal solutions are worked out, and the usefulness of the method, permitting the expression of results as asymptotes, is emphasized with examples. The theory is applied to consideration of the radiation from a single slot, the matching of impedance along a slotted guide, and the relation between amplitude distribution and load. The characteristics of a typical low-radiating-angle radar antenna, using a nonresonant slotted guide, are given.

621.396.67 2642

U.R.S.I.—I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951.—(Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) Summaries are given of the following papers:

15—Low-Frequency Antennas—P. S. Carter.
16—Antenna Systems for Radio Direction Finding—E. C. Jordan.
17—Electrically Small Antennas and the Low-Frequency Airborne Antenna Problem—J. T. Bolljahn.
18—Antennas in Conducting Media—R. K. Moore.
29—On the Theory of Antenna Beam-Shaping—A. S. Dunbar.
30—Beam Shaping in Doubly Curved Reflector Systems employing Quasi-Point Sources—A. E. Marston.
31—Lack of Uniqueness in Antenna Pattern Synthesis Methods and the Related Energy Storage Considerations—T. T. Taylor.
32—Feed Problems in Broad-Band Antenna Arrays—W. R. LePage and R. F. Gates.
33—Second-Order Beams of Slotted Waveguide Arrays—H. Gruenberg.
42—Aperture Phase Errors in Microwave Optics—K. S. Kelleher.
43—Review of Spherical Reflector Research at A.F.C.R.L.—J. E. Walsh.
44—Review of Recent Metal Lens Research at A.F.C.R.L.—J. Ruze.

621.396.67 2643
Wrotham Aerial System: Part 2—The Main Transmission Line and the A.M./F.M. Combining Filter—C. Gillam. (*Wireless World*, vol. 57, pp. 279–282; July, 1951.) The transmission line which conveys rf power to the multiple-slot radiator and distribution system (see Part 1: 2340 of November) is described. FM and AM transmissions can be radiated simultaneously from the single antenna system.

621.396.67 2644
Improved Anti-Fade Transmitting Aerial—H. Brückmann. (*Funk u. Ton*, vol. 5, pp. 5–16; January, 1951.) Full paper of which a shorter version in English was abstracted in 1863 of 1950.

621.396.67 2645
Impedance Characteristics of Thick Cylinders—S. Zisler. (*Onde élect.*, vol. 31, pp. 133–143; March, 1951.) The input impedance of cylindrical antennas, and in particular of thick cylinders, is calculated by an extension of two-wire transmission line theory. The equivalent coaxial line is assumed to have damping resistance and reactance uniformly distributed along its length, to account for power loss due to radiation from the antenna, and to provide the same current distribution in the line as is found in the antenna. Results are quoted, with curves and diagrams, for cylindrical $\lambda/4$ and $\lambda/2$ antennas for a length-to-diameter ratio between 2 and 50.

621.396.67 2646
Aperiodic Aerials: Use with Vertical-Incidence Ionospheric Recorders—R. Bailey. (*Wireless Eng.*, vol. 28, pp. 208–214; July, 1951.) An analysis is made of the performance of several types of simple resistance-terminated traveling-wave antennas, of practical height and shape, for transmission and reception in a vertical direction above a perfectly flat earth. The results are used to find the conditions for best signal-to-interference ratio for ionospheric echoes between about 0.5 and 20 mc, using the minimum number of antennas to cover the whole range. The results may also have application to short-distance sky-wave radio communication.

621.396.67.014.1 2647
Current Distributions on Helical Antennas—J. A. Marsh. (Proc. I.R.E., vol. 39, pp. 668–675; June, 1951.) Measurements were made over the frequency range 600 to 1,700 mc on two uniform circular helices, of diameter 8.61 cm, pitch angle 12.6°, and different lengths, using a sampling probe. The distributions of current amplitude and phase are shown graphically, and a satisfactory interpretation is given in terms of three traveling-wave modes. The relative phase velocities in the modes are deduced. When the T_1 mode predominates, there is good agreement between measured and calculated velocities and the value required for optimum end-fire directivity.

621.396.671+621.396.11 2648
Application of the Compensation Theorem to Certain Radiation and Propagation Problems—G. D. Monteath. (*Proc. IEE*, vol. 98, pp. 319–320; July, 1951.) Digest of IEE monograph. The compensation theorem is extended to deal with a continuous system in which a solution of Maxwell's equations must satisfy certain boundary conditions. In particular, the change in mutual impedance between two antennas is considered when changes are made to other parts of the system, e.g., the surface of the earth. The result is expressed in terms of the change in surface impedance.

621.396.671 2649
The Patterns of Antennas Located near Cylinders of Elliptical Cross Section—G. Sinclair. (Proc. I.R.E., vol. 39, pp. 660–668; June, 1951.) The patterns of small dipole and loop antennas mounted on or near a perfectly conducting cylinder of elliptical cross section, are calculated by determining the open-circuit terminal voltages of the antennas when receiving plane waves. The antennas are assumed to be such that they cause negligible distortion of the field when they are open circuited, thus reducing the problem to that of calculating the diffraction of a plane wave around the cylinder. The results have applications in the design of aircraft antennas and slotted-cylinder antennas.

621.396.671:778.3 2650
Photo Radiation Patterns—G. W. Goebel. (*Electronics*, vol. 24, p. 89; May, 1951.) A simple and economical method is described for exhibiting two-dimensional patterns of two or more interfering radiators of the same frequency and polarization.

621.396.676.029.6 2651
A Very-High-Frequency—Ultra-High-Frequency Tail-Cap Antenna—L. E. Raburn. (Proc. I.R.E., vol. 39, pp. 656–659; June, 1951.) An account of the development of an omnidirectional zero-drag aircraft antenna for vertical polarization. By insulating the vertical stabilizer from the remainder of the tail structure, a suitable radiator was obtained, covering the frequency band 100 to 400 mc. The principal radiation patterns are illustrated, and some details of mechanical design are included.

621.396.677 2652
Wide-Angle Metal-Plate Optics—J. Ruze. (Proc. I.R.E., vol. 39, p. 697; June, 1951.) Correction to paper abstracted in 1345 of 1950.

621.396.677 2653
Factors affecting Performance of Directional Antennas—A. E. Cullum, Jr. (*Broadcast News*, no. 63, pp. 43–51; March/April, 1951.) The horizontal and vertical radiation patterns for various phase and space relations of a two-element antenna show the distribution of the radiation between sky and ground waves. The influence of the shape and bonding of the towers, guy layout, and insulation and ground screens on the radiation pattern, is considered. Methods of sampling the radiated fields are given.

621.396.677.3 2654
On the Response of a Directive Antenna to Incoherent Radiation—F. W. Schott. (Proc. I.R.E., vol. 39, pp. 677–680; June, 1951.) A general expression is derived for the power gain of a directive antenna over a nondirective antenna, when receiving radiation comprising components with various directions of arrival and random phases. The usual method of expressing signal strength in db relative to a particular μv/m level may give rise to error in this case; only the received power can be stated. Results obtained by measurement with parabolic antennas at 9.375 kmc show that for widely scattered radiation the antenna aperture has little effect on received power.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.7 2655
Improved Pulse Stretcher—J. F. Craib. (*Electronics*, vol. 24, pp. 129–131; June, 1951.) A delay line, charged by crystal rectifiers in parallel, is used to stretch pulses up to 25 times their original length. The effects produced by superposing pulses within the stretch time are discussed. Design equations are presented.

621.316.729 2656
Synchronization of Relaxation Oscillations at Fundamental and Subharmonic Frequencies of an Applied Electromotive Force: Part 4—V. V. Vitkevich. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1245–1256; October, 1950.) In continuation of previous papers (see *Elektrosvyaz*, no. 11, 1940; *Zh. Tekh. Fiz.*, vol. 14, nos. 1/2, 1944; and vol. 15, no. 11, 1945), a general case is considered in which several electromotive forces are applied to the oscillator and, in addition, the critical voltage value of the nonlinear element is affected by these forces. Formulas are derived for determining the amplitudes and synchronization bandwidths at the fundamental and subharmonic frequencies of the complex applied emf, and a general theory of the synchronization of asymmetrical and symmetrical oscillators is presented. A geometrical interpretation of the conclusions is given.

621.316.8 2657
The Problem of a Non-ohmic Resistor in Series with an Impedance—E. B. Moullin. (*Proc. IEE*, vol. 98, pp. 87–96; March, 1951.) An experimental determination of the rms voltage required to maintain a given current through a silicon-carbide resistor in series with an ohmic resistor or a reactor. An appendix shows how the observed results can be predicted from the static characteristic by simple algebraic methods. The observed waveforms are also discussed.

621.318.4.042.13 2658
Cylindrical-Coil Theory applied to Calculation of Reluctance or Resistance between the End of a Straight Core and Infinity—P. M. Prache and R. Cazenave. (*Câbles & Trans.* (Paris), vol. 5, pp. 60–67; January, 1951.) A theoretical study is made of the reluctance along a tube of force through an air-cored cylindrical coil. An expression for the reluctance between the end of the coil and infinity is obtained as the difference between the total reluctance of the tube of force and that of its part inside the coil. Its value is nearly independent of the flux distribution within the winding. The expression is useful in the calculation of discontinuous magnetic circuits. The results obtained for reluctance may be translated into terms of resistance for estimating the value of resistance between the end of a cylindrical body and infinity.

621.318.42 2659
Design of Input (Regulation Control) Chokes—N. H. Crowhurst. (*Electronic Eng.* (London), vol. 23, pp. 179–181; May, 1951.) A design chart is presented for determining the optimum specification for a choke to cover a given range of current and output voltage. It is intended to be used in conjunction with the charts noted in 1096 of June.

621.318.572 2660
The Cascaded Binary Counter with Feedback—G. F. Montgomery. (*Jour. Appl. Phys.*, vol. 22, pp. 780–782; June, 1951.) A cascade of binary counters having arbitrary feedback arrangements introducing additional pulses, is analyzed. For a given count, or scale, a feedback arrangement requiring a minimum number of connections can be found.

621.318.572:621.385.3:546.289 2661
Some Transistor Trigger Circuits—P. M. Schultheiss and H. J. Reich. (PROC. I.R.E., vol. 39, pp. 627–632; June, 1951.) A survey of possibilities. Each of five circuits shown has operated successfully with at least one transistor. Difficulties encountered in reproducing results with other transistors or altered parameters are discussed.

621.385.3:546.289 2662
Duality, a New Approach to Transistor Circuit Design—R. L. Wallace, Jr. (PROC. I.R.E., vol. 39, p. 702; June, 1951.) Outline of principles applied. See 2369 of November.

621.385.3:621.385.5 2663
Valve Operating Conditions—(*Wireless Eng.*, vol. 28, pp. 195–196; July, 1951.) A triode with capacitive load and without feedback, may be cut off by a negative-going input voltage which can be amplified when no capacitance is present. The operation of triodes and pentodes under these conditions is discussed, and the case of an inductively-loaded pentode is also considered.

621.392 2664
A Network Theorem and its Application—F. W. Bubb, Jr. (PROC. I.R.E., vol. 39, pp. 685–688; June, 1951.) The theorem here proved is useful in analyzing the response of an hf circuit to an am signal by relating it to the response of an equivalent lf circuit to the signal envelope. The cases of a single-tuned circuit and a two-stage stagger-tuned amplifier are calculated as examples.

621.392.5 2665
Generalized Impedance Circle Diagrams in the Analysis of Coupled Networks—C. Ghosh. (*Indian Jour. Phys.*, vol. 24, pp. 223–231; May, 1950.) A practical method of developing circle diagrams for a representative "T" network is described for cases in which the circuit parameters vary.

621.392.5 2666
Generalized Representation of a Valve Circuit by an Iterative Matrix—U. Kirschner. (*Arch. elekt. Übertragung*, vol. 5, pp. 190–196; April, 1951.) The circuit is considered as an active quadripole, and matrix methods are developed, enabling its performance to be predicted. In illustration, the calculation is carried out for four different pentode circuits. See also 837 of May.

621.392.5 2667
Constant-Resistance Networks of the Linear Varying-Parameter Type—L. A. Zadeh. (PROC. I.R.E., vol. 39, pp. 688–691; June, 1951.) Explicit expressions are derived for the transfer function of such networks. See 834 of May.

621.392.5 2668
Initial Conditions in Linear Varying-Parameter Systems—L. A. Zadeh. (*Jour. Appl. Phys.*, vol. 22, pp. 782–786; June, 1951.) The response of the initially excited system to a given input is considered. Mathematically, the problem involves the solution of a linear differential equation, with time-dependent coefficients, subject to prescribed initial conditions. These conditions may be satisfied by superposing upon the given input a linear combination of delta functions, and treating the system as if it were initially at rest. Using the concept of a system function, a simple general expression for the response is developed. The result is similar in form to that obtained by the use of conventional Laplace transformation techniques in the case of a linear differential equation with constant coefficients.

621.392.52 2669
Input Admittance, Output Admittance and Voltage Transformation at the Centre of the Pass Band in Variable Three-Stage Filters—W. Pfost. (*Arch. elekt. Übertragung*, vol. 5, pp. 77–80; February, 1951.) General theory for the electrical design and construction of these filters was given in 843 of 1950. Formulas are now derived for the input and output admittances, and for the voltage transformation ratio at the center of the pass band. The values found are shown graphically and discussed. As required by the theory, the amplification is independent of direction of operation, and is nearly independent of bandwidth.

621.392.52 2670
The Double-T RC Filter—J. Thouzery. (*Radio franç.*, nos. 1–4, pp. 6–11, 17–20, 20–22 and 19–22; January to April, 1951.) The theory of the double-T filter is discussed at length, after a preface on tensor analysis. The relation between the ratio volts-input to volts-output, and the constants of the network, i.e., the transfer equation, is investigated for the general case and for practical parameters. Expressions are derived for the resonance frequency of the network for both general and particular cases, and frequency response curves are shown. Practical examples are discussed to illustrate applications of the theory.

621.392.6 2671
Representation of a Network or Transmission System by a Number of Superposed Balanced Circuits—L. J. Collet. (*Câbles & Trans.* (Paris), vol. 5, pp. 3–24; January, 1951.) Following on an earlier paper on superposed circuits (592 of April), a generalized treatment is given of the problem of balanced ac circuits. In any network, those circuits which can be defined by real coefficients are designated as real. In this sense it is shown that a network having a number of terminals exceeding by n the number of its component loops, can always be resolved into n real and mutually balanced circuits. Hence, a network including two sets of terminals each characterized by the same number n, can generally be resolved into n electrically independent quadripoles. Thus a transmission line constituted by n wires and the ground can be resolved into n balanced circuits.

621.396.6:061.4 2672
1951 Components Exhibition, Paris—J. Rousseau. (*T.S.F. pour Tous*, vol. 27, pp. 100–102 and 162–167; March and April, 1951.) Short illustrated review of exhibits. For other accounts see *Radio prof.* (Paris), vol. 20, pp. 16–19, February, 1951; *Télévision*, no. 12, pp. 75–78, March/April, 1951; *Radio franç.*, no. 2, pp. 13–15, February, 1951, and *Toute la Radio*, no. 154, pp. 99–107, March/April, 1951.

621.396.611.1 2673
The Influence of Initial Phase Angle on the Excitation of Subharmonic Oscillations in a Nonlinear Circuit—G. J. Elias, S. Duinker, and Tan Soen Hong. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 69–84; March, 1951.) Report of an experimental investigation of oscillations excited in a circuit comprising capacitor and iron-cored inductor by a sinusoidal-voltage input. Graphs are used to show the initial values of phase and amplitude for which relaxation and subharmonic oscillations occur, for various values of capacitance. The effect on subharmonic response of additional resistance is considered. The phase-angle control-switch circuit used is described in an appendix.

621.396.611.3 2674
The Pull-In Effect in Coupled Circuits—W. Herzog. (*Arch. elekt. Übertragung*, vol. 5, pp. 81–84; February, 1951.) To provide a physical picture of the pull-in effect, the simple case is considered of a bridge oscillator with two resonator circuits in parallel, the capacitance in one circuit being variable. The reactances of the two circuits are plotted on the same graph against frequency, and the dependence of the oscillation on the difference between the two reactances is demonstrated.

621.396.615 2675
Oscillators—W. Herzog. (*Arch. elekt. Übertragung*, vol. 5, pp. 169–180; April, 1951.) A comparative survey is given of the most useful oscillator circuits. Formulas determining amplitudes and frequencies are established.

621.396.619.13.018.78† 2676
Distortion in Linear Passive Networks—L. Riebman. (PROC. I.R.E., vol. 39, pp. 692–697; June, 1951.) A development of the theory of distortion in an FM system, based on a series solution of the superposition integral, provides a method of calculating the distortion produced by practical receiver circuits when the circuit transfer function is known in pole-zero form.

621.396.622.63 2677
The Diode Rectifier—D. Geist. (*Z. angew. Phys.*, vol. 3, pp. 32–35; January, 1951.) A theoretical treatment of the crystal-diode-capacitor circuit. The current-voltage characteristic is taken as linear, with different slopes in the forward and reverse regions, a finite value of resistance being assumed in the reverse region. The time constant, efficiency, etc., can be calculated from the formulas derived. In the special case of infinitely high reverse resistance, the formulas apply also to the tube rectifier.

621.396.645.018.422† 2678
Simplified Q Multiplier—H. E. Harris. (*Electronics*, vol. 24, pp. 130–132, 134; May, 1951.) A narrow-band amplifier incorporating a negative-resistance circuit and having high stability and ease of control, is described. The parallel-tuned circuit whose Q is to be increased, is connected to the grid of a cathode follower. The cathode is connected via a variable resistance to a tap on the tuned circuit. A stable Q of 30,000 can be obtained.

621.396.645.018.424† 2679
Notes on Wide-Band Amplifiers—L. Bernardi. (*Poste e Telecomun.*, vol. 18, pp. 261–274; July, 1950.) Practical design data for amplifiers used in carrier-frequency telephony or television are presented. Phase shift and amplification loss in voltage amplifiers at low and high frequencies are considered, and suitable compensating networks discussed. At hf, various two- and four-terminal networks are available. Cathode coupling of power amplifiers to coaxial cable is considered. A worked example illustrates the method.

621.396.645.029.3 2680
Extended Class-A Audio—H. T. Sterling. (*Electronics*, vol. 24, pp. 101–103; May, 1951.) Description of a push-pull af power amplifier, using a triode and tetrode connected in parallel in each half, and delivering nearly 50 w from four Type 807 tubes. At low levels, the tetrodes are cut off, and class-A triode operation is obtained with low distortion and power consumption. At high levels, the tetrodes conduct and deliver the full output required.

621.396.645.029.3 2681
Amplifier of Variable Output Impedance—R. Yorke and K. R. McLachlan. (*Wireless Eng.*, vol. 28, pp. 222–225; July, 1951.) Positive and negative feedback are used to provide an output impedance variable from −10 to +20Ω. The amplifier characteristics were investigated experimentally, and the principal results are outlined.

621.396.645.215 2682
Improvement of the Frequency Response of Output Transformers by use of Filter Coupling—R. Seidelbach. (*Funk u. Ton*, vol. 5, pp. 27–31; January, 1951.) The leakage inductance of the transformer is incorporated in a π-type filter, and a selective L-type network is coupled directly to this The system provides a high-frequency cutoff at a frequency below 25 kc, and is useful in an anode-modulation circuit in which frequencies of 25 to 60 kc give

rise to distortion. Frequency attenuation figures are tabulated for a typical set of component values.

621.396.645.3 2683
Maximum Permissible Value of Grid-Leak Resistance—O. Schmid. (*Arch. elekt. Übertragung*, vol. 5, pp. 85–88; February, 1951.) The value of the grid-leak resistance for an amplifier tube is required to be smaller the nearer the tube is worked to the zero-grid-voltage point, and the longer the grid remains at low values of negative potential. Hence, care is needed in fixing the value of this resistance for dc, af, and pulse amplifiers, in order to avoid nonlinear distortion. Where the grid voltage is changing rapidly, there is no danger of distortion if the value of the grid-leak resistance is fixed according to the mean values of grid current and grid voltage.

621.396.645.35 2684
D.C. Amplifiers—J. Schérer. (*Toute la Radio*, no. 154, pp. 93–96; March/April, 1951.) Traces the evolution of dc amplifiers, and indicates modern practice.

GENERAL PHYSICS

537.311.5+538.569.4 2685
U.R.S.I.-I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951—(PROC. I.R.E., vol. 39, pp. 716–720; June, 1951.) Summaries are given of the following papers:

8—Validity and Limitations of the Van Vleck-Weisskopf Equation for Atmospheric Microwave Absorption—T. F. Rogers.
41—Current Distributions on Large Reflecting Cylinders—H. J. Riblet.

537.523.3+537.525.3 2686
Starting Potentials of Positive and Negative Coronas with Coaxial Geometry in Pure N_2, Pure O_2, and Various Mixtures at Pressures from Atmospheric to 27 mm—C. G. Miller and L. B. Loeb. (*Jour. Appl. Phys.*, vol. 22, pp. 740–741; June, 1951.)

537.525.6:538.56]:621.396.615.142 2687
Electron Plasma Oscillations—Wehner. (*See* 2875.)

537.528:621.315.61 2688
The Formation of Conducting Bridges in Suspensions of Conductors and Semiconductors in Dielectrics: Part 2—L. G. Gindin, L. M. Moroz, I. N. Putilova, and Ya. I. Frenkel. (*Zh. Tekh. Fiz.*, vol. 21, pp. 143–148; February, 1951.) The behavior of 0.1 per cent suspensions of Al in benzene under strong electric fields was investigated, and, in particular, the formation of conducting "bridges" made up of Al particles between the electrodes. The effect of various factors on the formation of these bridges is discussed.

538.221 2689
Ferromagnetism: Magnetization Curves—E. C. Stoner. (*Rep. Progr. Phys.*, vol. 13, pp. 83–179; bibliography, pp. 180–183; 1950.) In this report, which is complementary to that abstracted in 3395 of 1948, an account is given of experimental and theoretical work over a period of about 16 years, on the behavior of ferromagnetic materials in low and medium fields. The complex of behavior represented by the magnetization curves of ferromagnetics is regarded as arising from small perturbations. The investigations selected for consideration are those providing quantitative information about these elementary processes, or exemplifying general methods of attacking the problems or illustrating general principles.

538.3:530.145 2690
The Reciprocity Theory of Electrodynamics—H. S. Green and K. C. Cheng. (*Proc. Roy. Soc. Edinb. A*, vol. 63, part 2, pp. 105–138; 1949/1950.) Application of the principle of reciprocity to problems of classical and quantum electrodynamics. The first step is the formulation of a reciprocally invariant Lagrangian function for a system of electrons in interaction with the em field. A study of the unaccelerated motion of an electron is extended to the case of arbitrary motion. The Hamiltonian energy of electron and field is determined, and this makes it possible to express the theory in a quantum form which avoids the usual divergence difficulties.

538.311:513.647.1:621.318.423 2691
Waves Guided by Helical Circuits—É. Roubine. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1748–1750; May 7, 1951.) A more general analysis than that presented in 1350 of July, applicable to any circuit whose limiting surface is generated by the helical motion of a curve. The electric field may be expressed in terms of a function Φ, which is the solution of an equation bearing to the Goldstein equation the same relation as the wave equation bears to the Laplace equation. The theory is applicable to (a) delay lines, (b) traveling-wave tubes, and (c) helical antennas.

538.56:535.13 2692
On the Nonspecular Reflection of Electromagnetic Waves—V. Twersky. (*Jour. Appl. Phys.*, vol. 22, pp. 825–835; June, 1951.) The nonspecular reflection of plane electromagnetic waves of arbitrary polarization by perfectly conducting surfaces, composed of either semi-cylindrical or hemispherical bosses on an infinite plane, is analyzed. Solutions for the problem of the single boss are given and extended to cover patterns of bosses. The results for the various cases are compared, and expressions are obtained for the ratios of the components of the reflected wave.

538.56:535.42 2693
Diffraction of Centimetre Electromagnetic Waves by Metal and Dielectric Disks—H. Severin and W. V. Baeckmann. (*Z. angew. Phys.*, vol. 3, pp. 22–28; January, 1951.) The field distribution in the axial region of a metal disk was investigated experimentally. Methods using a fixed disk and a movable disk led to similar results. The latter method, applied in the case of a dielectric disk, showed that the diffracted wave differs from that for a metal disk only by a constant factor which is identical with the optical reflection coefficient.

538.56:535.42 2694
Theory of the Diffraction of Electromagnetic Waves—H. Severin. (*Z. Phys.*, vol. 129, pp. 426–439; April 2, 1951.) Green's tensors are used to derive boundary-value formulas for electromagnetic diffraction, by means of which the field in the space can be calculated from the tangential components of the electric or magnetic field-strength over the boundary.

538.566 2695
Huyghens' Principle for an Electromagnetic Wave—R. de Possel and C. Pouget-Michel. (*Compt. Rend. Sci.* (Paris), vol. 232, pp. 1819–1821; May 16, 1951.) Huyghens' principle is expressed in terms of electric and magnetic doublets similar to those used by Larmor for the general solution of Maxwell's equations in a vacuum. Love's potentials are introduced. The method of deducing from the derived expression the classical formulas involving charges and currents, is indicated.

538.566 2696
Wave Packets, the Poynting Vector, and Energy Flow: Part 2—Group Propagation through Dissipative Isotropic Media—C. O. Hines. (*Jour. Geophys. Res.*, vol. 56, pp. 197–206; June, 1951.) A theoretical paper. For pulses propagated in infinite media or through slabs having appreciable absorption, the speed of transmission to be expected is $c/(dkn_1/dk)$, where n_1 is the real part of the refractive index at frequency $kc/2\pi$. Part 1: 1881 of September.

538.566 2697
Wave Packets, the Poynting Vector, and Energy Flow: Part 3—Packet Propagation through Dissipative Anisotropic Media—C. O. Hines. (*Jour. Geophys. Res.*, vol. 56, pp. 207–220; June, 1951.) "Formulas are developed giving the velocity of packet propagation in dissipative anisotropic media for both homogeneous and inhomogeneous waves." Part 2: above.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72+523.85]:621.396.822 2698
Radio Astronomy—M. Ryle. (*Rep. Progr. Phys.*, vol. 13, pp. 184–245; bibliography, pp. 245–246; 1950.) A survey, complementary to those of Hey (91 of 1950) and Pawsey (104 of February). Details of experimental observations are dealt with relatively briefly, while discussions at greater length are devoted to the fundamental limitations of the various experimental techniques, and to some of the theoretical implications of the observations already made.

523.72+523.85]:621.396.822 2699
Radio Astronomy—(*Wireless World*, vol. 57, pp. 275–278; July, 1951.) A brief survey of modern equipment and techniques developed for the study of radio stars, meteors in the upper atmosphere, and solar and galactic noise.

523.72:621.396.822 2700
Radio Emission from the Sunspot of Central Meridian Passage 1950 June 14—A. Maxwell. (*Observatory*, vol. 71, pp. 72–74; April, 1951.) Observations made at Jodrell Bank indicated that the radiation intensity on a wavelength of 3.7 m ($10^{-19}w/m^2$ per cps) was unusually high for a sunspot group of this area (600 millionths of the sun's disk) which was, according to optical observations, declining. The magnetic field of the spot (3,600 gauss) was insufficient to account for this intense radiation. The high intensity persisted for 6 days, being asymmetrical about the time of central meridian passage, but coinciding in time with a brighter-than-normal patch of flocculi in the spot.

523.74 2701
A Forecast of Solar Activity—W. Gleissberg. (*Jour. Geophys. Res.*, vol. 56, pp. 294–295; June, 1951.) Earlier predictions (1121 of 1944) are vindicated. New predictions indicate that the next sunspot cycle will probably again be one of intense activity, with an unusually shallow minimum.

523.746"1951.01/.03" 2702
Provisional Sunspot-Numbers for January to March 1951—M Waldmeier. (*Jour. Geophys. Res.*, vol. 56, p. 288; June, 1951; *Z. Met.*, vol. 5, p. 192; May/June, 1951.)

523.85:621.396.822 2703
Some Observations of the Variable 205 Mc/s Radiation of Cygnus A—C. L. Seeger. (*Jour. Geophys. Res.*, vol. 56, pp. 239–258; June, 1951.) U.R.S.I. September, 1950. General Assembly paper. A report on observations made between October, 1948 and May, 1950. Contrary to the results of Bolton and Stanley (2514 of 1948), the intensity of the radiation was found to vary with a mean period of about $\frac{3}{4}$ minute, which agrees with observations at lower frequency. Strong variations occur at all altitude angles of the source. These may be caused by ionospheric scattering.

550.38:621.396.11 2704
Relation between the Earth's Magnetism and the Propagation of Radio Waves between Washington and Bagneux—Lejay, Ardillon, and Bertaux. (*See* 2806.)

550.38 "1950.10/.12" 2705
International Data on Magnetic Disturbances, Fourth Quarter, 1950—J. Bartels and J. Veldkamp. (*Jour Geophys. Res.*, vol. 56, pp. 283–287; June, 1951.)

550.38 "1951.01/.03" 2706

Cheltenham [Maryland] Three-Hour-Range Indices *K* for January to March, 1951—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 56, p. 288; June, 1951.)

550.384.4 2707

The Daily Magnetic Variations in Equatorial Regions—A. T. Price and G. A. Wilkins. (*Jour. Geophys. Res.*, vol. 56, pp. 259–263; June, 1951.) A new analysis of the *Sq* field for the Polar Year 1932–1933 indicates that the maximum daily variation of *H* in equatorial regions occurs between the magnetic and dipole equators in South America and Africa, but to the south of both these equators in the Far East. The line of maximum variation swings seasonally in the direction opposite to the sun.

550.385/.386 2708

Sudden Commencements and Sudden Impulses in Geomagnetism: Their Hourly Frequency at Cheltenham (Md), Tucson, San Juan, Honolulu, Huancayo, and Watheroo—V. C. A. Ferraro, W. C. Parkinson, and H. W. Unthank. (*Jour. Geophys. Res.*, vol. 56, pp. 177–195; June, 1951.) During the period 1926–1946, the local time variations in the hourly frequency of sudden commencements appeared to be small; their frequency may be greatest around 1300 local time. The hourly frequency of sudden impulses was minimum at around 0800 and 2000 local time. The diurnal variations in frequency of sudden commencements and sudden impulses with a preliminary movement in the opposite direction, exhibit a local time effect with an afternoon maximum. Their frequency may depend on longitude. The curves for the annual mean number of sudden commencements and sudden impulses combined, and for annual mean sunspot numbers, show striking similarities.

550.385 2709

Principal Magnetic Storms [Sept. 1950–March 1951]—(*Jour. Geophys. Res.*, vol. 56, pp. 289–291; June, 1951.)

550.385:551.594.52 2710

The Southward Shifting of the Auroral Zone during Intense Magnetic Storms—T. Nagata. (*Jour. Geophys. Res.*, vol. 56, pp. 292–294; June, 1951.) Author's reply to comment noted in 873 of May (Ferraro).

551.510.5:621.396.9 2711

Observation of Radar Echoes coming from a Cloudless Region—J. Broc. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 2034–2036; May 28, 1951.) With radar equipment operating on a wavelength of 3.2 cm and a peak power of 30 kw, echoes of unexplained origin were observed on March 20, 1951, at Monaco. The echoes came from an apparent distance of 15 to 20 km, and from a height $\gg$1 km, the sky being clear.

551.510.535+551.594.5 2712

U.R.S.I.—I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951. (Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) Summaries are given of the following papers:

2—Martyn's Theory of Magnetic Storms and Auroras—H. G. Booker.

9—On the Definition of Virtual Height—J. Shmoys.

11—Fluctuations of F_2 Region between Stations separated by 100 to 150 Miles—H. W. Wells.

24—The Experimental and Theoretical Study of Ionospheric Absorption at 150 kc/s—A. H. Benner.

35—Southern Extent of Aurora Borealis in North America—C. W. Gartlein and R. K. Moore.

36—A V.H.F. Propagation Phenomenon associated with the Aurora—R. K. Moore.

38—Magneto-ionic Multiple Splitting determined with the Method of Phase Integration—W. Pfister.

551.510.535 2713

Evidence for Ionosphere Currents from Rocket Experiments Near the Geomagnetic Equator—S. F. Singer, E. Maple, and W. A. Bowen, Jr. (*Jour. Geophys. Res.*, vol. 56, pp. 265–281; June, 1951.) Two rockets carrying total-field magnetometers were fired from sites 60 miles apart. In one case, the recorded magnetic field at distances between 20 and 105 km decreased according to expectations for a simple dipole field. In the second case, a discontinuity was observed in the altitude range 93 to 105 km. The results establish the existence of a current system in the *E* region causing the diurnal variation of the earth's magnetic field at sea level, and support the dynamo theory of the daily magnetic variation proposed by Balfour Stewart and Schuster. A brief report of the same work is given in *Phys. Rev.*, vol. 82, pp. 957–958; June 15, 1951.

551.510.535 2714

The Theory of Magneto-ionic Triple Splitting—O. E. H. Rydbeck. (*Chalmers Tekn. Högsk. Händl.*, no. 101, 40 pp.; 1951. In English.) The coupling coefficients between the magneto-ionic modes of propagation are calculated, and the method of excitation of the *z* mode analyzed in detail. The transformation and reflection coefficients of the modes are deduced as functions of the collisional frequency. Poeverlein's graphical method is applied to determine the refractive indexes and ray paths. Data from Kiruna are used to illustrate the main results of the theory. Meek's results (3221 of 1948) are also explained. The possibility that scattering discontinuities can excite the *z* mode at lower latitudes is discussed briefly. See also 1147 of June and *Onde Élect.*, vol. 31, pp. 70–81 and 153–156; February and March, 1951.

551.510.535:621.396.11 2715

Change in the Nature of Medium-Wave Propagation at Sunset—Houtsmuller. (*See* 2809.)

551.594.5:621.396.11 2716

Radio Observations of the Aurora on November 19, 1949—N. C. Gerson. (*Nature* (London), vol. 167, pp. 804–805; May 19, 1951.) Using amateur reports of aurorally maintained reception, the position of the auroral reflecting belt is calculated, assuming that the height of reflection is 100 km, and that the ray paths are tangential to the earth. Possible causes of the "auroral modulation" are discussed. It is probable that this effect is due to changes of ionization density with position and time.

551.594.6:621.317.7.087:621.396.821 2717

An Atmospherics Waveform Receiver—W. J. Kessler. (Proc. I.R.E., vol. 39, p. 676; June, 1951.) The waveform is displayed on a cro, the horizontal sweep being triggered by the signal. This is delayed by 18-μs so that the leading edge is not lost. 50- or 100-μs markers are displayed on a separate sweep to avoid confusion with the signal. The display rate is limited to about 8 waveforms per second. Three sample waveforms showing multiple echoes are reproduced.

551.594 2718

The Flight of Thunderbolts [Book Review]—B. F. J. Schonland. Publishers: Clarendon Press, Oxford, and Oxford University Press, London, 1950, 152 pp., 15s. (*Nature* (London), vol. 167, pp. 787–788; May 19, 1951.) "An elementary but authoritative account" covering the development of knowledge regarding lightning discharges, methods of protection against lightning, the mechanism of electrification of thunderclouds, and radio atmospherics and their use in locating thunderstorms.

LOCATION AND AIDS TO NAVIGATION

621.396.9 2719

Port Approach and Berthing in Fog—F. J. Wylie. (*Jour. Inst. Nav.*, vol. 4, pp. 156–164; April, 1951.) A discussion of the role of shore-based radar in the berthing of ships in fog, and of the manner in which information obtained from radar or other aids should be presented to the master of a ship.

621.396.9:621.396.616 2720

Optimum Operation of Echo Boxes—W. M. Hall and W. L. Pritchard. (Proc. I.R.E., vol. 39, pp. 680–684; June, 1951.) The dependence of the optimum *Q* and maximum ring time of an echo box system on coupling method, pulse length, unloaded cavity *Q*, and transmitter-receiver properties, is examined theoretically. The results are shown graphically. The power taken by a direct-coupled box is also plotted as a function of pulse length, and the ratio of loaded to unloaded *Q*.

621.396.93:621.396.67 2721

U.R.S.I.—I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951. (Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) A summary is given of the following paper:

16—Antenna Systems for Radio Direction Finding—E. C. Jordan.

621.396.933 2722

Description and Evaluation of 100-Channel Distance-Measuring Equipment—R. C. Borden, C. C. Trout, and E. C. Williams. (Proc. I.R.E., vol. 39, pp. 612–618; June, 1951.) By combining the principles of operation of two separate earlier schemes for 50-channel equipment, a 100-channel system has been developed using only 20 distinct frequency channels (10 for interrogation and 10 for reply). These occupy a total bandwidth of 50 mc in the region of 1 kmc, together with 10 coded temporal relations between the interrogation and reply pulses. Possible identification schemes are discussed, and flight tests are described.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 2723

Photoconductivity of Caesium-Antimony Films—P. G. Borzyak. (*Zh. Tekh. Fiz.*, vol. 20, pp. 923–927; August, 1950.)

535.215 2724

Preparation of Photoconducting Cadmium Sulphide—R. E. Aitchison. (*Nature* (London), vol. 167, pp. 812–813; May 19, 1951.) Uniform layers of photoconducting CdS having any desired area, with thicknesses up to 5×10^{-4} cm, can be prepared by the slow evaporation in vacuo ($p \leqq 10^{-6}$ mm Hg) of CdS precipitated from an aqueous solution.

537.226 2725

Present-Day Theories of Ferroelectricity—H. Lumbroso. (*Rev. Sci.* (Paris), vol. 88, pp. 239–247; October/December, 1950.) A general discussion, with particular reference to the properties of Rochelle salt and $BaTiO_3$. A bibliography of 18 items is appended.

537.311.33 2726

Transistors: Part 1—J. Malsch. (*Arch. elekt. Übertragung*, vol. 5, pp. 139–148; March, 1951.) A survey paper discussing the fundamental physical processes of conduction in semiconductors such as Si and Ge.

537.311.33 2727

Valence Semiconductors—F. Stöckmann. (*Naturwiss.*, vol. 38, pp. 151–154; April, 1951.) A survey paper, with numerous references, in which the conduction mechanism in semiconductors is considered from the point of view of developing materials with properties suitable for practical applications. Mixed-crystal (valence) semiconductors have the advantage that the concentration of impurity centers depends only on the composition of the material, and is thus controllable and thermally stable.

537.311.33:061.3 2728

The Seventh All-Union Conference on the

Properties of Semiconductors—(*Zh. Tekh. Fiz.*, vol. 21, pp. 231–242; February, 1951.) Report on a conference held by the Academy of Sciences of the U.S.S.R., and by the Academy of Sciences of the Ukrainian S.S.R., in Kiev, on the 14th to the 21st of October, 1950. Thirty-nine papers were read, under the following headings: general; theory of semiconductors; photoelectric phenomena in semiconductors; new types of semiconductors; surface and contact phenomena; thermal properties of semiconductors; technical application of semiconductors. Summaries of the papers and of the discussions are given.

537.311.33:546.3 2729

The Formation of Barrier Layers with the aid of Chemical Combinations in Welds—S. G. Kalashnikov and L. N. Erastov. (*Zh. Tekh. Fiz.*, vol. 21, pp. 129–134; February, 1951.) A simple method is described for obtaining a barrier layer by welding two metals which form a layer of high specific resistance in the weld. In this way, barrier layers were obtained between Sb and Mg and between Sb and Zn. The technique is described, and the results of an experimental investigation, including the voltage-current characteristics, are presented. A theoretical interpretation is given of the rectification produced. It is not suggested that the method is suitable for practical applications, since the rectifying properties of the welded pairs apparently deteriorate rapidly.

537.311.33:546.841-3 2730

Semiconductor Properties of Thoria in a Vacuum—G. Mesnard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1744–1746; May 7, 1951.) Experiments were carried out on thoria cylinders deposited by electrophoresis on pairs of parallel tungsten wires serving as heater and electrode systems. Properties investigated included electrical and thermal conductivity, and the rectification at the tungsten-thoria interface. The observed variation of conductivity with temperature gives support to the views of Loosjes and Vink (3208 of 1950) on the electronic processes involved, while the values obtained are comparable with those of Danforth and Morgan (2539 of 1950). Departure of the observed thermoelectric effect from values given by existing semiconductor theory, are explained by an interface, pd, which increases with rise of temperature. Rectification occurs as a result of the interface, pd, and the greater abundance of electrons in regions of higher temperature.

538.221 2731

International Conference on Ferromagnetism and Antiferromagnetism, Grenoble [July 3rd to 7th, 1950]—(*Jour. Phys. Radium*, vol. 12, pp. 149–508; March, 1951.) Full text of 49 papers presented at the conference.

538.221 2732

The Origin of Intermittent Activation in Ferromagnetic Materials—R. Forrer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1746–1748; May 7, 1951.)

538.221 2733

The Magnetic Structure of High-Coercivity Alloys: Part 4—Magnetostriction Hysteresis in Alnico and Vicalloy Alloys—D. A. Shturkin and Ya. S. Shur. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1393–1399; November, 1950.) The results of an experimental investigation are analyzed on the assumption that these ferromagnetic materials consist of finely dispersed plate-formations insulated from one another, and that there is only one region of spontaneous magnetization within each formation.

538.221 2734

Magnetic Viscosity of Permalloy in Sinusoidally and Aperiodically Varying Fields—R. V. Telesnin and S. Z. Ushakov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1366–1371; November, 1950.) Magnetic viscosity, i.e., the time delay of variations in magnetization with respect to variations in the intensity of the magnetic field, has previously been investigated by a number of authors separately for sinusoidally and for aperiodically varying fields. The present paper reports measurements made on the same sample for both types of field. A number of conclusions regarding magnetic viscosity are drawn.

538.221 2735

The Ferromagnetism of the β_1 Phase of Co/Zn Alloys—A. J. P. Meyer and P. Taglang. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1914–1916; May 21, 1951.) An experimental study of a series of alloys in which the proportion of Zn was varied from 42 per cent to 60 per cent.

538.221 2736

Results of Measurements on High-Permeability Ferrite Cores—M. Kornetzki. (*Z. angew. Phys.*, vol. 3, pp. 5–9; January, 1951.) The permeability characteristics of ferrite cores are discussed and compared with those of high-permeability stacked and wound cores. The ferrite cores may have a μ up to 3,500, nearly independent of frequency up to the gyromagnetic limiting frequency, between 1 and 6 mc. Above this frequency, μ falls. Due to capacitive eddy currents, a resonance occurs which may lead to an apparent increase in μ. The Curie point is lower the higher the permeability.

538.221:539.382 2737

Measurement of the Young's Modulus of Ferrites—L. Weil and L. Bochirol. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1807–1809; May 16, 1951.) The method used depends on finding the resonance frequency of magnetostrictively excited mechanical oscillations in annular specimens.

539.23 2738

Thin Films of Hydrocarbons formed by Electron or Ion Bombardment—H. König and G. Helwig. (*Z. Phys.*, vol. 129, pp. 491–503; April 28, 1951.) The mechanism by which films of solid hydrocarbons are formed from gas on the walls of experimental vacuum tubes, is described. Disturbing effects and useful applications of the film are discussed.

539.234:546.59 2739

The Crystallization of Very Thin Gold Films—A. Colombani and G. Ranc. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1344–1346; April 2, 1951.) Report of an experimental investigation of films deposited by evaporation on to an amorphous (e.g., plexiglas) or crystalline (e.g., NaCl) support. The nature of the film at different stages of development was deduced from the variations of its conductivity. Results are presented in graphs and discussed briefly.

546.287:537.533.8 2740

Silicone Oil Vapour and Secondary Electron Emission—A. Lempicki and A. B. McFarlane. (*Nature* (London), vol. 167, pp. 813–814; May 19, 1951.) Measurements of the "sticking" potential of a fluorescent powder, and of the secondary-emission coefficient for various surfaces, lead to the view that silicone oil vapor, currently used instead of mercury vapor in diffusion pumps, will produce obnoxious results on any surface bombarded by electrons.

549.514.51:621.396.611.21 2741

Temperature Dependence of Quartz Resonators—R. Bechmann. (*Arch. elekt. Übertragung*, vol. 5, pp. 89–90; February, 1951.) A note complementing earlier papers (3332 of 1942 and 2187 of 1943) and giving values of oscillation coefficients, piezoelectric coefficients, and capacitance constants and their temperature coefficients, for some bar and plate resonators.

621.3.011.5:[546.131+546.131.02 2742

The Dielectric Properties of Solid HCl and DCl—C. S. E. Phillips. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1924–1927; May 21, 1951.) Measurements made at frequencies ranging from 80 cps to 6 mc are reported.

621.314.63 2743

The Theory of Direct-Current Characteristics of Rectifiers—P. T. Landsberg. (*Proc. Roy. Soc. A*, vol. 206, pp. 463–477; May 22, 1951.) Image force can be taken into account very simply by making minor adjustments in the diffusion and diode theories. Expressions are obtained enabling detailed comparisons to be made with experimental data on the dc reverse characteristics of Cu_2O, Se, and Ge rectifiers, assuming (a) a Mott barrier, and (b) a Schottky barrier. Agreement is satisfactory. The effect of the image force on the charge-carrier distribution is illustrated, and a general relation between the two theories is established.

621.314.63 2744

Contributions to the Theory of Heterogeneous Barrier-Layer Rectifiers—P. T. Landsberg. (*Proc. Roy. Soc. A*, vol. 206, pp. 477–488; May, 22, 1951.) Assuming an arbitrary distribution of space charge in the barrier layer, the general form of the current-voltage relation is derived from both diode and diffusion theory. A connection, valid for most barriers, between this characteristic and the capacitance-voltage curve is pointed out. The Sachs breakdown voltage can be deduced from the latter characteristic. Assumption of a barrier, in which the distribution of impurity centers is established by a diffusion process, leads to more likely values for the rectifier parameters.

621.314.63 2745

Theory of the Contact between Two Semiconductors with Different Types of Conductivity—A. I. Gubanov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1287–1301; November 1950.) Various phenomena observed when a voltage is applied between two semiconductors with different types of conductivity (hole and electron) are discussed. To explain these phenomena it is suggested that a barrier layer of a type different from that considered by other investigators is formed at the contact. The theory of this layer is presented, and it is shown that it must possess high resistivity and exert considerable rectifying action. A formula is derived for determining the current through the contact. I/V characteristics, for both forward and backward directions, plotted from this formula show good agreement with experimental results.

621.315.61 2746

Dielectrics Made to Order—A. R. von Hippel. (*Electronics*, vol. 24, pp. 126–128; June, 1951.) Study of the synthesis of dielectric materials based on investigation of dielectric breakdown.

621.315.61:621.3.015.5 2747

Electrical Breakdown over Insulators in High Vacuum—P. H. Gleichauf. (*Jour. Appl. Phys.*, vol. 22, pp. 766–771; June, 1951.) Account of a practical investigation, in continuation of previous work (2454 of November). Breakdown voltage was measured for various electrode and insulator materials. Data establishing the relation between insulator length and breakdown voltage were obtained. The effect of the degree of roughness of the insulator surfaces in contact with the electrodes was investigated, and experiments were made in which one of the electrodes was separated from the insulator.

621.315.616.9:621.3.015.5 2748

The Intrinsic Electric Strength of Polyvinyl Alcohol and its Temperature Variation—I. D. L. Ball. (*Proc. IEE*, vol. 98, pp. 84–86; March, 1951.) Measurements were made at temperatures from $-195°$ C to $90°$ C. Comparison with observations made on other high polymers [see 1696 of 1949 (Bird and Pelzer) and 764 of 1947 (Austen and Pelzer)] suggests that the temperature variation of the intrinsic electric strength of these materials is determined by the dipoles present rather than by the physical structure.

621.318.4.042.15 2749

Investigations into the Possibility of using Suspensions of Ferromagnetic Particles in Liquid Dielectrics in place of Powder Cores—H. H. Rust. (*Arch. elekt. Übertragung*, vol. 5, pp. 67–76; February, 1951.) Plastic suspensions of, e.g., carbonyl iron in insulating oil were investigated. Coils were embedded in the suspension, and determinations were made of the resistance of the suspension with and without addition of a dipolar substance to the dispersive medium, and of the Q and self-inductance of the coil as a function of temperature and sedimentation time. The results are compared with corresponding values for powder-core coils. Variation of the effective permeability of the suspension due to thixotropic effects resulting from mechanical stress was investigated, and a number of practical applications are indicated.

621.319.45 2750

The Oxide Layer on Aluminium and the Temperature Dependence of the Capacitance of the Electrolytic Capacitor—S. S. Gutin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 135–142; February, 1951.) A report on an experimental investigation, the main conclusions of which are as follows: (1) the oxide layer is of porous structure; (2) the temperature dependence of capacitance is determined by the action of the electrolyte in the pores; (3) covering the layer with a thin film of a solid dielectric reduces the temperature dependence of capacitance and ensures linear variation within the working range of temperatures.

669.018 2751

Resistance Alloys—H. Thomas. (*Z. Phys.*, vol. 129, pp. 219–232; February 13, 1951.) Measurements were made on Ni/Cr, Fe/Al, Ni/Al, Fe/Si, Ni/Cu/Zn, and Ni/Cu alloys. The resistivity/temperature curve is S-shaped, and the resistance is increased by heat treatment at low temperatures and is decreased by cold working. A physical explanation is advanced.

MATHEMATICS

517.94 2752

U.R.S.I.—I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951. (Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) A summary is given of the following paper:

40—Asymptotic Solution of Maxwell's Equation—M. Kline.

518.3 2753

The Smoothing of Experimental Curves—P. Vernotte. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1897–1899; May 21, 1951.)

518.3:514.6 2754

Nomogram and Slide-Rule for Solution of Spherical Triangle Problems found in Radio Communication—D. V. Dickson. (*Jour. Geophys. Res.*, vol. 56, pp. 163–175; June, 1951.)

681.142 2755

The Diagnosis of Mistakes in Programmes on the EDSAC—S. Gill. (*Proc. Roy. Soc. A*, vol. 206, pp. 538–554; May 22, 1951.) Description of methods, applicable generally to high-speed computers, for rapid diagnosis of mistakes due to faulty programming, including details of checking routines applicable particularly to the EDSAC.

517.53 2756

Equations Intégrales et Transformation de Laplace [Book Review]—M. Parodi. Publishers: Ministère de l'Air, Paris, France, 1950, 125 pp. (*Ann. Télécommun.*, vol. 6, p. 60; February, 1951.) Addressed primarily to mathematicians, but of interest also to engineers and physicists concerned with manipulating integral equations or using the symbolic calculus.

MEASUREMENTS AND TEST GEAR

529.786 2757

Accurate Time for Broadcast Studios—J. H. Greenwood. (*Electronics*, vol. 24, pp. 97–99; June, 1951.) Details of a system using a 60-cps standard frequency controlled by a 240-cps tuning fork to drive up to 10 synchronous clocks. Indications may be adjusted to agree with wwv signals to an accuracy within ±0.1 second.

621.316.726.078.3 2758

Frequency Stabilization: Ultra-High-Frequency Discriminator—G. Pircher. (*Onde élect.*, vol. 31, pp. 144–152; March, 1951.) Methods of stabilizing uhf oscillators and various types of discriminator, are outlined. A detailed description is given of a simple discriminator, based on modifications of the amplitude and phase of the stationary wave pattern in a guide terminated by a load tuned approximately to the resonant frequency to be controlled. Figures are given of an actual equipment, using a klystron oscillator, covering the waveband 30 to 36 mm, and applications in the field of uhf measurements are indicated.

621.316.8:621.317.3.029.55 2759

A Resistor for High-Frequency Measurements—F. Lappe and K. B. Westendorf. (*Z. angew. Phys.*, vol. 3, pp. 29–32; January 15, 1951.) The resistor consists of a large number of thin parallel wires stretched along the surface of a cylinder and connected in parallel. The return path is provided by a coaxial metal cylinder inside or outside the wire cage. The time constants for such arrangements are calculated, and details are given of suitable constructions for particular applications.

621.317.32 2760

Arrangement for the Compensation and Measurement of Small Direct Voltages—G. Paldus. (*Arch. elekt. Übertragung*, vol. 5, pp. 135–138; March, 1951.) A device suitable for balancing out the steady component of an applied voltage, and thus making the whole of the galvanometer scale available for measuring the variable component, consists basically of oppositely connected thermocouples with separate heaters connected for differential control by a potentiometer circuit. Variants of the basic arrangement are described.

621.317.32.029.63 2761

Modern Methods of High-Frequency Voltage Measurement up to about 10,000 Mc/s—M. J. O. Strutt. (*Arch. tech. Messen*, no. 182, pp. T32–T33; March, 1951.) Discussion of sources of error in a diode voltmeter, and description of this and other measurement systems, the latter depending on current variations.

621.317.324:621.396.611.1 2762

Methods of Measuring Adjacent-Band Radiation from Radio Transmitters—N. Lund. (Proc. I.R.E., vol. 39, pp. 653–656; June, 1951.) Three methods are described using, respectively, two modulation tones of equal amplitude, normal modulating signals, or thermal-noise modulation. Good agreement between the last two methods shows the usefulness of thermal noise measurements, which require considerably less equipment. The first method normally gives only a qualitative estimate, but distortion measurements can be used to calculate the adjacent-band radiation, which is in reasonable agreement with the values found by the other methods.

621.317.335.3.029.63† 2763

The Measurement of Permittivity and Power Factor of Dielectrics at Frequencies from 300 to 600 Mc/s—J. V. L. Parry. (*Proc. IEE*, vol. 98, pp. 303–311; July, 1951.) A resonance method developed from that of Hartshorn and Ward (351 of 1937) for use at higher frequencies. is described. The variable capacitive elements are mounted in a re-entrant cavity which serves as a wavemeter for capacitance calibration. The accuracy for permittivity measurement is estimated to be within ±1 per cent, and for power-factor measurement, to within $\pm 2\times 10^{-5}$ for values below 0.01, and $\pm 5\times 10^{-4}$ for higher values.

621.317.336.029.62/.63 2764

U.R.S.I.-I.R.E. Spring Meeting, Washington, D.C., April 16–18, 1951. (Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) A summary is given of the following paper:

3—Impedance Measurements in the 50 to 1,000 Mc/s Range—F. J. Gaffney.

621.317.42 2765

The Electron Multiplier as an Indicator of Weak Magnetic Fields—N. V. Krasnogorskaya. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1257–1266; October, 1950.) Experiments were conducted which show that an electron-multiplier tube with permanent magnets can be used as a magnetometer. It can be easily connected to an oscillograph and, being practically inertialess, it follows rapid variations of weak magnetic fields.

621.317.42 2766

Magnetic Field Measurements with Peaking Strips—J. M. Kelly. (*Rev. Sci. Instr.*, vol. 22, pp. 256–258; April, 1951.) The method is based on the oscilloscopic observation of voltage pulses induced in a coil wound round a length of high-permeability wire (i.e., a "peaking strip") arranged parallel to the magnetic field. Field strengths of the order of a kilogauss can be measured accurately to within ±0.02 gauss.

621.317.725 2767

The Theory and Design of the Reflex Voltmeter—A. E. Budarov and L. E. Leykhter. (*Zh. Tekh. Fiz.*, vol. 21, pp. 77–91; January, 1951.) Range and scale-linearity of tube voltmeters are improved by the use of negative feedback. A detailed theoretical discussion of the operation of the voltmeter is presented, followed by a report on an experimental investigation confirming the theoretical findings. Two alternative circuits are described, and design details are considered. The various advantages of this type of voltmeter are enumerated. In addition to those already mentioned, they include: all-mains supply; great stability; very high input impedance; very long life of the measuring tube owing to the considerable reduction in the filament voltage; simplicity in production and operation; small weight and overall dimensions.

621.317.755.087.5 2768

The Speed of Recording of Cathode-Ray Traces by External Photography—V. P. Shasherin. (*Zh. Tekh. Fiz.*, vol. 21, pp. 92–103; January, 1951.) An attempt is made to provide a theoretical basis for determining the maximum speed of recording. The main factors involved are the conditions under which the oscillograph is used, the conditions under which the photographs are taken, the inertia of the cro screen, the method of scanning, and the characteristics of the photosensitive layer of the negative. Each of these factors is discussed in detail, and formulas are derived taking them into account. Some experimental data are also given.

621.317.763:621.396.615 2769

Two 'Grid Dip' Oscillators—C. Guilbert. (*Toute la Radio*, no. 154, pp. 82–87; March and April, 1951.) Description, with full constructional details, of two small tube-oscillator wavemeters. One uses a single "magic-eye" tube for both oscillator and indicator, and covers the range 3.3 to 95 mc. The second has a normal indicating instrument, and covers the range 3.4 to 250 mc.

621.317.772.089.6 2770
Precision Calibrator for L.F. Phase-Meters—M. F. Wintle. (*Wireless Eng.*, vol. 28, pp. 197–208; July, 1951.) The calibrator provides a fixed-phase reference output and a calibrated variable-phase output. To improve accuracy, the phase shifting is performed at a harmonic of the output frequency, and the output is taken after frequency division. Details are given of the master oscillator, the main phase-shifting control, the frequency dividers, and selective amplifiers, together with examples of calculation of typical component values.

621.317.784 2771
Absolute Measurement of Microwave Power in Terms of Mechanical Forces—A. L. Cullen. (*Nature* (London), vol. 167, p. 812; May 19, 1951.) A radiation-pressure method is described for measuring the power in a low-loss cavity. Fox's rotary phase shifter (1255 of 1948) can be adapted to measure power absolutely without introducing reflection effects in the cavity.

621.385.001.4 2772
Valve Testing Methods and Apparatus of the French P.T.T. Administration—M. Ganet. (*Câbles & Trans.* (Paris), vol. 5, pp. 68–75; January, 1951.) A description of factory test methods and apparatus ensuring the reliability and uniformity of tubes supplied in bulk for use in telephone and telegraph line repeaters.

621.385.001.4 2773
Radio Valve Life Testing—R. Brewer. (*Proc. IEE*, vol. 98, pp. 269–274; July, 1951.) Tube failure may be due to emission failure, or to mechanical or electrical defects. The prediction of life is usually based on sample testing. The tests made may be for the purpose of quality control, pilot production, type establishment, or special applications, and vary accordingly. Sampling and test procedures, and suitable equipment are described. Evidence from life tests is summarized, and possible ways of speeding up procedure are outlined.

621.396.611.21 2774
The Measurement of the Amplitude of Oscillation of Quartz Crystals by the Interference Method—L. N. Borodovskaya and A. E. Salomonovich. (*Zh. Tekh. Fiz.*, vol. 21, pp. 221–224; February, 1951.) The method was used for measuring the oscillation amplitude of quartz crystals oscillating longitudinally, and for determining the relation between this amplitude and the voltages applied to, or taken from, the crystal. The theory of the method and the experimental results are discussed.

621.396.612.7.029.64 2775
A Spark Transmitter for Lightly Damped Centimetre Waves of Continuously Variable Frequency—H. Anders. (*Z. Phys.*, vol. 129, pp. 45–55; January 29, 1951.) A Hertzian dipole oscillator inside an almost closed cavity resonator is used as a continuously tunable generator over the 1- to 3-cm waveband. Since this arrangement produces a spectrum of oscillations, the radiation is passed through a second resonator whose spectrum coincides with that of the first at only one frequency. A nearly monochromatic radiation results, with intensity adequate for measurement purposes, and a logarithmic decrement of the order of 0.01.

621.396.615.015.7 2776
Narrow-Pulse Generator—S. F. Pearce and D. C. G. Smith: C. S. Fowler. (*Wireless Eng.*, vol. 28, pp. 225–226; July, 1951.) Further discussion of the difference between the measured and theoretical pulse lengths as described in 694 of April (Fowler). The discrepancy is ascribed to the finite operating time of the thyratron.

621.396.645.35:621.383 2777
A High-Sensitivity All-Mains Amplifier for Photocell Measurements—H. Tronnier and H. Wagener. (*Funk u. Ton*, vol. 5, pp. 1–4; January, 1951.) Modifications which improve the stability of the Etzold-type circuit (1324 of 1948) are discussed. The circuit described includes additional voltage stabilizers and two barretters, one to prevent overloading the main stabilizer, the other in the heater supply to the measurement circuit.

621.397.6.001.4 2778
Simple Test Generator for Television Sets using 441 and 819 Lines—J. Bergonzat. (*TSF pour Tous*, vol. 27, pp. 120–122; March, 1951.) Apparatus providing a test signal for adjusting and demonstrating television sets consists of a Colpitts oscillator generating frequencies from 40 to 60 mc. The various signals required to give a square-grid display are formed in four multivibrators. The circuit diagram, with component values, is given.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

522.615:522.2 2779
Servo Guider for Solar Telescopes—F. E. Fowler and D. S. Johnson. (*Electronics*, vol. 24, pp. 118–122; May, 1951.) Circuits incorporating photocells are used to correct the telescope direction for both hour angle and declination errors.

538.563:621-526 2780
Phase-Sensitive Detector Circuit having High Balance Stability—N. A. Schuster. (*Rev. Sci. Instr.*, vol. 22, pp. 254–255; April, 1951.) High stability of the detector balance, of great importance; e.g., in measurements of nuclear resonance where signal to noise ratio is low and an hour or more is required for signal measurements, is here achieved by using only one tube for the signal input, and switching its plate load by means of the reference signal.

621.316.7.076.7 2781
Improving Industrial Control Design—E. H. Vedder. (*Electronics*, vol. 24, pp. 104–106; May, 1951.) Based on a paper given in full in *Proc. NEC* (Chicago), vol. 6, pp. 471–479; 1950. Differences between requirements for industrial electronic apparatus and those for receivers are indicated. Long tube life and high reliability are important.

621.316.7.076.7 2782
The Electro-analogue, an Apparatus for Studying Regulating Systems: Part 1—Components and Functions—J. M. L. Janssen and L. Ensing. (*Philips Tech. Rev.*, vol. 12, pp. 257–271; March, 1951.) A universal model constituted by a network representing the process to be regulated, is combined with another model for the automatic controller which may act continuously or discontinuously. Oscillograms of typical response functions are given.

621.317.083.7:621.396.621 2783
Linear Discriminator for F.M. Telemetering—G. S. Sloughter and R. T. Ellis. (*Electronics*, vol. 24, pp. 113–115; June, 1951.) An asymmetrical flip-flop circuit is used to provide a stable FM discriminator for deviation ratios from zero to ±20 per cent, with subcarrier frequencies between 400 cps and 70 kc. The discriminator is linear to within 1 per cent over a deviation range of ±7.5 per cent. The particular application described is for the ground recording of flight characteristics of aircraft.

621.317.083.7:621.396.621.015.7 2784
Pulse-Width Discriminator—A. A. Gerlach and D. S. Schover. (*Electronics*, vol. 24, pp. 105–107; June, 1951.) Description of a simple high-stability pulse-width discriminator, handling pulses between 20 and 100 ms wide, used for channel identification at the receiving end of a three-channel telemetering system with pulse-width channel coding. Using pulse widths of 30, 60, and 90 ms, respectively, a tolerance of ±5 ms was satisfactory in the presence of noise.

621.365.029.63:537.523.5 2785
The Electronic Torch and Related High Frequency Phenomena—J. D. Cobine and D. A. Wilbur. (*Jour. Appl. Phys.*, vol. 22, pp. 835–841; June, 1951.) Description of an electronic torch operating at 1 kmc, and of some of the characteristics of the gaseous discharge produced. Power from a magnetron is coupled into a cavity, in turn coupled to a coaxial line through which gas is passed, and which is terminated by the torch section. The flame produced by polyatomic gases is capable of melting many refractory materials, due to the heat of association of the dissociated molecules. For a shorter account, see *Electronics*, vol. 24, pp. 92–93; June, 1951.

621.365.54† 2786
R. F. Induction Heating of Metals—M. Krüger and A. Koppenhöfer. (*Elektron Wiss. Tech.*, vol. 5, pp. 35–44; February and March, 1951.) A survey of the principles and apparatus used, with indication of specific applications.

621.383.001.8:621.798.4 2787
New Photoelectric Register Controls—G. M. Chute. (*Electronics*, vol. 24, pp. 92–97; May, 1951.) Description of three circuits for controlling packaging machinery. Thyratrons are used for feeding the correcting motor, and a three-photocell arrangement for obtaining high accuracy of register control.

621.384.6 2788
An Electronic Ram—W. Raudorf. (*Wireless Eng.*, vol. 28, pp. 215–221; July, 1951.) A simple method for accelerating electrically-charged particles is suggested, based on the transfer of the energy of a moving electric charge to a small fraction of that charge. A tubular arrangement is described in which the axial velocity of an electron beam is controlled by a longitudinal magnetic field. An extremely short and powerful em pulse is generated.

621.384.6 2789
Cyclic Accelerators—J. H. Fremlin and J. S. Gooden. (*Rep. Progr. Phys.*, vol. 13, pp. 295–349; 1950.) A review of the development of the more important particle accelerators requiring magnetic fields to control the particle orbits, with a discussion of the applications for which each is best suited economically.

621.384.612.2† 2790
The Synchrocyclotron at Amsterdam—F. A. Heyn. (*Philips Tech. Rev.*, vol. 12, pp. 241–256; March, 1951.) General description of the installation, with its oscillator and modulator, which produces in continuous operation deuterons with an energy of 28 mev.

621.385.3:578.08 2791
Wide Range Mechano-Electronic Transducer for Physiological Applications—S. A. Talbot, J. L. Lilienthal, Jr., J. Beser, and L. W. Reynolds. (*Rev. Sci. Instr.*, vol. 22, pp. 233–236; April, 1951.) A modified version of the RCA 5734 triode described in 3413 of 1947 (Olson) is used to measure forces of 100 mg to 25 kg.

621.385.83 2792
Energy Loss from Fast Electrons on passing through Foils (Multiple Scattering)—W. Schultz. (*Z. Phys.*, vol. 129, pp. 530–546; April 28, 1951.)

621.385.833 2793
Theory of the Independent Electrostatic [electron] Lens with Elliptical Central Electrode—É. Regenstreif. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1918–1920; May 21, 1951.) Explicit formulas are established for the gaussian trajectories in such a lens.

621.385.833 2794

Trajectories of Charged Particles Deflected Slightly from the Original Direction by an Electrostatic Field—B. M. Rabin and A. M. Strashkevich. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1232–1240; October, 1950.) Equations are derived determining the trajectories of the particles. Using these equations, it is possible to investigate the electronoptical properties of focusing and deflecting systems in three dimensions.

621.385.833 2795

The Optical Strength of Short Electron Lenses—O. I. Seman. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1180–1193; October, 1950.) It is shown that in determining the focal length of an electrostatic electron lens from the usual formula (1), an error as large as 100 per cent may be introduced. More accurate formulas, (8) and (21), are derived. The error introduced in determining the focal length of a magnetic lens from the usual formula, (2), is also estimated.

621.385.833:537.533.72 2796

The Significance of the Concepts 'Focus' and 'Focal Length' in Electron Optics, and Strong Electron Lenses with Newtonian Image—Formation Equation: Part 2—W. Glaser and O. Bergmann. (*Z. angew. Math. Phys.*, vol. 2, pp. 159–188; May 15, 1951.) Continuation of 2505 of November.

621.387:621.396.645.35 2797

Improved D.C. Amplifier for Portable Ionization Chamber Instruments—N. F. Moody. (*Rev. Sci. Instr.*, vol. 22, pp. 236–239; April, 1951.) The requirements of zero stability, adequate overall gain, and low battery consumption, are reconciled by the use of a two-stage reflex amplifier circuit. The anode load of the first (electrometer-tube) stage is returned to the screen grid of the second tube, a pentode cathode follower. The two stages are effectively cascaded, and a voltage gain >100 can readily be attained.

621.398:621.396.61 2798

A General-Purpose Remote Control System for Radio Transmitters—V. J. Tyler. (*Marconi Rev.*, vol. 15, pp 61–81; 2nd quarter, 1951.) The general requirements of remote-control systems using land lines are discussed, and the advantages of a standard design having wide applicability are indicated. A brief description is given of a system which has been adapted for the control of three widely different transmitters, viz., a small communications transmitter, a pair of large broadcasting transmitters, and an aircraft navigation aid beacon transmitter.

PROPAGATION OF WAVES

538.566+621.396.11 2799

U.R.S.I.-I.R.E. Spring Meeting, Washington, D. C., April 16–18, 1951. (Proc. I.R.E., vol. 39, pp. 716–720; June, 1951.) Summaries are given of the following papers:

1—Recent Mathematical Developments in the Theory of Tropospheric Propagation—B. Friedman.
4—Air-to-Air Tropospheric Radio Propagation—G. B. Fanning and F. P. Miller.
5—The Effect on Propagation of an Elevated Atmospheric Layer of Nonstandard Refractive Index—L. H. Doherty.
6—Experimental Discrimination of the Factors in V.H.F. Radio Wave Propagation—A. W. Straiton and C. W. Tolbert.
7—Some Characteristics of 92.9-Mc/s Propagation observed at a Distance of 127 Miles—R. J. Wagner, Jr.
10—Dispersion of F_2-Layer Critical Frequencies—M. L. Phillips and H. S. Moore.
11—Fluctuations of F_2 Region between Stations separated by 100 to 150 Miles—H. W. Wells.
12—Angle of Arrival and Polarization at Fort Chimo—J. E. Hogarth.
13—Use of Long-Distance Back-Scatter to Determine Skip Distance and Maximum Usable Frequency—W. Abel.
14—Application of Punch Cards to the Analysis of Multipath Ionospheric Pulse Propagation Records—W. C. Moore.
19—Wave Propagation over Rough Surfaces—W. S. Ament.
20—Simultaneous Mobile Measurement of the Field Strengths of Two V.H.F. Radio Stations over Irregular Terrain—R. S. Kirby.
21—Suppression of Waves by Zonal Screens—H. E. Bussey.
22—The Effect of Low-Level Atmospheric Conditions on Overwater Interference Patterns at Microwave Frequencies—V. R. Widerquist and J. E. Boyd.
23—V. H. F. Tropospheric Recording Measurements of Plane and Circular Polarized Waves in the Great Lakes Area—J. S. Hill and G. V. Waldo.
25—The Measurement of the Polarization of Ionospheric Reflections at Low Frequencies—R. A. Helliwell, A. J. Mallinckrodt, D. A. Campbell, and W. Snyder.
26—Theoretical and Experimental Investigation of the Polarization of Long Waves reflected from the Ionosphere—J. M. Kelso and H. J. Nearhoof.
27—Polarization Measurements of Low-Frequency Echoes—E. L. Kilpatrick.
28—A Method for obtaining the Wave Solutions of Ionospherically Reflected Long Radio Waves including All Variables and their Height Variation—J. J. Gibbons and R. J. Nertney.
34—Very-High-Frequency Propagation in the Equatorial Region—O. P. Ferrell.
36—A V.H.F. Propagation Phenomenon associated with the Aurora—R. K. Moore.
37—Phase Velocity of Vertically Polarized Electromagnetic Waves in the Diffraction Region at the Surface of a Sphere—H. Lisman.
39—Electromagnetic Energy Density and Flux—C. O. Hines.

538.566 2800

The Radiation from a Magnetic Dipole in a Spherically Stratified Atmosphere—G. Eckart. (*Arch. elekt. Übertragung*, vol. 5, pp. 113–118; March, 1951.) German version of 2009 of 1950.

621.396.11+621.396.671 2801

Application of the Compensation Theorem to Certain Radiation and Propagation Problems—Monteath. (*See* 2648.)

621.396.11 2802

The Diurnal Variation of Signal Propagation [time] and Frequency between America and Europe—A. Stoyko and N. Stoyko. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1817–1818; May 16, 1951.) A study involving 6,000 recordings, was made of the reception at Paris of time signals from Annapolis (*NNS*$_8$, 23.75 m) over the period 1944 to 1949. Variation of propagation time with hour of day was observed, signals arriving in the morning and evening taking less time than those in the middle of the day. Recordings of signals from WWV show diurnal variations of received frequency with higher values during the hours 0500 to 1500 U.T. than during the rest of the day. The contradiction between these two sets of results is removed, if it is assumed that the number of hops in the propagation path changes suddenly when the F_1 layer reaches a certain height. This assumption is supported by angle-of-arrival observations.

621.396.11 2803

Variation of the Velocity of Propagation of Radio Waves—A. Stoyko. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1916–1918; May 21, 1951.) Recordings of time signals from Annapolis at Paris, and from Pontoise at Washington, during the period 1944 to 1949, yield a value of 271,317 km for the apparent velocity, as against the value of 290,000 km deduced from observations of round-the-world signals. The relation between propagation time and sunlit length of path is investigated, and observations of sw reception, recorded during 1931 to 1939, are re-examined, taking this relation into account. Another mean value of 273,000 km is obtained for the apparent velocity. It is concluded that the apparent velocity of sw signals in the ordinary case is less than for round-the-world signals, and is variable.

621.396.11:531.74 2804

A Phase-Comparison Method of Measuring the Direction of Arrival of Ionospheric Radio Waves—W. Ross, E. N. Bramley, and G. E. Ashwell. (*Proc. IEE*, vol. 98, pp. 294–302; July, 1951.) The apparatus described uses two pairs of spaced, coaxial loop antennas at a separation of 100 m. "The signals from the antennas in a pair are amplified by means of matched receivers. The phase difference between the output signals from these receivers is displayed direct on a cathode-ray tube as the angle of inclination of the trace. With pulsed signals emitted from a suitable transmitter, and with corresponding timing equipment in the receiver, the individual rays making up the total ionospheric signal may be separated from each other. The apparatus covers the frequency band 4 to 15 mc, and the rms error of phase measurement is about 1°. Site errors, however, set a more severe limit to the accuracy of the directional measurements than do instrumental errors. In practice it is found that, for example, over an oblique path corresponding to a range of 700 km, bearings can be measured with an accuracy of about 1°, while the angle of elevation can be measured with an accuracy better than about 1½° so long as it exceeds 30°. These limitations mean that angles of elevation of *E*-layer reflections cannot be measured accurately at long range. It is possible, however, to obtain measurements of useful accuracy of the angle of elevation of *F*-layer reflections at ranges up to 1,000 km or more. Bearings can be measured accurately at all ranges and for all reflections."

621.396.11+538.566]:550.38 2805

The Effect of the Earth's Magnetic Field on Short-Wave Communication by the Ionosphere—G. Millington. (*Proc. IEE*, vol. 98, pp. 314–319; July, 1951.) Digest of IEE monograph. A re-presentation of the magneto-ionic theory for oblique transmission through the ionosphere. The quartic equation of the theory is solved graphically, and simple expressions are derived by the method of stationary phase for the differential coefficients for the group time, lateral deviation, and specific attenuation of the ray path. Numerical integration along the path is discussed with special reference to the relation of oblique to vertical transmission, and some preliminary results are given.

621.396.11:550.38 2806

Relation between the Earth's Magnetism and the Propagation of Radio Waves between Washington and Bagneux—P. Lejay, J. M. Ardillon, and G. Bertaux. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1975–1976; May 28, 1951.) Records of reception from WWV made at the Laboratoire National de Radioélectricité have been studied in conjunction with records from the magnetic observatory at Chambon-la-Forét. Close correlation is established between received field strength and magnetic activity. Graphs for February and July, 1950, show that the ionospheric disturbances, as indicated by reduced field strength, last for several days after the end of the magnetic disturbances, both in winter and summer. For weak magnetic disturbances, the field-strength fluctuations on 15 and 20 mc are parallel in winter, but opposed in summer. This difference is attributed to a reduction of muf.

The results are supported by observations of WWV 10-mc transmissions made at Turin.

621.396.11:550.385 2807

Enhanced Trans-equatorial Propagation following Geomagnetic Storms—O. P. Ferrell. (*Nature*, (London), vol. 167, pp. 811–812; May 19, 1951.) Radio-amateur reports of abnormal communication in the 6-m band during 1949 show that maximum usable frequencies are abnormally high during or immediately following the longer magnetic storms. The equatorial positive phase (891 of 1950) may continue for up to one day after the end of the storm.

621.396.11:551.510.535 2808

The Mechanism of *F*-Layer Propagated Back-Scatter Echoes—A. M. Peterson. (*Jour. Geophys. Res.*, vol. 56, pp. 221–237; June, 1951.) U.R.S.I.-I.R.E. 1950 Meeting paper. The results of multifrequency observations agree with theory if scattering is assumed to occur at the earth's surface rather than in the *E* region. The leading edge of the scatter echo corresponds to energy returned from beyond the edge of the skip-zone, while energy from the skip-zone edge returns with greater delay, the time difference becoming larger as the *F*-layer vertical incidence critical frequency is approached. The rapid buildup of echo amplitude which follows a sharply defined minimum time delay is explained as a focusing phenomenon.

621.396.11:551.510.535 2809

Change in the Nature of Medium-Wave Propagation at Sunset—J. Houtsmuller. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 85–114; discussion, pp. 115–116; March, 1951.) Investigations made under the auspices of the International Broadcasting Organization during the latter part of 1947 are reported. They provide important information regarding the influence of soil conditions at transmitter, receiver, and, in the case of double-hop transmissions, earth reflection point, and regarding the dependence of ionosphere properties on the sun's position. The observations indicate that electrons recombine not only with positive ions, but also with neutral atoms and molecules.

621.396.11:551.594.5 2810

Radio Observations of the Aurora on November 19, 1949—Gerson. (*See* 2716.)

621.396.11.029.6 2811

A Study of Tropospheric Scattering of Radio Waves—A. W. Straiton, D. F. Metcalf, and C. W. Tolbert. (Proc. I.R.E., vol. 39, pp. 643–648; June, 1951.) The following results, substantiating the scattering theory of Booker and Gordon (1757 of 1950), were observed at frequencies in the neighborhood of 100 mc: (a) at great distances, for horizontal polarization, energy was received over a considerable range of elevation angle for weak signals, but was restricted to the horizontal direction for strong signals; (b) while for vertically polarized waves, attenuation with distance agreed with the refracted-wave theory, for horizontally polarized waves, attenuation was less than expected. At about 9 kmc, the cone of received energy was restricted to a few degrees above the horizon, again in conformity with the scattering theory.

621.396.81 2812

Prediction of Short-Wave Propagation Conditions based on Ionosphere Observations—K. Rawer. (*Arch. elekt. Übertragung*, vol. 5, pp. 154–167; April, 1951.) A survey is given of methods practiced by the French government service. The upper and lower propagation frequency limits as determined, respectively, by ionosphere penetration and *D*-layer absorption, are discussed, and their relation to propagation path examined. The methods of ionosphere prediction take account of geographical position, daily and seasonal variations, and sunspot cycle. Forty literature references are included.

621.396.81 2813

Sunspots or Position of the Sun?—R. G. Sacasa. (*Rev. Telecomun.* (Madrid), vol. 5, pp. 11–22; June, 1950.) N.B.S. predictions are compared with those made by the method of 2614 of 1950. Graphs of field strength at Madrid of 10-, 15- and 20-mc transmissions from Washington for January, 1949, and January, 1950, are given.

621.396.81.029.6 2814

Predicting Performance of U.H.F. and S.H.F. Systems—E. A. Slusser. (*Electronics*, vol. 24, pp. 116–121; June, 1951.) Aids provided to simplify signal-strength calculations for uhf radiating systems for a wide range of conditions include: nomograms for (a) the power gain of antenna systems, (b) field strength as a function of distance, frequency, and antenna constants, (c) spatial attenuation between given antennas, (d) effect of ground reflections, (e) calculation of optical and radio horizons, Fresnel zone radii, and shadow losses. Receiver noise and atmospheric absorption are considered, and a detailed example of the calculations for a complete system is given.

621.396.812:551.510.535 2815

Fading of Short Wireless Waves due to the Interference between the Magneto-ionic Components—S. N. Mitra. (*Indian Jour. Phys.*, vol. 24, pp. 197–206; May, 1950.) Periodic fading of a downcoming wave may be due to interference between the two magneto-ionic components when the difference between their equivalent heights of reflection is changing at a uniform rate. A twin-loop antenna system that accepts only one of the components, is described. Its use in investigations of this type of fading is illustrated, and the effect of ionospheric irregularities with random motion and with steady drift, is discussed. See also 442 of 1950.

621.396.812.3 2816

Comparison of Field-Strength Fluctuations of the Swiss Broadcasting Stations in the First Fading Zone—C. Glinz. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 29, pp. 1–25; January 1, 1951.) The measurements made at St. Gall on transmissions from Beromünster, reported in 2036 of 1949 (Gerber and Werthmüller), are compared with measurements for the Sottens and Monte Ceneri stations over the period 1936 to 1950. The seasonal variations of amplitude for Sottens are in phase with those for Beromünster, while those for Monte Ceneri are opposed in phase. For all three stations, the amplitude of fluctuation varies inversely as sunspot activity. The results are used to calculate the absorption in the *D* layer shown in graphs. Stratification of the *E* layer is discussed. Correlation of the results with variation of the sw attenuation/frequency curve for different values of sunspot activity is attempted, but cannot be accomplished without knowledge of the nocturnal critical frequencies of the lowest ionosphere layers. After March, 1950, when the wavelength of Monte Ceneri was changed from 257 m to 539 m, the field-strength fluctuations for this station resembled those for Beromünster.

RECEPTION

621.396.621 2817

Up-to-Date High-Fidelity Receiver: Cathode-Coupled Frequency Changer—J. Rousseau. (*TSF pour Tous*, vol. 27, pp. 112–113; March, 1951.) Certain modifications to the receiver described earlier (982 of May) are explained, and the advantages of using separate tubes for frequency changer and oscillator, with cathode coupling, are discussed in detail. Circuits and component values for use with various tubes are given.

621.396.621 2818

The T.R.153 [receiver]—B. Morisse. (*Toute la Radio*, no. 153, pp. 56–59; February, 1951.) Description of a high-quality broadcast receiver, incorporating the cathamplifier circuit [78 of February (Parry)] in the audio amplifier.

621.396.621:621.396.615 2819

A High-Stability Directly-Calibrated Oscillator—G. L. Grisdale. (*Marconi Rev.*, vol. 15, pp. 89–96; 2nd quarter, 1951.) The oscillator described was designed for use in a series of communication receivers, and covers the frequency range 2 to 4 mc. It is hermetically sealed and has a frequency/temperature coefficient $<5\times10^{-6}$ per degree C. A direct-reading frequency scale allows accuracy of tuning to within ±1 kc, and the frequency changes by <1 part in 10^6 for a 1 per cent change in tube anode or filament voltage.

621.396.621:621.396.619.13 2820

F.M. Modulation in U.S.W. Broadcasting Receivers—A. Nowak. (*Telefunken Ztg.*, vol. 23, pp. 139–153; December, 1950.) Three different methods of demodulation are considered, the discussion being illustrated by reference to the basic circuits of a series of receivers produced by the Telefunken company. The operation of the ratio detector is studied in detail. An expression is derived for the optimum amplitude limiting on which the design of the diode coupling circuit can be based. The effect of adding resistance to the hf and dc circuits to stabilize the output, is discussed. The simpler demodulation method of using an AM detector, with operating point on the side of the resonance curve, is applied in the cheaper receivers.

621.396.621:621.396.619.13 2821

Commercial U.S.W. Receivers—G. Vogt. (*Telefunken Ztg.*, vol. 23, pp. 155–166; December, 1950.) Requirements for lf fidelity, hf selectivity, and sensitivity in different FM receivers for use in rebroadcasting stations, are discussed. Performance figures and details are given for four types of receiver. A monitor receiver with frequency range 87 to 110 mc has an intermediate frequency of 10.7 mc. A general-purpose unit for reception of programs for relaying has two intermediate frequencies of 16 mc and 4 mc, respectively, giving improved selectivity and general performance. Other receivers on this principle are used for multi channel reception, operating on a fixed carrier frequency, and incorporating FM feedback to the second detector through a phase-equalizing network.

621.396.621:621.396.65 2822

Design Considerations for a Radiotelegraph Receiving System—J. D. Holland. (*Proc. IEE*, vol. 98, pp. 253–262; July, 1951.) Difficulties encountered in the reception of frequency shift and AM telegraph signals over radio links are discussed. Greater transmission efficiency can be obtained by the former system. Fading and interference affect both systems and give rise to distortion. The three main components of distortion are defined, and bandwidth requirements before and after demodulation are discussed. The need for exceptional frequency stability is stressed in relation to the reduction of errors for both methods. Diversity operation is considered, and an AM diversity method is described.

621.396.621.5 2823

A Dual-Diversity Frequency-Shift Receiver—D. G. Lindsay. (Proc. I.R.E., vol. 39, pp. 598–612; June, 1951.) Reprint. See 1761 of 1950.

621.396.622 2824

Sensitivity and Operating Condition of Mixing Circuits using Diodes—H. F. Mataré. (*Arch. elekt. Übertragung*, vol. 5, pp. 57–66 and 119–124; February and March, 1951.) Mixing problems in the decimeter-wave region are discussed. On the assumption that the noise is determined mainly in the if stages, a simple circuit is derived for the equivalent noise source, from which the operating condition and corresponding sensitivity can be found. The inherent feedback of the diode is taken into account. The "equivalent conversion noise resistance" is introduced to explain the behavior when using harmonics for the mixing step. Transit-time effects are considered. A method is presented for predicting the sensitivity. Calculated values are compared with results measured at a wavelength of 55 cm, and good agreement is found.

621.396.6/.8 2825
Reception Radiophonique, Parasites (Broadcast Reception, Interference) [Book Review]—Y. Angel. Publishers: Eyrolles, Paris, France, 1950, 166 pp., 880 fr. (*Ann. Télécommun.*, vol. 6, p. 59; February, 1951. *Rev. gén. élect.*, vol. 60, p. 88; March, 1951.) Complementary to an earlier work on broadcast receivers (738 of 1950). This volume deals with interference and its suppression, and antennas and long-distance reception.

STATIONS AND COMMUNICATION SYSTEMS

621.3.015.7(083.7) 2826
Standards on Pulses: Definitions of Terms—Part I, 1951. (PROC. I.R.E., vol. 39, pp. 624–626; June, 1951.) Reprints of this Standard 51 IRE 20 S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York, N. Y., at $0.50 per copy.

621.39.001.11 2827
Adaptation of Message to Transmission Line: Part 2—Physical Interpretation—B. Mandelbrot. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 2003–2005; May 28, 1951.) A cybernetically reversible complex transmission process is investigated by breaking it down into elementary differentiation and integration processes, corresponding to the quantization of the message into words, followed by its reconstruction. The operation of the system is considered in terms of entropy variations.

621.39.001.11 2828
A Functional Equation of Information Theory—J. Ville. (*Câbles & Trans.* (Paris), vol. 5, pp. 76–83; January, 1951.) To estimate the quantity of information delivered by a given source of independent signals of unequal probabilities of occurrence, the "transmission cost" of each signal is considered. An expression is obtained for the transmission cost of a signal as a function of the number of messages transmitted. The frequencies of occurrence of the signals being known, a minimum transmission cost will be attained when the signals are actually transmitted at these frequencies. The functional equation thus permits calculation of transmission-cost deviation, and of its lower limit. For a given occurrence frequency, the amount of information delivered is closely related to its transmission cost; hence, an estimate may be obtained which depends not upon the number of possible values ascribable to the signal, but only upon the relative frequencies of these values.

621.39.001.11:621.396.619 2829
Limited-Spectrum Signals and Modulation Theory—J. Oswald. (*Câbles & Trans.* (Paris), vol. 5, pp. 54–59; January, 1951.) Amplitude modulation may be defined by means of linear "commutation" operators, which are associated with the product of any function having an integrable square, multiplied by a periodic function of limited bandwidth. By the use of these operators, the spectral composition of a signal may be defined for any modulating function of finite bandwidth. Analogous but nonlinear operators may be defined for the case of frequency modulation. See also 210 of February.

621.395.43:621.395.822.1 2830
Crosstalk in Amplitude-Modulated Time-Division-Multiplex Systems—J. E. Flood and J. R. Tillman. (*Proc. IEE*, vol. 98, pp. 279–293; July, 1951.) Mathematical analysis of the crosstalk occurring when energy proper to one channel arrives during the time allocated to another. This may arise from coupling and decoupling circuits, stray capacitance, the use of cable links, and low-pass filter networks. Both Fourier analysis and operational calculus are used, and experimental results confirming the analysis are presented. Improvements in crosstalk ratio may be obtained by inductance compensation in amplifiers. When many distorting networks are present, improvement may result from the use of nonrectangular pulses.

621.395.64 2831
Portable Repeater used for Broadcast Programmes—F. A. Peachey, G. Stannard, and C. Gunn-Russell. (*Electronic Eng.* (London), vol. 23, pp. 162–166; May, 1951.) Describes the principles and details of apparatus used by the B.B.C. for relaying over G.P.O. lines. Equalizers are connected at the input to the repeater amplifier, and in its negative feedback path. The complete assembly comprises two equalizer-amplifier units (one spare), one switching, telephone and dc testing unit, and batteries.

621.395.66 2832
Level Regulation and Telesignalling on the Paris-Brive-Bordeaux-Toulouse Coaxial Cable—R. Sueur, F. Job, and F. Le Guen. (*Câbles & Trans.* (Paris), vol. 5, pp. 40–53; January, 1951.) The system includes both automatic and tele-regulated repeaters, the equipment comprising temperature compensators, voltage-actuated distortion regulators, telecontrol and telesignalling apparatus. Circuit diagrams and explanations of operation are given.

621.396.41 2833
Note on the Coding of a Multiplex Transmission—F. H. Raymond. (*Ann. Télécommun.*, vol. 6, pp. 55–57; February, 1951.) Rules are formulated for deriving arithmetically a series of code numbers corresponding to given combinations of signals in m channels. It is shown that the simplest procedure is to derive the code signals for the m channels of a time-division multiplex system.

621.396.619.13:621.317.35 2834
The Component Theory of Calculating Radio Spectra with Special Reference to Frequency Modulation—N. L. Harvey, M. Leifer, and N. Marchand. (PROC. I.R.E., vol. 39, pp. 648–652; June, 1951.) To find the response of a filter to an FM signal, a vectorial representation of phase is used. On a plane rotating at the filter response frequency, a fixed "filter-response" vector is drawn. The FM signal vector rotates relative to this. The instantaneous angular difference between the vectors being θ, it is assumed that the filter current is proportional to the signal and to cos θ, and that the "filter-response" vector will adjust its phase to extract maximum energy from the signal during each modulation cycle. The filter current is then calculable, and the results for sinusoidal modulation agree with previous methods of analysis.

621.396.619.13:621.396.813 2835
Nonlinear Distortion in Frequency Modulation—E. Kettel. (*Telefunken Ztg.*, vol. 23, pp. 167–174; December, 1950.) Analysis, particularly of amplitude distortion in different demodulator circuits, and of frequency distortion in the transmission and reception circuits handling the FM signal. Theoretical distortion factors for harmonics of different orders are calculated in these cases. The actual distortion effects occurring in practice are due to combination tones, and their relative magnitudes are often measured instead of distortion factor. In FM systems, the relation between these two factors varies. Methods of distortion measurement for FM systems are outlined.

621.396.65 2836
Telecommunication Networks using U.H.F Radio Links—L. U. Marin and J. J. R. Moral. (*Rev. Telecomun.* (Madrid), vol. 5, pp. 23–44; June, 1950.) Part 1 critically surveys the evolution and present day state of radio-link technique, giving, in tabular form, brief details of 23 existing or projected links. Part 2 makes concrete proposals for the establishment of four links branching from Madrid, to connect the twenty principal Spanish towns. Frequency modulation is advocated. The method of determining the sites and calculating signal to noise ratio is outlined, with numerical examples in illustration. An estimate is made of the equipment required and the cost of the installation and maintenance of the system.

621.396.65:621.394.3 2837
Teleprinter System using Radio Links (Tor)—H. C. A. van Duuren. (*Tijdschr. ned. Radiogenoot.*, vol. 16, pp. 53–67; March, 1951.) A two-way system is described in which, on reception of a mutilated character, the receiver stops operating and the local transmitter signals the fault back to the remote station. The remote receiver actuates its associated transmitter to repeat the message, starting with the mutilated character. A seven-unit code is used for the transmission, with conversion from and to the usual five-unit code.

621.396.712.029.6(43) 2838
U.S.W. Broadcasting—H. Bredow. (*Telefunken Ztg.*, vol. 23, pp. 123–126; December, 1950.) Advantages of usw technique are outlined, and details of the German network operating in the 90-mc band are shown.

654.01 2839
Optimum Junction Points in Telephone and Communication Networks—K. Dorr. (*Arch. elekt. Übertragung*, vol. 5, pp. 125–134 and 197–201; March and April, 1951.) Vector analysis is used to determine the location of junction points in star and mesh systems so as to give the most economical arrangements. Mechanical projection apparatus is described for dealing with particular problems. Several examples are considered, including the broadcasting of television from aircraft, and the location of relay stations in usw radio link systems.

SUBSIDIARY APPARATUS

621.314.63 2840
Measurement of Characteristics of Rectifier Disks for Different Harmonic Contents—F. Jerrentrup. (*Z. angew. Phys.*, vol. 3, pp. 14–21; January, 1951.) The behavior of dry rectifiers under different operating conditions is investigated experimentally. Measurement circuits used for operation with steady dc, pulsed ac, etc., are shown, and the conductance characteristics of Cu_2O, Fe-Se, and Al-Se disks, are compared. Forward and backward conductance are treated separately. Typical parameters for the three types are tabulated.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 2841
Television Transmission in Local Telephone Exchange Areas—L. W. Morrison. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 280–294; March, 1951.) The use of lines for transmitting video signals over short distances between pickup camera, local studio, and points on intercity networks, is discussed. The physical and electrical properties of the various types of cable, and the design and performance of video amplifiers and equalizers now in use are described.

621.397.3.083:621.392.53 2842
Storing Video Information—A. L. Hopper. (*Electronics*, vol. 24, pp. 122–125; June, 1951.) A fused silica bar operating on 54 mc is used to provide a delay of 63.5 μs, the time required to

scan one television line, thus making possible a comparison of signal amplitudes along adjacent lines. Inclusion of a filter with a frequency response inverse to that of the bar, enables the effective pass band to be doubled. A bandwidth of 14 mc at the 3-db points is obtained, the signal to noise ratio exceeding 54 db.

621.397.5:535.62 2843
The Physiological Basis of Colour Television—Y. Le Grand. (*Onde élect.*, vol. 31, pp. 173–177; April, 1951.) Physiological considerations suggest that the seeing of color television depends on the fact that the ability of the human eye to distinguish detail is less when the color of the picture alone changes than when the level of illumination varies, rather than on the possibility that the resolving power of the eye may be lower in the blue than in the other parts of the visible spectrum.

621.397.5:535.623 2844
Field-Sequential Color Companion—E. Cohen and A. Easton. (*Electronics*, vol. 24, pp. 110–114; May, 1951.) An adapter unit for connection by a lead and plug to the video-stage tubeholder of a black-and-white-television receiver, for reception of the C.B.S.-system color transmissions. Scanning is at 144 frames per second, interlaced, and a three-color rotating disk is arranged in front of the tube face. Time-base generators, power supplies, and motor synchronizing circuits, are incorporated.

621.397.5(083.74) 2845
Conversion of Television Standards—A. Cazalas. (*Onde élect.*, vol. 31, pp. 178–183; April, 1951.) The possibility of using television signals conforming to a first standard to provide a service conforming to another standard, is examined. Difficulties arising from modification and mixing of the signals generated, and those due to interlaced scanning and phasing effects, are set out. The technique is suggested of using the transmission by the first standard to produce an electrical image on an auxiliary screen, which is then used to produce the transmission conforming to the second standard. Apparatus for achieving this is described.

621.397.7 2846
Rack-Mounted Flying-Spot Scanner—R. L. Kuehn and R. K. Seigle. (*Radio & Telev. News, Radio Electronic Eng. Suppl.*, vol. 16, pp. 7–9; April, 1951.) Description of self-containing scanning equipment designed for 2 inch ×2 inch standard slides. Circuit details and block diagrams of the cathode-ray tube scanner, and associated amplifier and photomultiplier units, are given.

621.397.6:535.88 2847
The RCA PT-100 Theater Television Equipment—R. V. Little, Jr. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56. pp. 317–331; March, 1951.) Description of commercial equipment.

621.397.6.001.4 2848
Simple Test Generator for Television Sets using 441 and 819 Lines—Bergonzat. (*See* 2778.)

621.397.61:621.396.615.142.2 2849
Five-kw Klystron U.H.F. Television Transmitter—H. M. Crosby. (*Electronics*, vol. 24, pp. 108–112; June, 1951.) Description of the G.E. Type Z 1891 klystron, which has an output >5 kw at frequencies between 475 and 890 mc, and its use in a high-power uhf television transmitter. Using this klystron as video amplifier, a power gain of 50 is obtained, with low noise, over a band of 5 mc with a response flat to within 1 db.

621.397.621.2:535.62 2850
Constructing the Tricolor Picture Tube—D.G.F. (*Electronics*, vol. 24, pp. 86–88; May, 1951.) The cathode-ray tube for receivers using the R.C.A. color-television system is described. Details are given of photoengraving processes suitable for the manufacture of the apertured metal mask, and also of the method of preparing the stencil for the phosphor-dot screen in accurate alignment with the mask so as to ensure that only the appropriately colored dots are scanned by any one gun.

621.397.621.2:535.88 2851
Projection Kinescope 7NP4 for Theater Television—L. E. Swedlund and C. W. Thierfelder. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 332–342; March, 1951.) The features of this 7-inch, 80-kv kinescope include (a) high-efficiency, low-color-shift, white fluorescent screen; (b) adequate high-voltage insulation; (c) electrostatically focused high-current beam; (d) magnetic deflection through a narrow angle to conserve deflection power and maintain focus over the picture area.

621.397.621.2.001.4 2852
Picture-Tube Performance—K. A. Hoagland. (*Electronics*, vol. 24, pp. 123–125; May, 1951.) The current required in the focusing coil of television-receiver cathode-ray tubes under given pattern conditions is discussed. It is found that wide tolerances are necessary for this current if the position of the aid is defined relative to the tube neck instead of relative to the gun. Corresponding considerations apply in the case of tubes with es focusing. The vertical resolution of picture tubes is also investigated, and on the basis of subjective tests, an optimum value is found.

TRANSMISSION

621.316.726:621.396.61 2853
Communications Technique in French Aviation—Babin. (*Onde élect.*, vol. 31, pp. 161–172; April, 1951.) An account of the methods used for frequency stabilization, both for continuous ranges and for fixed frequency channels. Three distinct techniques for using the stable frequencies generated by quartz oscillators are indicated. The quartz may be used directly, or with the interposition of a discriminator, or with locking of the controlled circuits, as in a multivibrator. Several types of equipment actually in use are described.

621.396.61/.62 2854
Army Walkie-Talkie in Mass Production—R. J. (*Electronics*, vol. 24, pp. 98–100; May, 1951.) A portable sixteen-tube FM vhf transmitter-receiver, giving speech communication over a range of 5 miles. By use of subminiature tubes and Ge diodes, the weight and size have been reduced to half that of the equipment used in the 1939 to 1945 war.

621.396.61:534.861 2855
Justification for raising the Standard of Broadcast Transmissions—Beurtheret. (*See* 2622.)

621.396.61:621.396.615.16 2856
Specification for U.S.W. F.M. Broadcasting Transmitters set up by the Broadcasting Corporations of the [Western German] Federated Republic—F. Gutzmann. (*Telefunken Ztg.*, vol. 23, pp. 127–130; December, 1950.) General discussion of harmonics, interference, modulation, audio frequency range, carrier-frequency constancy, and distortion.

621.396.619.13:621.396.615 2857
A Simple Method of Producing Wide-Band Frequency Modulation—H. Rakshit and N. L. Sarkar. (*Indian Jour. Phys.*, vol. 24, pp. 207–222; May, 1950.) Full description of the principles and experimental arrangement including a phase-discriminator afc system for carrier stability. See 2362 of 1949.

621.396.619.13:621.396.615.16 2858
The Series of Transmitters for U.S.W. Broadcasting—W. Burkhardtsmaier. (*Telefunken Ztg.*, vol. 23, pp. 131–138; December, 1950.) Illustrated description of the construction, design, and operating characteristics of various transmitter units developed for the German FM broadcasting service. The complete installation provides an output of 10 kw in the frequency range 87.5 to 100 mc. Frequency swing is 75 kc at 100 per cent modulation, distortion factor about 1 per cent. The two self-contained preamplifier stages are designed to feed a 60-Ω asymmetric load, and can be coupled to an antenna for radiation at powers of 250 w or 3 kw. Specially designed tubes are used [see 2872 (Rothe et al.) below]. Details are given of the completely enclosed, single-tube construction of the 10-kw stage.

TUBES AND THERMIONICS

535.215.4 2859
Investigations of the External Photoelectric Effect in Cuprous and Cupric Oxide—L. Meyer-Schützmeister. (*Z. Phys.*, vol. 129, pp. 148–160; February 13, 1951.) *I/V* characteristics for semiconductor and metal photocathodes were compared, using a photocell of specially adapted construction. For metal cathodes, the negative anode voltage at which cutoff occurs is independent of the material. For semiconductors, the cutoff voltage does depend on the material. The results suggest that the energy levels of the electrons released by the external photoelectric effect are lower in semiconductors than in metals, by about 0.1 v for Cu_2O, and by about 0.55 v for CuO.

621.383.4 2860
Critical Review of Semiconductor-Photo-Resistance Cells—H. Helbig. (*Elektron Wiss. Tech.*, vol. 5, pp. 57–62; February and March, 1951.) An attempt to define the useful field of application of such cells, with special reference to the CdS type. Mainly useful where intense illumination is to produce a current of the order of a milliampere. Where changes do not take place rapidly, they can be used in relay circuits, for sound-film recording, and for qualitative indication of X radiation. They are unsuitable for accurate measurements.

621.383.4 2861
Operating Limits of Photoresistances—H. Müser. (*Z. Phys.*, vol. 129, pp. 504–516; April 28, 1951.) The useful range of operation is limited by the resistance noise in the semiconductor; the magnitude of this is calculated. It is generally greater than the thermodynamic fluctuations due to the temperature radiation of the surroundings. This noise limitation cannot be overcome by using devices such as long-period galvanometers, narrow-band amplifiers, etc., to suppress the noise.

621.383.4 2862
The Characteristics and Properties of Lead Sulphide Photoresistors—B. T. Kolomiets. (*Zh. Tech. Fiz.*, vol. 21, pp. 3–11; January, 1951.) A report on an extensive experimental investigation. Results are presented in tables and curves. The main conclusion is that these resistances are highly suitable for a number of practical applications.

621.385 2863
Design of Robust, Shock-Proof Electronic Valves—G. Lewin. (*Le Vide*, vol. 6, pp. 974–978; March, 1951.) The effect of mechanical shock on a tube is analyzed. Its magnitude depends on the rigidity of the assembly. On this basis, design fundamentals are developed. Data are given regarding the physical properties of the metals used, and the correct use of fragile materials such as glass.

621.385.001.4 2864
Radio Valve Life Testing—Brewer. (*See* 2773.)

621.385.029.64/.65 2865
New Travelling-Wave Valves for Microwaves—H. H. Klinger. (*Arch. elekt. Über-*

tragung, vol. 5, pp. 167–168; April, 1951.) Two tubes are briefly described, whose operation is based on direct exchange of energy between electrons and electromagnetic wave. Both use arrays of metal dipoles of delay-lens type to slow down the waves. One of the tubes is a modified drift-space klystron; the other uses a reflector to bend the electron beam, which is injected at an angle to the delay field. Inherent noise is low; useful wavelength range is probably limited to 1 to 2 cm.

621.385.029.64 2866
On the Theory of Electron Wave Tubes—O. E. H. Rydbeck and S. K. H. Forsgren. (*Chalmers Tekn. Högsk. Händl.*, no. 102, 29 pp.; 1951. In English.) Equations are derived first for the waves produced in a single-beam tube with strong axial magnetic focusing. Both beam waves and guide waves are present. When the beam is shot into a neutral ionized gas, strong traveling waves may be excited. The effect of reducing the magnetic field is investigated. Analysis for a two-beam tube indicates that klystron waves, plasma waves, and interaction waves, are all present, the distribution of energy between them depending on the manner of excitation. The theoretical results are in agreement with measurements on a 3-kmc two-beam tube developed at the Chalmers Research Laboratory.

621.385.032.212 2867
A New Modulated Light Source—J. A. Darbyshire. (*Electronic Eng.* (London), vol. 23, pp. 167–169; May, 1951.) The Ferranti GMC 6 "crater lamp" is described and compared with the older MAC 4. It has a higher light output for given current, and better linearity. Coating of the viewing window by evaporated barium has been eliminated. Operating conditions for particular applications are stated.

621.385.15 2868
Secondary Electron Multipliers—N. Schaetti. (*Z. angew. Math. Phys.*, vol. 2, pp. 123–158; May 15, 1951.) A survey paper reviewing secondary emission phenomena, the production of suitable emitting surfaces, types and operating characteristics of photo-multipliers, and applications in various fields. A bibliography comprising 80 items is provided.

621.385.18 2869
Controllable Gas Diode—E. O. Johnson. (*Electronics*, vol. 24, pp. 107–109; May, 1951.) The object of the device, which is called a "plasmatron," is to combine the continuous control afforded by the vacuum tube with the low impedance of the thyratron. It is operated at an anode potential too low to cause ionization. A discharge is set up between the anode and an auxiliary cathode, the discharge current being controlled by a small, conventional hard tube. The current flowing between anode and main cathode can be controlled by varying the auxiliary discharge current with fairly good linearity, a current ratio of 90:1 having been obtained.

621.385.2 2870
Thermodynamics of a Two-Cathode System—H. Dormont. (*Le Vide*, vol. 6, pp. 979–984; March, 1951.) The results obtained by Champeix (2586 of November) are shown to be valid only for the limiting case where the temperature difference between the two electrodes is small compared with their mean temperature. Discussion of the source of thermionic emission in oxide cathodes is continued. Further experimental data are required.

621.385.2:537.525.92 2871
The Space-Charge Smoothing Factor—C. S. Bull. (*Proc. IEE*, vol. 98, p. 278; July, 1951.) Discussion on 2058 of September.

621.385.3/.4 2872
U.S.W. Transmitting Valves—H. Rothe, W. Engbert and H. Kraft. (*Telefunken Ztg.*, vol. 23, pp. 175–182; December, 1950.) For the final power stages of transmitters operating in the 100-mc band (a) the grounded-grid triode, and (b) the grounded-cathode tetrode, are of particular interest. To maintain the electrode connections as lossless as possible, disk seals are most suitable. The high power necessarily drawn by the control circuit in (a) is largely available as useful power at the anode. In (b) the much lower power required increases with rising frequency due to the inductance of the cathode leads. The construction and operating characteristics of two 10-kw triodes and three tetrodes for powers of 0.25 kw, 1 kw, and 3 kw, are shown.

621.385.38 2873
Grid Current and Grid Emission Studies in Thyratrons—the Trigger-Grid Thyratron—L. Malter and M. R. Boyd. (PROC. I.R.E., vol. 39, pp. 636–643; June, 1951.)

621.385.38:537.315 2874
The Distribution of the Electric Field in a Three-Electrode Gas-Discharge Tube with a Large Back Voltage—V. D. Andreev, L. E. Levina, and B. G. Mendelev. (*Zh. Tekh. Fiz.*, vol. 21, pp. 149–154; February, 1951.) The potential distribution in the plasma between the anode and grid of a Hg-vapor tube during the reverse half-cycle was investigated by taking oscillograms of probe currents. On the basis of the results obtained, the division of the voltage between the electrodes is discussed.

621.396.615.142:[537.525.6:538.56 2875
Electron Plasma Oscillations—G. Wehner. (*Jour. Appl. Phys.*, vol. 22, pp. 761–765; June, 1951.) "Electron plasma oscillations are excited by a beam of fast electrons in a stabilized low-pressure mercury discharge. Probe measurements reveal that the uhf fields are localized in thin layers, the plasma density and frequency of which follow Langmuir's law. The beam of fast electrons traversing such an oscillation layer becomes velocity modulated, and excitation conditions result from drift time and bunching considerations similar to those in a klystron. A sealed-off tube described covers a frequency range between 800 and 4,000 mc (five modes) without changing or matching any resonance circuit."

621.397.61:621.396.615.142.2 2876
Five-kw Klystron U.H.F. Television Transmitter—Crosby. (*See* 2894.)

621.396.822 2877
The Nature of Currents producing the Main Component of Radio Valve Noise—A. R. Shul'man. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1505–1508; December, 1950.) A noise component of frequency between 50 and 100 cps usually appears in the anode circuit of a tube with an indirectly heated cathode. Experiments were conducted to determine the origin of this component, and it is concluded that it is due to current between the heater and the cathode produced by thermal emission of electrons from the heater on to the cathode, and sometimes from the cathode on to the heater. The value of the current is affected by impurities, and can be considerably reduced by a suitable treatment of the heater material.

621.396.822 2878
Induced Grid Noise in Triodes—A. van der Ziel. (*Wireless Eng.*, vol. 28, pp. 226–227; July, 1951.) The Llewellyn-Peterson theory of tube operation at uhf (2578 of 1944) is applied to extend Bell's formula (1833 of 1950) for induced grid noise in triodes and pentodes. Better agreement with experimental results is thus obtained.

621.396.822:621.396.645.3.029.42 2879
Experimental Investigation of Fluctuation Phenomena Limiting the Operation of Direct-Voltage and Very-Low-Frequency Amplifiers—K. Kronenberger. (*Z. angew. Phys.*, vol. 3, pp. 1–5; January, 1951.) Results are presented of noise measurements on tube amplifiers under different operating conditions. Fluctuations of heater and anode supply voltages constitute the most important source of noise. When these voltages are nearly constant, i.e., vary by $<10^{-4}$ per cent, in the oxide-cathode tube operating below 0.1 cps, the effect of emission-drift is the most serious, and above 0.1 cps it is the flicker effect. For the tungsten cathode, the anomalous flicker effect determines the sensitivity. Since the noise voltage of an unwanted signal is generally higher at the low-frequency end of the spectrum, advantages are gained by mechanical conversion of a direct voltage to high-frequency voltage before amplification.

621.385.832 2880
Le Tube à Rayons Cathodiques [Book Review]—L. Chrétien. Publishers: Chiron, Paris, France, 1950, 191 pp., 585 fr. (*Ann. Télécommun.*, vol. 6, p. 58; February, 1951.) A manual for practicians and students, dealing with the use of the cathode-ray tube in measurement, radar, and television applications.

MISCELLANEOUS

621.396.001.5 2881
Engineering Research of the British Broadcasting Corporation—(*Engineering* (London), vol. 171, pp. 793–795; June 29, 1951.) A brief survey of investigations in progress, including work on high-definition and color television, interference between transmitters sharing a frequency channel, impulsive interference, and methods of economizing on bandwidth; field-strength mapping; design and performance of transmitting antennas, using models for much of the work; development of broadcast and special-purpose receivers; electroacoustics. See also *Elec. Rev.* (London), vol. 148, pp. 1341–1343; June 29, 1951, and *Elec. Times*, vol. 119, pp. 1098–1100; June 28, 1951.

621.396.001.57 2882
Technical Development and Research by Telefunken during the War—H. Lux. (*Telefunken Ztg.*, vol. 23, pp. 11–26; September, 1950.) A brief review is given of advances in the following branches: tube development, including hf, metal/ceramic and tr switching tubes, high-power types with thoria cathodes, and transit-time types; medium- and high-power transmitters and measuring equipment; decimeter-wave technique in communications, including radio links; radar and radio navigation; piezoelectric crystals; hf heating technique; general research on propagation in the wavelength range 20 m to 10 cm, on antennas, on receiver sensitivity, and on hf measuring techniques; af techniques and electroacoustics; and techniques related to television.

621.38 2883
Electronic Fundamentals and Applications [Book Review]—J. D. Ryder. Publishers: Prentice-Hall, New York, N. Y., 806 pp., $9.00. (*Radio & Telev. News, Radio Electronic Eng. Suppl.*, vol. 16, p. 26; April, 1951.) "Written to give the student a knowledge of . . . the physical principles underlying electron tubes, the characteristics of the tubes themselves, and the electrical circuits in which they are used . . ."

621.39 2884
Telecommunications Principles [Book Review]—R. N. Renton. Publishers: Pitman, London, Eng., 450 pp., 37s. 6d. (*Electrician*, vol. 146, p. 1131; April 6, 1951.) Covers the City and Guilds examination syllabus, Grades I, II, and III. Many examples are worked out. "The work can be most highly recommended."

PROCEEDINGS OF THE I.R.E.

Published Monthly by
The Institute of Radio Engineers, Inc.

Index to Volume 39—1951

The Institute of Radio Engineers, Inc.
1 East 79 Street
New York 21, N.Y.

TABLE OF CONTENTS

GENERAL INFORMATION

The Institute

The Institute of Radio Engineers serves those interested in radio and allied electronics and electrical-communication fields through the presentation and publication of technical material.

Membership has grown from a few dozen in 1912 to approximately 30,000. There are several grades of membership, depending on the qualifications of the applicant, with dues ranging from $5.00 per year for Students to $15.00 per year for Members, Senior Members, Fellows, and Associates of more than five years' standing.

The PROCEEDINGS is sent to members of record on the date of publication. IRE Standards, as listed on Cover IV, are in stock and are available for those who wish to buy them at the cost shown opposite the name of the Standard.

The PROCEEDINGS

The PROCEEDINGS has been published without interruption from 1913, when the first issue appeared. Over 4,000 technical contributions have been included in its pages, portraying a currently written history of developments in both theory and practice. The contents of every paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Textual material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

The first issue of the PROCEEDINGS was published in 1913. Volumes 1, 2, and 3 comprise four issues each. Volume 4 through volume 14 contain six numbers each, and each succeeding volume is made up of twelve issues.

In 1939, the name of the PROCEEDINGS of The Institute of Radio Engineers was changed to the PROCEEDINGS OF THE I.R.E. and the size of the magazine was enlarged from six by nine inches to eight and one-half by eleven inches.

Subscriptions

Annual subscription rates for the United States of America, its possessions, and Canada, $18.00; to college and public libraries when ordering direct, $13.50; other countries, $1.00 additional for postage.

Contents of Volume 39—1951

Volume 39, Number 1, January, 1951

Proceedings of the I.R.E.

Volume 39, Number 2, February, 1951

Proceedings of the I.R.E.

Volume 39, Number 2, February, 1951 (Cont'd.)

Volume 39, Number 3, March, 1951

Proceedings of the I.R.E.

Volume 39, Number 3, March, 1951 (Cont'd.)

Volume 39, Number 4, April, 1951
PROCEEDINGS OF THE I.R.E.

Volume 39, Number 4, April, 1951 (Cont'd.)

Volume 39, Number 5, May, 1951
PROCEEDINGS OF THE I.R.E.

Volume 39, Number 5, May, 1951 (Cont'd.)

Volume 39, Number 6, June, 1951

Proceedings of the I.R.E.

Volume 39, Number 6, June, 1951 (Cont'd.)

Volume 39, Number 7, July, 1951

Proceedings of the I.R.E.

Volume 39, Number 8, August, 1951
Proceedings of the I.R.E.

Volume 39, Number 9, September, 1951
Proceedings of the I.R.E.

Volume 39, Number 9, September, 1951 (Cont'd.)

Volume 39, Number 10, October, 1951
Proceedings of the I.R.E.

Volume 39, Number 10, October, 1951 (Cont'd.)

Volume 39, Number 11, November, 1951

PROCEEDINGS OF THE I.R.E.

Volume 39, Number 11, November, 1951 (Cont'd.)

Volume 39, Number 12, December, 1951

PROCEEDINGS OF THE I.R.E.

INDEX TO AUTHORS

Numbers refer to the chronological list. Light-face type indicates papers, **bold-face type indicates discussions and correspondence,** and *italics refer to books and book reviews.*

INDEX TO BOOK REVIEWS

INDEX TO SUBJECTS

This listing includes technical, sociological, economic, and general papers. Numbers refer to chronological list.

O

P

Q

R

S

NONTECHNICAL INDEX

IRE Professional Groups

Miscellaneous

Obituaries

Report of the Secretary

Representatives in Colleges

Representatives on Other Bodies

Sections

Standards

Back Copies

The Instit ndeavors to keep on hand a supply of back copies (PROCEEDINGS for sale for the convenience of thos(do not have complete files. However, some issues at of print and cannot be provided.

All availalck issues of the PROCEEDINGS OF THE I.R.E. are p at $2.25 per copy to individual non-members; $1.65 per copy to public libraries, colleges, and subscription agencies; and $1.00 per copy to Institute members. Price includes postage in the United States and Canada; postage to other countries is ten cents per copy additional.

1913–1915 Volumes 1–3 Quarterly

1913 Vol. 1 January (a reprint)

1916–1926 Volumes 4–14 Bimonthly

1920	Vol. 8	it, October
1921	Vol. 9	it, October, December
1922	Vol. 10	June, August, October
1923	Vol. 11	ary, April, October
1924	Vol. 12	August, October
1925	Vol. 13	April, June, October, December
1926	Vol. 14	All 6 issues

1927–1938 Volumes 15–16 Monthly

1927	Vol. 15	October, December
1928	Vol. 16	ary to December, inc.
1929	Vol. 17	May, June, November
1930	Vol. 18	to December, inc.
1931	Vol. 19	y to November, inc.
1932	Vol. 20	y, March to December, inc.
1933	Vol. 21	All 12 issues
1934	Vol. 22	All 12 issues
1935	Vol. 23	All 12 issues
1936	Vol. 24	January, February, March
1937	Vol. 25	February to December, inc.
1938	Vol. 26	All 12 issues

1939–1950 Volumes 27–39 Monthly

New Format—Large size

1939	Vol. 27	y, February, March, May, December
1940	Vol. 28	issues
1941	Vol. 29	issues
1942	Vol. 30	y to May, inc., October to December, inc.
1943	Vol. 31	ary to December, inc.
1944	Vol. 32	y, April, August to December, inc.
1945	Vol. 33	y to June, inc., September, November, De-
1946	Vol. 34	January, February, April to December, inc.
1947	Vol. 35	January to July, inc., September to December, inc.
1948	Vol. 36	All 12 issues
1949	Vol. 37	January to May, inc., July to December, inc.
1950	Vol. 38	All 12 issues
1951	Vol. 39	All 12 issues

Cumulative es covering the periods 1913–1942 and 1943–1947 are available at the same prices as back issues of PROCEEDI

Proceedings Binders

Sturdy bir of blue Spanish-grain fabricoid with gold letterin ich will serve either as temporary transfers or an anent holders for the PROCEEDINGS may be obtai the large size only (1939 to date) at the following s: $3.00 for plain; $3.50 for Vol. No. or year, or your name imprinted; $4.00 for Vol. No. and name or name and year imprinted. Postage is prepaid in United States or Canada; postage to other countries, 50 cents additional.

Membership Emblems

The IRE em is available to members in three useful forms; apel button, the pin, and the watch charm. Each 14-karat gold, enameled in color to designate the of membership.

Gr	Background Color
Fellow	Gold
Senior ber	Dark Blue
Membe	Light Blue
Associa	Maroon
Student	Green

The lapel button has a screw back with jaws which grip the cloth of the coat. These are priced at $3.00 each. The pin, at $3.50 each, is provided with a safety catch. The watch charm, enameled on both sides and equipped with a suspension ring for attaching to a watch charm or fob, is $5.50.

Prices on emblems are the same for all grades of membership and include Federal tax, postage, and insurance or registered-mail fee.

Remittance should accompany your order.

Please do not send cash except by registered mail.

'E INSTITUTE OF RADIO ENGINEERS, INC.

1 East 79 Street, New York 21, New York

Current IRE Standards

In addition to the technical papers published in the PROCEEDINGS OF THE I.R.E., Standards on vario is subjects have bnted. These are available at the prices listed below. 20% discount will be allowed on orders for 100 or more copies mailed to oress.

Price

48 IRE 2, 11, 15. S1 **Standards on Antennas, Modulation Systems, and Transmitters: Definitions of Terms, 1948.**
(vi+25 pages, 8½×11 inches) $0.75

48 IRE 2. S2 **Standards on Antennas: Methods of Testing. 1948. (ASA C16.11-1949).**
(vi+18 pages, 8½×11 inches) $0.75

42 IRE 3. S1 **American Recommended Practice for Volume Measurements of Electrical Speech and Program Waves. (ASA C16.5-1942).**
(8 pages, 7¾×10⅝ inches) $0.20

50 IRE 4. S1 **Standards on Circuits: Definitions of Terms in Network Topology, 1950.**
Reprinted from the January, 1951, PROCEEDINGS (pp. 27–29) $0.50

38 IRE 6. S1 **Standards on Electroacoustics, 1938. Definitions of Terms, Letter and Graphical Symbols, Methods of Testing Loud Speakers**
(vi+37 pages, 6×9 inches) $0.50

42 IRE 6. S1 **American Recommended Practice for Loud-speaker Testing. (ASA C16.4-1942).**
(12 pages, 7¾×10⅝ inches) $0.25

51 IRE 6. S1 **Standards on Electroacoustics: Definitions of Terms, 1951.**
Reprinted from the May, 1951, PROCEEDINGS (pp. 509–532) $1.00

50 IRE 7. S1 **Standards on Electron Tubes: Definitions of Terms, 1950.**
Reprinted from the April, 1950, PROCEEDINGS (pp. 426–438) $0.75

50 IRE 7. S2 **Standards on Electron Tubes: Methods of Testing, 1950.**
Part I reprinted from August, 1950, PROCEEDINGS (pp. 917–948); Part II reprinted from September, 1950, PROCEEDINGS (pp. 1079–1093) $1.25

50 IRE 8. S1 **Standards on Electronic Computers: Definitions of Terms, 1950.**
Reprinted from the March, 1951, PROCEEDINGS (pp. 271–277) $0.75

42 IRE 9. S1 **Standards on Facsimile: Definitions of Terms, 1942.**
(vi+6 pages, 8½×11 inches) $0.20

43 IRE 9. S1 **Standards on Facsimile: Temporary Test Standards, 1943**
(iv+8 pages, 8½×11 inches) $0.20

49 IRE 12. S1 **Standards on Radio Aids to Navigation: Definitions of Terms, 1949.**
Reprinted from the December, 1949, PROCEEDINGS (pp. 1364–1371) $0.60

45 IRE 14. S1 **Standards on Piezoelectric Crystals: Recommended Terminology, 1945.**
(iv+4 pages, 8½×11 inches) $0.20

49 IRE 14. S1 **Standards on Piezoelectric Crystals, 1949.**
Reprinted from the December, 1949, PROCEEDINGS (pp. 1378–1395) $0.80

49 IRE 16. S1 **Standards on Railroad and Vehicular Communications: Methods of Testing, 1949.**
Reprinted from the December, 1949, PROCEEDINGS (pp. 1372–1375) $0.50

47 IRE 17. S1 **Standards on Radio Receivers: Methods of Testing Frequency-Modulation Broadcast Receivers, 1947 (ASA C16.12-1949).**
(vi+15 pages, 8½×11 inches) $0.50

48 IRE 17. S1 **Standards on Radio Receivers: Methods of Testing Amplitude-Modulation Broadcast Receivers, 1948.**
(vi+24 pages, 8½×11 inches) $1.00

49 IRE 17. S1 **Tests for Effects of Mistuning and for Downward Modulation.**
1949 Supplement to 47 IRE 17. S1. Reprinte the December, 1949, PROCEEDINGS (pp. 1376–1 ... $0.25

51 IRE 17. S1 **Standards on Radio Receivers: ield Method of Measurement of Spurious Radrom Frequency Modulation and Television Bro Receivers, 1951.**
Reprinted from the July, 1951, PROCEEDING803–806) $0.50

51 IRE 20. S1 **Standards on Pulses: Definitions ns—Part 1, 1951.**
Reprinted from the June, 1951, PROCEEDING524–626) $0.50

51 IRE 20. S2 **Standards on Transducers: Des of Terms, 1951.**
Reprinted from the August, 1951, PROCEEDIN897–899) $0.50

48 IRE 21. S1 **Standards on Abbreviations, Grapym-bols, Letter Symbols, and Mathematical 1948.**
(vi+21 pages, 8½×11 inches) $0.75

49 IRE 21. S1 **Standards on Designations forical, Electronic and Mechanical Parts and thebols, 1949.**
Reprinted from February, 1950, PROCEEDIN118–124) $0.60

51 IRE 21. S1 **Standards on Abbreviations of Electronic Terms, 1951.**
Reprinted from the April, 1951, PROCEEDIN397–400) $0.50

47 IRE 22. S1 **Standards on Television: Methodsting Television Transmitters, 1947.**
(vi+18 pages, 8½×11 inches) $0.75

48 IRE 22. S1 **Standards on Television: Methodsting Television Receivers, 1948. (ASA C16.13-**
(vi+32 pages, 8½×11 inches) $1.00

48 IRE 22. S2 **Standards on Television: Definitiorrms, 1948.**
(iv+4 pages, 8½×11 inches) $0.50

50 IRE 23. S1 **Standards on Television: Methoeasurement of Television Signal Levels, Res and Timing of Video Switching Systems, 195**
Reprinted from the May, 1950, PROCEEDIN551–561) $0.70

50 IRE 23. S2 **Standards on Television: Methoeasurement of Time of Rise, Pulse Width,ulse Timing of Video Pulses in Television, 195**
Reprinted from the November, 1950, PROCE (pp. 1258–1263) $0.75

50 IRE 23. S3 **Standards on Television: Metholeasurement of Electronically Regulated Powplies, 1950.**
Reprinted from the January, 1951, PROCE (pp. 29–35) $0.75

50 IRE 24. S1 **Standards on Wave Propagationitions of Terms, 1950.**
Reprinted from the November, 1950, PROCE (pp. 1264–1268) $0.60

45 IRE 24. S1 **Standards on Radio Wave PropaDefinitions of Terms Relating to Guided Wa45.**
(iv+4 pages, 8½×11 inches) $0.20

Normas Sobre Receptors de Radio, 1938.* A -language translation of "Standards on Radiivers, 1938," by the Buenos Aires Section of Theite of Radio Engineers. (vii+64 pages, 6×9 inch.... .... **Two Argentios (Postpaid)**

* Not carried in Stock at IRE Headquan New York. Obtainable only from Seior Domingo Arbóor of *Revista Telegrafica*, Peru, 165, Buenos Aires, Argentina.

PRICES ARE NET AND INCLUDE POSTAGE TO ANY COUNTRY.
INCLUDE REMITTANCE WITH ORDER AND ADDRESS.

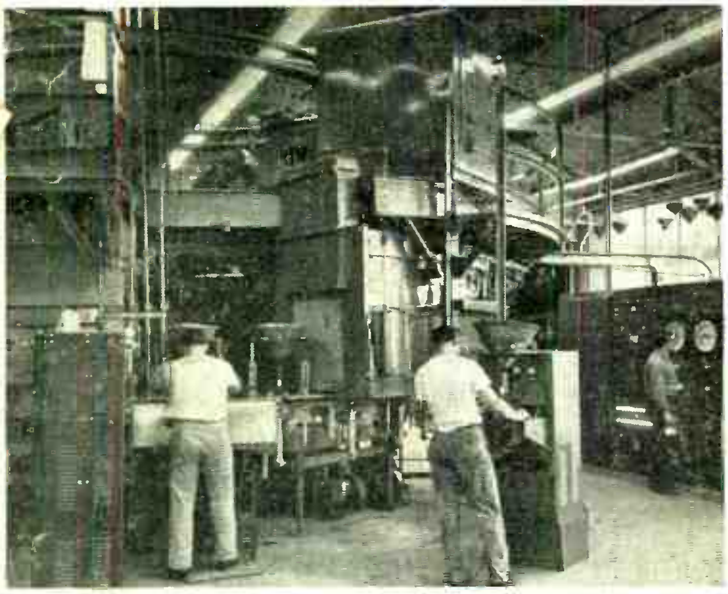
THE INSTITUTE OF RADIO ENGINEERS, IN

1 East 79 Street, New York 21, New York

Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

June, 1951
Volume 39 Number 6



Sylvania Electric Products Inc.

LARGER MACHINERY FOR LARGER TUBES

Television picture tubes are produced on an automatic rotary baking and exhausting machine 30 feet in diameter, and provided with a service cat-walk 10 feet above floor level. Oil diffusion pumps attached to the 24 heads operate continuously.

PROCEEDINGS OF THE I.R.E.

Table of Contents, Indicated by Black-and-White Margin, Follows Page 32A

The Standards on Pulses: Definitions of Terms—Part I, 1951, appear in this issue.

The Institute of Radio Engineers

for every application

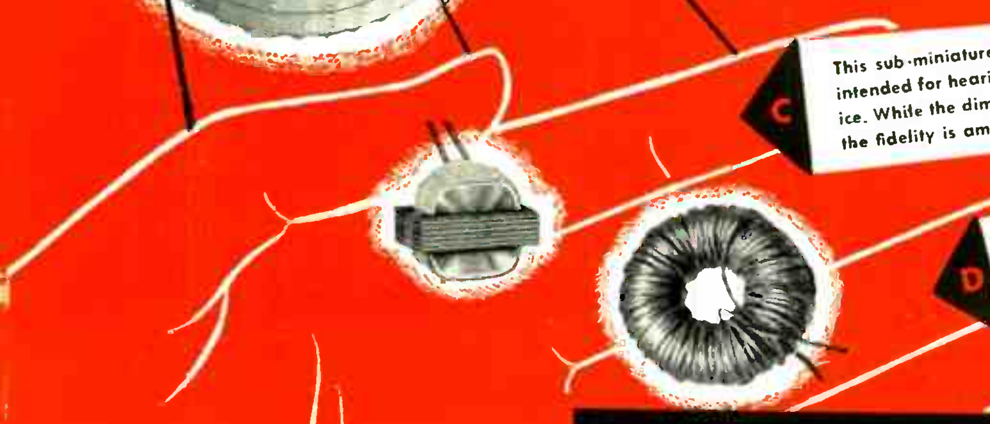
While the catalogue line of UTC components covers a wide variety of applications, many people are not familiar with the full range of products produced by UTC. It is impossible to describe the thousands of special UTC designs as they become available. The illustrations below are intended to indicate some of the range in size of these special products.

A This 100 cubic foot modulation transformer is for 50 Kw. broadcast service. Frequency response flat from 30 to 20,000 cycles.

B The high Q toroid coil shown is 12" in diameter. It operates in a 50 Kw. circuit at supersonic frequency.

C This sub-miniature (.18 cubic inch) output transformer is intended for hearing aid and other extreme compact service. While the dimensions are only 7/16" x 9/16" x 3/4", the fidelity is ample for voice frequency requirements.

D This sub-miniature (.18 cubic inch) permalloy dust core toroid is available in a wide range of inductances, and for frequencies from 1,000 cycles to 50 Kc.



United Transformer Co.

150 VARICK STREET • NEW YORK 13, N. Y.

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

Hermetically-sealed CUP CERAMIC CAPACITORS for Precision tuned Circuits

With SPRAGUE HERLEC Dependability

Top-notch capacitance stability, high Q, and excellent retrace characteristics are only part of the story on Sprague-Herlec Cup Ceramic Capacitors.

In addition, they are small, hermetically-sealed, and easy to mount securely against the effects of vibration and shock. Their extreme stability and compactness make them unexcelled for rigid frequency control applications and as reference capacitance standards in either laboratories or electronic circuits. Low self-inductance likewise makes them valuable in v-h-f bypass applications.

With Sprague-Herlec metal cup ceramic capacitors in precision circuits, it is often possible to control the capacitance tolerance within ± 1% and the temperature coefficient tolerance within ± 10 parts per million!

For complete details write for Engineering Bulletin 603 to either Sprague or Herlec

HERLEC CORPORATION

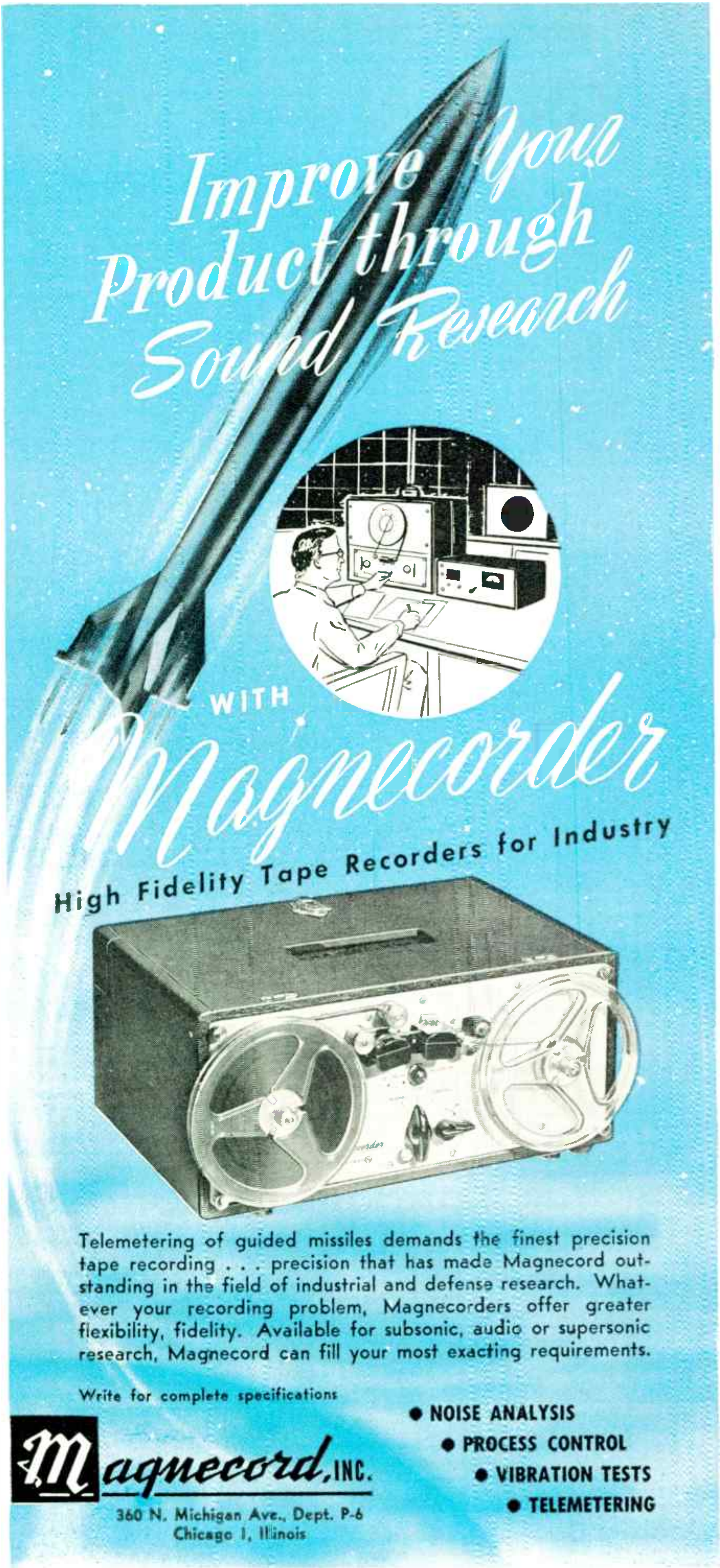
422 N. 5th Street, Milwaukee, Wisconsin; A wholly-owned subsidiary of the

SPRAGUE ELECTRIC COMPANY

North Adams, Massachusetts

Proceedings of the I.R.E. June, 1951, Vol. 39, No. 6. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of Contents will be found following page 32A



Important Meetings!

June 20-22
IRE 7th Regional Conference, Seattle, Washington

Information: Chairman, Mr. A. E. Harrison, University of Washington Campus, Seattle, Washington.

June 21-22
Electron Devices Conference, University of New Hampshire Durham, New Hampshire

Information: Chairman, Prof. Herbert Reich, Dunham Laboratory, Yale University, New Haven, Conn.

June 25-29
1951 Summer General Meeting of AIEE, Royal York Hotel, Toronto, Canada.

June 27-28
1951 Annual IAS Summer Meeting, 7660 Beverly Blvd., Los Angeles, California.

June 28-30
Institute of Navigation National Meeting, Hotel New Yorker, New York, New York.

August 22-24
1951 IRE West Coast Convention and Pacific Electronic Exhibit, Civic Auditorium, San Francisco, California.

Heckert Parker, Exhibits Manager
600 Beach St., San Francisco 9

September 10-14
Instrument Society of America Convention and Instrument Show
(Houston IRE Section Participating)
Civic Auditorium
Houston, Texas

Information: Mr. Richard Rimbach, Exec. Secr., 921 Ridge Ave., N.S. Pittsburgh 12, Pennsylvania

October 22-24
1951 National Electronics Conference, Edgewater Beach Hotel Chicago, Illinois

Exhibits: R. M. Soria, American Phenolic Corp., 1830 South 54th Ave. Chicago 50, Illinois

October 29-31
Radio Fall Meeting, King Edward Hotel Toronto, Ont., Canada

November 1-3
Audio Fair, Hotel New Yorker New York City, New York

Exhibits: Mr. Harry Reizes, 342 Madison Ave., New York 17

IRE Regional Meetings and Shows Speed Electronic Progress

Complete Coverage!

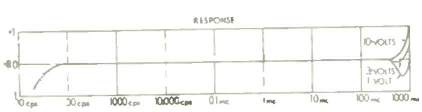
2 to 700,000,000 cps

-hp- precision voltmeters for every ac voltage measuring need!

From 2 cps to 700 mc, there's an accurate, easy-to-use -*hp*- voltmeter for any voltage measuring job. You can choose from 5 precision instruments (including a battery-operated portable unit) the dependable -*hp*- voltmeter that exactly fills your need. Each gives you familiar -*hp*- operating characteristics of high sensitivity, wide range, broad applicability, time-saving ease of operation. -*hp*- also provides a complete line of voltmeter accessories—voltage dividers, connectors, shunts and multipliers—to extend the useful range of your equipment. For complete details, see your -*hp*- sales representative or write direct.

New -hp- 410B Vacuum Tube Voltmeter

Gives same wide range and flat response performance as -*hp*- 410A voltmeter, but sets new standard of mechanical convenience, ease of operation, minimum bench space. Readily detachable probe leads fit in handy compartment in new, compact, streamlined case. Special diode probe design places capacity of approximately 1.3 $\mu\mu$fd across circuits under test. Shunt impedance is extremely high—10 megohms at low frequencies—thus circuits under test are not disturbed and true voltage readings are assured. New -*hp*- 410B provides 1 db accuracy from 20 cps to 700 mc; and may be used as a voltage indicator up to 3,000 mc. Also serves as audio or dc voltmeter or ohmmeter.



Response, -*hp*- 410B Voltmeter

INSTRUMENT	PRIMARY USES	FREQUENCY RANGE	VOLTAGE RANGE	INPUT IMPEDANCE	PRICE
-hp- 400A	General purpose ac measurement	10 cps to 1 mc	.005 to 300v 9 ranges	1 megohm 24 $\mu\mu$fd shunt	$185.00
-hp- 400B	Low frequency ac measurements	2 cps to 100 kc	.005 to 300v 9 ranges	10 megohms 24 $\mu\mu$fd shunt	$195.00
-hp- 400C	Wide range ac measurements High sensitivity	20 cps to 2 mc	.0001 to 300v 12 ranges	10 megohms 15 $\mu\mu$fd shunt	$200.00
-hp- 404A	Portable, battery operated	2 cps to 50 kc	.0005 to 300v 11 ranges	10 megohms 20 $\mu\mu$fd shunt	$185.00
-hp- 410B	Audio, rf, VHF measurements, dc voltages; resistances	20 cps to 700 mc	0.1 to 300v 7 ranges	10 megohms 1.3 $\mu\mu$fd shunt	$245.00

-hp- 400C Vacuum Tube Voltmeter

General purpose precision voltmeter offering wide range, high sensitivity, high stability. Quick-reading linear meter scale shows RMS volts or dbm direct from −72 dbm to +52 dbm. Broad usefulness includes direct noise or hum measurements, transmitter and receiver voltages, audio, carrier or supersonic voltages, or power gain. Also may be used as 54 db amplifier to increase signal level to oscilloscopes, recorders, power amplifiers, etc.

-hp- 404A Battery-Operated Voltmeter

Precision vacuum tube instrument for general voltage measurement where ac power is not available. Compact, portable, splash-proof—ruggedly constructed for field operations. Wide voltage range permits all types of measurements including remote broadcast line and carrier checks, strain gauge system tests, telemetering and geophysical circuit measurements, etc. In the laboratory, offers completely hum-free measurements of very low noise level.

HEWLETT-PACKARD CO.

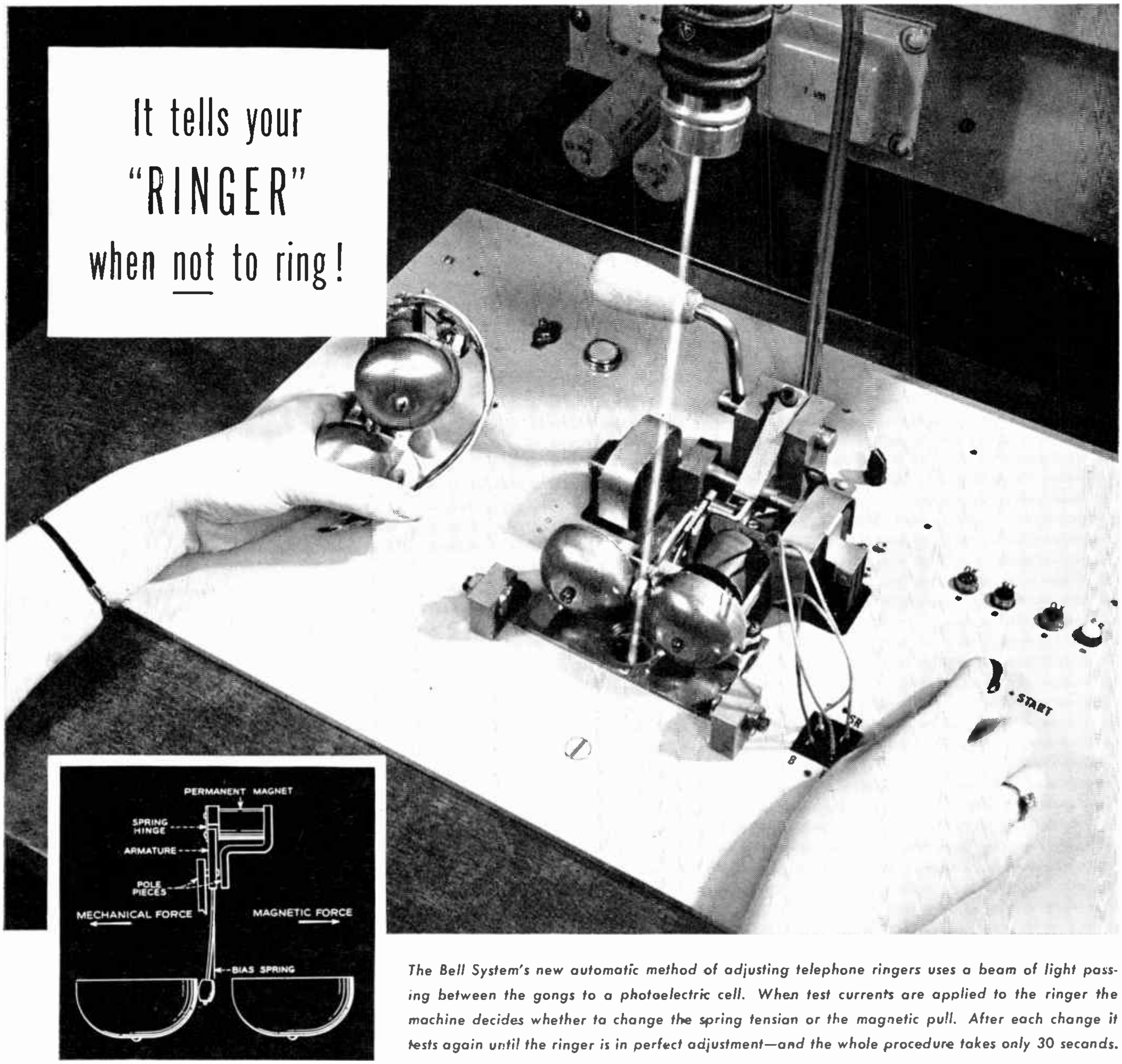
2251D Page Mill Road • Palo Alto, Calif., U.S.A.

Sales representatives in principal areas

Export: Frazar & Hansen, Ltd.
San Francisco • New York • Los Angeles

HEWLETT-PACKARD hp INSTRUMENTS

It tells your "RINGER" when not to ring!



The Bell System's new automatic method of adjusting telephone ringers uses a beam of light passing between the gongs to a photoelectric cell. When test currents are applied to the ringer the machine decides whether to change the spring tension or the magnetic pull. After each change it tests again until the ringer is in perfect adjustment—and the whole procedure takes only 30 seconds.

To you, it's your familiar telephone bell. To telephone engineers, it's a "ringer." And it has two jobs to do. It must ring, of course, when someone calls you. And it must overlook the numerous electrical impulses which do not concern it, such as those sent out by your dial.

Ability to respond to some impulses, to ignore others, requires exact adjustment between the pull of a magnet and the tension of a spring. If they are out of balance your telephone might tinkle when it oughtn't, or keep silent when it should ring.

In the past, adjustment was made by hand, little by little until the proper setting was reached. It took time. But now Bell Laboratories engineers have developed a machine which adjusts new ringers perfectly, before they leave the Western Electric Company plants where they are made. And the operation takes just 30 seconds.

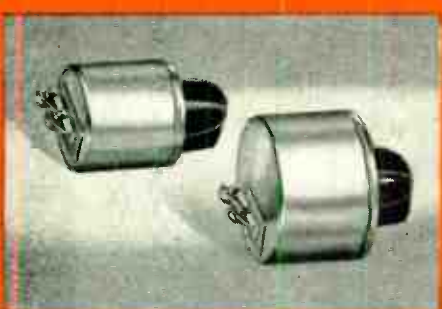
This is another example of how the Laboratories work constantly to improve every phase of telephony—keeping the costs low while the quality of service grows higher and higher.

BELL TELEPHONE LABORATORIES

WORKING CONTINUALLY TO KEEP YOUR TELEPHONE SERVICE ONE OF TODAY'S GREATEST VALUES

OHMITE RHEOSTATS

MEET REQUIREMENTS OF JOINT ARMY-NAVY SPECIFICATION JAN-R-22



Models H (enclosed) and J (enclosed)
Also AN 3155 (AN-R-14a)

TYPE	OHMITE MODEL	WATT RATING
RP10	H	25
RP11	H enclosed	12.5
RP15	J	50
RP16	J enclosed	25
RP20	G	75
RP25	K	100
RP30	L	150
RP35	P	225
RP40	N	300
RP45	R	500
RP50	T	750

OHMITE RHEOSTATS MEET THESE RIGID TESTS:

★ 5-Hour Vibration Test (Required for RP 10-11-15-16-20-25)
★ 50-Hour Salt-Spray Corrosion Test
★ 150-Hour 95% Humidity Electrolysis Test
and other tests as prescribed in Specification JAN-R-22

By meeting these severe Joint Army-Navy requirements, Ohmite Rheostats have proved what industry has long accepted as true—that they can be depended upon for unfailing performance under the toughest operating conditions. All-ceramic construction . . . a smoothly gliding metal-graphite brush . . . uniform windings locked in place by vitreous enamel . . . insure close control throughout years of trouble-free service. It will pay you to standardize on Ohmite Rheostats for your product.

Be Right with

OHMITE®

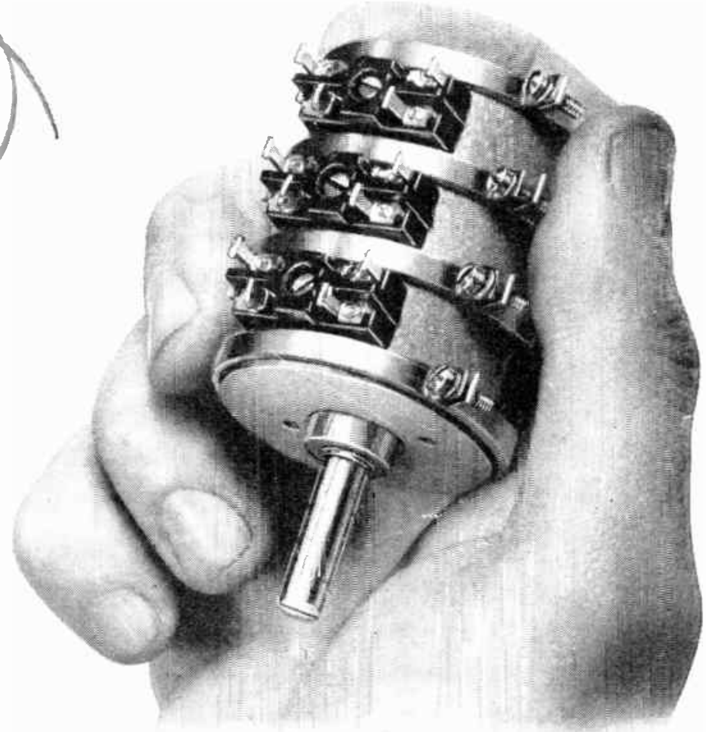
RHEOSTATS • RESISTORS • TAP SWITCHES

OHMITE MFG. CO., 4862 Flournoy St., Chicago 44, Ill.

Here's Fairchild's Newest Potentiometer!

TYPE-746
PRECISION POTENTIOMETER OFFERS:

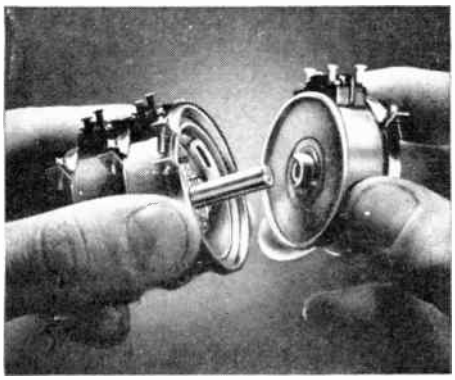
- Low Torque
- Accurate Phasing
- Quick Replacement
- Ganging up to 20 on a shaft



The finest we've ever built! That's our idea of the new "746". It's got lower torque, a new more accurate phasing adjustment, and a new method of ganging that makes it easy to put as many as twenty cups on a single shaft. Individual cups in a gang are easily replaced if necessary.

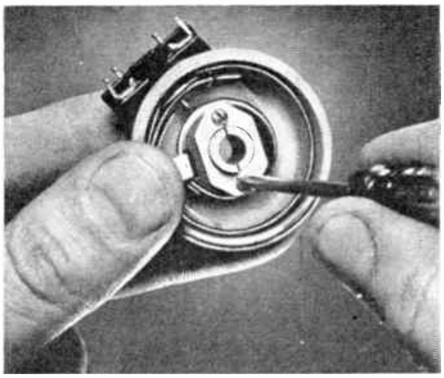
The new potentiometer is available with linear or non-linear windings to meet your specifications. Its attractive case is made of grey anodized aluminum.

The "746" is just one of the complete Fairchild family of precision potentiometers. What are your requirements? Write, giving details, to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, N.Y.* Dept. 140-13H1.



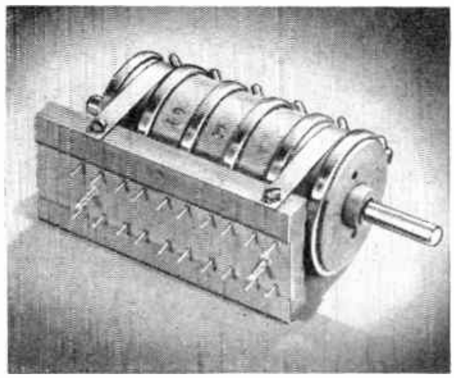
EASY REPLACEMENT

To replace a unit in a "746" gang, loosen connecting-band screws, remove "cup," slip new "cup" under bands, and tighten screws. This feature pays off in experimental work where circuit elements are changed periodically.



ACCURATE PHASING

A new type phasing adjustment is simpler and more accurate. A retainer plate clamps shaft to wiper arm. To adjust for phasing, loosen two screws, set the arm to the correct position, then tighten screws.



FLEXIBLE DESIGN

Typical of the special consideration Fairchild gives to its customers' special requirements is this plug-in version of the "746." Where fast servicing is a must, the advantages of this "quick-change" unit are quite apparent.

SPECIFICATIONS

Accuracy (overall resistance)—±.5% (linear), ±1.0% or better (non-linear)

Mechanical accuracy—
concentricity (shaft to pilot)—.0015 in. FIR max.;
radial play—.0009 in. FIR max.;
shaft—centerless ground stainless steel to .2500 diam. (+.0000, —.0005 in.);
pilot hub—machined to .5000 (+.0000, —.0005 in.)

Torque—1.5 oz-in.

Dimensions—diameter 1.750 max.; length (1 cup) .800 in. ±.009 in.; added length per unit ganged .580 in. ±.002 in.

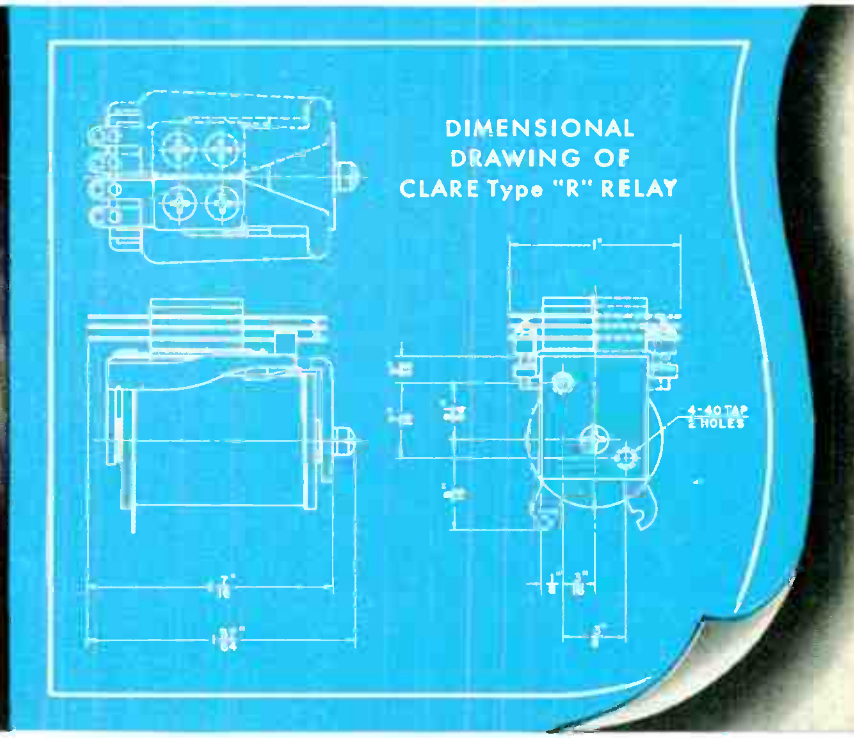
Case—grey anodized aluminum

FAIRCHILD
PRECISION POTENTIOMETERS

A NEW CLARE RELAY...

the Type "R" combines extremely small size with unusual sensitivity and long life

Approximately Actual Size



SPECIFICATIONS

SIZE
Length: 1 7/16" — Height: 1 3/64" — Width: 1"

WEIGHT
Approximately 2 ounces

COIL
Single or double-wound

OPERATING VOLTAGE
Up to 230 volts d-c

ARMATURE
Single or double arm

CONTACT ASSEMBLY
Form A to C. Maximum of 10 springs in each pileup.

MOUNTING
Two #4-40 tapped holes in end of heelpiece

This new CLARE Type "R" d-c Relay embodies many features of the famous CLARE Type "K" Relay, which was the first to combine the advantages of a telephone type relay with the small size, light weight and resistance to vibration required to meet the rigid demands of aircraft service.

In appearance, the Type "R" resembles the Type "K", but, through hardly noticeable structural differences, CLARE has given the new Type "R" even greater sensitivity and operating range. Both relays use the same contact springs, but the Type "R" coil is longer and of larger diameter, to provide greater winding space. Life expectancy of the new relay has been not only increased but multiplied.

The CLARE Type "R" Relay retains in an improved form the reed armature suspension which discerning engineers have come to recognize as one of the subtler reasons for the superior performance of CLARE Type "K" Relays over other relays of comparable size and somewhat similar appearance.

The Type "R" is available as either an open or hermetically sealed relay. Clare sales engineers are located in principal cities to give you firsthand information on this new relay and to cooperate with you on any complex relay problem. Call them or write to C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials Ltd., Toronto 13. Cable Address: CLARELAY.

Write for CLARE Bulletin No. 115

CLARE RELAYS

First in the Industrial Field

It's a fact that

✓ Very few customers object to our suggestions on redesign. For those who object, we make their designs exactly as they want them. Most customers appreciate the close co-operation given and gladly use our fifty years of specialized experience.

Here are the reasons:

1. *Makes best use of favorable ceramic characteristics, avoids misapplication.*
2. *Designs are kept within practical manufacturing limits.*
3. *Subsequent assembly operations frequently are speeded.*
4. *Several components can sometimes be combined into one.*
5. *Maximum economy.*

Diversified equipment which is far beyond the reach of the average ceramic manufacturer permits us to use the best production method for the job.

Illustrated is a redesign which saved about 60% in original price, saved assembly time, and gave improved performance. Our engineers and production men will make recommendations on request. Send blue prints or samples with data on operating conditions and tolerance requirements. By working together we can serve you better.

AMERICAN LAVA CORPORATION

50TH YEAR OF CERAMIC LEADERSHIP

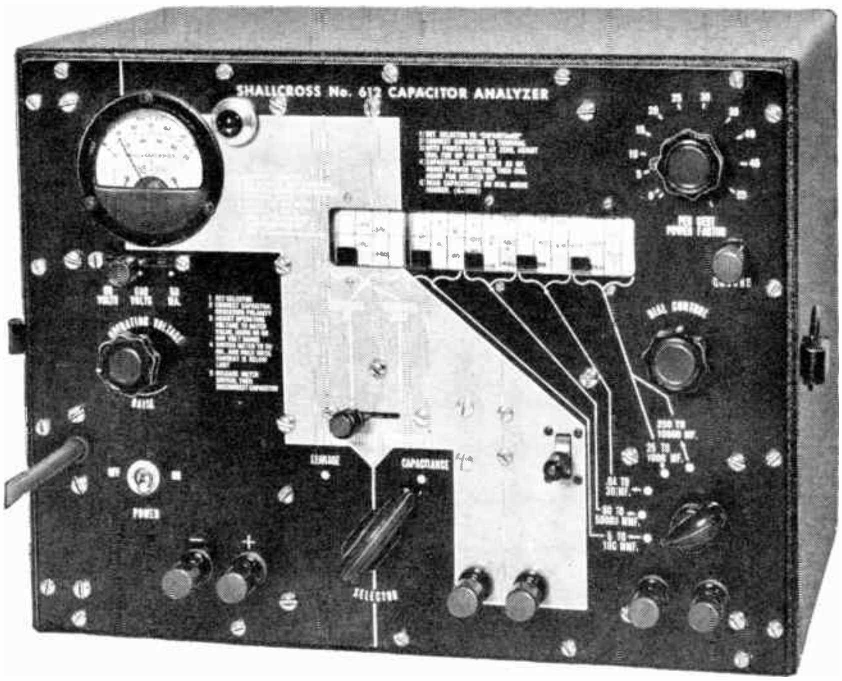
CHATTANOOGA 5, TENNESSEE

OFFICES: METROPOLITAN AREA: 671 Broad St., Newark, N. J., Mitchell 2-8159 • CHICAGO, 228 North LaSalle St., Central 6-1721 PHILADELPHIA, 1649 North Broad St., Stevenson 4-2823 • LOS ANGELES, 232 South Hill St., Mutual 9076 NEW ENGLAND, 38-B Brattle St., Cambridge, Mass., Kirkland 7-4498 • ST. LOUIS, 1123 Washington Ave., Garfield 4959

SHALLCROSS

612 CAPACITOR ANALYZER

ACCURATE • WIDE RANGE • DIRECT READING



Typical of the high quality, completeness and utility of Shallcross instruments, this analyzer provides a convenient means of determining the following capacitor characteristics:

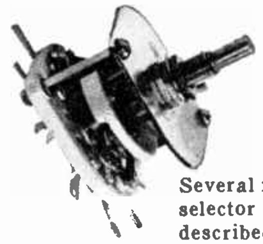
TEST	RANGE	ACCURACY
Leakage	0-60 V d-c	±3% of full scale
	0-600 V d-c	±3% of full scale
	0-50 MA. d-c	±3% of full scale
Insulation	1.1-120 megohms	±20%
Resistance	110-12,000 megohms	±20%
Capacitance	5-100 mmf	±3 mmf
	80-50,000 mmf	±5%
	0.04-30 mfd	±5%
	25-1,200 mfd	±10%
	250-12,000 mfd	±20%
Power Factor	0.50%	±10%

SHALLCROSS MANUFACTURING COMPANY

Collingdale, Pa.

PRECISION RESISTORS • D-C BRIDGES • DECADE RESISTANCES • HIGH-VOLTAGE MEASURING INSTRUMENTS • GALVANOMETERS • LOW-RESISTANCE TEST SETS • ROTARY SELECTOR SWITCHES • TRANSMISSION TEST SETS • CUSTOM-BUILT ELECTRONIC SPECIALTIES

Something New



NEW OVAL SELECTOR SWITCHES

Several new oval rotary selector switches are described in Bulletin L13 just issued by the Shallcross Manufacturing Co., Collingdale, Pa. Six basic plates and three rotor types produce switches having from one to three poles per deck or gang and with other desired mechanical and electrical details. As many as 18, 9 or 6 positions may be obtained in single-, double-, or triple-pole types respectively. These may be single-, double-, or triple-pole decks exclusively or a combination of different types.

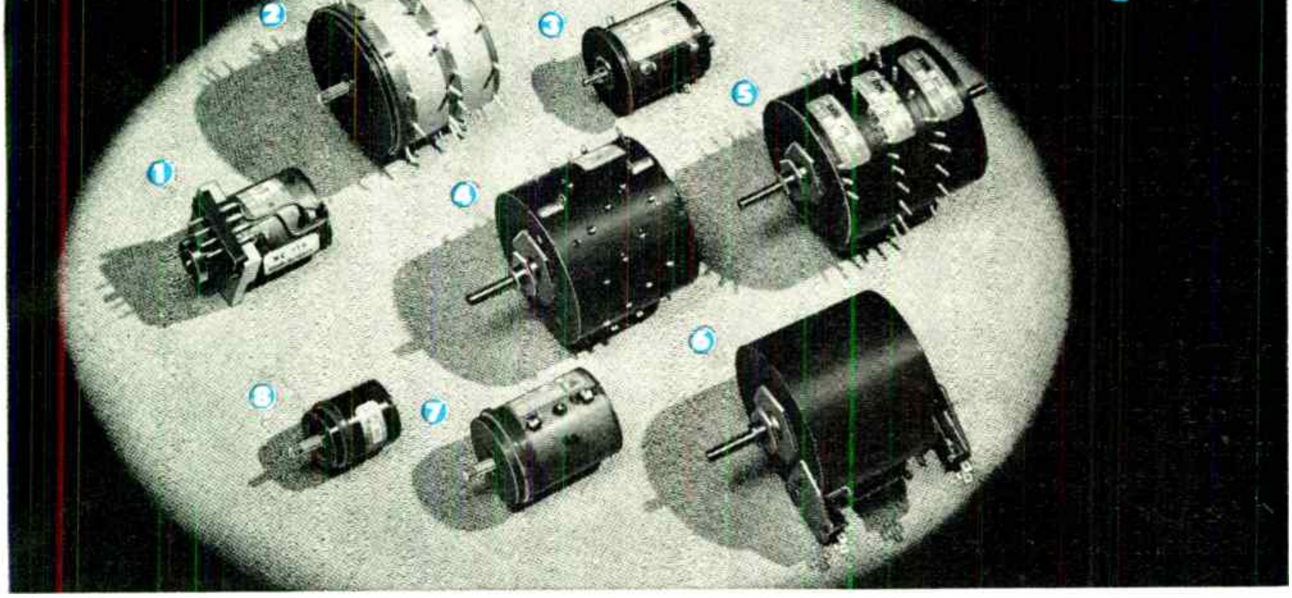
VERTICAL STYLE PRECISION RESISTORS FOR JAN USES

Improved vertical style precision wire-bound resistors for use where mounting requirements make it desirable to have both terminals at the same end of the resistor have been introduced by the Shallcross Manufacturing Co., Collingdale, Pa. These units provide a longer leakage path from the mounting screws to the terminals. Known as Shallcross Types BX120, BX140, and BX160, they are designed to meet JAN requirements for styles RB40B, RB41B and RB42B respectively. For commercial uses, the resistors carry somewhat higher ratings than for JAN applications. Wire leads instead of terminals can be furnished if desired. Complete details will gladly be sent on request to the manufacturer.

FLAT, METAL-ENCASED WIRE-WOUND RESISTORS

Flat, metal-encased, Type 265A wire-wound power resistors introduced by the Shallcross Manufacturing Company, Collingdale, Pa., are space wound, have mica insulation, and are encased in aluminum for mounting flat against a metal chassis. At 175° C. continuous use they are conservatively rated for 7½ watts in still air and 15 watts when mounted on a metal chassis. Write for Bulletin 122.

Typical of the TOUGH POTENTIOMETER JOBS solved by Helipot



Precise Accuracy + Maximum Versatility + Space-saving Compactness

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

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

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

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

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

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

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

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

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

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

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

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

THE Helipot CORPORATION, SOUTH PASADENA 6, CALIFORNIA

BUSINESS IN MOTION

To our Colleagues in American Business . . .

Never has industry turned out more goods than during the past few years of unprecedented customer demands. In endeavoring to meet these, Revere has developed new techniques, established new plants, installed the newest equipment and modernized the old, and stepped up its training program, including the development of some new ideas in relation to safety. Throughout the country, similar steps have been taken by manufacturers generally. This is the response of free enterprise to the stimulus of a free and growing market. It is fortunate that American industry was not only willing but able to do this, because now it is evident that these facilities and these skills must be devoted more and more to the defense of our freedom.

Defense Orders or "DO's" are being issued, and their volume is bound to increase. Already prime contractors are seeking sub-contractors, and sub-sub-contractors are receiving orders too, down to small local firms operating only a few machines. Perhaps few people realize the importance of the "small shop"; the fact is that these establishments have a tremendous total capacity supplementing that of the great corporations, which practically never make everything that is needed for a finished product such as a tank, a plane, a ship, radar equipment. The "smalls" are just as vital as the "bigs."

Revere knows that when the time of trial comes, it is more important than ever to increase production efficiency. This makes complete information essential to those who have taken on DO contracts. Revere pledges its full cooperation, and will gladly provide all it knows about its metals.

This knowledge is made available in two principal ways. First, there are many booklets containing technical data, including physical properties, and also in many cases suggestions as to recommended fabrication practices. In addition to the booklets, which are distributed on request, Revere either reproduces or summarizes them in the various Sweet's Files, Chemical Engineering Catalogue, Marine Catalogue, Refinery Catalogue. This printed material is therefore available freely to all who will ask for it, or look it up. The second way in which Revere's knowledge and skill is made available is through the Technical Advisory Service, a group of capable men whose collective experience covers practically all applications of copper and copper alloys, and aluminum alloys. In war and peace, these men have rendered invaluable service, collaborating closely on such matters as selection of the proper metal, temper, width, gauge, and in helping to solve production problems. As a result, scrap has been reduced, rejects lessened, production increased, money and materials saved. The services of the Technical Advisors are obtainable through the Revere Sales Staff which also has wide experience in the selection and application of Revere Metals. If you have orders whose specifications include non-ferrous metals, Revere will gladly place its information at your disposal.

If you purchase and work with other materials, Revere suggests that it should be realized that not only is American productive capacity tremendously greater, but that there has been a likewise large growth in knowledge about materials of all kinds. So it is recommended that no matter what you make now, or are called on to make in the future, you ask your suppliers to share their knowledge with you. It will make you and our country stronger.

REVERE COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

150th YEAR OF SERVICE TO AMERICA

Executive Offices:

230 Park Avenue, New York 17, N. Y.

SEE "MEET THE PRESS" ON NBC TELEVISION EVERY SUNDAY

EXTRA-DEPENDABLE

LONG-LIVED

WON'T SHORT

WITHSTAND HIGH HEATER-CYCLING

SHOCK-RESISTANT

G-E 5-STAR ARINC TUBES ARE 5 WAYS BETTER!

Take your design cue from airmen who praise the superiority of ARINC types!

"Records kept by our company show replacement of ARINC tubes to be only 2% as compared with an average of 49% replacement for 13 types of regular tubes over the same period."

W. W. LYNCH, *System Communications Superintendent,* Pan American World Airways System.

"Our company is using ARINC reliable tubes wherever possible. Experience has shown that equipment using these tubes seldom causes delays from tube failure."

J. H. CARMICHAEL, *President,* Capital Airlines.

"The effect of using ARINC tubes in our equipment became immediately apparent. Off-schedule removals of airborne equipment due to tube failures have been materially reduced."

J. R. CUNNINGHAM, *Director of Communications,* United Air Lines.

"During a six months' carefully controlled comparison service test, ARINC tubes required only ¼ as many replacements as first quality standard brand tubes."

FRANK R. WAGNER, *Supervisor of Radio, Electrical and Instrument Engineering,* TWA

"We have four VHF Navigation Receivers which have been in service for more than 1,000 hours each. Of the 104 ARINC tubes used, we have not had a single failure."

J. LANE WARE, *Supervisor of Communications Engineering,* National Airlines.

GL-5654

GL-5670

GL-5686

A pioneer in manufacturing ARINC tubes, G.E. offers nine 5-star types which set new standards of dependability and long life. Specify these tubes in electronic circuits now on your boards, to increase safety factor . . . reduce upkeep costs . . . *build reputation* for your equipment!

G-E 5-star ARINC tubes are a joint achievement of Aeronautical Radio, Inc., and General Electric Co. Built with exacting care—individually tested—they accent the reliability of altimeters, radio compasses, h-f receivers, other apparatus that guides air passengers safely to their destinations.

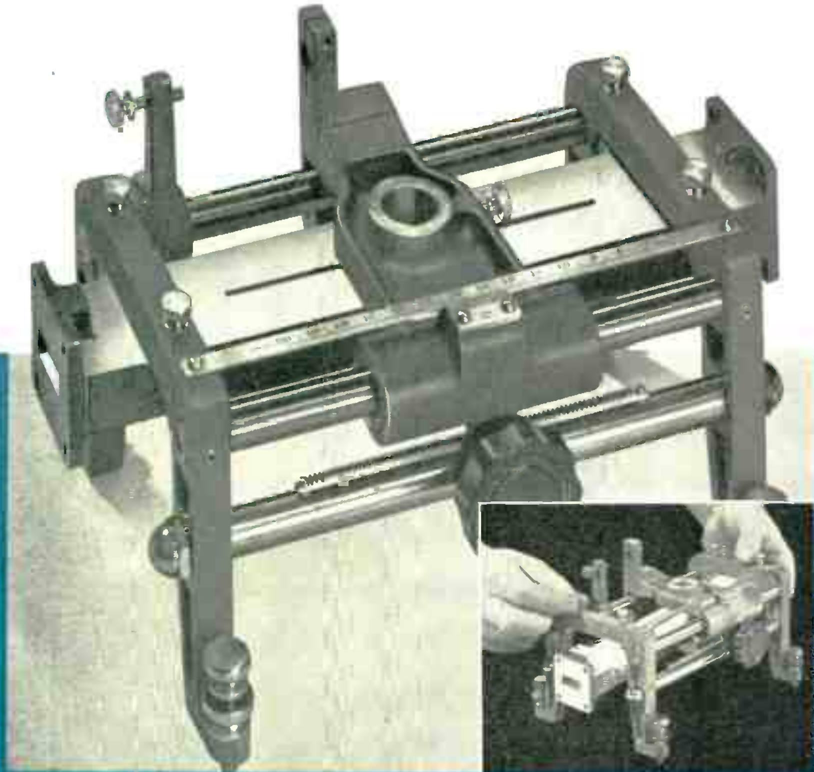
Learn more about G-E 5-star tubes! Write for Bulletin ET-B29A, which tells the story of their design, manufacture, and testing . . . describes and rates the various 5-star types . . . shows how *you,* as equipment builder or designer, can profit from their superior performance! *General Electric Company, Electronics Department, Schenectady 5, New York.*

GL-5654 —Sharp-cutoff r-f pentode
GL-5670 —H-f twin triode
GL-5686 —Power-amplifier pentode
GL-5725 —Semi-remote-cutoff r-f pentode
GL-5726 —Twin diode
GL-5749 —Remote-cutoff r-f pentode
GL-5750 —Pentagrid converter
GL-5751 —High-mu twin triode
GL-5814 —Medium-mu twin triode

GENERAL ELECTRIC

185-K4

IMPEDANCE



COMPLETE COVERAGE!

CONTINUOUS microwave coverage, 10 mc to 12,400 mc. High mechanical stability. Simple operation. Broad applicability. Precision accuracy. Compact size!

New *–hp–* microwave equipment gives you *complete coverage* for VHF, UHF and SHF impedance measurements. Instrumentation includes VHF Bridges as well as the slotted coaxial and waveguide sections which are fundamental tools in impedance measurements. These instruments can be used to measure load or antenna impedance, system flatness, connector reflection, percentage of reflected power, standing wave magnitude or phase, characteristics of coaxial transmission lines or rf waveguide systems, characteristics of rf chokes, resistors, condensers.

For complete details see your *–hp–* sales representative or write direct.

HEWLETT-PACKARD COMPANY

2160D Page Mill Road · Palo Alto, California

Sales representatives in principal areas.
Export: Frazar & Hansen, Ltd.
San Francisco, New York, Los Angeles

–hp– 809B UNIVERSAL PROBE CARRIAGE
with *–hp–* 810A WAVEGUIDE SLOTTED SECTIONS

Now—a single probe carriage operates with up to 3 different slotted sections—waveguide or coaxial! This means important savings in time; lower investment in instrumentation. The new -hp- 809B Universal Probe Carriage mounts slotted sections covering frequencies from 3,000 to 12,400 mc (see table on opposite page)—and you can *interchange sections in 30 seconds or less!*

-hp- 809B Carriage is accurately calibrated in mm. for readings as low as 0.1 mm. Dial gauge may be readily mounted if more accurate readings are needed. Carriage travels on a new 3-point ball-bearing suspension system, and operates in conjunction with -hp- 442A Broad-Band Probe and -hp- 440A Coaxial Detector combination; or with -hp- 444A Untuned Probe. The extremely broad usefulness of this new Universal Carriage means far greater flexibility and lower cost for complete microwave instrumentation.

–hp– 417A VHF DETECTOR

For use with *–hp–* 803A VHF Bridge. A super-regenerative (AM) receiver covering all frequencies 10 to 500 mc in 5 bands. Offers approx. 5 μv sensitivity over entire band; quick, easy operation, direct-reading frequency control. Thoroughly shielded, suitable for general laboratory use including approximate frequency checks, measurements of noise, interference, etc. $200.00 f.o.b. factory.

–hp– 415A STANDING WAVE INDICATOR

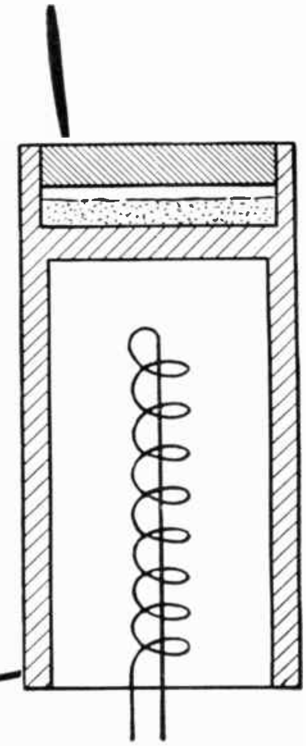
Designed for use with all waveguide or coaxial slotted sections, to give direct reading of standing wave ratio in VSWR or db. Consists of high gain amplifier with low noise level, operating at fixed frequencies between 300 and 2,000 cps. (Normal frequency 1,000 cps., plug-ins for other frequencies available). Input circuits for use with crystal detector or bolometer. $200.00 f.o.b. factory.

HEWLETT-PACKARD *hp* INSTRUMENTS

Announcing

the First Basically New Cathode in Over 25 Years

Providing High Emission Density with Long Life



PHILIPS L-CATHODE

Illustrated above is the first basically new cathode in over 25 years—the new PHILIPS L-CATHODE. It offers these practical advantages to users of klystrons, disc seal triodes, iconoscopes, magnetrons, special cathode-ray tubes and other types where a high degree of reliability is essential:

1. *High emission current density with long life*
2. *Can be made in great variety of shapes and to close tolerances*
3. *Not damaged by high electrostatic forces*
4. *Reactivates after arcing*
5. *Can be reactivated after exposure to air*
6. *Withstands severe ionic and electronic bombardment*
7. *Maintains uniform emission characteristics during its life*

The PHILIPS L-CATHODE originated in the Philips Research Laboratories in Holland, and was further developed by Philips Laboratories, Inc., here. It was first described in the June 1950 *Philips Technical Review*. Leading electronic research laboratories have already used development types of various shapes with outstanding results.

Samples of PHILIPS L-CATHODES for experimental use can be supplied in limited quantities to those interested in producing greatly improved tubes of high emission density with longer life.

PHILIPS LABORATORIES, INC.

Dept. P6, Irvington-on-Hudson, New York

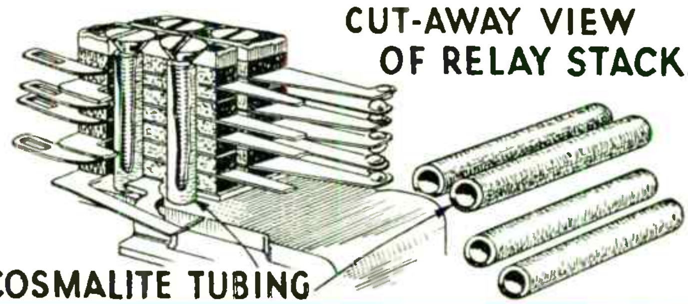
ALLIED CONTROL RELAYS

built with

Cleveland PHENOLIC TUBES

ensure

- ★ DEPENDABILITY
- ★ PRECISION PERFORMANCE
- ★ LONG LIFE
- ★ ECONOMY



The Allied Control Co. has built a long and enviable record as a quality supplier of control relays to both private industry and governmental services.

Their S K Relay shown above, is typical of the various Allied Relays in which CLEVELAND CONTAINER tubing provides excellent service.

It is likewise the answer for hundreds of other problems of manufacturers in the electrical industry.

CLEVELITE* and COSMALITE*

Laminated Phenolic Tubing

combines electrical and physical properties to meet the most exacting requirements.

CLEVELITE is produced in six grades . . .
COSMALITE in four grades . . .
A grade for EVERY need.

Cleveland Container's large production facilities further ensure minimum production costs as well as dependable deliveries.

Why pay more? . . . for the BEST call CLEVELAND.

*Trade Marks

Write today for our new descriptive brochure. Also ask for quotations and samples to meet your exact specifications.

The CLEVELAND CONTAINER *Co.*

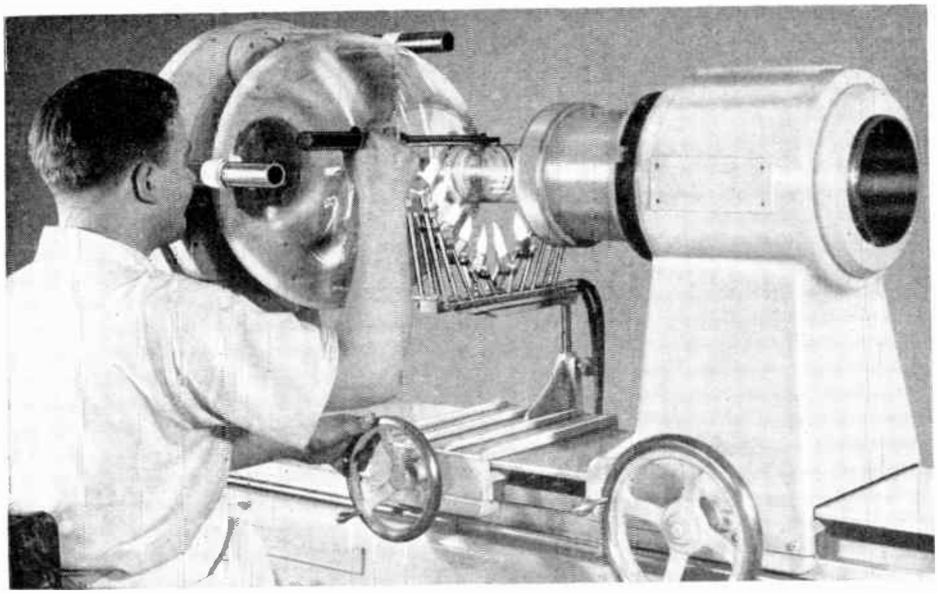
6201 BARBERTON AVE. CLEVELAND 2, OHIO

PLANTS AND SALES OFFICES at Plymouth, Wisc., Chicago, Detroit, Ogdensburg, N.Y., Jamesburg, N.J.
ABRASIVE DIVISION at Cleveland, Ohio
CANADIAN PLANT: The Cleveland Container, Canada, Ltd., Prescott, Ontario

REPRESENTATIVES

NEW YORK AREA R. T. MURRAY, 604 CENTRAL AVE., EAST ORANGE, N. J.
NEW ENGLAND R. S. PETTIGREW & CO., 10 N. MAIN ST., WEST HARTFORD, CONN.
CHICAGO AREA PLASTIC TUBING SALES, 5215 N. RAVENSWOOD AVE., CHICAGO

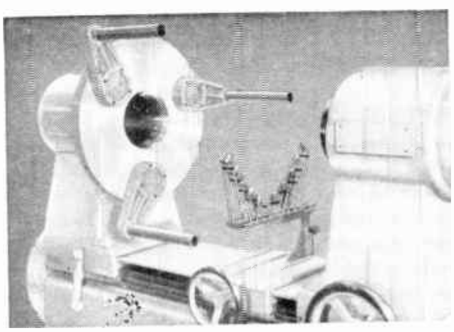
LITTON INDUSTRIES NEWS



LITTON GLASSWORKING LATHES SPEED PRECISION ASSEMBLY OF TV KINESCOPE, VACUUM TUBES

Modern vacuum tubes have extremely close alignment tolerances. Often, sub-assemblies must be separately aligned before junction. During sealing, both assemblies must be rotated in perfect phase to maintain this alignment.

Versatile, adaptable Litton Glassworking Lathes meet these requirements. They are built on a normalized cast iron bed, with precision ground ways and axial alignments of highest accuracy and positive phase. The lathes will chuck and hold units such as copper anodes to runouts of .001″.



Close-up of spindle head, Litton Model K lathe, showing exceptionally large diameter opening of universal chuck

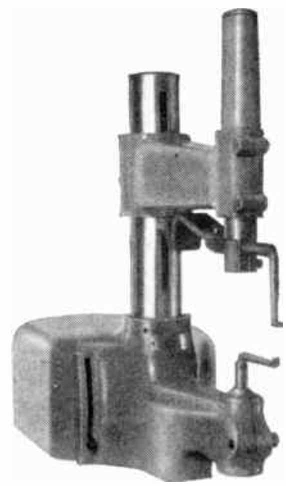
Air passages are arranged to avoid contamination, yet permit use of neutral gasses when sealing glass to metal. Burners provide the narrowest possible heating area commensurate with ample total heat. Continuously variable spindle speed, which makes possible much glassworking without blowing, is optional on all models. Foot pedals control the air or neutral gas supply, and the oxygen and gas volume. Convenient hand controls govern carburetion and air intake to the spindles.

Leading TV tube makers use Litton Glassworking Lathes to speed production of kinescope tubes 10″ to 27″ in diameter. Manufacturers find that the speed and handling ease of Litton lathes enable glassworkers to seal tube funnels to domes in minimum time—with complete control of glass distribution. Since most manufacturers align sub-assemblies on the lathe, the accurate phasing of Litton spindle heads is also an important factor.

Reliable Litton Glassworking Lathes are adaptable to the widest possible variety of glassworking jobs. Five models offer a choice of radial clearance ranging from 4″ to 17½″. and axial working lengths from 20¾″ to 75½″.

LITTON SPOTWELDERS OFFER HIGH POWER, EXTREME FLEXIBILITY FOR PRECISION JOBS

Litton Model A Precision Spotwelder offers broad applicability of use in the manufacture of vacuum tubes. It makes possible the quick altering of production set-ups. Rated 2 kva continuous duty, it efficiently handles average sized



or very precise jobs. Accurate alignment and absence of side play permit butt welding or parallel welding of small wires without rolling. Foot pedals and switches control top mandrel and power supply. Model A spotwelder has 6½″ throat depth and extension jaws can be added.

SPOTWELDER TIMER

A new timer for the Litton Model A Spotwelder has been developed by Litton Industries and will be available for delivery soon. The timer employs two simple controls. One adjusts weld time in steps of 1, 2, 3, 5, 7, 10, 15, 25 and 60 cps. The other adjusts heat control in 6 steps. Proper adjustment of these controls makes possible precision welds up to the 2 kva rating of the welder.

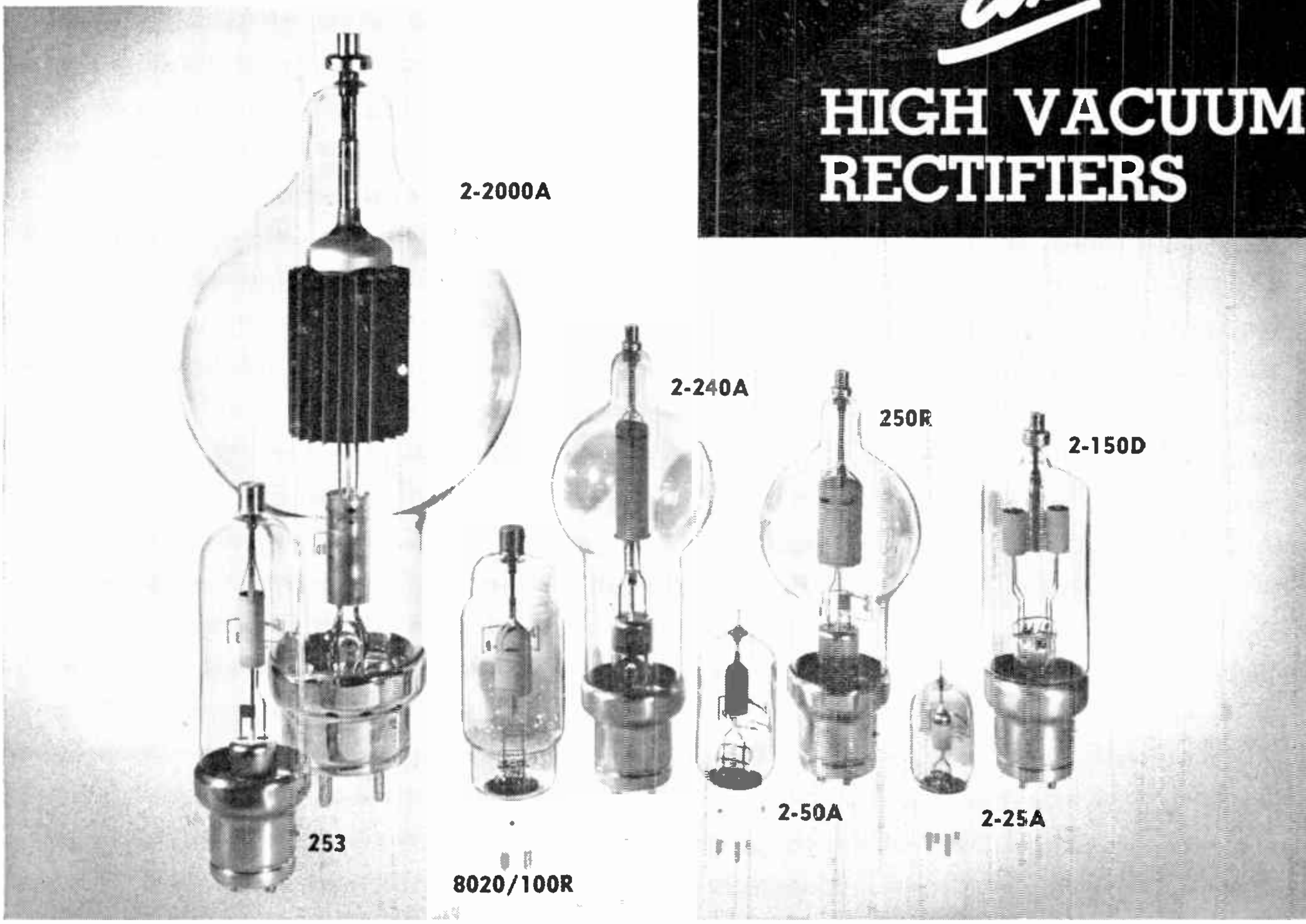
LITTON INDUSTRIES

SAN CARLOS, CALIFORNIA, U. S. A.

DESIGNERS AND MANUFACTURERS of: Glassworking Lathes and Accessories, Vertical Sealing Machines, Burner Equipment, Precision Spotwelders, Oil Vapor Vacuum Pumps, Glass Baking Ovens, Vacuum Tubes and Tube Components, Magnetrons, High Vacuum Molube Oil, Microwave Equipment.

Eimac

HIGH VACUUM RECTIFIERS



Eimac's comprehensive series of vacuum rectifiers permits a choice of "per-tube" ratings of d-c plate current from 50 ma. to 750 ma. and a choice of inverse voltage ratings from 15,000 to 75,000 volts.

These are ruggedly constructed tubes built to withstand more than normal abuse in rectifying and voltage multiplying circuits or as diode clippers. Their design incorporates many of the features long associated with the famous Eimac transmitting tubes . . . Pyrovac plates . . . thoriated tungsten filaments . . . no troublesome internal insulating materials . . . and, of course, a "hard" vacuum.

Put Eimac high vacuum rectifiers to work for you. Write today for detailed data sheets giving complete operating information and application notes.

EITEL-McCULLOUGH, INC.
San Bruno, California

Export Agents: Frazar & Hansen 301 Clay St., San Francisco, California

284

TYPE	DESCRIPTION	MAXIMUM DIMENSIONS Length Inches	Diameter Inches	AVERAGE PLATE CUR. Ma.	PLATE DISSIPATION Watts	PEAK INVERSE VOLTAGE Volts	FILAMENT Volts	Amps
2-25A	High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	4.5	1.5	50	15	25,000	6.3	3.0
2-50A	High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	5.75	2	75	30	30,000	5.0	4.0
2-150D	High vacuum rectifier. High voltage medium current. Instant heating thoriated tungsten filament. Radiation cooled pyrovac plate.	8.38	2.75	150	90	30,000	5.0	13.0
2-240A	High vacuum rectifier. High voltage, high current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	11.25	4	500	150	40,000	7.5	12.0
2-2000A	High vacuum rectifier. High voltage, high current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	18	8.25	750	1,200	75,000	10.0	25.0
250R	High vacuum rectifier. High voltage medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	10.25	4	250	150	60,000	5.0	10.5
253	High vacuum rectifier. High current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	9	2.75	350	100	15,000	5.0	10.0
8020/100R	High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate.	8	2.38	100	60	40,000	5.0	6.5

Better Circuit Reliability

Without

OPERATOR ADJUSTMENT

Sync Generator set up with compact GPL Control Console

POWER SUPPLY *Built in*

the New GPL SYNCHRONIZING GENERATOR

SMALLER THAN EXISTING UNITS

EASY MAINTENANCE

BUILT-IN POWER SUPPLY

STANDARD RELAY PANELS
Easy to Rack Mount

The GPL Synchronizing Pulse Generator provides circuit reliability superior to that of comparable studio equipment. Operator adjustments are now eliminated by means of advanced circuit design, including binary counting circuits, delay-line-controlled pulse width – all operating from a stable master oscillator. The generator provides standard RTMA outputs with automatic termination of unused outputs. The AFC circuit is readily set to operate at mid-range when locked to the line.

Since the unit is smaller than existing equipment, even with its self-contained power supply, it is ideal for field operation. Swing-down panels simplify maintenance. Components are mounted on standard relay panels, facilitating studio rack mounting.

Typical of other GPL developments, the Synchronizing Generator is designed for maximum quality, operating efficiency, and dependability. Write for literature and operating information.

Write, Wire or Phone for Details

General Precision Laboratory
INCORPORATED
PLEASANTVILLE NEW YORK

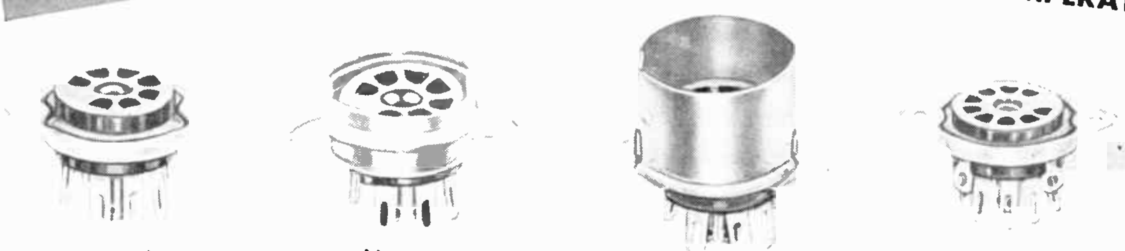
TV Camera Chains • TV Film Chains • TV Field and Studio Equipment • Theatre TV Equipment

MYCALEX

low loss miniature TUBE SOCKETS

OFFER ALL THESE ADVANTAGES:

- CLOSER TOLERANCES
- LOWER DIELECTRIC LOSS
- HIGH ARC RESISTANCE
- HIGH DIELECTRIC STRENGTH
- GREAT DIMENSIONAL STABILITY
- IMMUNITY TO HUMIDITY
- HIGH SAFE OPERATING TEMPERATURE



-cost no more than PHENOLIC TYPES

These glass-bonded mica sockets are produced by an exclusive MYCALEX process that reduces their cost to the level of phenolic sockets. Electrical characteristics are far superior to phenolics while dimensional accuracy and uniformity exceed that of ceramic types.

MYCALEX miniature tube sockets, available in 7-pin and 9-pin types, are injection molded with great precision and fully meet RTMA standards. They are produced in two grades, described as follows, to meet diversified requirements.

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

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

MYCALEX TUBE SOCKET CORPORATION
Under Exclusive License of
MYCALEX CORPORATION OF AMERICA
30 ROCKEFELLER PLAZA, NEW YORK 20, N.Y.

SINCE 1919
MYCALEX
THE INSULATOR
TRADE MARK REG. U.S. PAT. OFF.

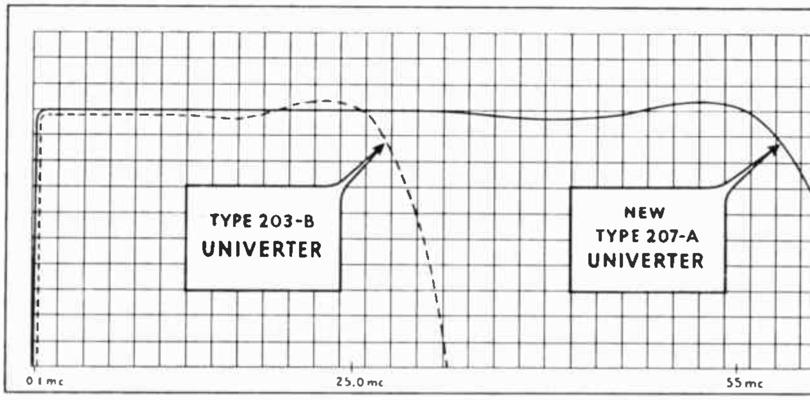
CORPORATION OF AMERICA
"Owners of 'MYCALEX' Patents"
Executive Offices: 30 Rockefeller Plaza, New York 20 • Plant and General Offices: Clifton, New Jersey

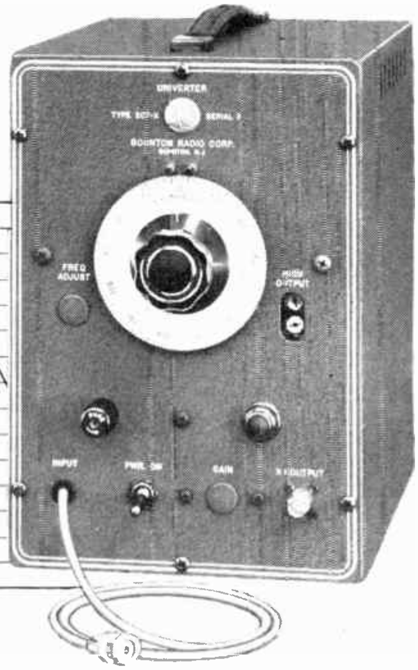
Wide Band UNIVERTER

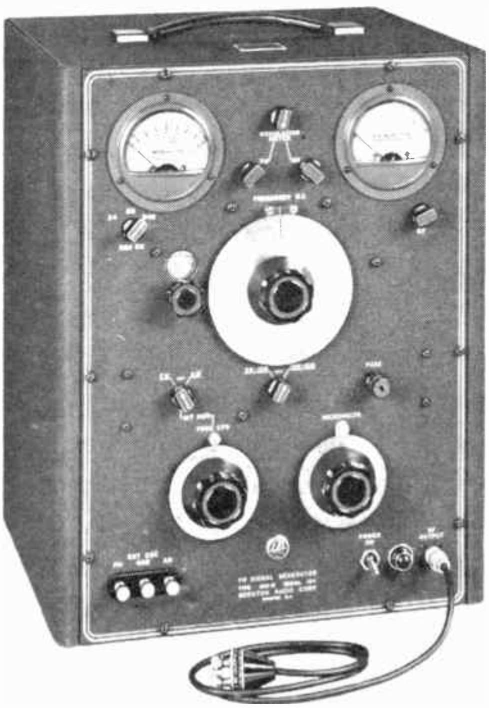
for complete frequency coverage

when used with the

FM-AM SIGNAL GENERATOR TYPE 202-B







UNIVERTER

TYPE 207-A

The Univerter Type 207-A provides a continuous extention of the frequency range of the 202-B FM-AM Signal Generator down to 0.1 mc. The two instruments may be used over a continuous frequency range of 0.1 mc. to 216 mc. The Univerter Type 207-A subtracts 150 mc. from a signal obtained from the 202-B and provides outputs between 0.1 mc. and 55 mc. without change of signal level. Negligible spurious signals are introduced and modulation of the signal is unaffected. Small incremental changes can be made in frequency to allow the study of band pass characteristics of very narrow band receivers. A regulated power supply prevents change of gain or frequency with line voltage.

SPECIFICATIONS (*When used with 202-B*)

FREQUENCY RANGE: 0.1 mc. to 55 mc. (0.3 mc. to 55 mc. with 200 kc. carrier deviation).

FREQUENCY INCREMENT DIAL: Plus or minus 300 kc. calibrated in 5 kc. increments.

FREQUENCY RESPONSE: Flat within ± 1 db over frequency range.

FREQUENCY ADJUST: Front panel control allows calibration with 202-B output.

OUTPUT: Continuously variable, at XI jack from 0.1 microvolt to 0.1 volt across 53 ohms by use of 202-B attenuator.

HIGH OUTPUT: Uncalibrated approximately 1.5 volts from 330 ohms into open circuit.

DISTORTION: No appreciable FM distortion at any level.
No appreciable AM distortion at carrier levels below 0.05 volt and modulation of 50%.

SPURIOUS RF OUTPUT: At least 30 db down at input levels less than 0.05 volts.

Write for complete information
(In Canada, direct inquiries to R.C.A. Victor Co., Ltd., Montreal)

FM-AM SIGNAL GENERATOR

TYPE 202-B

- The standard signal source for the FM and TV industry.
- Univerter 207-A extends frequency range down to 0.1 mc. without change in signal level or modulation characteristics below.

SPECIFICATIONS:

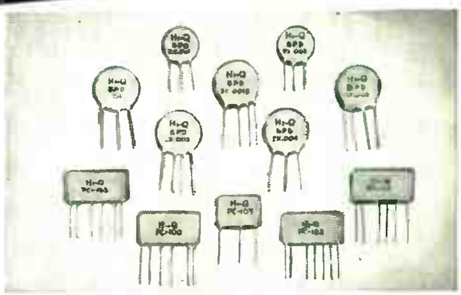
RF RANGES: 54-108, 108-216 mc.
FREQUENCY DEVIATION: 0-24 kc., 0-80 kc., 0-240 kc.
FM DISTORTION: Less than 2% at 75 kc. deviation
AMPLITUDE MODULATION: Continuously variable 0-50%.
RF OUTPUT VOLTAGE: 0.1 microvolt to 0.2 volt.

BOONTON RADIO Corporation
BOONTON, N. J.

Home entertainment has changed

. . . and in producing better and better TV MODERN ELECTRONICS LOOK TO HI-Q*

Capacitors • Trimmers • Choke Coils • Wire Wound Resistors



HI-Q DISKS AND PLATES

High dielectric by-pass, blocking or coupling capacitors for use where their geometrical shape makes them more adaptable than tubular components. Essentially similar, other than shape, except that in multiple units, **HI-Q** Plates do NOT have to have a common ground, as is the case with the Disk type.

The fast development of the television industry since World War II has been matched, stride for stride, by **HI-Q.** For TV producers were quick to recognize this organization as their most dependable source for the ceramic components they needed in such profusion. They quickly learned that **HI-Q** engineers were competent and resourceful in developing new components to meet new needs as they arose.

Now, though **HI-Q** output has reached several million capacitors, trimmers, choke coils and wire wound resistors each month, never once have the original precision standards or strict adherence to specifications and tolerances been shaded — or the rigid system of inspection of each individual unit at each stage of production been relaxed. The **HI-Q** engineering staff is just as ready as ever to cooperate with your engineers in the production of special components for special requirements.

BETTER 4 WAYS

- PRECISION
- UNIFORMITY
- DEPENDABILITY
- MINIATURIZATION

JOBBERS—ADDRESS: 740 Belleville Ave., New Bedford, Mass.

HI-Q*

*Trade Mark Registered, U. S. Patent Office

Electrical Reactance Corp.
OLEAN, N. Y.

SALES OFFICES: New York, Philadelphia, Detroit, Chicago, Los Angeles

PLANTS: Olean, N. Y., Franklinville, N. Y. Jessup, Pa., Myrtle Beach, S. C.

Unique!

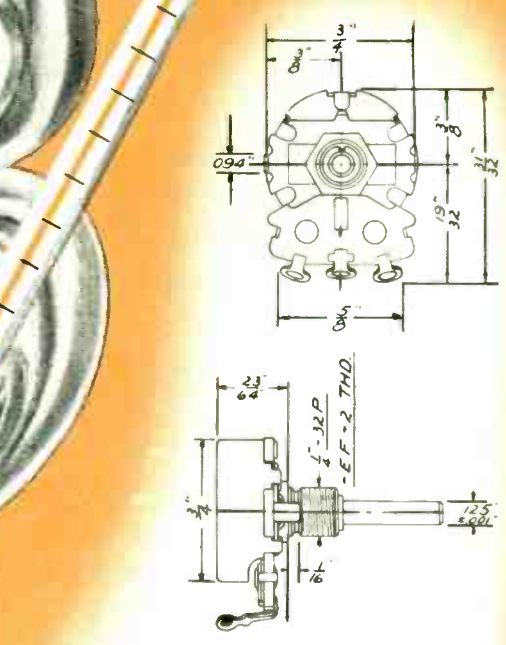
MINIATURIZED

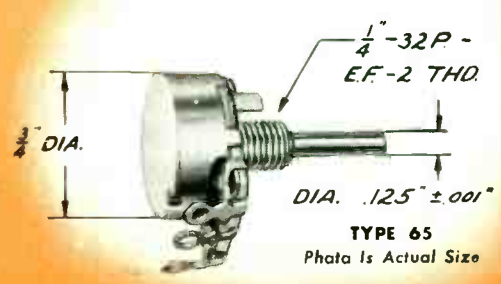
VARIABLE RESISTOR for MILITARY USE

Manufactured from specially developed materials, Type 65 variable composition resistor is a unique, successful combination of miniature size and unprecedented high temperature and humidity stability. For example, Type 65 delivers top performance in fast military airplanes flying in tropical areas where both temperature and humidity vary tremendously from ground level to extreme altitudes.

Exceptionally good delivery cycle due to tremendous precision mass production facilities.

Please give complete details on your requirements when writing or phoning for further information.





TYPE 65
Photo Is Actual Size

VARIABLE RESISTORS (COMPOSITION and WIRE WOUND) WITH or WITHOUT ASSOCIATED SWITCH COMBINATIONS

Specialists in Precision Mass Production of Variable Resistors

FOUNDED 1896

CHICAGO TELEPHONE SUPPLY *Corporation*

ELKHART • INDIANA

REPRESENTATIVES

S. J. Hutchinson, Jr.
401 North Broad Street
Philadelphia 8, Pennsylvania
Phone: Walnut 2-5369

W. S. Harman Company
4057 McClung Drive
Los Angeles 8, California
Phone: Axminister 3-6219

In CANADA

C. C. Meredith & Co.
Streetsville, Ontario

SOUTH AMERICA

Jose Luis Pontet
Buenos Aires, Argentina
Montevideo, Uruguay
Rio de Janeiro, Brazil
Sao Paulo, Brazil

OTHER EXPORT

Sylvan Ginsbury
8 West 40th Street
New York 18, N. Y.

JAN Type RA 20A
2 Watt (CTS Type 252)

JAN Type RA 20B
2 Watt (CTS Type GC-252)

JAN TYPE RA 25A
3 Watt (CTS Type 25)

Meets JAN-R-19

VARIABLE WIRE WOUND RESISTORS TO ALL SPECIFICATIONS IN JAN-R-19

TYPES NOW AVAILABLE

JAN TYPE	CTS TYPE		WATT RATING
RA 20A	252	(without switch)	2
RA 20B	GC-252	(with switch)	2
RA 25A	25	(without switch)	3
RA 25B	GC-25	(with switch)	3
RA 30A	25	(without switch)	4
RA 30B	GC-25	(with switch)	4

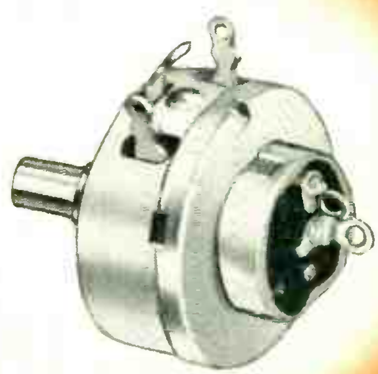
Exceptionally good delivery cycle due to tremendous precision mass production facilities.

Please give complete details on your requirements when writing or phoning for further information.

ILLUSTRATIONS ARE ACTUAL SIZE.

JAN Type RA 25B
3 Watt (CTS Type GC-25)

JAN Type RA 30A
4 Watt (CTS Type 25)



JAN Type RA 30B
4 Watt (CTS Type GC-25)

CTS

CHICAGO TELEPHONE SUPPLY *Corporation*

ELKHART • INDIANA

REPRESENTATIVES

S. J. Hutchinson, Jr.
401 North Broad Street
Philadelphia 8, Pennsylvania
Phone: Walnut 2-5369

W. S. Harmon Company
1638 South La Cienega Blvd.
Los Angeles, California
Phone: Axminister 3-6219

IN CANADA

C. C. Meredith & Co.
Streetsville Ontario

SOUTH AMERICA

Jose Luis Pontet
Buenos Aires, Argentina
Montevideo, Uruguay
Rio de Janeiro, Brazil
Sao Paulo, Brazil

OTHER EXPORT

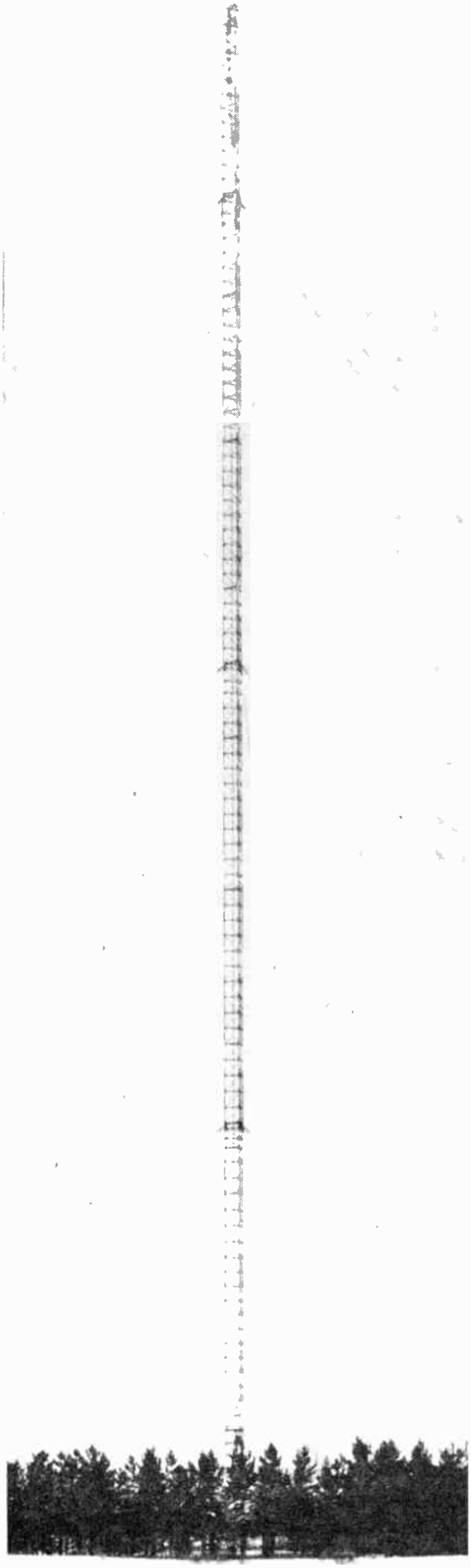
Sylvan Ginsbury
8 West 40th Street
New York 18, N. Y.

A COMPLETE LINE OF VARIABLE RESISTORS (COMPOSITION AND WIRE WOUND) FOR MILITARY APPLICATIONS.

Specialists in Precision Mass Production of Variable Resistors

FOUNDED 1896

TRUSCON...a name you can build on



Truscon Builds

World's Tallest Radio Tower

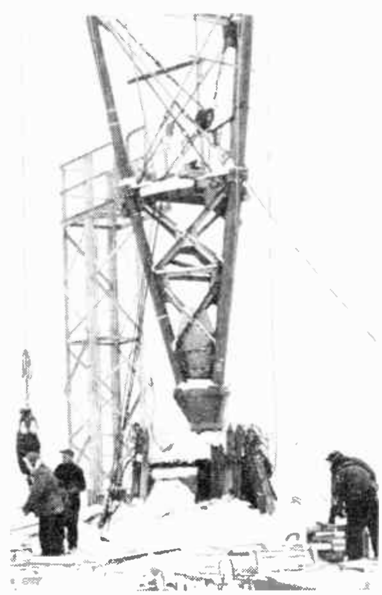
Truscon has fabricated for the United States Government a 1,212-foot tower recently erected near Rome, New York.

The 1,212-foot giant overshadows every other man-made structure in the world, except the 1,250-foot high Empire State building. It is truly a tribute to the skill of the men at Truscon who designed, engineered, and produced it. Not many years ago a tower of this type and height would have been considered impractical to build. Work on the structure began on Truscon's drawing boards in the spring of 1948. Erection was started in September 1950.

Nearing completion last November, the tower had its first test of consequence during the storm that brought record snow and 125-mile-an-hour winds to the eastern section of the country. In this blow the tower swayed approximately seven-tenths of the seven feet it is calculated to sway in a 150-mile-an-hour hurricane.

Requiring 772 tons of fabricated steel, the great structure is supported by 4 miles of guy cables, most of which are anchored almost a quarter of a mile away from the base. The new tower will be used by scientists at the Griffiss Air Force Base near Rome (N. Y.) for the study of Loran, a radio navigation aid first developed during the last war.

Today, rising skyward in many nations are many hundreds of Truscon-built "fingers of steel" over which pour communications for the attentive ears and eyes of the world's people. Your phone call or letter to any convenient Truscon district office, or to our home office in Youngstown, will bring you immediate, capable engineering assistance on your tower problems. Call or write today.



A construction photograph shows giant tower's base. More than 1400 cu. yds. of concrete was used to form base pier and guy anchors.

TRUSCON
MARK OF MERIT PRODUCTS

TRUSCON® STEEL COMPANY
Subsidiary of Republic Steel Corporation
Youngstown 1, Ohio

CHOICE OF THE MARINES FOR *Dependable Controls*

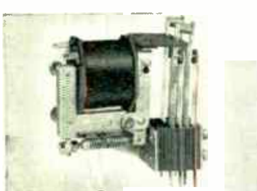
GUARDIAN RELAYS

"From the halls of Montezuma" to marine testing laboratories — technicians are singing the praise of Guardian Relays. The Series 335 D.C. Relay shown above is but one of a most comprehensive line of Guardian Relays *approved* by the Marine Corps and all branches of the Armed Forces. Available with open-type construction or *hermetically sealed* — Guardian's Series 335 D.C. Relay or any Government Approved Guardian control unit used singly, in combination, or in completely packaged controls, will serve you faithfully and well.

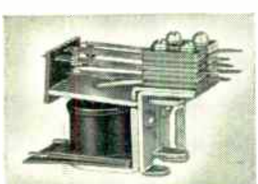
The Series 335 D.C. Guardian Relay shown above is available with the A.N. Connector Plug, Octal Plug and Lug Header hermetic seal *containers. Packs loads of power over a wide operating range, withstands the rigors of dust, moisture, salt air, temperature changes, vibration and impact.*

A.N. CONNECTOR PLUG HERMETIC SEALED CONTAINER

NEW CATALOG on Hermetically Sealed Guardian Relays with various containers is yours for the asking, cost-free.

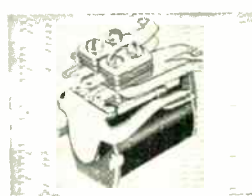


Series 30 A.C.

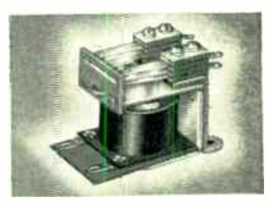


Series 210 A.C.—215 D.C.

Series 220 A.C.



Series 595 D.C.



Series 610 A.C.—615 D.C.

WRITE OR WIRE... FREE CATALOG, SPECIFIC RECOMMENDATIONS, NO OBLIGATION.

GUARDIAN G ELECTRIC

1628-G W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

There's a TIC® Potentiometer for every application

RVP3* High Precision machined aluminum base Potentiometers . . . available in models for either linear or non-linear functions with standard resistance values up to 200,000Ω. Linearity to ±0.1%. Eleven gang assembly shown — example of TIC's potentiometers multi-ganged with TIC's adjustable clamp ring. Can be supplied to meet various mounting requirements — single hole, 3 tapped hole mounting or servo mounting as desired.

Miniaturization of precision potentiometers is keeping pace with the increased demand for smaller assemblies and compact design. Now you can minimize wasted space with TIC's outstanding, new RV7/8* and RV1-1/16* Miniature Potentiometers.

In spite of their thumbnail size the RV7/8 and the RV1-1/16 are precision, high linearity variable resistors (or adjustable trimmers) of high stability — achieving a standard of performance hitherto unavailable in such miniature potentiometers.

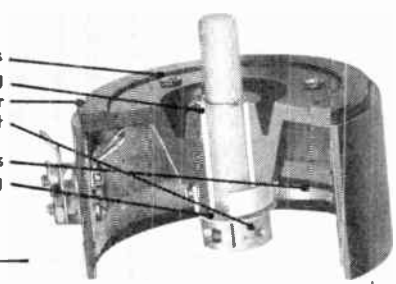
Construction features include: precision machined aluminum base . . . low torque . . . all soldered connections except sliding contacts . . . paliney contacts . . . can be sealed to withstand all humidity, salt spray and altitude specifications. Ganging if desired with TIC adjustable clamp ring.

RV7/8* available with linear resistance elements *only* — nine standard resistance values from 100 to 25,000 ohms. Power rating 6 watts at 25°C. Illustration shows RV7/8 with threaded bushing . . . servo mounting available if desired.

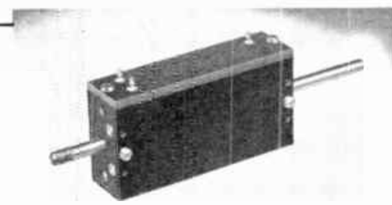
RV1-1/16* available with linear or non-linear resistance elements — nine standard resistance values from 100 to 50,000 ohms. Illustration shows RV1-1/16 with 3 tapped hole mounting . . . servo mounting or threaded bushing if desired.

Sine and cosine potentiometers available in RVP3* and RV2* bases.

Type RV1⅝* and RV2* High Precision Potentiometers . . . semi-standardized types of precision machined aluminum base potentiometers with exceptionally high electrical accuracy and mechanical precision. For both linear and non-linear functions. Designed for precision instrument, computer and military applications. Accurate phasing of individual units possible with clamp-ring method of ganging. Ball bearing models available.



Type RVT Translatory Potentiometers . . . actuated by longitudinal instead of rotary motion providing linear electrical output proportional to shaft displacement. Used as a position indicator, high amplitude displacement type pickup and for studying low frequency motion or vibration. Features exceptionally high linearity and resolution. Available in various lengths and resistance values.



Type RV3* Bakelite Base Precision Potentiometers . . . available in models for either linear or non-linear functions. Stock resistance values ranging from 100Ω to 200,000Ω and power ratings of 8 and 12 watts. 360° mechanical rotation or limited by stops as desired. Potentiometers of this type available to widely varying accuracy requirements (linearity to ±0.25%) — see TIC Bulletin RV3-250. Special models available for high humidity applications.

TECHNOLOGY INSTRUMENT CORP.

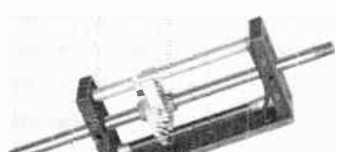
531 Main Street, Acton, Mass.

Telephone: Acton 600

Engineering Representatives

Cleveland, Ohio — PRospect 1-6171
Chicago, Ill. — UPtown 8-1141
Rochester, N. Y. — Monroe 3143
Canaan, Conn. — Canaan 649
Dayton, Ohio — Michigan 8721

Manhasset, N. Y. — Manhasset 7-3424
Boonton, N. J. — Boonton 8-3097
Cambridge, Mass. — ELiot 4-1751
Hollywood, Cal. — HOllywood 9-6305
Dallas, Texas — DIxon 9918



**Numbers refer to diameter of bases.*

Another RCA FIRST!

The new RCA 21" metal rectangular kinescope

featuring electrostatic focus

Now . . . RCA electrostatic focusing in a super-large 21" kinescope . . . combining all the advantages of this new RCA technical achievement with the superior constructional features of metal-shell picture tubes. The tube provides television pictures of equal or better quality than magnetically focused types and is a lasting, long-range engineering accomplishment which goes beyond the present program to conserve critical materials.

The new RCA-21DP4 has a maximum over-all length of only $22\frac{5}{8}''$, and a picture area of $18\frac{3}{8}''$ x $13\frac{15}{16}''$. The frosted Filterglass face is made of high-quality glass, provides improved contrast, and minimizes specular reflection. Since the tube utilizes the structural strength of steel, and weighs substantially less than a comparable all-glass tube, it can be safely shipped in the receiver.

The RCA-21DP4 employs an electron gun of improved design that provides good uniformity of focus over the entire picture area. Focus can be maintained automatically with variation in line voltage and with adjustment of picture brightness. Because the focusing electrode draws very little current, the voltage for the focusing electrode can be provided easily and economically. Design-center maximum voltage rating is 18 kilovolts, diagonal-deflection angle 70°, and horizontal-deflection angle 66°.

RCA Application Engineers are ready to consult with you on the application of the RCA-21DP4 and its associated components to your specific designs. For further information, write RCA, Commercial Engineering, Section 47FR, Harrison, N. J., or your nearest RCA field office.

FIELD OFFICES: (EAST) Harrison 6-8000, 415 S. 5th St., Harrison, N. J. (MIDWEST) Whitehall 4-2900, 589 E. Illinois St., Chicago, Ill. (WEST) Trinity 5641, 420 S. San Pedro St., Los Angeles, California.

The Fountainhead of Modern Tube Development is RCA

RCA

RADIO CORPORATION of AMERICA

ELECTRON TUBES **HARRISON, N. J.**

PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 39 — *June, 1951* — NUMBER 6

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

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

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

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

Allan B. Oxley

REGIONAL DIRECTOR, 1951-1952

Allan B. Oxley, chief engineer of the home instrument division, RCA Victor Company, Ltd., Montreal, was born in London, England, in 1901. After attending night school for 13 years, he took a postgraduate physics and atomic theory course at the University of Toronto in 1929.

He began his career as an apprentice instrument maker, and became a full-fledged instrument maker in 1922; at that time he engaged in the manufacture of microscopes, spectrometers, and fine scientific instruments. From 1924 to 1926 he was mechanician to the department of physiology, University of Toronto, and assisted Sir Frederick Banting in the development of insulin for diabetics. Later, when the Oxley and Meredith Company was created, Mr. Oxley was also engaged in the organization of the IRE Canadian Section. In 1927 he was appointed manager and chief engineer of the King Radio Corporation, and in 1929 became chief engineer of the Williams Piano Company, which had absorbed the King Corporation. In 1933, as chief engineer of the Philco Radio and Television Corporation, he was sent to England to organize a London factory. He returned to Canada in 1935 to become, first, quality engineer, and later, chief engineer for the RCA Victor Company, Ltd., in Montreal.

During World War II he was responsible for a number of important electronic research developments for all branches of the British, Canadian, and U. S. governments.

An Associate of the Institute in 1925, Mr. Oxley assisted in the organization of the Montreal Section. He became a Member in 1933, and a Senior Member in 1946. In 1933 he served as Chairman of the Toronto Section, in 1939 and 1949 as Chairman of the Montreal Section, and is now Regional Director of the Canadian Region.

Mr. Oxley was chairman, also, of the Engineering Committee of the Radio Manufacturers Association of Canada, in 1939, 1940, 1941, 1944, 1945, 1946, and 1949. He was general co-ordinator of the Canadian Radio Technical Planning Board from 1944 to 1948, and has been Canadian representative on the United States RTMA Receiver Committee since 1939. He is a member of the Royal Canadian Institute.

Why Awards

ERNST WEBER

In an age when the long-standing antagonism between far-visioned idealism and allegedly "practical" materialism is especially intense, the following address becomes a particularly significant and encouraging "confession of faith." Its author is director of the Microwave Research Institute, and head of the electrical engineering department at the Polytechnic Institute of Brooklyn, and is also Chairman of the Subcommittee on Tutorial Papers of the IRE Education Committee.—*The Editor.*

One of the highlights at the Banquet during the 1951 IRE National Convention was the bestowal of five specific prizes and awards upon members of the Institute who have contributed in an outstanding manner to the advancement in the art and science of radio and electronics. Three of these awards are in themselves memorials to outstanding late members of the Institute, Harry Diamond, Morris N. Liebmann, and Browder J. Thompson; the fourth is the Editor's award, and the fifth, the highest award, is the Medal of Honor.

In addition, the Institute confers the honorary Fellow grade upon a varying number of members who have distinguished themselves both professionally and through service to the Institute.

Having witnessed several of these memorable occasions and having been asked to respond for the newly elected Fellows at the most recent Convention, I began to reflect seriously about the basic reasons for this tradition. What is at the root of these honorary awards, and what is it that the Institute obtains in return?

Surely all of us have wondered from time to time just what makes an organization grow and expand. Or perhaps, since organizations are composed of people, what motivates a person to belong to a group, to join in an effort transcending his own realm? There are some cynics who pride themselves with superior intelligence, and who hold the view that man does *only* what benefits him as an individual or what gives him personal pleasure—and that there is no such thing as caring for others, or working because of a sense of obligation to mankind. At least as a teacher, I must object most strenuously to such a viewpoint—it is contrary to all basic tenets of educational philosophy! Such a viewpoint is as ruinous for the family as it is for the nation. We certainly would never have this Institute of Radio Engineers if personal motivation were as empty and barren as these cynics would have us believe.

Any social institution *is* what its members make it! If its members do not care, if they let "the others" worry, let "the others" be the fools to work—then that institution, however great it might appear at the moment, is *doomed;* it will crumble and disappear. If on the other hand, there are enough members who *actively* support the institution, who freely devote time and effort to its tasks and obligations, who share in the belief of a common goal and of common interests in higher achievements, then that institution will grow strong and prosper.

Looking back at the growth of The Institute of Radio Engineers, one cannot but conclude that it must have had a strong core of devoted men who made it what it is today, who gave, and still are giving unstintingly of their time and effort for the promotion of its services to the members. These men have created the meaning of honors and awards conferred by the Institute, they have made the "Fellowship" in this great professional Institute worthy of its distinguished connotation!

I take it, then, that awards are the tangible expression of encouragement to continue the individual professional contributions with renewed and greater effort in order to serve better this Institute, our profession, and thus, our country.

What the Engineers Have Done to Television*

ROBERT·C. SPRAGUE†

MY APPEARANCE here tonight reminds me or an experience I had when learning to ski a few years ago. On looking down the precipitous slope I asked myself, "What am I doing here?" I might ask myself a similar question tonight. "What is a parts manufacturer doing addressing a gathering of engineers?"

Of course I realize that the invitation to speak here resulted from the fact that in a weak moment last June I accepted the job of president and chairman of the board of the Radio-Television Manufacturers Association. Fortunately, I will soon unload the responsibilities of presidency on the shoulders of another and younger man.

I do not need to tell you that the job of heading an Association which represents an industry as volatile, dynamic, and explosive as the radio-television industry is not a moonlight cruise on calm seas. The hazards of accepting such a responsibility, like those of public speaking or letter writing—the latter of which have been recently illustrated at the White House—are many and the rewards, if any, are uncertain.

However, I would be less than frank if I did not admit that there are many compensations, and most of us in this situation are like the farmer who said that he was going to town Saturday night to get drunk and complained, "Oh how I hate it."

I do feel at home, however, in speaking to such a representative group of our industry's engineers although for some years I have lived on the management's side of the railroad tracks. While there may be a wide difference of opinion as to whether management, sales, and promotion or technological developments are chiefly responsible for the remarkable development of radio and television in the last quarter of a century, I note that the radio engineers were at least smart enough to organize first. I refer to the fact that The Institute of Radio Engineers was founded in 1912, while RTMA was not formed until 1924.

Seriously, I believe that the radio-television industry's achievements may be credited to three major factors: research and development, production efficiency, and alert management and salesmanship. Together they have made possible the mass production of high-quality radio and television receivers, as well as other electronic equipment so important to our Armed Forces and to other industries.

The subject I had adopted tonight "What the Engineers Have Done to Television," may sound facetious. I assure you that it is not. It is obvious that were it not for the engineers, there would be no television. Less widely known is the fact that there would be no television as we know it today except for the unselfish work of a large number of engineers who have consistently sought to protect the public interest, as well as the interests of the industry, by resisting experiments conducted at the expense of the consumer's pocketbook. It is certain that if the action of these engineers, and of the industry generally, had been only self-seeking, or faulty, or erroneous—as has been recently implied by a member of the Federal Communications Commission—that the public would have rebelled and television would not be the multibillion dollar industry it is today.

In a supplemental opinion issued by Commissioner Robert F. Jones in connection with the FCC's preliminary findings in the color television case, the "good faith, truth, and veracity" of prominent industry engineers were questioned. Commissioner Jones stated that the hearing had made it "crystal clear that the industry's engineers were unsound analysts," and in 1946 and 1947 had given testimony which was "completely worthless" because "their economic interests blinded their engineering judgment." He added, moreover, that there was "grave doubt" that these engineers could "be relied upon to predict the potential performance of any system whose adoption might prejudice their economic interest." The Commissioner also charged that the industry attempted to "confuse the Commission" by "sham engineering testimony."

A judicial examination of these charges made by a responsible government official raises the following interesting questions affecting professional engineers and their relationships to their industrial employers:

Are honest differences of engineering opinion legitimate?

Do engineers understand more fully systems and equipment at the later stages of development than in earlier stages?

Accordingly, may earlier, and honest engineering opinions, require subsequent and legitimate modification?

Is it justifiable for professional engineers (as it is for medical doctors) to require detailed technical knowledge and extensive tests (clinical experience) before approving and widely adopting something new?

Should economic factors (that is, the cost of new and replacement apparatus, obsolescence of older apparatus, high manufacturing costs resulting from changeovers) be considered by engineers in determining the value of a system or equipment? If not, how should engineers be "insulated" in an "economic vacuum?"

In general, are engineers less sound analysts than statesmen, politicians, other professional men, industrialists, and so on? If they are, what is the reason for their inadequacy?

Are engineers more prone to offer fallacious, sham, questionable, or worthless conclusions than any other of the previously mentioned groups? If so, why?

I do not think it necessary for me or any one else to come to the defense of our industry's engineers. Their record for more than a quarter of a century speaks for itself. However, I would like to take this opportunity to deny categorically the charges which have been hurled at our industry and to comment briefly on what may properly be expected from the engineering fraternity.

Engineers are highly individualistic persons who often do not agree with each other or their commercial colleagues. They are influenced in their decisions by facts and figures, and not by pressure from whatever source.

It is doubtless true that engineers recognize economic factors and hesitate to spend money recklessly, whether it comes from the ultimate consumer or their employer, and I admit that this is rather a novel characteristic in these times.

Our industry engineers generally strive to offer to the public the fruits of their technological developments at the lowest possible cost, but they seek to avoid foisting on the public either systems or equipment which have not been thoroughly tested.

It is true that engineers are sometimes brutally frank in expressing opinions, whether they speak as individuals or representatives of industry or engineering groups, but I think that any fair-minded person would concede that they are honest and fair in their ultimate judgments.

Perhaps a brief review of the early history of television and the role played by engineers in its development will be timely. After early experiments, black-and-white television had developed to a point early in 1940 where the Radio Corporation of America made plans to market television receivers and to place commercial television stations on the air. The Federal Communications Commission at that time regarded such action as tending to freeze television standards and thus block its development and discourage further research and experimentation. The Commission in its ruling stated that there should be no commercial TV broadcasting until, and I quote, "the probabilities of basic research have been fairly explored."

The FCC added that as soon as the majority of engineers in industry were prepared to approve any one of the competing systems of broadcasting as standard, the Commission would authorize commercial television operations.

As a result of a meeting between FCC Commissioner James Fly and W. R. G. Baker, director of engineering for many years for RMA and RTMA, and a former President of IRE, the first National Television System Committee was formed. The monumental work of that committee is a matter of record. In six months this committee developed standards and made a report to the FCC. Its technical panels of experts produced reports and minutes totaling 600,000 words, devoted 4,000 man hours to meetings and an equal time to travel, and witnessed 25 technical demonstrations. As you know, the standards for black-and-white television proposed by

* Decimal classification: R583.17×R071. Original manuscript received by the Institute, March 21, 1951; delivered, Annual Banquet, 1951 IRE National Convention, Waldorf-Astoria Hotel, New York, N. Y.
† Radio-Television Manufacturers Association, Washington, D. C., and Sprague Electric Company, North Adams, Mass.

these industry engineers were adopted and television was off to a successful start, only to be delayed several years by our entry into World War II.

Perhaps the best measure of success of this almost unparalleled example of co-operation among industry engineers is that several foreign countries have adopted television standards based upon those of the United States with only minor variations.

In 1941 the same group of engineers who recommended standards for black-and-white television considered color television, and reported that the state of the art was such that it was undesirable at that time to attempt to set color standards. Five proposed color systems, intended to operate in a 6-mc channel, were analyzed by the NTSC. Four of these were variations of the CBS system with the principal differences in lines, frames, and type of interlace.

Much has been made of the fact that members of four of the NTSC committees were overwhelmingly in favor of color television as demonstrated by the Columbia Broadcasting System. Overlooked, or at least underemphasized, are some other opinions expressed by these engineers at the same time. By substantial majorities these engineers voted against adoption of color standards at that time and expressed the belief that black-and-white standards should not be influenced by color television considerations.

At the risk of digressing, I would like to comment that the question of whether any one prefers color in television is as elementary as "do you like a sunny day?" Of course people like color, and engineers, I believe, are usually regarded as people. But when it comes to a question as to whether color television is feasible, economical, and workable, there is no one better qualified to voice an opinion than the engineer who can consider color television in relation to all of the complicated and technical aspects of the transmission and reception of information by means of electronic equipment.

Following the war, in 1946 the first formal request that color television standards be established was made to the Federal Communications Commission. At this time the Radio Technical Planning Board—another and newer committee of industry engineers—stated that "adequate standards for color television for a 6-mc channel cannot be established at this time." The FCC apparently concurred in this opinion and denied the request that color television standards be adopted.

Again in 1948 the Commission reopened hearings on color television, and another industry committee—the Joint Technical Advisory Committee, sponsored by RMA and IRE—stated that in its opinion, based on evidence submitted by various subcommittees, it was "impracticable to set up commercial standards for color television in the present state of the art."

The most recent FCC hearing on color television in 1948–1949, I am sure, is fresh in your mind and the events need not be recounted. It should be noted, however, that a majority of the industry's engineers were in agreement that a compatible, acceptable color system of television could be developed within a reasonable length of time. This testimony was not accepted by the Commission, as it adopted the CBS field-sequential system instead. Now, of course, actual color broadcasting on the CBS system has been delayed by court action, and the present industrial mobilization program with its attendant shortage of critical materials has, at least for the time being, made color an academic question.

Thus, as I have pointed out, the engineers have been in the center of the color-television controversy for the past ten years, and until recently their opinions have been respected and their judgment followed. The great contribution television is making and will continue to make in the fields of entertainment, education, and commerce is sufficient evidence of the soundness of this engineering judgment. When practical color television is ready for the public, the industry's engineers will again be the key men in bringing this additional service to the public at the lowest cost and with a minimum of dislocation.

If wholesale indictment of your great profession, which I quoted a few minutes ago, were not so damaging and unfair, it would be ridiculous. You have given to the world so much in the field of electronics that I am confident the public does not concur in the views expressed by this government official. Your work, individually and collectively, and that of your great Institute, utterly refutes these unjust charges. I only urge you to continue to be guided in the future, as in the past, by your own high standards, and I venture to offer you this credo and code for engineers:

Differences of engineering opinion can and should honestly exist. Unanimity of engineering judgment is not necessarily desirable.

Changes in engineering opinion resulting from the passage of time and the gaining of experience are healthy, and normally to be expected.

It is improper and futile to ask engineers to reach final and valid conclusions at too early a stage in the development of new systems or equipment.

It is not only ethically correct but essential in the public interest that engineers shall duly weigh the economic factors involved in their planning work. Engineers are charged not only with the production of devices, but also with the selection of the least expensive devices which, under existing and likely future circumstances, will adequately serve the public at the lowest cost. In other words, engineering includes common sense.

There is conclusive evidence that engineers as a group and individually have at least as high moral, ethical, professional, and mental standards and qualifications as any other useful group of citizens.

There is also unchallengeable evidence that engineering societies are openminded and fair in their solicitation and selection of papers for publication in their journals, in their committee membership from various companies, and in their honorary awards.

We in the RTMA are rightly proud of what our industry has done for America in terms of communications, radio broadcasting, television broadcasting, military security, and host of other services. We know that the national welfare and prosperity have been furthered by our efforts and achievements.

You in The Institute of Radio Engineers have labored mightily and successfully to make your aims and ours come true. You properly share our pride of accomplishment in the public interest.

May both our organizations long continue to co-operate effectively for the happiness, comfort, and safety of our nation.

Before closing, I would like very briefly to sum up how it appears to me the present mobilization program will affect the radio-television industry. Our latest information indicates that military electronics production will reach a peak rate of $2.5 billion in the fall of 1952, and thereafter will decline to an annual rate of about $1.5 billion. In this connection, it should be borne in mind that military production dollars have about half the impact on our industry as civilian production dollars. This is for a variety of reasons, but particularly because a considerable portion of special and elaborate mechanical gear is obtained from manufacturers not generally considered a part of our industry.

These figures on military electronics production indicate that our industry will not be so heavily loaded with military contracts but rather it will be able to maintain a substantial amount of civilian production, even at the peak of the military output—except in the unfortunate event of an all-out war. Apparently, only the shortages of certain critical materials will prevent manufacturers from turning out as many radio and television sets as their plant facilities and military orders will permit. It is, therefore, highly important that government officials make provision in their planning for the healthy continuation of our civilian economy, for we do not know when all of the present manpower and production facilities in our industry may be needed for the nation's defense.

A Dual-Diversity Frequency-Shift Receiver*

D. G. LINDSAY†

Continuing the planned program of publication of valuable tutorial material, there is here presented a paper of this character. It has appeared in the *Proceedings of the Institution of Radio Engineers, Australia*, and is available to these PROCEEDINGS OF THE I.R.E. because of the close and co-operative arrangements existing between these two Societies. It has also received the approval of the Subcommittee on Tutorial Papers of the IRE Educational Committee.—*The Editor.*

Summary—**Major improvements have been made in radio-telegraph transmission over a period of years, but the improvement resulting from signaling by frequency shift instead of interruption of the carrier wave has been achieved in practice only recently. These matters are discussed, and associated with a number of references to other work in this field. A description is given of dual-diversity receiving equipment, employing audio-frequency filters for separation of "mark" and "space" signals. The equipment is intended for use on direct-printing, commercial telegraph circuits.**

I. Introduction

IN THE EARLY days of radio telegraphy, the method of operation involved aural reception and hand-key transmission by skilled operators at average working speeds seldom in excess of 20 words per minute, even under the best transmission conditions. Because of the discriminating ability of the human ear, experienced operators could copy very weak signals, marred by noise or interference, with remarkable accuracy. The radio circuit systems available adequately fulfilled the service necessary in those days, but requirements soon arose for greater efficiency of use in many radio circuits: for instance, higher speed or more direct methods of conveying the message such as by direct-printing methods became desirable. To this end, the use of methods and apparatus borrowed from wire telegraph technique quickly revealed that the speeding up caused by the removal of the human limitation in Morse speeds, removed also one of the most valuable of the operator's contributions, that of discrimination between signal and noise. To be of much practical use for commercial purposes, a radio-telegraph circuit employing any of the "mechanical" substitutes for operators was found to require, by comparison, a much better signal-to-noise ratio. These requirements involved the development of improved apparatus, the use of high transmitting power, better aerial systems, and so on, resulting in increased costs for equipment and maintenance.

A definite limitation to the increase of signaling speeds on the low frequencies then in general use, resulted from the increased bandwidth required for proper transmission of the signaling characters and the difficulty of obtaining this, due mainly to the low decrement of the transmitting aerial systems. At about this time the possible implications of scientific theories and some practical investigation, into the propagation of radio waves over the earth's surface, gave rise to ideas that frequencies above 1,500 kc, previously considered useless, might in fact prove valuable, especially for long-distance communication. It was realized, at once, that the use of very much higher frequencies would almost eliminate the restriction on signaling speeds and development of these new ideas into practical apparatus and communication circuits proceeded at a remarkable rate. The provision of direct radio circuits between England and Australia was among the first of the major communication projects to employ the high-frequency range, and due to the foresight and skill of the engineers concerned proved to be remarkably successful from its inception.

The new communication medium was found to have many vagaries in transmission conditions and it seemed almost impossible to predict the reliability of any given communication circuit, although the practical results were generally very good, showing great improvement over the previous technique in economy and speed. Further careful investigation and analysis of practical results revealed many methods for further improvement, such as directive aerial systems (in use on the England-Australia circuit from its inception), diversity reception, transmission employing frequency diversity by small variations of the transmitted frequency, general improvement in apparatus, and the application of much information on the correct choice of operating frequencies. These developments were made, and applied to commercial systems progressively, over a period of years.

By this time the use of machine telegraph methods had gained wide acceptance for commercial circuits, the improvements in technique and apparatus providing signals sufficiently good in quality for this purpose, despite the further vagaries of the transmission medium as compared with the earlier circuits on low frequencies, while the search for further improvements went on continually. The keying methods in use up to this time were, almost universally, the transmission of the carrier wave for the "mark" code. It had been realized for a long time that transmission of the required intelligence by varying the frequency of the emitted wave would be possible, but it was not appreciated that some advan-

* Decimal classification: R361.107. Original manuscript received by the Institute, May 2, 1950. This paper comprises the substance of a lecture presented before the Radio Engineering Convention arranged by the Institution of Radio Engineers, Australia, at Sydney, November, 1948, and later published in *Proc. IRE* (Aust.), 11, 29, February, 1950.

† A.W.A. Radio-Electric Works, Ashfield, N.S.W., Australia.

tage may accrue by so doing. In this way the method of telegraph signaling by shifting of the carrier-wave frequency, instead of interrupting it, now gaining wide acceptance, may be compared with its near relation in the telephony field, frequency modulation, the advantage of which has taken a long time to be appreciated.

In the very early days of radio telegraph transmission, signaling by means of frequency shift was employed because of difficulties in interrupting the carrier wave. This was the case in Poulsen arc transmitters and the keying method involved the transmission of the normal frequency for the "mark" condition and a shift of frequency, by altering the tuned circuit values with a relay or similar device, for the "space" condition. In this case, the "space" signal was not used to convey intelligence and only created undesirable interference, although it has been stated that some operators were sufficiently skilled to be able to copy the "space" signal, when the "mark" or normal frequency was subject to bad interference, often by the "space" signal of some other transmitter.

The basic difference between the modern methods of employing frequency-shift transmission and this very early example is that the frequencies corresponding to "mark" and "space" in the former are both used to convey intelligence.

The first major use of frequency-shift radio-telegraph appears to have been made during World War II (Sprague, 1944, Buff, 1946, and Vanderlippe, 1947). The lack of trained code operators and the desire for expediting communications made the advantages of the frequency-shift system attractive and, generally, good results were achieved. The large improvement obtained, however, tended to force the application unduly and resulted in apparatus originally designed for on-off keying being modified by the addition of various units, often with unsatisfactory results requiring frequent adjustments. This was because of the unsuitability of the original equipment, mainly with respect to frequency stability. Endeavors to overcome this by various expedients in the frequency-shift apparatus itself did not prove entirely successful.

II. General

Principle of Frequency-Shift Keying

As used in the frequency range between 2 and 30 mc, frequency-shift keying involves the transmission of two slightly different frequencies for the "mark" and "space" signaling conditions. The amount of frequency shift employed may vary with desired applications; one value that has been frequently used is a total shift of 850 cycles, that is, plus and minus 425 cycles from the assigned frequency, the "mark" signal, by the generally accepted convention, being the higher frequency of the two. In comparing the method with on-off keying it may be said in general terms that the two methods are special examples of the two fundamental modulation methods, amplitude modulation and frequency modulation. On-off keying is equivalent to 100 per cent amplitude modulation of a carrier with square waves and, similarly, frequency-shift keying of the type described is equivalent to frequency modulation of a carrier with square waves.

General Methods of Transmission and Reception

As might be inferred from the details given above, frequency shift transmitters and receivers require a high degree of stability in the frequency determining elements, both for the amount of frequency shift and the nominal value of the carrier frequency. In practice, depending on the particular arrangement adopted for the apparatus, the value of frequency shift may be allowed to change a few hundred cycles due to variations of the shift itself, or the carrier frequency, or both, unless automatic-frequency control or bias-correcting circuits are used, in which case variations in carrier frequency of a relatively slow nature, up to perhaps 2,000 cps, may be tolerated. Where it is desired to keep the channel width small total variations of ± 200 cps or less are desirable. These requirements are reflected in the design of both transmitting and receiving equipment, as the achieving of such stability is by no means an easy matter, although it can be obtained by the application of existing technique.

Of several methods that have been developed for use in the frequency-determining portions of a frequency-shift transmitter, one particular method has been widely used. This method involves the direct-frequency modulation of a low-frequency oscillator on a frequency of about 200 kc by the signaling code, the mixing of this output with a crystal oscillator and selection of the upper sideband frequency of the combination, followed by frequency multiplication and amplification to reach the desired final frequency and power output. Apart from the keying and frequency-control sections, the transmitter is generally quite conventional and equipment previously used for on-off keying is frequently employed, sometimes with a small reduction in power output because of the continuous duty-cycle. Both low-frequency oscillator and crystal oscillator are ordinarily temperature-controlled and a stability of a few cycles per megacycle can be maintained over periods of many hours with such equipment, accompanied by small variations of the amount of frequency shift in the order of a few cycles. Descriptions of frequency-shift transmitter exciter units using this general method are given in several of the references listed at the end of this paper.

Numerous methods have been developed for the reception of frequency-shift signals and it is safe to say that many more will be developed; differences in some cases are of a detailed nature but nevertheless may be important in obtaining the best possible performance in practice. The technique can be compared in some respects, again, to the developments in frequency-modulation receivers for telephony, where the variety of methods is ever increasing. In frequency-shift reception there is as yet no ideal system which makes all others obsolete; rather there are a number of methods, some differing only in detail., which have their own advantages and disadvantages, the choice of any one of these depending largely on the desired application. The choice of systems, for instance, may well be different for high-speed telegraph circuits using tape recorders, from a circuit in which it is desired to use direct-printing teletypewriters where the signaling speed is much lower.

The actual receiver units employed for frequency-shift reception are usually fairly conventional and not greatly different from those employed in on-off telegraph reception. Good frequency stability is a basic requirement; other design differences may result from the differing requirements of operating frequencies and similar factors. It is usual to employ two receivers in space diversity, that is, operating from aerials spaced apart five wavelengths or more, employing common oscillators for frequency determination, a common automatic-gain-control system and some means of combining the final outputs in the succeeding detector apparatus. It has been found that an improvement of the order of 10 db is obtained from the use of two receivers in diversity where noise is a limiting factor, and greater gain where multipath fading is a limitation (Petersen, Atwood, Goldstine, Hansell, and Shock, 1946). Also, there is only a small amount of further improvement by using three receivers in diversity; in this respect, frequency-shift systems are different from on-off systems, as in the latter a considerable improvement is effected by using three receivers in diversity. A number of points of interest in connection with the design of receiver units are covered in a later section of this paper where a complete equipment which has been developed is described.

It is in the means employed for dealing with the frequency-shift signals after passage through the receiver units that the main differences in the various methods appear. It is convenient, for the objectives of this paper, briefly to describe and contrast two methods which have found considerable application. These are referred to as the "audio filter" and "discriminator" methods.

In the former method, the receiver units are so designed that audio-frequency outputs are obtained from the receivers at a suitable frequency, spaced, of course, by the amount of frequency shift. If we take as an example a shift of 850 cycles, the two frequencies could be 2,125 cycles ("mark") and 2,975 cycles ("space"). These frequencies are applied to a filter for noise reduction, to a limiting amplifier to eliminate level variations, and are then applied to separating filters which comprise narrow band-pass filters centered around the "mark" and "space" frequencies, respectively. Further amplifier and rectifier stages (one each for "mark" and "space") then deal with the separated signals and derive the desired output for operation of the telegraph machine.

In the "discriminator" method the signal, varying in frequency by the "mark-space" difference, is usually taken from the receiver at the intermediate frequency and is then amplified and converted to a lower frequency, in the range 30 to 50 kc, further amplified and limited, and finally applied to a linear discriminator of the type often used for frequency-modulation telephony. The output of the discriminator is followed by a low-pass filter, chosen to suit the keying speed, and comprises nearly square waves of opposite polarity for mark and space, which are arranged to derive the output signals for operation of the telegraph machine.

Although the method in which the audio filters are used is commonly referred to as the "audio-filter" method, it must be appreciated that these audio filters constitute a frequency discriminator and the main difference between this and the so-called "discriminator" method is that the former is a *nonlinear discrminator* while the latter is a *linear discriminator*.

An advantage of the "audio-filter" method is that it can be so designed that moderate amounts of frequency drift do not cause telegraph bias. Provided the mark and space frequencies fall within the pass bands of their respective filters, it does not matter whether they are exactly centered or not. Further, during a period in which the communication circuit is idle and the transmitter is radiating a continuous mark condition, as is the normal convention, the output of the mark circuit is maintained and will hold the recording apparatus in a nonoperating condition. One of its disadvantages lies in the fact that it must be designed for a definite value of frequency shift. There is some latitude in the shift which can be handled, but it will be realized that the centers of the mark and space filters will normally be arranged to be separated by the expected shift. If designed for a shift of 850 cps the filters would not be optimum for a shift of, say, 500 cps. Most transmitter arrangements employed for frequency-shift circuits can be adjusted readily for various values of frequency shift but the inflexibility of the receiver in this respect is nevertheless a disadvantage. A second disadvantage is that the filter cutoff characteristics place a limit on the keying speed due to "ringing" effects. This method was originally designed to work with a fixed shift of 850 cps and for a definite telegraph speed in the vicinity of 50 bauds, corresponding to a speed of about 60 words per minute, the usual maximum working speed on teletypewriter circuits employing the special 5-unit code for these systems. For this particular service there is no doubt that the method gives very satisfactory results.

The linear discriminator method was initially developed to accommodate transmitter and receiver drift which could not be handled by the nonlinear discriminator system. A linear discriminator can be designed so that it has a long linear characteristic sufficient to accommodate the frequency shift plus the expected drift, assuming the receiver portion has an appropriate selectivity characteristic. The difficulty with this method is that when drift does occur, some standing direct current appears in the discriminator load network, directly proportional in amount to the drift from the true center frequency of the network, thus causing a bias to either "mark" or "space." Various circuit arrangements adopted to eliminate this bias tend to introduce other difficulties, one of which is a risk of false operation due to noise during idle periods, when the transmitter is signaling a continuous "mark." There is some evidence that the linear discriminator system is actually less tolerant of frequency variations than the "audio filter" system, even where bias correction is applied. Automatic-frequency-control arrangements are not usually applied in this system and are, in fact, somewhat difficult to apply, whereas such systems are frequently used, and can be more easily arranged, with the "audio-filter" system. The major advantages of the linear discriminator method are first, that a wide range of frequency-shift values can be accommodated, and second, there is no particular limitation on the telegraph speed.

It is emphasized that the details given above are a considerably over-simplified picture of the two methods; it is not completely fair to the merits of either, as detail variations that have been used in each, particularly in diversity-combining circuits, make it difficult to draw a sharp line of demarcation between them. There is, as yet, insufficient evidence of either a theoretical or practical nature to say definitely that any one system is the better; each has its advantages and disadvantages, as stated earlier in this section. Moreover, it is not the purpose of this paper to discuss these matters in full detail. Some of the references given at the end of the paper treat this portion of the design problem in greater detail, although there is some divergence of opinion, indicating that only further development of more extensive practical tests will clarify the position. Considerations such as ease of operation, ability to change readily existing on-off keying equipment and similar items, can also influence the choice of a particular method.

III. Improvement from the Use of Frequency-Shift System

Considerations of Signal, Noise, and Fading

The frequency-shift system does not show any advantage over on-off keying for operation over radiotelegraph circuits that provide stable conditions and adequate signal levels similar to those on wire-telegraph circuits. In the case of long-distance radio-telegraphy in the high-frequency range, however, frequency shift shows a marked advantage. It is not possible to express quantitatively, by a single figure-of-merit, just how much better is the frequency-shift system in practice. This is because of the rapid fading and high noise conditions, multiple-path fading effects and so on, which commonly prevail in the high-frequency range. The results of theoretical treatment, laboratory tests, and field trials yield a unanimous verdict that there is a major improvement, but it is only possible to express this quantitatively under certain assumed conditions.

The result of extensive field trials are reported in a paper by Petersen, Atwood, Goldstine, Hansell, and Schock (1946). These tests were conducted over a considerable period, employing error counts to determine quality of a particular circuit, and adjustment or extrapolation of transmitter power to equality of error count was employed to give a comparison in terms of transmitted power, or field strength. These results are, of course, an average, not representing any one specific condition. It was found that a dual-diversity frequency-shift system, as compared with a three-receiver diversity system using on-off keying, showed an equivalent power gain of 11 db. The results also indicated that the gains caused by diversity reception (as compared with a single channel), amounted to approximately 10 db for both frequency-shift and on-off keying (two and three receivers, respectively) when noise is the limitation, with considerable diversity gain in the frequency-shift case where multipath distortion and not noise is a limitation. The gain caused by diversity reception where the limitation is multipath distortion cannot be expressed in terms of a power gain, as errors independent of power are caused.

In a paper by Ruddlesden, Forster, and Jelonek (1947), an interesting theoretical treatment is given of the improvement in signal-to-noise ratio by using the frequency-shift system. With the help of a number of simplifying assumptions, values of 10 and 14 db are obtained for the improvement, using audio-filter and discriminator-conversion methods, respectively, for a single channel. If it is inferred that double-diversity reception eliminates most of the effects caused by fading, the agreement between these figures and those quoted above for field trials is of interest.

A very detailed paper by Davey and Matte (1948) gives figures on the improvement caused by the frequency-shift system where noise is the limiting factor, of the order of 3 to 6 db, again for a single-channel system. These results do not agree closely with those quoted above. Greater improvements are indicated, however, for smaller amounts of signal distortion.

It is in dealing with rapidly fading signals, such as are usually encountered in hf telegraph circuits, that the circuits and the frequency-shift system shows to its greatest advantage. In considering this it is necessary for purposes of description to distinguish between non-selective or "flat" fading and selective fading, sometimes referred to as "multipath propagation." Selective fading implies propagation by such paths that two or more signal elements can arrive separated in time by amounts up to two milliseconds or more. This is significant since phasing conditions for slightly different frequencies can cause wide variations in level. Non-selective fading means that large level variations may occur without appreciable differences in path length so that frequencies a few hundred cycles apart do not fade by greatly different amounts. In practice, of course, both types often occur in combination over periods applying to communication.

Assuming, for the moment, that diversity reception is not being employed, considering nonselective fading and the on-off keying system, the effect of level variations, slow compared with the signaling speed, can be reduced by suitable automatic-gain-control (agc) circuits. In the case of rapid variations, the automatic-gain-control circuits can do little to assist and as the on-off system is relatively critical to level changes in the detected output, unless some compensating arrangement is adopted, failure will occur in hf circuits where the level variations are often extreme and rapid. Moreover, the failure will occur regardless of transmitted power. In a frequency-shift system, the agc circuits can have much smaller time constants, because of the presence at all times of either "mark" or "space" signals; efficient limiting amplifiers can be used and the system as a whole is not particularly critical to level changes, the frequency change being the significant factor. Some telegraph bias is produced if extreme level changes occur at the moment of frequency change between "marking" and "spacing" conditions, but in general the frequency-shift system is much more tolerant of nonselective fading.

With selective fading, again assuming that diversity reception is not used, the effect of large differences in path length in the normal on-off system is usualy a filling in of the spacing intervals with resulting telegraph bias; in many cases the distortion of characters is very severe. In frequency-shift systems, large time differences in the propagation paths (2 to 5 milliseconds) give rise to disturbances at the beginning of each signal element, caused by simultaneous arrival of "mark" and "space" signals resulting in a beat tone of frequency equal to the amount of frequency shift employed. As the frequency shift is generally several times greater than the highest, significant, modulating frequency which results from the keying signals, the action of the normal types of frequency-shift, detection arrangements largely reduces these effects. There is some divergence of opinion concerning the merits of frequency shift in relation to selective fading, but a recent paper by Bray, Lillicrap, and Owen (1947) gives a great deal of valuable data on this point, these workers employing a "fading machine" to perform tests under controlled laboratory conditions. Their evidence indicates that distortion of the signal elements in the usual frequency-shift systems can be much smaller than with on-off keying and the results are supported by field trials to which reference has previously been made (Petersen and others, 1946).

Diversity reception is commonly employed in hf radio-telegraph circuits, usually of the space-diversity type, as frequency diversity is wasteful in channel width. Diversity methods can improve greatly the performance of both on-off keying and frequency-shift keying systems, provided the diversity combining or selective method has the proper action and speed of response. This is especially important for frequency-shift keying (Lyons, 1946). With on-off keying the diversity action also helps to average rapid level changes to which this system is sensitive; for this reason, an appreciable improvement results from increasing the number of receivers providing the diversity action, from two to three. Since a frequency-shift system is tolerant of level changes, it could be expected that a smaller improvement would be obtained in changing from double to triple diversity and this is borne out in practice.

Bandwidth

It is generally agreed that for a given signaling speed there is no great difference in receiving bandwidth requirements for on-off or frequency-shift keying. The "audio filter" method of frequency-shift reception does not appear to require a wider band than the "discriminator method" (Jones and Pfleger, 1946), but the difference is not of great practical significance as the failure point caused by noise does not seem to be affected.

For transmitters, the channel width occupied by a communication circuit operating at a given telegraph speed should be as narrow as possible; this matter is becoming of increasing importance in an already overcrowded spectrum and the interference created by any circuit must be reduced to a minimum. Contradictory though it may appear at first sight, the frequency-shift system is more economical in channel width and creates less interference than conventional on-off systems operating at equivalent speeds. With either system, keying by signal elements which are actually square waves would result in the occupation of a wide band and the creation of much undesirable and unnecessary interference. It has been found that considerable departure from the theoretical square-wave signal elements can be tolerated, provided the circuits pass harmonics up to the third of the effective modulating frequency (proportional to keying speed). Shaping of the signal elements by filter networks is employed for this purpose. In conventional on-off transmitters, however, it is usual to carry out the keying in low-level stages, allowing the successive higher power stages to cut off as the driving energy is removed. These later stages are ordinarily operated at the highest possible efficiency, that is under class-C conditions. In these circumstances well-shaped keying signals can become considerably distorted in their passage through such amplifier stages, giving rise to undesirable steep-sided transients which spread the band over which interference is created. The obvious alternatives of carrying out the keying and shaping of signal elements at a high power point, or employing class-B linear stages after a low-power stage which is keyed, are not at all attractive from the practical viewpoint. It is reasonable to state that insufficient attention has been given in the past to these factors and that valuable channels are not being used to the best advantage.

By contrast, the frequency-shift system enables the proper shape of keying signals to be preserved in passage through successive stages, since there is always a signal present and the small variations in frequency from "mark" to "space" pass easily through circuits as normally used.

Thus, it appears reasonable to expect a considerable reduction in side-frequency amplitude with frequency-shift systems, as compared with conventional on-off systems. A paper by Wickizer (1947) gives both theoretical and practical data on this matter; reduction in side-frequency amplitudes measured at one kilocycle from the carrier frequency of at least 20 db, are indicated. Similar conclusions are reached in a paper by Davey and Matte (1948). These considerations and data disclose a further important advantage for frequency-shift keying.

In telegraph communication circuits of either type relatively narrow bands can and should be used. Frequency stability of both transmitter and receiver can have a large effect, therefore, on the effective bandwidth occupied. There is every reason to demand the best possible stability in either on-off or frequency-shift systems so that the band occupied shall be no wider than necessary. Frequency-shift systems are much less tolerant of small frequency changes than are on-off systems, but these more stringent requirements can be met by the use of well-known techniques. What might be considered a disadvantage of the frequency-shift system, therefore, is in fact in its favor because it forces the attainment of better stability and further emphasizes the advantages in respect to reduced spreading of the band by keying, as outlined above.

General Comments

The above information is in the nature of a critical review of technical opinion, as expressed in available data, together with comments and explanation. While there is some divergence of opinion about the advantages offered by frequency-shift keying, and difficulty of expression in quantitative terms and in an unqualified manner, there seems little doubt that there are many advantages. There are competing alternative systems, such as the use of a number of telegraph channels, possibly combined with telephony channels, in a single-sideband system, but the frequency-shift system appears attractive for the requirements of many communication circuits, especially where the period of reliable working per 24 hours is to be as long as possible. There also seems no reason why time-division multiplex systems cannot be applied to obtain multichannel facilities (Bray, Lillicrap, and Owen, 1947, and Petersen, and others, 1946). There is considerable evidence to indicate that many radio-telegraph circuits in use overseas are being changed progressively to frequency-shift working and this process is likely to continue, assisted by development work which is also continuing in an effort to improve further existing apparatus and technique. Present apparatus for frequency-shift systems is considerably different from that used in on-off systems, particularly at the receiving end, but is not much more complicated or expensive, and may be simplified as the result of further development.

IV. Description of a Frequency-Shift Receiver

Initial Requirements

The equipment to be described is intended for use in radio-telegraph teleprinter services over varying distances employing frequencies in the range 1.5 to 30 mc. Teleprinters employ a five-unit start-stop code in which all signal elements occupy nearly equal time intervals, approximately 22 milliseconds, resulting in a relatively low signaling speed of about 50 bauds corresponding to 60 wpm. For this purpose the "audio filter" method of separating the "mark" and "space" signals is suitable, resulting in the necessity for choice of a nominal value of frequency shift, given as 850 cycles in this instance. Other requirements giving a basis for design were:

(a) Equipment to be self-contained and rack-mounted, preferably in a single rack not more than 7 feet 6 inches high.
(b) Operation from single-phase power mains of 200 to 260 volts, 40 to 60 cycles.
(c) Double-diversity receivers required with provision for connection to balanced aerials of 200 ohms impedance, or 75 ohms unbalanced.
(d) Equipment to operate with a minimum of attention once set up for a particular circuit. Arrangements for automatic-frequency control desirable.
(e) Receivers to provide for three pre-set frequencies in the range 1.5 to 30 mc, with crystal control of the frequency-changing oscillator; or alternatively, selection of a variable oscillator for use in rapidly setting up a new circuit for which a suitable crystal may not be available.
(f) Test equipment to be provided for checking frequency drift and for carrying out tests on the audio-frequency portions of the equipment.

General Arrangement

As indicated in the previous section, it was desired to mount the equipment in a rack, using standard 19-inch panel-mounting units. This involves division of the equipment into a number of units each of which can be mounted at convenient points in the rack structure. The arrangement adopted is shown in block-schematic form in Fig. 1 and the disposition of the units in the cabinet-type rack in Fig. 2. The units indicated in these figures are separately described in later sections; this section deals only with their general arrangement and functions.

Referring to Fig. 1, the aerial coupling unit, two receivers and the main oscillator are seen on the left side of the figure. The aerial coupling unit provides terminations for the aerial feeders (balanced or coaxial) and coaxial connectors for output to the receivers. Where balanced aerials are used, switches enable the selection of coupling transformers, providing the 200- to 75-ohm matching and balance-to-unbalance connections required. Two coupling transformers for each receiver are fitted, covering the frequency ranges 1.5 to 6.7 mc and 6.7 to 30 mc.

The receivers are arranged for an unbalanced 75-ohm input circuit and when the aerial feeders are 75-ohm coaxial cables, the aerial coupling-unit switching arranges a direct connection to the receiver input circuits. As is common practice for diversity receivers, the frequency-changing oscillator is common to the two receivers and the unit containing this oscillator (termed the main oscillator unit), also includes the beat-frequency oscillator and its afc circuits, again common to the two receivers. Apart from this arrangement of oscillators, the receivers are, for the most part, of conventional superheterodyne type.

Output from the receivers at audio frequency (2,125 and 2,975 cycles for "mark" and "space," respectively) is taken to separate limiter-detector units for each channel. These units each contain an input filter, limiting amplifier, "mark" and "space" separating filters followed by a further amplifier stage, rectifier and relay valves, and a differential noise-suppressor circuit.

The output from these units, comprising dc pulses corresponding to "mark" and "space" signals, connects with the relay unit where the two channels combine at the coil of the keying relay. The keying relay is arranged to send dc to the output line derived from the rectifiers and a power transformer included in the relay unit.

The unit termed "frequency deviation indicator and vf keying unit" provides for indication on the center-zero meter of the deviation of "mark" and "space" frequencies from their nominal values of 2,125 and 2,975 cycles and a separate section for supplying test "mark" and "space" signals at these frequencies. It is employed for initial setting up of the equipment on a particular circuit and for observation during operation of any drift in center-frequency adjustments.

The other units shown in Fig. 1, comprising meter panel, jack strip, power switching panel, and terminal panel provide for test measurements of voltages, currents and audio-frequency levels, patching of unit inputs and outputs, control of mains supply circuits, and connections to external circuits, respectively.

Each unit which requires power-supply arrangements contains its own power transformer, rectifier valve, and filter circuits, and the mains input is controlled by separate switches on the power switching panel.

Fig. 2 shows the arrangement of the various units in the rack assembly, and also indicates the vertical space occupied by each. All major units are of recessed-chassis construction in which valves and large components project horizontally from the rear of the chassis, the open side of which faces the front, contains all minor components, and is covered by a readily removable panel. The rack is a metal cabinet, approximately 14 inches deep, with a hinged door over the full height at the rear. This method of construction of the individual units and assembly into the rack results in an equipment of good appearance and accessibility. The construction and components are such as to ensure reliable operation under extreme climatic conditions.

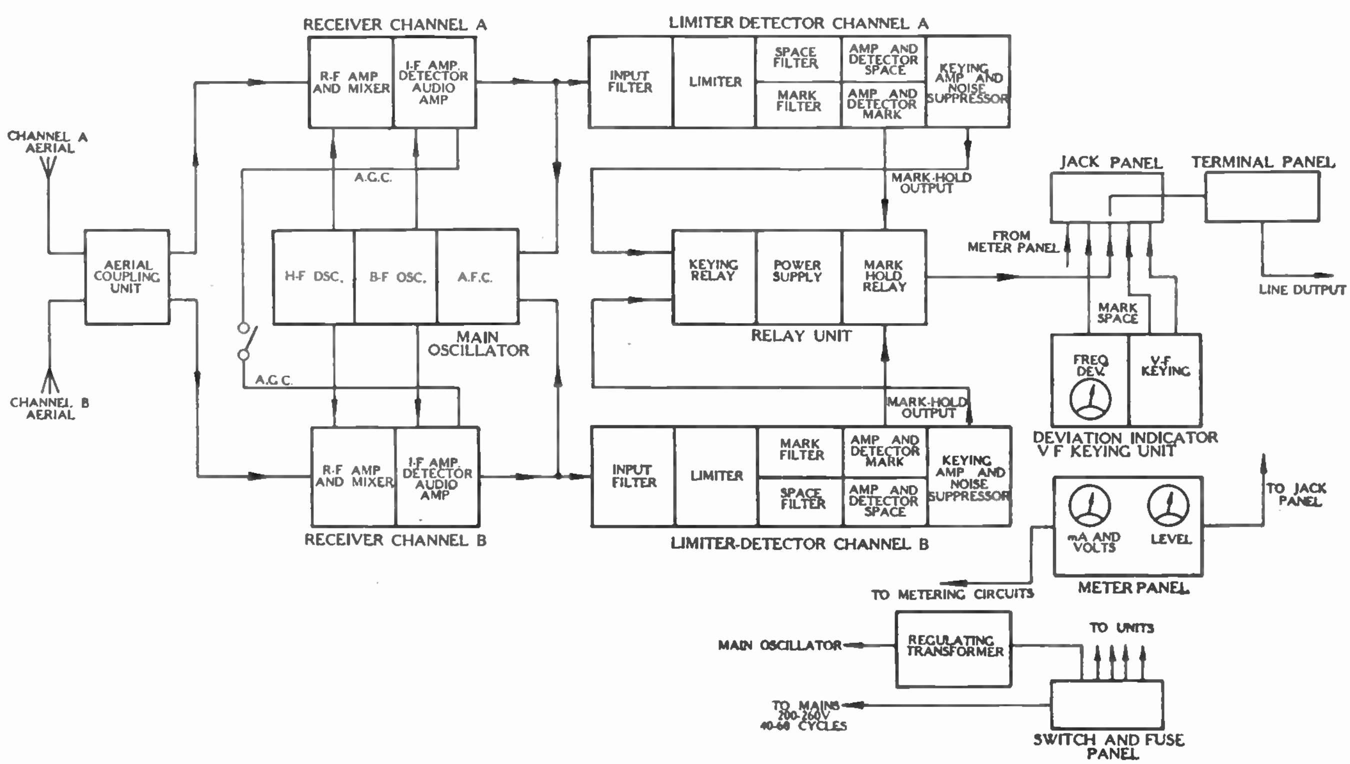


Fig. 1—Block-schematic diagram of frequency-shift receiving equipment.

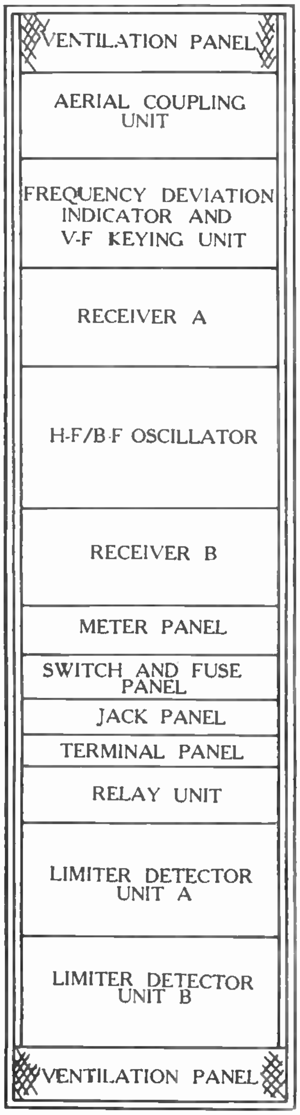


Fig. 2—Dual-diversity receiving equipment rack layout.

Receivers

Fig. 3 is a block-schematic diagram of the stage arrangement employed in each of the receiver units. Miniature valves are used throughout, except for the rectifier providing the high-voltage supply, because of their convenience and improved characteristics. Two amplifier stages at signal frequency are employed to achieve good signal-to-noise ratio and image response, and the full frequency coverage of 1.5 to 30 mc is divided into three ranges, normally 1.5 to 5 mc, 5 to 15 mc, and 15 to 30 mc, the desired range being selected by a switch on the front panel. There are no tuning controls of the conventional type, each coil circuit covers the range indicated with adjustments accessible from the rear of the unit. It is intended that the receivers will be present to the desired operating frequencies anywhere in the range indicated and the particular frequency required selected by the switch referred to above. If more than one of the desired operating frequencies falls in one of the ranges indicated, coil assemblies can be readily varied accordingly.

Two rf amplifier stages are followed by a mixer, in which the heterodyne frequency is taken from the main oscillator unit, giving an intermediate frequency of 455 kc, which is amplified in two stages employing a total of five double-tuned transformers. A switch is provided on the front panel for control of the selectivity given by the IF amplifier and it has three positions termed Wide, Normal, and Crystal. These positions give selectivity characteristics for 1 drop in response of approximately 3 db of ± 3.5 kc, ± 2 kc, and ± 0.9 kc, respectively.

A separate amplifier circuit and rectifier is used to derive agc voltages, which are applied to all preceding stages. This is done for two reasons; first, because this

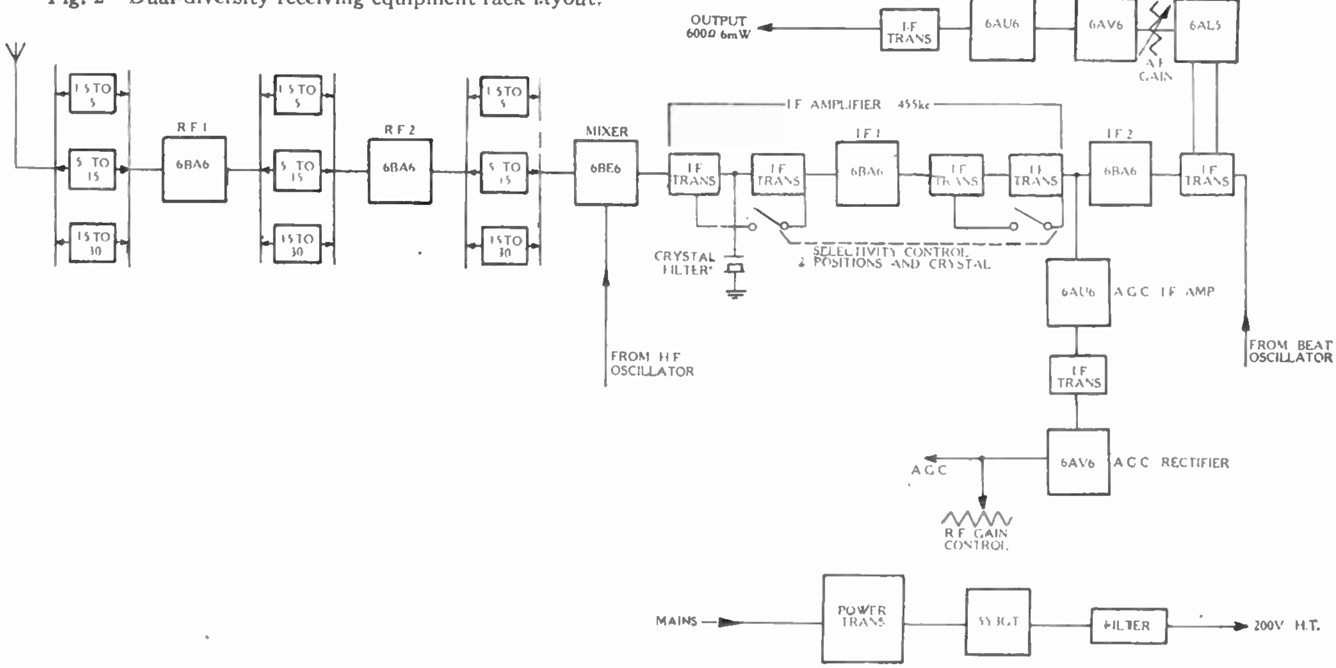


Fig. 3—Block-schematic diagram of the receiver.

method enables very good agc characteristics to be obtained, and second, because it ensures freedom from undesired effects on the agc caused by stray coupling from the beat oscillator. An rf gain control is provided, operating on the agc circuit in such a manner that the shape of the agc curve is not appreciably altered when the rf control is adjusted. An over-all reduction in sensitivity results only in a shift of the whole agc curve towards increasing input voltage for the beginning of action. Provision is also made for common connection of the agc circuits of the two receivers employed in the complete equipment. This is arranged through a switch enabling the common agc to be disconnected if desired.

A balanced detector circuit employing a 6AL5 double diode is used, enabling connection of the beat-frequency oscillator in such a manner that common coupling between the two receivers is minimized.

The two audio-frequency amplifier stages, employing 6AV6 and 6AU6 (triode-connected) valves, are quite conventional in arrangement and the output transformer is arranged to suit a 600-ohm load circuit. Provision is made for the connection of meters into the mixer circuit and both the agc and signal diode circuits, for assistance in initial alignment to a particular frequency.

High-Frequency and Beat-Frequency Oscillator

Fig. 4 is a block-schematic diagram of this unit and it will be noted that there are three portions shown as independent of one another. The portion designated "power supply section" provides high-voltage and filament supply for the other two portions, termed "hf oscillator section" and "beat oscillator section," respectively. These latter two are actually independent sections electrically and are combined mechanically for convenience, as each operates in conjunction with the two receivers. The whole unit is referred to as the "main oscillator" in Fig. 1 and in the previous text.

The hf oscillator provides for selection of any one of three crystals, or of a variable tuned circuit covering the whole range in three steps. It is intended that the variable oscillator should be used only for initial alignment or in cases where a new circuit has to be set up quickly and no suitable crystal is available. Crystals used have individual temperature-controlled ovens, a complete crystal unit being approximately the size of a "G-T" valve. These crystals, the variable oscillator circuit, and the oscillator valves are contained in a further temperature-controlled oven, resulting in very close temperature control of the crystals (better than $\pm 1°C$), and control to about $\pm 3°C$ of the variable oscillator circuit. Provision is also made for slight adjustments to the crystal frequencies (a range of about 0.01 per cent), by means of variable capacitors across each crystal, the adjustments being available on the front panel of the unit. The variable oscillator also has the main and vernier tuning controls on the front panel. Other circuits of the hf oscillator, outside the oven, comprise an untuned buffer stage and a tuned output stage, both employing 6AU6 valves.

The beat oscillator section provides for three conditions of operation: manual, in which the beat frequency can be manually adjusted; afc, in which it is controlled by afc circuits; and crystal, where it is maintained at 452–450 kc by a quartz crystal. The manual position is intended for use mainly in setting up a circuit, although it can be used to compensate for the inability of a crystal in the hf section to be adjusted precisely to the required frequency, as long as the variations are small enough to enable passing of the intermediate frequency thus created through the selective circuits of the IF amplifier. If the transmitter with which the receiver is working has adequate frequency stability, so that the variations in frequency caused by both transmitter and receiver over a working period are likely to be less than

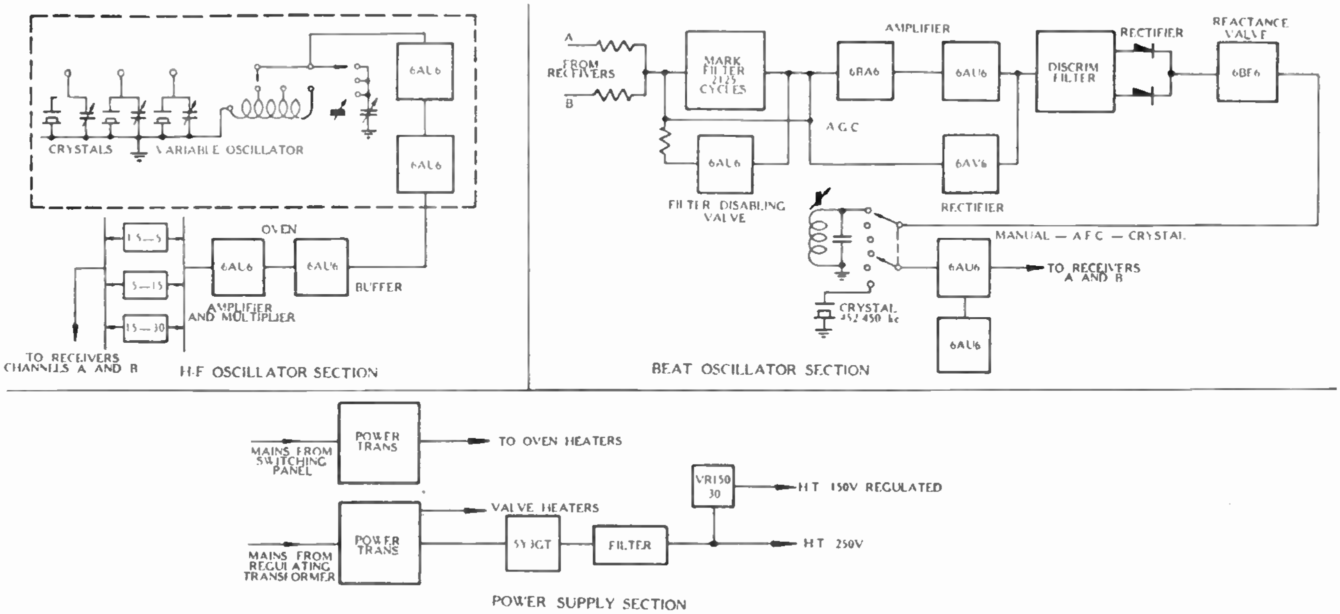


Fig. 4—Block-schematic diagram of the high-frequency oscillator and the beat-frequency oscillator.

the variation allowed by the audio-frequency filters (that is, within about ±300 cycles), then the crystal position may be used. It is also useful during the initial setting up of a circuit. If the frequency variation caused by both transmitter and receiver is likely to be much greater than ±300 cycles and the rate-of-change of frequency is fairly small, then the afc position may be used. The afc circuits will compensate for drift up to about ±2,000 cycles if the circuit is set up correctly, so that an initial manual adjustment, referred to the crystal and frequency deviation indicator, gives the correct mark frequency. The afc circuit employs a mark filter for its input network (see next section and Figs. 6 and 7), deriving its input signal from both receivers through bridging pads. The circuit time constants are slow so that the correcting output voltage is maintained during keying of the transmitter, and the mark filter prevents any response from the space signaling condition. During periods in which no message is actually being transmitted, it is usual for the transmitter to radiate a continuous mark signal and, in these circumstances, the afc circuit continues to operate and correct for frequency drift. If, however, during a working period the transmitter ceases to radiate for some time and when it does recommence the frequency drift which has occurred is greater than 300 cycles, the mark filter would prevent the afc circuit from operating if it were not for the special by-pass circuit provided. The circuit uses a 6AU6 valve which is termed a "filter disabling valve" and is so arranged that a correcting signal is derived despite the wide variation in frequency, but once the frequency is again centered in the range of the filter, the valve is prevented from amplifying appreciably by a bias derived from a slow-acting, agc circuit. A range of ±2,000 cycles (approximately) is covered by this means. False operation of the afc circuit can be caused by interfering signals, or by inadvertent transmision of a space signal for a long period; in the first case, a complete failure of communication may occur; in the second, a recovery period comprising a few seconds of mark signal is required. Where the transmitter is of adequate stability and a crystal of proper frequency is available for the hf oscillator of this equipment, there would be no necessity for the use of the afc position. It can be very useful, however, if the transmitter is not sufficiently stable, or immediate operation on a new frequency is required, calling for the use of the variable hf oscillator.

The power supply section provides high-voltage and low-voltage power for operation of the two sections described above. A VR150/30 gas regulator valve is used to stabilize high potentials to the oscillators and the mains input is stabilized by a regulating transformer (mounted in the rack) to obtain the best possible stability from the variable hf oscillator which is suceptible to filament voltage changes. The heater supply to the temperature-controlled crystals and the main oven is provided from a separate transformer having an individual mains switch, so that the ovens may be left running continuously and other units switched off as desired. Pilot lamps are provided in the oven circuits to indicate operation of the thermostats.

Provision is made for metering circuits to provide for measurement of important voltage and current values in the unit, using the meter panel mounted on the rack.

Limiter-Detector Unit

The general arrangement of the limiter-detector unit is shown in Fig. 5, applying to channel *A*. The unit for channel *B* is identical; the keyed dc outputs of the two channels combine at the relay unit and there is a common connection associated with the noise-suppressor circuit.

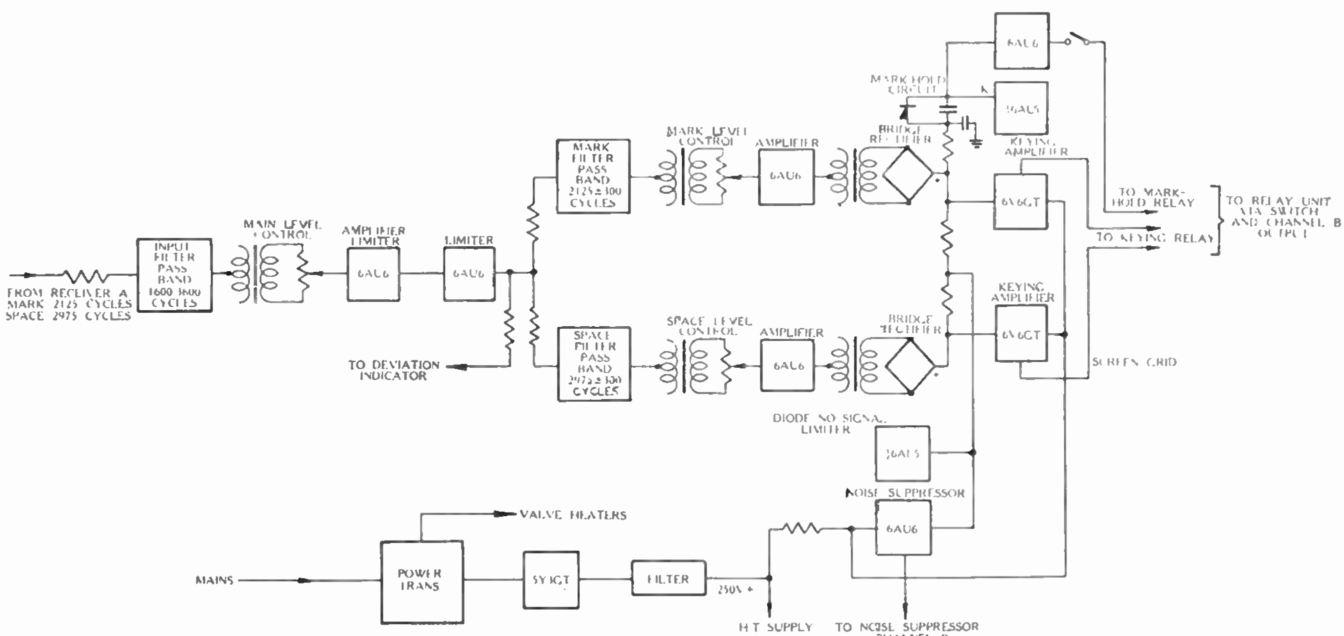


Fig. 5—Block-schematic diagram of the limiter-detector unit (channel A).

In the receiver units previously described there are not many features which are peculiar to the frequency-shift system; these receivers would be suitable for the on-off keying system if the initial requirements were similar. Conversely, receivers originally designed for on-off keying could be used with some modification, if their frequency stability were sufficiently good. The major differences in technique appear in the limiter-detector unit and its associated relay unit.

The limiter-detector unit deals with mark and space signals from the receiver output on nominal frequencies of 2,125 and 2,975 cycles, respectively. The difference between these frequencies is, of course, the deviation as transmitted, but the reason for the choice of 2,125 cycles for the mark frequency is of interest. In this unit the mark and space signals pass first through an input filter which is intended to allow for variations in the order of 300 cycles in both mark and space frequencies. As the nominal value of frequency shift is 850 cycles, this means that the input filter must have a pass-band of approximately: $2[(850/2)+300]$ cycles $=1{,}450$ cycles (or 725 cycles on each side of a nominal center frequency).

After passing through this filter, the mark and space signals are applied to a two-stage limiting amplifier to reduce level variations. Fig. 6 shows the input-output characteristics of the limiting amplifier. The output of the limiting amplifier is then applied to the mark and space filters which perform the separation of mark and space signals. These filters must each have a pass-band

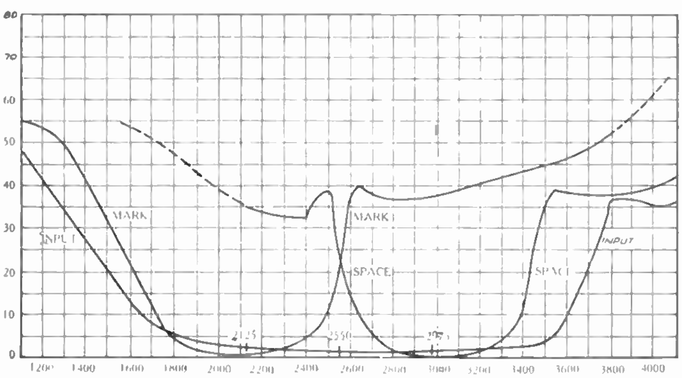


Fig. 6—Characteristics of filters.

approximately 600 cycles wide (± 300 cycles), but at the mean frequency (425 cycles above mark and 425 cycles below space), there must be considerable attenuation so that the separating process can be carried out properly. This mean frequency is chosen to be the third harmonic of 850 cycles, that is, 2,550 cycles, introducing considerable attenuation at this frequency. This fixes the center frequencies for mark and space at 2,125 and 2,975 cycles, respectively, and also indicates a suitable pass band for the input filter, 1,825 to 3,275 cycles. It will be recalled that multipath propagation can give rise to beats between the mark and space frequencies, that is 850 cycles, with production of harmonics of this frequency also likely. If the input filter is designed to cut off fairly rapidly below 1,825 cycles, it will obviously attenuate 850 cycles considerably and 1,700 cycles (second harmonic) to some extent; these facts, together with the noise reduction also attained, are the reasons for its use. The third harmonic falls at 2,500 cycles and is attenuated by mark and space filters; the fourth harmonic, 3,400 cycles, is appreciably attenuated by the space filter and the fifth is substantially eliminated. These are some of the considerations for the choice of audio-frequencies; others are, that any much higher frequency would introduce difficulties in aural checking and also require better percentage stability in the filters. These particular frequencies were chosen in some of the earlier equipment using an 850-cycle frequency-shift value and it is evident that a good choice was made.

The above nominal values for mark and space frequencies have been followed and result in practical filter characteristics as shown in Fig. 7. The circuit arrangement of each of the filters is shown in Fig. 8. It will be

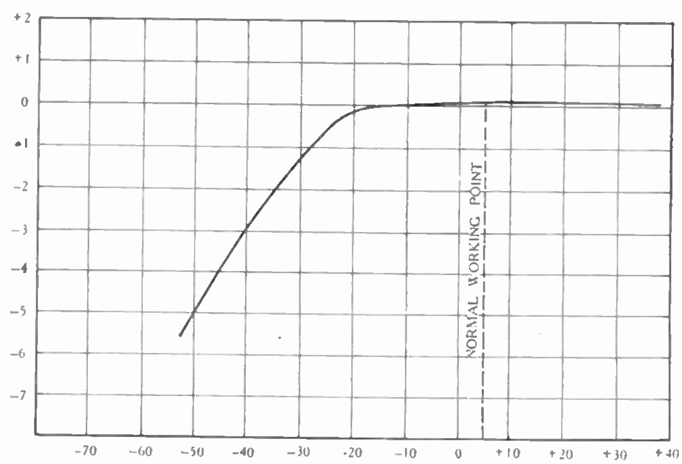


Fig. 7—Characteristics of limiter.

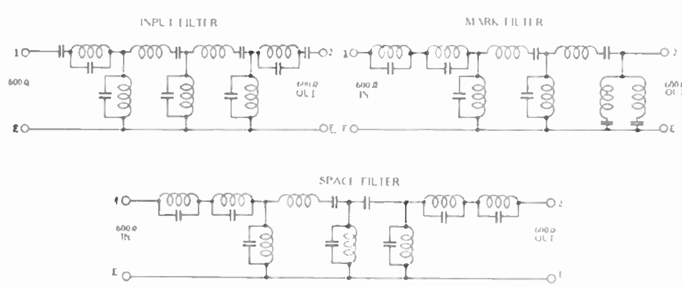


Fig. 8—Arrangement of filters.

noted that the slope of the attenuation curve outside the nominal pass-band is not particularly steep, and also that a relatively large number of sections is employed in each filter because the coils used are of relatively low-Q value (25 to 30). This has been found advantageous as there is only a slight tendency to "ringing" in the filters at keying speeds double the normal working speed; correct resistance termination of the filters also assists in this respect. In addition, the filters have been designed to enable compensation for appreciable variations in the value of individual components to facilitate production in fairly large quantities. The price paid for these desirable factors is that the filters are somewhat large physically. Each type is contained in a metal tank approximately $5\frac{1}{2}$ by $5\frac{1}{2}$ inches (base), and 9 inches in height.

The mark and space filters are each followed by an amplifier stage, preceded by a level-adjusting control to enable adjustment of mark and space levels to equality. The amplifier stage for each circuit is succeeded by a transformer and a bridge-connected selenium rectifier, with the output polarity positive with respect to earth.

The outputs of the bridge rectifiers connect to the signal grids of two 6V6GT valves which act as a keying amplifier. The circuit is essentially that of a symmetrical dc amplifier with the valve cathodes connected together and to earth (ht negative) through a fairly large resistor. Application of the dc signal to one causes the other to cut off and drives an output current in one direction through half the relay coil; application of signal to the second results in a reverse action.

Considering the two channels, it will be obvious that, with normal signals in both, the keying amplifier of each channel will make approximately equal contribution to the output current operating the relay for either mark or space. If the signal fades in one channel it will be replaced by an approximately equal noise voltage, caused by the action of the limiting amplifier and receiver agc circuits. Note, however, that this noise voltage appears in *both* mark and space output circuits of the channel concerned. This results in an increase in voltage at the common connection of two resistors, the other ends of which connect to the signal grids of the keying amplifier valves (see Fig. 5). This change of potential, caused by fading of the desired signal and increase in noise, is employed to produce a suppressing action on the contribution of the affected channel to the output relay current. The potential change (positive for increasing noise) is applied to the signal grid of 6AU6 valve (noise suppressor) and results in a reduction of the screen-grid potential of the keying amplifier. In addition, a differential action is produced by connecting the cathodes of the noise-suppressor valves of channel A and channel B together and through a high resistance to ht negative. Thus, an increase in noise on one channel decreases its contribution to the output current and increases that of the other channel. When both channels are equally noisy, the output current decreases slightly but remains equal for both channels. Should one channel develop a fault prior to the keying amplifier, resulting in the elimination of both signal and noise, a diode circuit ("diode no-signal limiter") using half a miniature double-diode (6AL5) is employed to prevent the differential action of the noise suppressor from blocking the other channel also. The unaffected channel can then do its best to maintain the communication circuit. The complete noise-suppressor system is designed in such a manner that very rapid action is possible; a mark signal might be accepted from channel A and the succeeding signal element from channel B with A making no contribution. The operating conditions of the keying amplifier valves are chosen to make the cut off, caused by reduction in screen potential, quite sharp and if one channel output is practically all noise the differential action results in a normal value of output current from the other channel (that is, its ordinary contribution is approximately doubled), and little or no output from the noisy channel. Assisted by the common agc circuits of the receivers this arrangement provides good diversity action and results in a marked improvement in communication reliability as compared with a single channel.

A further facility termed a "mark-hold circuit" is provided in each channel (see Fig. 5, near top of diagram). As indicated in an earlier section, it is the normal convention in frequency-shift systems for the transmitter to send a continuous mark signal when the circuit is "idle"; the first "space" is actually a "start" signal for the motor of the teletypewriter. However, when this convention is not followed, if the transmitter fails or is deliberately shut-down for a short period, it is possible that noise and interfering signals might cause false operation of the teletypewriter. To prevent this without requiring attention by either the receiver operator or the operator at the teletypewriter, a circuit is provided which allows normal operation but operates to signal a mark on the output line if normal keying, or the transmission of a continuous mark, ceases for a period in excess of about 0.25 second. The arrangement adopted involves the use of an integrating circuit associated with two rectifier diodes (one selenium rectifier and half a double vacuum-diode 6AL5—see above), and a 6AU6 valve, the anode current of which operates a relay. When the mark-hold relay drops out because of the failure of the signal, the relay contacts operate to send mark on the output loop, prevent the signal relay from chattering because of noise impulses, and operate a warning pilot lamp. The relay and pilot lamp are actually mounted in the relay unit and the anode currents of both mark-hold valves pass through the relay coil. A switch on each limiter-detector unit enables the mark-hold circuit to be made inoperative if desired; with very poor signals and high noise level, false operation is possible and extra attention from the receiver technician will be required to keep the circuit operating.

Provision is made for metering all important circuit voltages and currents using the meter panel mounted on the rack.

Relay Unit

This unit, shown in Fig. 9, contains the keying relay, the mark-hold relay, power supply circuits to provide the desired line current, and a metering switch and meter providing for measurements of all important current values. Two further switches provide for the selection and combining of A and B outputs and the selection of the desired arrangement for keying to line.

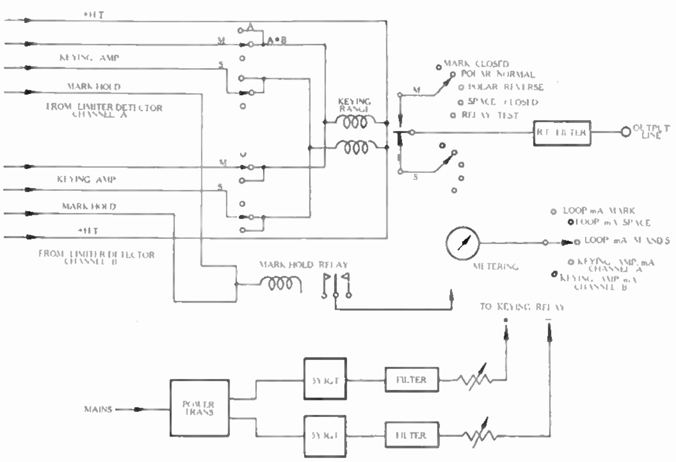


Fig. 9—Block schematic, relay unit.

The output from A and B channel, limiter-detector units combines at the keying relay and it is desirable to have a means of using each channel separately for checking purposes; also, a defective channel may need to be isolated, since certain possible faults in one channel can react on the other. These facilities are provided by the switch shown on Fig. 9 to the left of the keying relay.

The keying relay is a conventional "side-stabel" high-speed type used with the coils connected series-aiding and the keying amplifier high-voltage supplies connected to the center-point of the winding. Normal signalling current is approximately 15 ma from the keying amplifier of each channel. The contacts provide for keying of the dc power supply to line via spark suppression filters and a switch which enables polar keying or on-off keying to be selected. In the on-off keying positions the local dc supply is not used, the relay contacts key the line and the dc supply would be provided at the remote end. In addition, the switch gives reversal positions (that is, reversed polarity for mark and space in the polar keying condition and current to line for space signals instead of mark in the on-off keying position). It is easily possible with the frequency-shift system to obtain reversed-keying conditions and although the point where the reversal occurs must be found and the final correction made there, it is convenient to be able to obtain a temporary correction in this manner. The normal current sent to line is approximately 60 ma for a 500-ohm loop circuit. This switch has one further position which places a potential of about 6 volts 50 cycles on the relay coil and terminates the line output in a resistor, removing all normal circuit connections. This enables checking and adjustment of the relay. Because of the importance of the current values in this unit a metering switch and meter are included. These provide facilities for measuring the keying-amplifier currents of each channel, or the combined value (in conjunction with the channel-selecting switch) and the loop currents keyed to the output line.

The mark-hold relay is a conventional telephone relay having a 10,000-ohm coil and contacts performing the functions described in the above section.

Frequency-Deviation Indicator and VF Keying Unit

The arrangement of this unit is shown in Fig. 10. It provides for measurement of frequency deviations in one section and, in the other, for outputs at the mark and space frequencies to be used in checking and setting up the audio-frequency portions of the equipment.

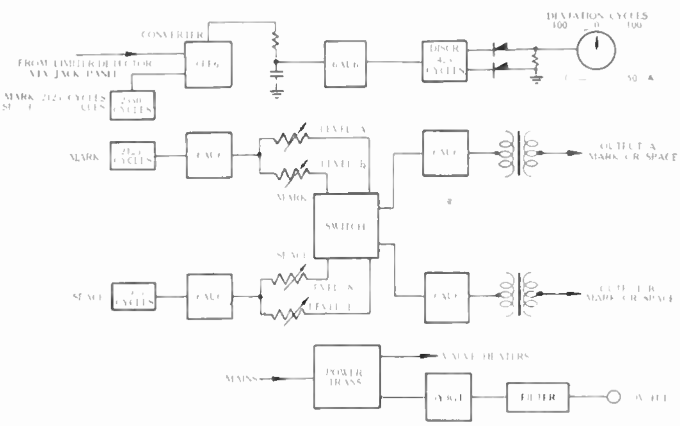


Fig. 10—Block schematic, frequency deviation indicator and voice frequency keying unit.

The deviation indicator derives its input signal from the output of the limiting amplifier in the limiter-detector unit, via the jack field, where the input circuit is normalled to channel A but can be patched to channel B, if desired. The first stage is a frequency converter employing a 6BE6 valve with the oscillator section arranged to operate at 2,550 cycles. If both mark and space signals are at their nominal frequencies, 2,125 and 2,975 cycles, respectively, the output of the converter stage derived from either will be at 425 cycles. A low-pass filter is employed to reduce spurious components at other frequencies, and the output from this filter is applied to an amplifier stage and then to a discriminator (Seeley-Foster type) with a dc center-zero meter in its rectifier circuit. The discriminator is arranged to have a reasonably linear characteristic over a range in excess of 100 cycles above and below a center-frequency of 425 cycles and the meter is calibrated over a 100-cycle range on each side of zero.

The deviation indicator is useful both in setting up a circuit and in checking operation during actual use. If the receiver is correctly tuned and the deviation of the transmitter is also correct, both mark and space indications will be zero. If an indication of 100 cycles, in opposite directions, is given for mark and space signals, this shows that the receiver is not correctly tuned and adjustment may then be made. If the mark signal reading is zero and the space reading 50 cycles, the indication is that the transmitter deviation is incorrect by that amount.

The vf keying unit comprises two oscillators, at 2,125 and 2,975 cycles, controls for level adjustment, a key-switch for selecting mark or space frequencies, and two amplifier stages, each delivering output to appropriate points in the jack field. Accurate adjustments of level may be made by using the level indicator in the meter panel, and fixed attenuator pads also provided in the jack field permit a wide range of levels to be applied to various points in the equipment. Tests may be made for correct operation of the limiter-detector units, including the action of noise-suppressor and mark-hold circuits, relay unit, and frequency-deviation indicator.

V. Performance Summary

Receiver

The following is a summary of the typical performance characteristics of the receivers, operating with the main oscillator unit.

Sensitivity and Signal-to-Noise Ratio

The figures were taken with input from signal generator modulated 30 per cent at 400 cycles through a 75-ohm resistor. Noise figures relate to the drop in output on removing modulation.

TABLE I

Frequency mc	*Output Microvolts*	*Output Milliwatts*	*Noise db*
3.15	1.0	10	12
5.05	0.5	10	9
7.2	0.5	10	7.5
14.3	1.0	10	10
27.0	1.0	10	9

Image ratio

mc	*Image ratio db*
3.15	100
5.05	73
7.2	76
14.3	60
27.0	38

Selectivity (measured on IF amplifier at 455-kc center frequency)

Input ratio		kc off resonance					
		Wide		Normal		Crystal	
Times	db	+	−	+	−	+	−
1.4	3	3.0	3.0	1.5	1.5	0.9	0.9
3	10	3.7	3.7	2	2	1.25	1.25
1,000	60	11	11	7.5	7.5	4.5	4.5

Automatic Gain Control

At any frequency in the range the agc characteristic is within 4 db for input signals between 10 microvolts and 0.1 volt.

TABLE II

Automatic Frequency Control

Control range (approximate)		Resultant error in mark frequency
Within filter pass-band	Maximum	
±300 cycles	±2,000 cycles	Less than 60 cycles

Tuning Ranges

Radio-frequency, 1.5 to 30 mc in three ranges
- (a) 1.5 to 5 mc
- (b) 5 to 15 mc
- (c) 15 to 30 mc.

Beat oscillator, manual tuning ±5 kc.

High-frequency oscillator
- (a) Variable, 1.5–30 mc, as above.
- (b) Crystal range of 0.01 per cent of frequency by vernier control for each of three crystal positions.

Frequency Stability

High-frequency oscillator crystals. For a variation in ambient temperature of ±20°C, the frequency change is not greater than 25 cycles for frequencies up to 15 mc and not greater than 50 cycles for frequencies above 15 mc.

Beat-frequency oscillator crystal. Within 0.01 per cent of 452.450 kc at any temperature in the range 10–60°C.

Variable-frequency hf oscillator. Better than 100 cycles per megacycle for temperature variations of ±5°C.

Limiter-Detector Unit

The following data are typical performance figures for the important characteristics of the limiter-detector unit:

Filter response. Details are given in Fig. 6.

Limiting amplifier. Details are given in Fig. 8.

Keying amplifier. Output to relay coil (per channel) 15 mA on mark or space signals.

Noise suppressor. Removal of the signal from one channel reduces the relay current from this channel to less than 4 ma, and increases that of the other channel by not less than 90 per cent in an action time not greater than 2,000 microseconds.

Mark-hold circuit. Operates normally to within 3 db of the keying relay failing to follow signals caused by noise.

Relay Unit

The rated characteristics of the relay unit are as follows:

Normal keying speed—50 bands (approximately 60 wpm on standard teletypewriter).

Maximum recommended keying speed—100 bauds.
Polar output—60-0-60 ma in 500 ohms.

Frequency Deviation Indicator and VF Keying Unit—
Performance figures are as follows:

Deviation indicator range—100-0-100 cycles, accuracy ±10 cycles.

Accuracy of 2,550-cycle oscillator—Within 10 cycles of nominal frequency.

VF keying unit outputs—Not less than ±10 dbm for each channel.

Accuracy of 2,125-cycle and 2,975-cycle oscillators—Within 10 cycles of nominal frequency.

Bibliography

1. R. M. Sprague, "Frequency-shift radiotelegraph and teletype systems," *Electronics*, vol. 17, pp. 126–131; November, 1944.
2. C. Buff, "Frequency-shift keying techniques," *Radio.*, vol. 30, pp. 14–17; August, 1946.
3. R. A. Vanderlippe, "Frequency-shift radio teletype in World War II," *Bell Lab. Rec.*, vol. 25, pp. 442–446; December, 1947.
4. H. D. Petersen, J. B. Atwood, H. E. Goldstine, G. E. Hansell, and F. E. Schock, "Observations and comparisons on radio telegraph signalling by frequency-shift and on-off keying," *RCA Rev.*, vol. 7, pp. 11–31; March, 1946.
5. R. Ruddlesden, E. Forster, and A. Jelonek, "Carrier-frequency-shift telegraphy," *Jour. IEE*, vol. 94, Part IIIA; 1947.
6. J. R. Davey and A. L. Matte, "Frequency-shift telegraphy-radio and wire applications," *Bell. Sys. Tech. Jour.*, vol. 27, pp. 265–304; April, 1948.
7. W. J. Bray, H. G. Lillicrap, and F. C. Owen, "The fading machine and its use for the investigation of the effects of frequency selective fading," *Jour. IEE*, vol. 94, Part IIIA; 1947.
8. W. Lyons, "Criteria for diversity receiver design," *RCA Rev.*, vol. 8, pp. 373–378; June, 1947.
9. T. A. Jones, and K. W. Pfleger, "Performance characteristics of various carrier telegraph methods," *Bell. Sys. Tech. Jour.*, vol. 25, pp. 483–531; July, 1946.
10. G. A. Wickizer, "Relative amplitude of side frequencies in on-off and frequency-shift telegraph keying," *RCA Rev.*, vol. 8, pp. 158–168; March, 1947.
11. L. E. Hatfield, "Frequency-shift radio transmission," Proc. I.R.E., vol. 36, pp. 116–121; January, 1948.
12. J. A. Smale, "Some developments in commercial point-to-point radio-telegraphy," *Jour. IEE*, vol. 94, Part IIIA; 1947.

Description and Evaluation of 100-Channel Distance-Measuring Equipment*

R. C. BORDEN†, C. C. TROUT†, MEMBER, IRE, AND E. C. WILLIAMS†

Summary—The common civil military program for navigational aids for use on and off the airways is based on a polar co-ordinate system from which a pilot may obtain his position relative to a known ground location. This navigational information is supplied in the form of azimuth and distance measurement from any preselected ground facility.

Under this program, azimuth information is supplied by the vhf omnirange which is now being installed on a country-wide basis. Development work on the final version of the other parameter of the navigation system, namely, the distance-measuring equipment, is nearly completed and plans are being made to add that equipment to the vhf omnirange facilities in the near future. Addition of the distance-measuring equipment to the instrument landing system (ILS) to provide continuous distance information to the runway touchdown point also is contemplated.

The distance-measuring equipment, commonly referred to as DME, employs pulse techniques and the interrogator-transpondor or challenge-reply principle so widely used in identification equipment during World War II. A portion of the frequency band, 960 to 1,215 mc, assigned to aids to air navigation, is being used for DME implementation. This report outlines the history and development of the present nationally adopted 100-channel DME, and describes the modifications necessary to convert earlier 50-channel equipment into equipment suitable for operating within a 100-channel system.

In operation, the aircraft unit of the DME, commonly called the interrogator, continuously transmits a train of short duration rf pulses. These pulses are received at the ground station, called the transpondor, and are shaped by suitable circuits into trigger pulses, which modulate the transmitter portion of the transpondor. The regenerated rf pulses transmitted by the transpondor, on a different frequency than the original received pulses, are in turn received by the airborne receiver section of the interrogator. Electrical measuring circuits within the interrogator measure the time elapsed between each pulse transmitted by the interrogator and the reception of its corresponding reply pulse from the transpondor. This time is directly related to the distance between the aircraft and the ground station and is converted into proper voltages for operating the pilot's distance indicator. The indicator is calibrated directly in miles. As many as 50 aircraft can use the same ground transpondor equipment simultaneously to obtain continuous distance information, without interference and with no reduction in performance.

The airborne interrogator must also be able to identify the ground transpondor, with which it is operating, as a check of the channel selecting mechanism. This is accomplished by the transmission of a third pulse, with a definite time relation to the distance measuring pulses from the transpondor. The interrogator recognizes the presence of this pulse and causes a 400-cps tone to be heard in the pilot's headset. With this arrangement several identification systems are possible, two of which are discussed.

I. Introduction

Many organizations have taken an active part in the development of distance-measuring equipment (DME). Among these are the Naval Research Laboratory, Canadian Research Council, Hazeltine Electronics Corporation, Federal Telecommunication Laboratories, U. S. Air Force, and the Civil Aeronautics Administration. The earliest operating DME was built by the Canadian Research Council in 1945, and was designed for operation in the frequency region of 200 mc. Early in 1946, an experimental DME designed for operation

* Decimal classification: R526.2. Original manuscript received by the Institute, February 19, 1951.

† CAA Technical Development and Evaluation Center, Indianapolis, Ind.

in the region of 1,000 mc was completed at the Naval Research Laboratory by the Combines Research Group. The primary purpose of constructing this model was to determine whether propagation in this frequency band would be suitable for DME use. With this question answered in the affirmative, the U. S. Air Force and the CAA awarded contracts to the Hazeltine Electronics Corporation and to the Federal Telecommunication Laboratories for the construction of development models of 1,000 mc distance-measuring equipment.

Upon delivery of these equipments, extensive laboratory and service tests were conducted by the Air Force and the CAA Technical Development and Evaluation Center.[1] The primary purpose of these tests was to evaluate the relative merits of two different types of DME channeling. At the time these equipments were ordered, the requirement existed that the DME system be capable of operation on any of 50 independent and noninterfering channels. It was intended to install a DME transpondor at each vhf omnirange[2] site to provide a complete azimuth-distance ground facility. There were to be 30 such omniranges with a 435-nautical mile square, each with a discrete frequency assignment, (these 30 frequencies would be repeated in each adjacent 500-mile square, noninterference being achieved by the geographical separation). In addition, DME transpondors were to be installed at each instrument landing site, there being 20 of these for each 435-nautical mile square, all with different frequency assignments. The means of obtaining the required 50 channels employed by Federal[3] and the means employed by Hazeltine[4] were different. The Federal equipment provided the 50 channels by assigning discrete radio frequencies for each channel, requiring a total of 100 radio frequencies with 2.5-mc separation. The Hazeltine approach to the problem consisted of selecting 26 radio frequencies, with adjacent spacing of 9.5 mc, within the assigned band and repeating each frequency four times. To prevent interference between channels on the same frequency, a system of double pulse transmissions, or pulse multiplex, was employed. In such a system, desired signals are selected and undesired signals rejected by proper design of video coding and decoding circuits.

The tests performed by the Air Forces and CAA were conducted under the auspices of RTCA subcommittee SC-40 (Standardization of DME Testing Procedures).[5] Both systems of channeling were found to be satisfactory, but the channeling problem became complicated further by a change in the operational requirement for the number of channels to be provided by the system. Re-examination of the over-all domestic short-range navigation requirements by RTCA subcommittee SC-22 led to the decision that a total of 100 channels must be provided rather than only 50.[6] Since separations of the order of 2.5 mc had been shown to be practical, and since the pulse-multiplex method of channeling had likewise proved its worth, the decision was made by SC-40 to combine the two techniques in order to obtain the 100 channels. Tests of the pulse-multiplex system indicated that the number of times an individual frequency could be employed might well be increased from four to ten; thus, 100 channels could be provided by only 20 radio frequencies (ten for interrogation and ten for reply). By employing an adjacent channel spacing of 2.5 mc, the entire DME system could be implemented within a total band of approximately 50 mc. Conservation of spectrum was held to be of extreme importance due to the requirements of RTCA subcommittee SC-31.[7] Among the requirements laid down by SC-31 was one prescribing that the 960 to 1,215-mc band must ultimately be used as a frequency medium for the entire short-range navigation system, including DME, omnirange, instrument landing, voice, and pictorial display.

A specification covering a system having the above channeling characteristics was prepared by SC-40. This specification was presented at the February, 1949, meetings of the International Civil Aviation Organization, and as a guide for DME development. Production equipment meeting the system requirements outlined by the SC-40 specification will be available in 1951. Detailed descriptions of the design and operation of 50-channel equipment of this type have been covered in literature,[1,3,4,8] and therefore, these details will not be repeated. This paper will be confined to those modifications required to expand the utility of the system to 100 operating channels.

II. Description of Equipment

Fig. 1 illustrates the basic operation of a DME system and indicates the proposed plan for installation of VOR transmitters and DME transpondors at the same site.

Figs. 2 and 3 are simple block diagrams of a basic DME interrogator and transpondor, respectively.

As a starting point, a summary of those equipment characteristics which differ in the two systems may be listed in Table I.

[1] R. C. Borden, C. C. Trout, and E. C. Williams, "UHF Distance-Measuring Equipment for Air Navigation," TD Report No. 114.
[2] G. C. Hurley, S. R. Anderson, and H. F. Keary, "The CAA VHF Omnirange," TD Report.
[3] H. Busignies, "High-stability radio distance-measuring equipment for aerial navigation," *Electrical Commun.*, vol. 25, pp. 237–243; September, 1948.
[4] C. J. Hirsch, "Pulse-multiplex system for distance-measuring equipment (DME)," Proc. I.R.E., vol. 37, pp. 1236–1242; November, 1949.
[5] RTCA Paper 121-48/DO-24, Report of SC-40 (DME System characteristics); December 15, 1948.
[6] RTCA Paper 76-48/DO-17, Report of SC-22 (Pairing of localizer, glide slope VHF omnirange, and DME frequencies); August 2, 1948.
[7] RTCA Paper 27-48/DO-12, Report of SC-31 (Air Traffic Control); May 12, 1948.
[8] J. Wesley Leas, "Some operational aspects of distance measuring equipment in the transition air navigation system," *Aero. Eng. Rev.*, vol. 9, pp. 31–36; January, 1950.

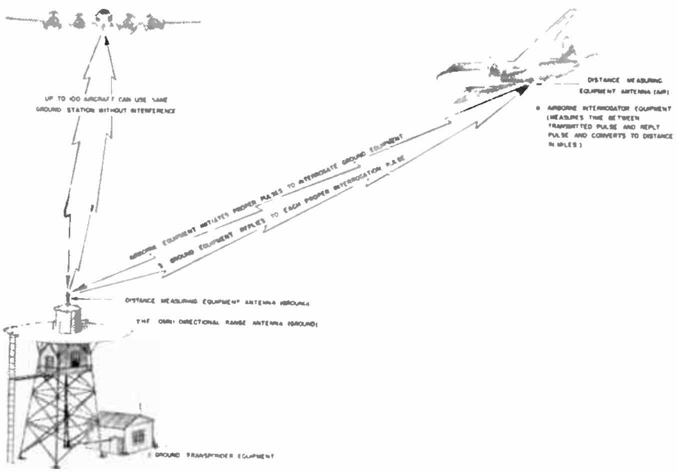


Fig. 1—DME system operation.

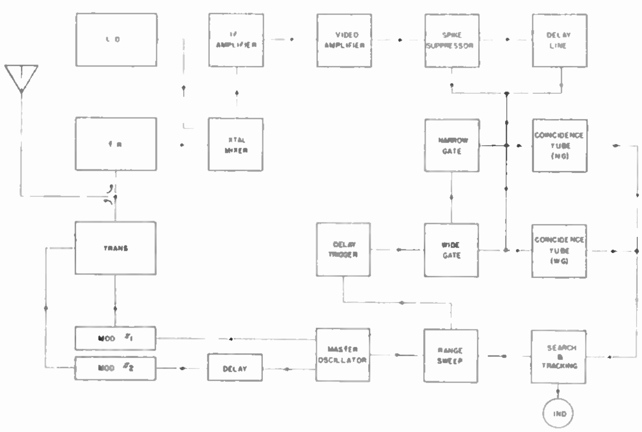


Fig. 2—Block diagram of the DME interrogator (airborne).

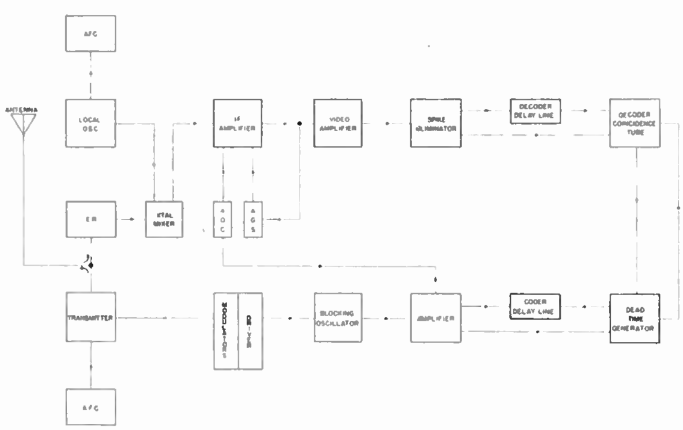


Fig. 3—Block diagram of the DME transpondor (ground).

III. Pulse Duration

Two methods of obtaining adjacent channel selectivity of the order of 70 db have been employed successfully in DME receivers. One of these, the spike-suppression circuit[1], requires the use of a wider pulse if the bandwidth is to be decreased. 2.5 μsec is sufficiently wide to permit satisfactory adjacent channel rejection with 2.5-mc channel separation. A typical response curve for a receiver employing spike suppression is shown as Fig. 4. Widening of the radio-frequency pulse from 1.5 to 2.5 μsec does not adversely affect the system in any other way, with the exception of the accompanying increase in duty cycle.

TABLE I

	50-channel	100-channel
1. Pulse duration	1.5 μsec	2.5 μsec
2. Pulse spacings	10, 15, 20, and 25* μsec or none†	14, 21, 28, 35, 42, 49 56, 63, 70, 77 μsec
3. Reply delay	75 μsec	115 μsec
4. Reply frequencies	One of 13* frequencies, or one of 51† frequencies in 1087.5-1,215-mc band	One of 10 frequencies in 1188.5- to 1211.0-mc band
5. Interrogation frequencies	One of 13* frequencies or one of 51† frequencies in 960- to 1087.5-mc band	One of 10 frequencies in 963.5- to 986.0-mc band
6. Identification	Gap coding	Third pulse
7. Traffic handling requirements	———	More stringent than 50-channel

* Hazeltine.
† Federal.

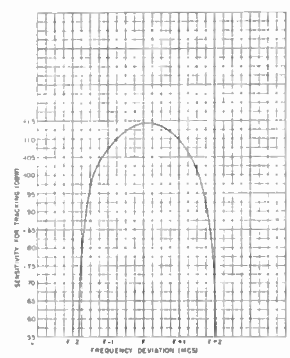


Fig. 4—Receiver bandwidth HEC interrogator (after spike suppression).

IV. Pulse-Spacings

The requirement for additional pulse-spacing in order to provide ten modes has necessarily resulted in a requirement for longer magnetostriction delay lines. These longer lines have presented no serious problems inasmuch as the attenuation with length of such a line is practically negligible. These ten pulse-spacings have been paired for interrogation and reply to form ten standard modes, as shown in Table II.

TABLE II

Mode	*Interrogation Spacing*	*Reply Spacing*
A	14	17
B	21	70
C	28	63
D	35	56
E	42	49
F	49	42
G	56	35
H	63	28
I	70	21
J	77	14

The performance of coders and decoders at the longer pulse spacings has been completely satisfactory.

V. Reply Delay

The increase in the reply delay from 75 to 115 μsec was required in order to accommodate the longer pulse-spacings employed in the 100-channel system. In the DME system, distance is measured by the interrogator as the time elapsed between the second interrogation pulse and the second reply pulse. Fig. 5(a) indicates the time relationship between these pulses when the aircraft is zero miles from a transpondor operating on Mode A. For a transpondor operating on Mode B, the delay line coil controlling the time of the first reply pulse, r is simply moved seven microseconds closer to r_2,

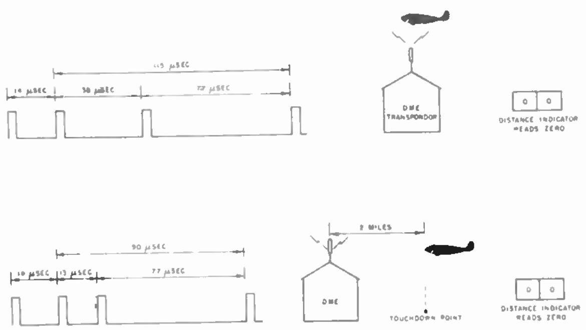


Fig. 5—Reply delay.

leaving the time interval between i_2 (the second interrogation pulse) and r_2 unchanged at 115 μsec. The additional 38 μsec are necessary to meet a requirement imposed when DME is used for instrument landing and the transpondor is not located at the touch-down point (where it is desirable for the aircraft distance indicator to display zero miles). By decreasing the time interval, i_2 to r_2, without changing the reply spacing, r_1 to r_2, this 38-μsec tolerance permits creation of an artificial zero up to approximately three miles from the transpondor. Fig. 5(b) shows the time relationship existent when the transpondor (still Mode A) is located approximately two nautical miles from the touch-down point. Note that the reply delay is adjusted for 90 μsec instead of 115 μsec. The 25-μsec difference is equivalent to a distance change of approximately two nautical miles.

VI. Interrogation and Reply Frequencies

In the case of the Federal DME, the frequency stability of the 50-channel units was sufficient to meet the 100-channel requirements; however, a considerable improvement in the reliability of the airborne transmitter AFC has been achieved in the 100-channel models. In order to meet the new channeling requirements, however, it was necessary to provide for cross-banding of the ten frequency pairs, whereas this was not required of the earlier 50-channel equipment. The total number of crystals required was nonetheless reduced and the pre-selector tracking problem was somewhat simplified due to the decreased receiver bandspread (25 mc as opposed to 127.5 mc).

In the case of the Hazeltine equipment, addition of an automatic-frequency-control system for the airborne equipment was required. This circuitry is similar to that employed for the ground transmitter of the 50-channel equipment. Airborne receiver stability was achieved through use of a single crystal, constant local oscillator frequency, and a variably tuned IF strip.

A contract has recently been awarded by the Civil Aeronautics Administration for a total of 450 100-channel DME transpondors. These transpondors will incorporate directly crystal-controlled transmitters. The output transmitter tube will be a type A-2328 triode operating as a power amplifier. Extensive tests performed at the manufacturer's plant have proven this tube to be a suitable power amplifier at the frequencies involved.

VII. Identification

Earlier DME systems employed gap-coding as a means of transpondor identification. Gap-coding provides an indentification signal in the aircraft by virtue of periodic interruption of the reply pulses (performed by a keyer unit located at the transpondor) in accordance with characters of the Morse code. Interruption of the reply in this fashion may be detected in the aircraft either visually or aurally. Under conditions of very low traffic density (approximately ten aircraft per transpondor), this system is adequate. As the traffic loading increases, two factors tend to break down this identification system. As the result of transpondor count-down, the code indications in the aircraft are likely to be interrupted, and as a result of high random pulse levels, the code indicating device in the aircraft is likely to register spurious signals.

For reliable operation in a pulse-multiplex system with ten transpondors operating on each of the ten reply frequencies, an improved means of identification was dictated. The system selected involves the periodic transmission of a three-pulse reply signal from the transpondor. This third pulse will always have a definite time relationship (10.5 μsec later) with respect to the second reply pulse. A special gate in the interrogator will recognize the absence or presence of this coding pulse. The coding circuit, when activated by the third pulse, will result in the insertion of a 400-cps tone (generated within the interrogator) into the pilot's headset. Through this same headset the pilot will also be receiving both a voice and a Morse code identification from the VOR to which he is tuned. By employing the same mechanical keyer to activate the third pulse reply at the transpondor as is used to key the VOR identification, a synchronous relationship between the VOR code and the 400-cps tone is established. There have been several proposed systems for utilizing a synchronized 400-cps tone generated within the airborne DME to indicate to the pilot that he is tuned to the ground transpondor associated with the vhf omnirange or instrument landing system facility being used.

In one system the mechanical keyer is arranged to initiate transmission of the third pulse immediately following the last character of the VOR Morse code. Simultaneously, a long dash will be transmitted (at 1,020 cps) on the VOR channel which will have a duration equal to the period of transmission of the third pulse (approximately 1.5 seconds). The result, as heard in the pilot's headset, will be a beat note between 400 and 1,020 cps. Should the pilot inadvertently tune to a DME transpondor not associated with the VOR whose identification he is hearing, he will not hear this beat note, but will hear separate and distinct 400- and 1,020-cps long dashes. This is assured by asynchronous operation of keyer motors at the different VOR sites. The proper time relationship between the 400- and 1,020-cps tones is indicated in Fig. 6 under System I, as well as a typical example of improper time relationship due to incorrect channel selection by either the VOR or DME airborne units.

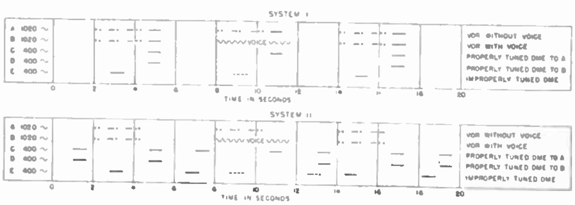


Fig. 6—VOR/DME identification system.

In another proposed system the transmission of the third pulse would be initiated for a period of about three-fourths second both before and after the transmission of the three Morse code characters from the VOR. The long dash would be eliminated from the VOR transmissions. This would result in the pilot hearing a 400-cps dash, followed by the VOR identification code at 1,020-cps and another 400-cps dash. In the event that improper channel selection should occur, the pilot will hear either one or the other of the 400-cps tones interfering with the VOR identification code letters. This system is illustrated in Fig. 6 under System II. It should be noted that this system does not cause any changes or additions to the VOR identification signals. Furthermore, such a signal, the long 1,020-cps dash, as heard in aircraft equipped with VOR only, may easily be confused with the letter "T."

On theoretical grounds, implementation of the 100-channel system, using only tested techniques (techniques employed in either or both of the 50-channel systems) presents no problems. Past experience, however, has indicated that until equipment meeting specific requirements has actually been built and operationally tested, it is sometimes difficult to predict performance with accuracy. For this reason, the placement of contracts for 100-channel equipment on a production basis was delayed until several developmental models became available. It was believed that satisfactory performance of such models under conditions which are to be encountered by future production equipment would be an excellent criterion of the soundness of 100-channel principles. Equipment manufacturers were urged to expedite delivery of 100-channel developmental models (both transpondors and interrogators), and simultaneously the Civil Aeronautics Administration initiated a project for combining two of the 50-channel interrogators (one pulse-multiplex and one high-stability type) in such a manner as to check various 100-channel characteristics. In addition to laboratory tests of 100-channel equipment, it was deemed essential to perform flight tests of the equipment under conditions of traffic density equivalent to that predicted for the foreseeable future.

Since the distance-measuring equipment, being an integral part of the Common Air Navigation and Traffic Control System defined by SC-31 of the RTCA, and now being developed under the direction of the Air Navigation Development Board, is a facility to be used by military as well as civil aircraft, a committee embracing both civil and military members was set up to direct the 100-channel DME evaluation program. This committee met on September 15, 1949, at the CAA Technical Development and Evaluation Center to discuss the various aspects of the test program. The test program was divided into four phases:

Phase I. Tests of the CAA combined pulse multiplex high stability 50-channel units.

Phase II. Laboratory tests of preproduction 100-channel units.

Phase III. Flight tests to determine the traffic handling capacity of 100-channel equipment.

Phase IV. Laboratory tests to determine the traffic handling capacity of 100-channel equipment.

Phases I and II consisted of rechecking equipment characteristics which are common to both 50- and 100-channel systems, and particularly the testing of combinations of 50-channel techniques required for implementation of the 100-channel system.[9]

Phases III and IV are considered of major significance inasmuch as they were set up for the purpose of providing more realistic justification that the 100-channel system, as designed, is capable of operating under the high traffic densities which ultimately are anticipated. Prior to the initiation of this test program, there was no suitable laboratory test equipment for determining traffic handling capacities of such system accurately, nor were there a sufficient number of ground stations geographically located so that an operational test could be arranged. Data of the type desired were previously limited to statistical analyses and to laboratory tests conducted with equipment not completely

[9] TD Report No. 119.
R. C. Borden, C. C. Trout, and E. C. Williams, "Evaluation of 100-channel DME."

capable of simulating actual operating conditions. Although the data from these two sources were in reasonable accord, it was believed highly desirable to perform tests under more realistic conditions. If 100 transpondors were used in a 435 nautical square mile area, then there would be a total of ten transpondors transmitting on a common frequency, i.e., each transpondor frequency must be repeated ten times. It is conceivable that each of these transpondors may, at a given time, be supplying service to as many as 50 aircraft. In such cases, all of the transmitted replies will be eligible for reception by all of the 500 aircraft involved, subject to line-of-sight propagation restrictions. Any given aircraft will receive only one of the ten transpondors with a proper reply spacing, but all ten will be receved at the correct frequency. Interlacing of the replies from the nine improperly moded transpondors will result in an appreciable number of artificial replies having the correct mode being produced. In this case, the ability of the aircraft decoding circuits to sort out the replies of proper spacing and reject all others is extremely important. Furthermore, the ability of the searching and tracking circuits of a particular interrogator to recognize and respond only to those properly moded replies initiated by its own transmitter also is imperative. Inserting the line-of-sight limitations and a more realistic picture of traffic distribution, SC-40 concluded that the ability of the system to operate properly in the face of a loading of 20 aircraft for each of the nine improperly moded transpondors, and 50 aircraft for the properly moded transpondor, would guarantee satisfactory system operation for many years to come. The purpose of Phases III and IV of this program was to determine whether or not this ability was inherent in equipment which can be built now. Only Phase III will be herein discussed, Phase IV activity still being in progress.

VIII. Phase III

Performance of Phase III of the test program necessarily was delayed until ten transpondors had been modified to meet 100-channel requirements. It was intended originally to employ this number of transpondors in the test, distributed as follows: Indianapolis, Ind., and vicinity, (4); Lafayette, Ind., (1); Terre Haute, Ind., (1); Dayton, Ohio, (1); and Wilmington, Ohio, (3). All of the transpondors were operated at the same reply frequency (1,201 mc), but with different reply spacings.

Due to transfer of the Air Force's All-Weather Flying Division activities from Wilmington, Ohio, to Wright-Patterson Air Force Base at Dayton, Ohio, it was not possible for that group to provide all of the transpondors assigned to them for the test; however, one of three was made available. In order to avoid dependence upon air-to-ground communication to eight different sites during progress of the test flights, special schedules were set up based on time synchronization. Flight tests were conducted on November 29, 1949, and repeated on the following day.

In order to assure that the test aircraft was within propagation range of all transpondor sites, a flight test area was selected which was within 100 nautical miles of all sites. This area is indicated on the map, Fig. 7, and was roughly 35 nautical miles east of the Center. In order to assure further that strong signals would be received from all transpondors, all flights were made at altitudes above 8,000 feet indicated (mean elevation above sea level of all transpondors was approximately 800 feet). In tabulating the flight test data, unless a synchronized reply was received from a given transpondor, the "fruit" from that transpondor was not considered.

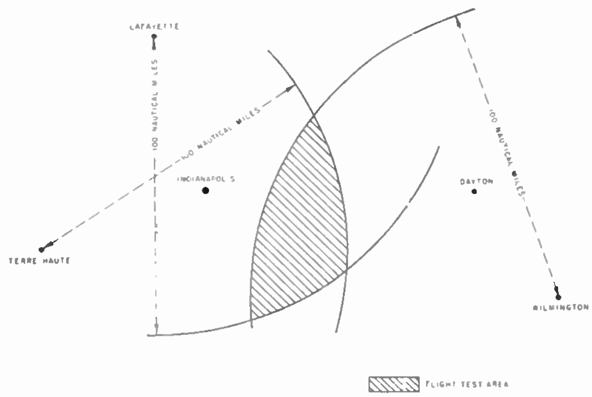


Fig. 7—Map of flight area.

Squittering of the transpondors was effected by turning up the automatic-gain-stabilization control while observing the transmitter current. A curve showing the relation of transmitter current and other transpondor characteristics to squitter rate is shown as Fig. 8. The squittering rates employed, ignoring the effect of count down, are roughly equivalent to the interrogation of each transpondor by 10, 20, 30, 40, and 50 aircraft, taken chronologically.

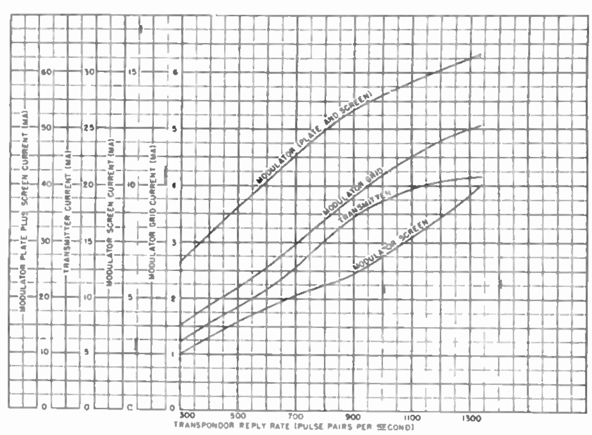


Fig. 8—Variation of transpondor parameters with reply repetition rate.

Two aircraft were employed during the first day's test, one from the Naval Air Test Center, Patuxent River, Md., and the other a CAA airplane N-181 based at the Technical Development and Evaluation Center. The test setup as installed on the Navy DC-4 airplane is shown in Fig. 9. Both test aircraft carried 100-channel airborne equipment. On returning to the ground, the flight engineers of both aircraft reported failure to receive replies from the Wilmington transpondor at all times during the flight. Replies were obtained from the remaining seven transpondors at all times.

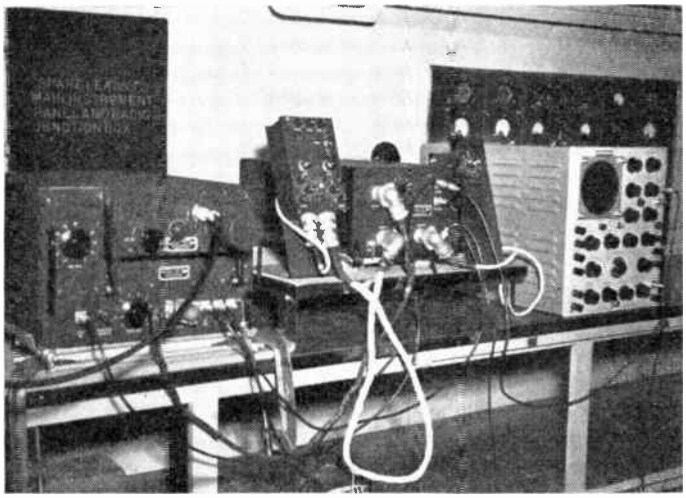


Fig. 9—Hazeltine 100-channel DME interrogator.

In the case of the Lafayette and Brownsburg, Ind., (Indianapolis area) transpondors, although strong synchronized replies were observed at the 1,500-pulse pair fruit level, it was not possible to "lock-on" the station. Examination of the oscilloscope carried in each airplane indicated that this failure was due to erratic transmission of the second reply pulse at high transmitter duty cycles. This is not a traffic handling capacity problem, but was a temporary malperformance of the transpondor. The fact that proper locking-on to the remaining five beacons was accomplished is evidence that excessive fruit was not responsible for this malperformance. It is significant that the Terre Haute, Ind., transpondor, farthest from the operating area, was locked-on at all fruit levels without difficulty. Furthermore, the Terre Haute transpondor was transmitting three pulses (two for reply and one for identification) at all times, thus adding to the total fruit level.

At no time during either flight was the search time of the airborne units appreciably lengthened, nor was the memory time of either unit extended. It is these two functions which generally fail first, under conditions of extremely high random pulse levels, i.e., fruit. One of the interrogators was adjusted for a 15-second memory time throughout the test.

Fig. 10—Federal 100-channel DME interrogator.

These findings support the statistical figures which had been previously presented. In view of the fact that the airborne equipments demonstrated the ability to operate through the highest levels of fruit reached during the tests, the upper limit cannot be stated. On the other hand, since the maximum fruit level produced during the period of the flight test was in excess of that demanded of the equipment by current specifications, there is no reason to believe that the equipment will prove other than completely satisfactory from a traffic handling point of view.

In order to ascertain that the results of November 29 were not representative of some intangible but favorable condition, the tests were repeated on the following day. On November 30, both aircraft repeated the flights of the preceding day with minor schedule changes. The results obtained were the same.

IX. Conclusions

The following conclusions may be drawn on the basis of these investigations of the transition from a 50-channel to a 100-channel DME system:

(1) The combination of 50-channel techniques to provide 100-channel equipment has been accomplished without detriment to the system.

(2) The traffic-handling capacity of present 100-channel equipment is adequate to accommodate any foreseeable future traffic densities.

(3) By compressing DME service into a relatively small portion of the total frequency spectrum allocated for air navigation aids, the way has been left open for system designers to employ considerable latitude in the design of future air navigation facilities complementary to the DME.

Single-Conductor Surface-Wave Transmission Lines*

GEORG GOUBAU†, ASSOCIATE, IRE

Summary—**This paper gives general information on single-conductor transmission lines, in particular on lines having a dielectric-coated conductor. A design chart is included and experimental results are presented.**

I. Introduction

The total field which is excited on any waveguide consists of two parts: the field of the regular wave and a supplementary field which takes care of the continuity of the field at places where the guide is excited, or, where its geometrical structure deviates from the ideal one. In the case of closed waveguides with only one propagating mode, the supplementary field is made up of nonpropagating wave modes with purely reactive power flow. On open guides the supplementary field is a radiation field. Consequently, the excitation of an open guide is connected with more or less radiation loss, and any kind of distortion of the field results in radiation loss. From this point of view the open guides appear inferior to the closed ones. However, the open guides have two advantages: they are less bulky and less expensive; and, since the field is not confined to a small area, the field density is low. This, in return, has the effect that, under comparable conditions, the resistive and dielectric losses of open guides are much smaller than those of closed guides.

The only open waveguide which has been thoroughly investigated so far is the two-wire line. No doubt, this line has a range of applications in which it is superior to cables because of low losses and lower costs. The dielectric guide, known since 1910,[1] received more interest after the dielectric rod antenna was discovered. Although a dielectric wire is capable of propagating a wave mode of very low attenuation[2,3] (the so-called "dipole mode"), its application for transmission lines is not very promising for mechanical reasons, and because the extension of the field is very large. This paper deals with the third type of open waveguides, namely, the single-conductor guide which shows very promising aspects with regard to practical application.

The single conductor as a means for propagating a nonradiating wave is of particular interest—for practical reasons since it appears to be the simplest waveguide; for theoretical reasons since it occupies a unique place among waveguides in that it would not be a guide at all under ideal conditions, that is, if the conductivity were infinite and the surface perfectly clean and smooth. This fact already indicates that an ordinary copper wire, with its high conductivity and rather smooth surface, is not suitable as a waveguide, except for extremely high frequencies where the effect of the finite conductivity becomes more apparent. Although it has been known for more than 50 years, from a paper by Sommerfeld[4], that a conductor of finite conductivity should guide a nonradiating wave mode, actually this wave mode has not been utilized for transmission lines. There are antennas with single-wire feeds, but, the conditions are such that the energy is transported by a radiating wave like on long wire antennas rather than by Sommerfeld's surface wave. In principle, this wave should appear on every antenna built from wires. Experience shows, however, that the field of such antennas is practically the same as the field of ideally conducting antennas and only slightly modified by the finite conductivity. This modification has nothing to do with Sommerfeld's wave, which is an independent solution of Maxwell's equations. The reason why Sommerfeld's wave does not enter into antenna problems is that the usual coupling devices excite chiefly the supplementary field, i.e., the radiating wave. As shown later, Sommerfeld's wave requires launching devices of very large dimensions, and this is because the field extends very far from the conductor; in other words, the field decreases slowly in the radial direction. The large extension of the field makes Sommerfeld's wave impractical not only because of the difficulties in exciting the wave, but also for the reason that a large clearance around the conductor is necessary in order to avoid severe distortion of the field. Furthermore, the large extension of the field makes the wave very sensitive to small bends or sags in the line. The wave would become practical only for frequencies above 10,000 mc.

It is of interest that the usual imperfection of the surface of a conductor can be sufficient to convert Sommerfeld's wave into another wave mode, a wave mode which is mainly determined by the surface conditions. This type of wave is of particular interest since the extension of the field can be controlled by modifications of the conductor surface. It is this type of wave which is considered in this paper.

II. Field of the Wave

The only wave mode which is guided by a single conductor is a radially symmetrical transverse magnetic mode having in cylinder co-ordinates the field components[5]

* Decimal classification: R117.1. Original manuscript received by the Institute, August 16, 1950; revised manuscript received, January 24, 1951. Presented, IX General Assembly URSI, Zurich, Switzerland, September, 1950. Also presented in a condensed version, 1950 IRE National Convention, New York, N. Y., March 8, 1950.

† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

[1] D. Hondros and P. Debye, "Elektromagnetische wellen an dielektrischen drähten," *Ann. Phys.*, vol. 32, p. 465; 1910.

[2] C. H. Chandler, "An investigation of dielectric rod as wave guide," *Jour. Appl. Phys.*, vol. 20, p. 1188; December, 1949.

[3] W. M. Elsasser, "Attenuation in a dielectric circular rod," *Jour. Appl. Phys.*, vol. 20, p. 1193; December, 1949.

[4] A. Sommerfeld, "Fortpflanzung elektrodynamischer wellen an einem zylindrischen leiter," *Ann. Phys. u. Chemie*, vol. 67, p. 233; December, 1899. (See J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., p. 527; 1941.

[5] Mathematically there exist also wave modes with cyclic periodicity around the cylinder (investigated by D. Hondros, "Uber elektromagnetische drahtwellen," *Ann. Phys.*, vol. 30, p. 905; 1909) which are damped out extremely fast and practically not excited.

$$E_r = jAhH_1(\gamma r)e^{j(\omega t - hz)}$$

$$E_z = A\gamma H_0(\gamma r)e^{j(\omega t - hz)} \qquad \text{with} \qquad \gamma = \sqrt{k^2 - h^2}$$

$$H_\phi = jA\sqrt{\frac{\epsilon}{\mu}}\, kH_1(\gamma r)e^{j(\omega t - hz)} \qquad k = \frac{2\pi}{\lambda};$$

(λ = free-space wavelength, h = propagation constant). H_0 and H_1 are Hankel functions of the first or second kind, depending on which root of γ is chosen. The parameter γ is a measure of the steepness of the field decrease in the radial direction. It also determines the reduction in phase velocity. If we call the diameter of a circle around the wire, within which a certain percentage (for instance, 90 per cent) of the energy is propagated, the diameter of the field, then we may say: γ determines the diameter or the extension of the field. The larger γ, the smaller the diameter of the field and the smaller the phase velocity of the wave.

The boundary conditions at the surface of the guide require the continuity of E_z and H_ϕ. They can be satisfied, as shown by Sommerfeld,[4] on a conductor with finite conductivity. Since the ratio between the tangential electric and magnetic field components at the surface of a good conductor is very small, γ is very small compared with k. Hence, the distance from the wire at which the field becomes negligible is a large number of wavelengths.

The boundary conditions can also be satisfied on conductors of infinite conductivity if the surface is coated with a dielectric or magnetic layer. This case, first considered by Harms,[6] is of particular interest. There, γ becomes purely imaginary, provided the losses in the layer are negligible. Since the thickness of this layer can be varied, the value of γ and therefore the extension of the field can be adjusted to any desired value. For frequencies above 100 mc, reasonable field diameters are obtainable under conditions for which the phase velocity of the surface waves is little smaller than the velocity of light and the energy propagated within the layer amounts to a small fraction of the total energy. These facts are illustrated by the curves in Fig. 1. They show the 90 per cent radius of the field ($\rho_{90\%}$), the reduction in phase velocity (δv), and the fraction of the energy propagated within the dielectric layer N_i/N, as functions of the thickness of this layer. The curves refer to a case where the wire radius is 0.1 cm and the frequency 3,000 mc. The dielectric layer is assumed to be enamel with a dielectric constant of 3. We see from the curves that very thin layers of dielectric already affect the extension of the field considerably while the phase velocity is reduced by less than 1 per cent. Since the part of the energy which is propagated within the layer is very small, the dielectric losses are almost negligible. It is of interest that within the frequency range above 100 mc a dielectric layer of 10^{-3} cm has more effect on the extension of the field than the conductivity of good conductors.

[6] F. Harms, "Electromagnetische wellen an einem draht mit isolierender zylindrischer hülle," *Ann. Phys.*, vol. 23, p. 44; 1907.

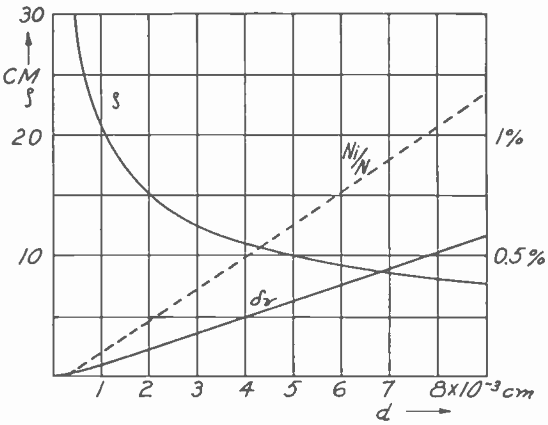


Fig. 1—Ninety-per cent power flow radius of the field $\rho_{90\%}$, reduction in phase velocity δv, and relative power propagated in the dielectric layer d as functions of the thickness of the layer (wire radius 0.1 cm; $\epsilon_i/\epsilon = 3$; $f = 3{,}000$ mc).

It should be mentioned that the boundary conditions required for the existence of the surface wave can also be satisfied as average conditions on conductors with rough surfaces like threaded or twisted wires. If the corrugations become very large, like the disk-loaded wires used in some traveling-wave tubes, the surface wave degenerates to an accompanying field and is no longer the main carrier of the energy.[7] Therefore, the losses are greatly increased.

III. Excitation of the Wave

The excitation device for any surface wave should simulate the field distribution within a cross plane of the guide as well as possible since any deviation from this distribution is compensated by the supplementary field and thus results in a radiation loss. Theoretically, the simulation of the field could be achieved by a layer of electric or magnetic dipoles—in our case, for instance, by axially directed electric dipoles covering the surface of a metal plane which is perpendicular to the wire, with a density proportional to the longitudinal electric field component of the wave. Since the field of the wave extends without limit, the dipole distribution should also extend to infinity. Since in the actual case the field is simulated only within a finite area, a certain radiation loss is unavoidable.

A practical method of simulating the field of the surface wave on a single conductor becomes evident if we compare the field of this wave with the field of a wave in a coaxial line. Consider first an ideal coaxial line with infinite conductivity. The wave which is propagated in such a line is a transverse electric and magnetic mode. If the inner conductor is coated with a dielectric layer, the wave obtains a longitudinal electric component like that of the surface wave, and can be described mathematically as the superposition of two waves: one which has a similar field distribution to the surface wave (described by Hankel functions) and another, the field of which decreases in the *inward* direction (described by a

[7] For a study of disk-loaded guides, see W. Rotman," A Study of Single Corrugated Guides," Air Force Cambridge Research Laboratories, Cambridge, Mass.; February, 1950.

Bessel function). The field of the latter may be considered as the image of that part of the first wave which would extend beyond the outer conductor of the line. If the diameter of this conductor becomes larger and larger, the amplitude of this field decreases more and more, and finally becomes negligible if the diameter is very large.

The effect of the diameter of the outer conductor on the power transmission is demonstrated by the curves in Fig. 2 which show the transmitted power N as a function of the ratio between outer b and inner a radii under the condition that the current I in the line is maintained constant. If we define N/I^2 as the impedance Z of the line, the curves give a plot of Z versus b/a for an ordinary coaxial line, and another coaxial line, the inner conductor of which is coated with a thin dielectric layer. In both cases the losses are neglected. The broken curve which refers to the ordinary coaxial line shows that the impedance approaches infinity if the outer conductor becomes larger and larger. This indicates that the outer conductor is essential even if its radius is already very large. The solid curve which refers to the line with dielectric-coated inner conductor approaches a finite value if the diameter of the outside conductor of the line is increased. Therefore, the outer conductor has no appreciable effect on the wave if its radius is sufficiently large; and the surface-wave transmission line can be considered as a dielectric-coated coaxial line, the outer conductor of which has such a large diameter that it has no effect on the field of the wave. If we would have considered the finite conductivity, the curve for the ordinary coaxial line would also approach a finite limit, but for a much larger ratio of b/a. The finite conductivity creates a wave mode which can be guided by the inner conductor only, namely, Sommerfeld's wave.

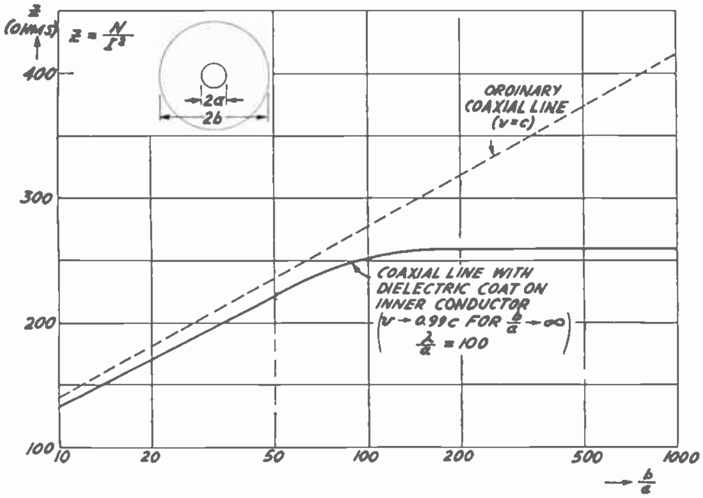


Fig. 2—Effect of the diameter of the outer conductor of a coaxial line on the power transmission if the inner conductor is coated with a dielectric layer.

The consideration of the surface waveguide as a coaxial line with a remote outer conductor immediately answers the question of how the surface wave can be launched (see Fig. 3.) Starting with a coaxial line, the inner conductor of which has a dielectric coat, the outer diameter is gradually increased until it is so large that it does not affect the field considerably. The field distribution then approaches that of the open line. Thus, as shown in Fig. 3, the wave can be launched by means of a metal cone which is connected to the outside conductor of a coaxial feed line. The inner conductor of that line is connected to the surface waveguide. The field lines shown in this figure give a rough idea of how the surface wave develops. The complete setup for a surface-wave transmission line is shown in Fig. 4.

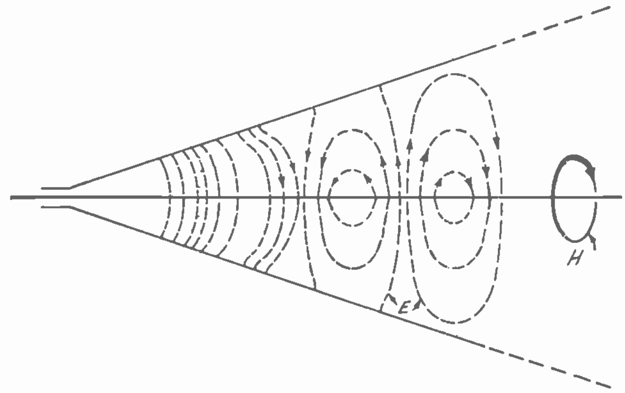


Fig. 3—Launching of the surface wave.

Fig. 4—Sketch of a surface-wave transmission line.

IV. Losses of a Surface-Wave Transmission Line

Losses of surface-wave transmission lines with a dielectric-coated wire consist of three parts: the conductivity loss in the wire, the loss in the dielectric coat, and the loss, effected by the launching device, due to a partial excitation of the radiating mode. In addition to these losses there are further losses, introduced by supports, bends, and sags, which shall be considered later. The conductivity loss and dielectric loss are proportional to the length of the line. They can be calculated from the field distribution of the wave. The launching loss is independent of the length of the line and can be determined with fair accuracy by the following consideration: At the receiving end, that portion of the wave energy will be received which travels within the area of the aperture of the horn. The wave energy outside this area will be lost. The ratio between the received energy and the total energy determines the efficiency of the receiving horn. Because of the reciprocity theorem the efficiency of the transmitting horn must be the same as the efficiency of the receiving horn.

Fig. 5 shows curves from which the three components of the loss can be determined. These curves are used for the actual design of transmission lines.[8] The formulas for the conductivity loss (for copper) and the dielectric loss are given in Fig. 5. P and Q are functions of a variable G, which, for thin dielectric layers, is proportional to the thickness of the layer. The curves ρ'/a' give the required radius of the horns ρ', measured in multiples of the guide radius a' (conductor radius plus dielectric layer thickness) if a loss of 1, 2, 3, $\cdots$ db for both horns together is allowed.

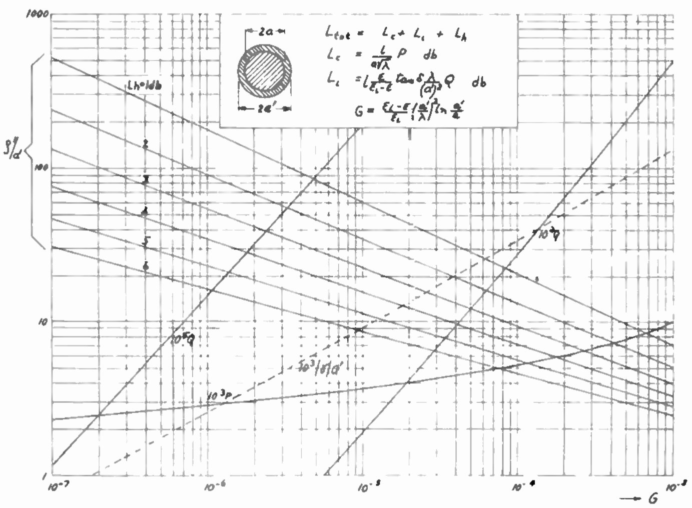


Fig. 5—Design curves (a, a', ρ', and λ in cm, length l of the line in feet).

The conductivity loss which is proportional to P changes little with increasing layer thickness. The dielectric loss rises much faster (see curve Q); however, since only a small fraction of the energy is propagated in the layer, the dielectric loss is usually small compared with the conductivity loss even for rather thick layers. The diameter of the horns for a given launching loss, or the launching loss for a given size of the horns decreases with increasing thickness of the insulation since the field is more concentrated.

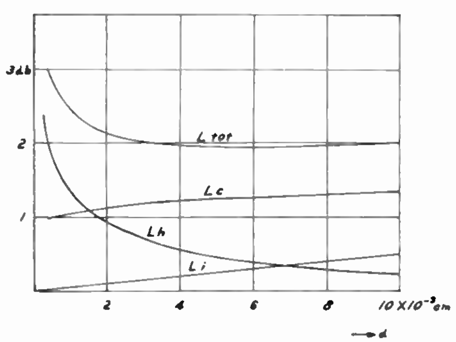


Fig. 6—Conductivity loss L_c, dielectric loss L_i, launching loss L_h, and total loss L_{tot}; as functions of the thickness d of the dielectric layer. (Length of the line 120 feet, wire diameter 0.2 cm, horn diameter 33 cm, $f = 3{,}000$ mc, $\epsilon_i/\epsilon = 3$.)

A numerical example for the relation between the losses and the thickness of the dielectric layer is shown in Fig. 6. The curves are calculated for a copper wire of 0.2 cm diameter and 120-foot length. The dielectric layer is assumed to be enamel ($\epsilon_i/\epsilon = 3$; $\tan \delta = 8 \times 10^{-3}$). The opening of the horns is 33 cm and the frequency 3,000 mc. If the thickness of the layer is increased, the total loss of the line L_{tot} first goes down because of the decreasing launching loss. After passing a flat minimum it rises since the dielectric losses become more and more influential.

A curve like this was verified experimentally. The measurements were made for a frequency of 1,600 mc and a wire diameter of 0.26 cm. The wire was outdoors for several months and had a rather thick corrosion layer. Under this condition the measured loss was 3 db. Then most of the corrosion was removed by sandpaper. The result was an increase of the loss to about 3.6 db due to the increased extension of the field. The expected loss for a wire with perfectly clean surface is much higher. Then several coats of polystyrene solution were applied by means of a brush. The loss went down more and more until a minimum of 1.7 db was reached. Further increase of the dielectric layer gave a slight increase of the loss. The thickness of the dielectric coat varied more than 1:10 along the wire; thus it was not possible to measure the average thickness. This test indicated that the wave is not sensitive to inhomogeneities of the guide.

The frequency response of a surface-wave transmission line is demonstrated in Fig. 7. In this case, an ordinary enamel wire of 0.2-cm diameter and 5×10^{-3} cm enamel thickness ($\epsilon_i/\epsilon = 3$; $\tan \delta = 8 \times 10^{-3}$) is assumed. The length of the line is 120 feet and the opening of the horns 33 cm. The conductivity loss L_c and the dielectric loss L_i increase with frequency while the launching loss L_h decreases. The total loss has a minimum of about 2 db. The actual minimum loss measured for this setup was 2.2 db.

Fig. 8 shows a comparison between the losses of coaxial cables, copper waveguides, and ordinary enamelled copper wires used for surface-wave transmission lines. The curves for the waveguides and the wires are calculated curves. The #12 wire (0.205-cm diameter) has an enamel coat of 0.00375-cm thickness and the #8 wire (0.325-cm diameter) a coat of 0.0025 cm. The dielectric

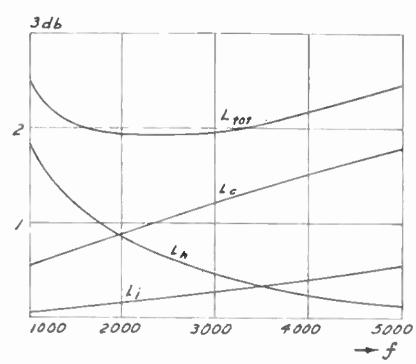


Fig. 7—Frequency dependence of the losses of a line. (Length, 120 feet; wire radius, 0.2 cm; dielectric layer thickness, 0.005 cm; $\epsilon_i/\epsilon = 3$, $\tan \delta = 8 \times 10^{-3}$, horn diameter, 33 cm.)

[8] For the theory on which these curves are based, see G. Goubau, "Surface waves and their application to transmission lines," *Jour. Appl. Phys.*, vol. 21, p. 1119; November, 1950.

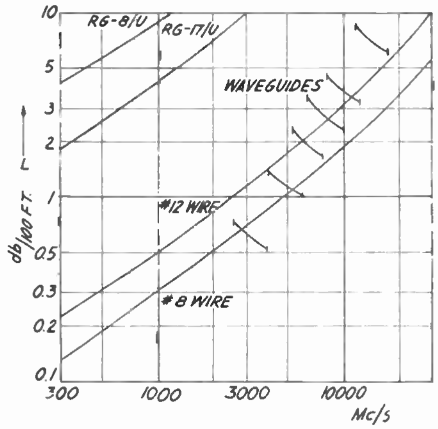


Fig. 8—Comparison between the losses of coaxial cables, copper waveguides, and ordinary enamel copper wires.

constant was assumed to be 3 and the power factor 8×10^{-3}. There is some uncertainty about the value of the power factor at very high frequencies; the value of 8×10^{3} is valid at low frequencies. Therefore, a measurement on the #12 wire was performed at 23,000 mc, where the dielectric loss is comparable with the conductivity loss. The measured loss per 100 feet was 9.5 db while the curve shows 7.5 db. Assuming the discrepancy is due entirely to a larger power factor of the enamel, its value would have to be 80 per cent higher at this frequency range than for low frequencies.

Since there is practically no limitation on the diameter of the conductors as far as other modes are concerned, the loss of surface waveguides can be made much smaller than indicated in Fig. 8. Fig. 9, for instance,

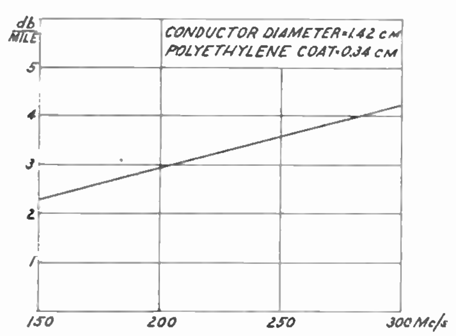


Fig. 9—Theoretical loss of a line for long-distance communication.

shows the theoretical loss of a line with a diameter of 1.42 cm and a polyethylene coat of 0.34 cm. There the loss is only a few db per mile.

What has been considered so far is only the regular loss which is present under ideal conditions where the conductor is perfectly straight and the field is not disturbed by any object in the vicinity of the conductor. Actually the wire has sags and small bends, and it needs supports if the length exceeds a certain limit. Furthermore, since it is in the vicinity of the earth the field cannot extend without limits, as it does in the theory. All these effects cause additional loss by radiation or absorption, and the question is how does this additional loss compare with the regular loss. Although this question requires further studies, it can be answered in part. In our experiments, which were made with lines of enameled wires having a loss of the order of one db per 100 feet, the usual sag and small bends effected by supports caused no measurable increase of the loss. For instance, a test was made with a line of 600-foot length, stretched 4 to 8 feet above the ground along the slope of a hill. The wire was 0.32 cm in diameter and had an enamel coat of 0.0025 cm. The horns had a 33-cm diameter and the frequency was 1,600 mc. The theoretical loss was 4.5 db and the measured loss about 5 db. The number of supports, which consisted of waxed strings, was varied and there was no measurable effect on the loss observed. The situation could be different in cases like that of Fig. 9 where the line has a very low attenuation and a large number of supports are required within a stretch of 1 db loss. Although there are, as yet, no experimental results available for such lines, approximate calculations show that even in such cases the supports should not increase the loss considerably, provided they are properly made. The proximity of the ground has under reasonable conditions no noticeable effect on the performance of the line.

Another cause of losses should be mentioned, namely, rain and ice. Rain is troublesome only if it accumulates on the conductor in the form of drops. These drops act like small radiating dipoles. A water film is of no consequence in its effect on the performance of the line. It merely enlarges the thickness of the dielectric layer and behaves like a rather good dielectric because the ratio, power factor to dielectric constant minus one, which determines the losses, is very small. The radiation loss of drops can be calculated approximately. It decreases with the second power of the conductor radius and the fourth power of the wavelength. In the frequency range below 1,000 mc the loss should be negligible. For higher frequencies some observations were made. On the 600-foot line mentioned before (diameter 0.32 cm) an increase of the loss by 1.5 db was measured for a frequency of 1,600 mc under conditions where parts of the wire were densely covered with big drops. The same increase occurred under similar conditions at 3,300 mc on the 120-foot line (diameter 0.2 cm). In case of vertically inclined antenna feed lines, where drops cannot stay settled, the effect of rain is much less. For experimental purposes surface-wave transmission lines were installed in a television relay link which was operated over a period of several weeks. The frequency was 2,000 mc. The lines showed no noticeable increase of loss during rain or snow fall.

The formation of ice on the lines is serious since ice forms thick layers with rather poor dielectric properties. However, the formation of ice in most locations can be easily prevented by electric heating.

V. Acknowledgments

These experiments were made at the Signal Corps Engineering Laboratories. The author wishes to express his appreciation to John Hessel; R. E. Lacy, Chief and

Deputy Chief, Radio Communication Branch; J. J. Egli, Chief, Radio Relay & Microwave Section; and especially to the group who co-operated in the measurements, C. E. Sharp, L. R. Battersby, E. A. Conover, and, particularly, A. Meyerhoff, who also assisted in the preparation of the material.

Standards on Pulses: Definitions of Terms—Part I, 1951*

COMMITTEE PERSONNEL

Standards Committee, 1950–1951

J. G. Brainerd, *Chairman*
L. G. Cumming, *Vice-Chairman*
A. G. Jensen, *Vice-Chairman*

M. W. Baldwin, Jr.	J. E. Keister	R. F. Shea
R. R. Batcher	E. A. Laport	J. R. Steen
H. G. Booker	Wayne Mason	Henry Tanck
F. T. Budelman	R. A. Miller	Bertram Trevor
W. G. Cady	L. S. Nergaard	W. N. Tuttle
P. S. Carter	A. F. Pomeroy	L. C. Van Atta
J. W. Forrester	George Rappaport	K. S. Van Dyke
A. G. Fox	H. E. Roys	D. E. Watts
R. A. Hackbusch	P. C. Sandretto	Ernst Weber
J. V. L. Hogan	E. S. Seeley	L. E. Whittemore

Definitions Co-ordinating Subcommittee

A. G. Jensen, *Chairman*
M. W. Baldwin, Jr.
P. S. Carter
E. A. Laport

Task Group on Pulse Definitions

C. J. Hirsch, *Chairman*

R. H. Daugherty, Jr.	W. M. Goodall	Nathaniel Rochester
Harold Goldberg	W. H. Huggins	W. G. Tuller

C. M. Wheeler W. T. Wintringham

I. INTRODUCTION

THE MEANINGS of commonly used terms in pulse work have often been a matter of disagreement. The IRE Standards Committee, faced with the fact that different technical committees proposed different definitions for the same term, and wishing to try to introduce uniformity where little has existed, set up a special task group with wide representation of special interests to propose standard definitions of terms concerned with pulses. Months of work by the task group and intensive critical review by the Standards Committee has led to results the first half of which are given below. These are necessarily compromises. The Standards Committee urges IRE members (a) to try to use the terms according to the definitions below, so that reasonable uniformity may be achieved, and (b) to take into account that in this particularly controversial region many compromises have been necessary so that favorite meanings and uses may appear not to have been considered, whereas in actuality it is unlikely that the very thorough review of the field has failed to unearth them for the Committee's consideration. The present set of definitions has been judged

* Reprints of this Standard, 51 IRE 20 S 1, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

sufficiently important to print at this time; it is hoped that the second half will be ready for printing within nine months.

Since many pulse shapes are possible, and a clear concept of the ones under discussion is desirable, it may be helpful to use drawings of pulse shapes, pulse times, magnitudes, durations, and the like, to show how these quantities apply.

In these definitions, linear superposition of pulses and possibly other waveforms is understood. Since it is possible to generate pulses whose characteristics may not seem to be adequately covered by the definition, it has been assumed that complex pulses can be analyzed in terms of more fundamental pulses and waveforms somewhat as a complex periodic wave can be considered as the sum of a fundamental and harmonics.

II. DEFINITIONS

Average Absolute Pulse Amplitude. The average of the absolute value of the instantaneous amplitude taken over the pulse duration.

Note—By "absolute value" is meant the arithmetic value regardless of algebraic sign.

Average Pulse Amplitude. The average of the instantaneous amplitude taken over the pulse duration.

Crest Factor of a Pulse. The ratio of the peak-pulse amplitude to the rms pulse amplitude.

Leading Edge Pulse Time. The time at which the instantaneous amplitude first reaches a stated fraction of the peak pulse amplitude.

Mean Pulse Time. The arithmetic mean of the leading edge pulse time and the trailing edge pulse time.

Note—For some purposes the importance of a pulse is that it exists (or is large enough) at a particular instant of time. For such applications the important quantity is the *Mean Pulse Time.* The *Leading Edge Pulse Time* and *Trailing Edge Pulse Time* are significant primarily in that they may allow a certain tolerance in timing.

Peak Pulse Amplitude. The maximum absolute peak value of the pulse excluding those portions considered to be unwanted, such as spikes.

Note—Where such exclusions are made, it is desirable that the amplitude chosen be illustrated pictorially.

Pulse. A variation of a quantity whose value is normally constant; this variation is characterized by a rise and a decay, and has a finite duration.

Note 1—The word "pulse" normally refers to a variation in time; when the variation is in some other dimension, it shall be so specified, such as "space pulse."

Note 2—This definition is broad so that it covers almost any transient phenomenon. The only features common to all "pulses" are rise, finite duration, and decay. It is necessary that the rise, duration, and decay be of a quantity that is constant (not necessarily zero) for some time before the pulse and has the same constant value for some time afterwards. The quantity has a normally constant value and is perturbed during the pulse. No relative time scale can be assigned.

Pulse Amplitude. A general term indicating the magnitude of a pulse.

Note 1—For specific designation, adjectives such as average, instantaneous, peak, rms (effective), etc., should be used to indicate the particular meaning intended.

Note 2—Pulse Amplitude is measured with respect to the normally constant value unless otherwise stated.

Pulse Amplitude, Average. See *Average Pulse Amplitude.*

Pulse Amplitude, Average Absolute. See *Average Absolute Pulse Amplitude.*

Pulse-Amplitude Modulation (PAM). Amplitude modulation of a pulse carrier.

Pulse Amplitude, Peak. See *Peak Pulse Amplitude.*

Pulse Amplitude, RMS (Effective). See *RMS (Effective) Pulse Amplitude.*

Pulse Bandwidth. The smallest continuous frequency interval outside of which the amplitude of the spectrum does not exceed a prescribed fraction of the amplitude at a specified frequency.

Caution—This definition permits the spectrum amplitude to be less than the prescribed amplitude within the interval.

Note 1—Unless otherwise stated, the specified frequency is that at which the spectrum has its maximum amplitude.

Note 2—This term should really be "Pulse Spectrum Bandwidth" because it is the spectrum and not the pulse itself that has a bandwidth. However, usage has caused the contraction and for that reason the term has been accepted.

Pulse Carrier. A pulse train used as a carrier.

Pulse Decay Time. The interval between the instants at which the instantaneous amplitude last reaches specified upper and lower limits, namely, 90 per cent and 10 per cent of the peak-pulse amplitude unless otherwise stated.

Pulse Duration. The time interval between the first and last instants at which the instantaneous amplitude reaches a stated fraction of the peak pulse amplitude.

Pulse-Duration Modulation (Pulse-Length Modulation) (Pulse-Width Modulation). A form of pulse-time modulation in which the duration of a pulse is varied.

Note—The terms "pulse-width modulation" and "pulse-length modulation" are also used to designate this system of modulation but the term "pulse-duration modulation" is preferred.

Pulse Duty Factor. The ratio of the average pulse duration to the average pulse spacing.

Note—This is equivalent to the product of the average pulse duration and the pulse repetition rate.

Pulse Frequency Spectrum. See *Pulse Spectrum.*

Pulse Interleaving. A process in which pulses from two or more sources are combined in time-division multiplex for transmission over a common path.

Pulse Interval. See *Pulse Spacing.*

Pulse-Interval Modulation. A form of pulse-time modulation in which the pulse spacing is varied.

Pulse-Length Modulation. See *Pulse-Duration Modulation.*

Pulse-Position Modulation (PPM). A form of pulse-time modulation in which the position in time of a pulse is varied.

Pulse, Radio-Frequency. See *Radio-Frequency Pulse.*

Pulse Regeneration. The process of restoring pulses to their original relative timings, forms, and/or magnitudes.

Note—In many devices, pulses may become distorted due to phase or amplitude distortion, limiting, or other processes. It is often desirable to restore the pulse to something resembling its original form before it has become so distorted that the original information which it contains is completely destroyed. This process is normally called pulse regeneration.

Pulse Repetition Frequency. The pulse repetition rate of a periodic pulse train.

Pulse Repetition Period. The reciprocal of the *Pulse Repetition Frequency.*

Pulse Repetition Rate. The average number of pulses per unit of time.

Pulse Rise Time. The interval between the instants at which the instantaneous amplitude first reaches specified lower and upper limits, namely, 10 per cent and 90 per cent of the peak-pulse amplitude unless otherwise stated.

Pulse Spacing (Pulse Interval). The interval between the corresponding pulse times of two consecutive pulses.

Note—The term "pulse interval" is deprecated because it may be taken to mean the duration of the pulse instead of the space or interval from one pulse to the next. Neither term means the space *between* pulses.

Pulse Spectrum (Pulse Frequency Spectrum). The frequency distribution of the sinusoidal components of the pulse in relative amplitude and in relative phase.

Note—The definition of this term was phrased to convey the idea that the spectrum is a complex (phasor) function of frequency and to express this function most nearly in a manner which corresponds to the method of measuring it (i.e., measuring amplitude and phase separately).

Pulse Spike. An unwanted pulse of relatively short duration superimposed on the main pulse.

Note—This term came into wide use in radar to define the first part of the pulse fed through a TR tube. This portion contains most of the pulse energy, has a duration about 10^{-3} that of the rest of the pulse, and an amplitude up to 10^6 to 10^9 times that of the rest of the pulse. Seen on a cathode-ray tube, it looks like a spike sticking up from the pulse. By extension, the term has come to be applied to any *unwanted* pulse of relatively short duration superimposed on the wanted pulse.

Pulse Time, Leading Edge. See *Leading Edge Pulse Time.*

Pulse Time, Mean. See *Mean Pulse Time.*

Pulse-Time Modulation. Modulation in which the time of occurrence of some characteristic of a pulse carrier is varied from the unmodulated value.

Note—This is a general term which includes several forms of modulation, such as pulse-duration, pulse-position, pulse-interval modulation.

Pulse Time, Trailing Edge. See *Trailing Edge Pulse Time.*

Pulse Train. A sequence of pulses.

Pulse, Unidirectional. See *Unidirectional Pulse.*

Pulse-Width Modulation. See *Pulse-Duration Modulation.*

RMS (Effective) Pulse Amplitude. The square root of the average of the square of the instantaneous amplitude taken over the pulse duration.

Radio-Frequency Pulse. A radio-frequency carrier amplitude-modulated by a pulse. The amplitude of the modulated carrier is zero before and after the pulse.

Note—Coherence of the carrier (with itself) is not implied.

Trailing Edge Pulse Time. The time at which the instantaneous amplitude last reaches a stated fraction of the peak pulse amplitude.

Unidirectional Pulse. A pulse in which pertinent departures from the normally constant value occur in one direction only.

Note—This is sometimes called "single-polarity" pulse, a term which is deprecated.

Condenser Microphone Sensitivity Measurement by Reactance Tube Null Method*

H. E. VON GIERKE† AND W. W. VON WITTERN†

Summary—A reactance tube circuit acting as a sinusoidally varying capacity in the frequency range from 20 to 200,000 cps is used to compensate the capacity variations of a condenser microphone in a high-frequency carrier circuit. The microphone is driven as in the electrostatic actuator method by a voltage of the same frequency, that modulates the reactance tube capacity. By adjusting amplitude and phase of the reactance tube modulation, the total capacity change of the circuit becomes zero. The sensitivity and phase lag of the microphone response can be determined from the actuator and modulating voltages when the circuit is properly calibrated. The circuit is advantageous in obtaining an equivalent pressure chamber calibration of condenser microphones up to 200 kc. The null method makes it useful for relatively unsensitive microphones.

Introduction

A CALIBRATED reactance tube circuit (RTC), designed to represent a variable condenser, is useful for the measurement of periodic capacity changes as in the calibration of condenser microphones.

Fig. 1 will be used to explain the principle of this calibration method. The microphone is driven as in the electrostatic actuator method.[1,2] The sinusoidal actuator voltage and the necessary polarizing dc voltage is applied between the membrane and the counter-electrode of the microphone, to produce periodic electrostatic forces which cause vibrations of the membrane producing capacity variations. The capacity of the microphone C and the inductance of the coil L form a circuit which is excited by a carrier frequency generator the frequency of which is situated on the side of the resonance curve of the circuit. A change of the capacity of the microphone changes the tuning of the circuit and thus the carrier frequency voltage across the circuit. This voltage is modulated by these capacity changes. When rectified and displayed on the oscilloscope, the curve represents the capacity change of the circuit. However, owing to the response of the LC circuit, the amplitude of this trace depends on the modulating frequency.

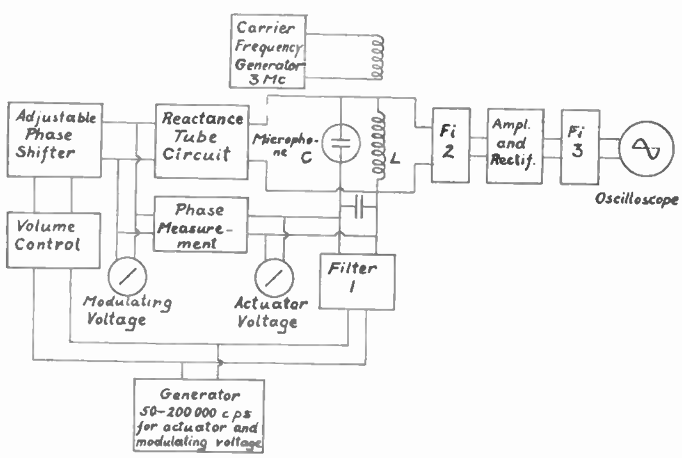


Fig. 1—Schematic diagram of the arrangement used for the microphone calibration by means of the reactance tube circuit.

Parallel to the LC circuit is the reactance-tube circuit (henceforth referred to as RTC). Its modulating voltage is taken from the same generator as is the actuator voltage for the microphone. By adjusting the amplitude and phase of this modulating voltage the capacity change of the RTC can be made equal and opposite in phase to the capacity change of the microphone. Under the condition that any series resistances in the microphone and RTC branch are negligible the total capacity change of the circuit then will be zero. This adjustment can be controlled by observing the oscilloscope. The ratio of the modulating voltage to the actuator voltage is proportional to the sensitivity of the microphone. The phase relation between these voltages corresponds to the phase relation between membrane-force and resultant capacity change. This phase relation is easy to measure because the magnitude of these voltages is of the order of several volts.

The filter networks 1 and 2 prevent the actuator voltage from influencing the oscilloscope directly by electrical cross talk. The output impedance of the filter 1 must be very low for the carrier frequency. Filter 3, sharply tuned to the frequency under measurement, is applied to decrease the noise level so that a very high amplification can be used.

The advantage of this null method is that the result is independent of the frequency response characteristics of filters and amplifiers and of the frequency response characteristic which results from transients of the LC circuit.

The Reactance Tube Circuit

For such measuring purposes the RTC must produce only reactance changes without resistance changes because both produce a carrier modulation in the circuit. A constant ohmic resistance parallel to the circuit would have no influence other than on the sensitivity of the measurement.

* Decimal classification: R254.2×R385.53. Original manuscript received by the Institute, May 8, 1950; revised manuscript received, October 30, 1950.
A preliminary report of this work was presented before the Acoustical Society of America in St. Louis, November 19, 1949, JASA 22 (1950) p. 85.

† Aero Medical Laboratory Engineering Division, Air Materiel Command, Dayton, Ohio.

[1] M. Gruetzmacher and E. Meyer, "Eine Schallregistriervorrichtung zur Aufnahme der Frequenz-Kurven von Telephonen und Lautsprechern," *Elekt. Nach. Techn.*, vol. 4, pp. 203–211; 1927.

[2] L. J. Sivian, "Absolute calibration of condenser transmitters," *Bell Sys. Tech. Jour.*, vol. 10, pp. 96–115; January, 1931.

A pure reactance exists between plate and cathode of an electron tube when the plate voltage is "fed back" to the grid with a 90° phase shift.[3,4] To realize this condition in the RTC (Fig. 2) we employ an adjustable phase shifting stage. To produce a periodically changing capacity the amplitude of the reactance tube (RT) plate current is modulated. Direct-grid modulation of the RT produces a varying plate resistance in parallel with the desired varying capacity. The modulation is therefore produced in a special stage. To avoid undesired variations of the plate resistance of the RT, the modulating voltage is removed by compensation and filtering following the "modulation stage."

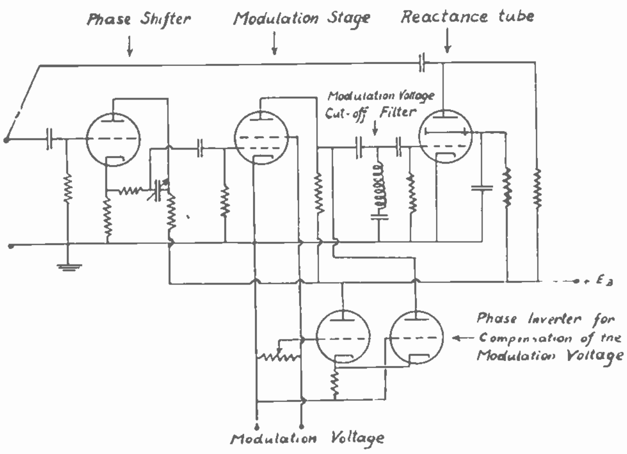


Fig. 2—Complete circuit of the reactance tube circuit without resistive component.

Phase Adjustment of the RTC

If in addition to the varying capacity, the RTC has a varying resistance component the carrier voltage across the tuned circuit LC (Fig. 1) is modulated when the carrier frequency is located on the side of the resonance curve, and also when the same carrier frequency is located at the peak of the resonance curve. But a small change in capacity alone will modulate the voltage only when the carrier frequency is located on the side of the resonance curve. Therefore, by switching off the microphone actuator voltage and tuning the LC circuit exactly to the carrier frequency, the phase of the RTC can be adjusted so that the modulation by the RTC disappears. This indicates that the circuit is acting as a pure reactance. Exact tuning of the circuit is obtained by adjusting for disappearance of the modulation produced by the microphone, when the modulating voltage of the RTC is cut off.

Microphone Calibration Procedure

Following the adjustment of the RTC, for calibration of the microphone, the LC circuit is slightly detuned with the carrier frequency maintained constant. The actuator voltage is now applied resulting in modulation of the voltage across the LC circuit by both the microphone and the RTC. By adjusting the phase and the amplitude of the modulating voltage to the RTC, the summed carrier modulation is reduced to zero. In this compensated state no sideband frequencies are present at the input of the RTC. This condition is essential because sideband frequencies would introduce errors, since the RTC is not adjusted to them. Now only the unmodulated carrier to which the RTC acts as a pure capacity is present.

Frequency Response Characteristic of the RTC

The capacity change of the RTC must be the same for all modulating frequencies. This means that the modulation percentage of the output current resulting from a certain modulating voltage must be independent of the modulating frequency. The frequency response characteristic of the circuit, between the modulating stage and the output stage, is therefore designed to be flat between the sideband frequencies of the modulated carrier. This is accomplished by resistance capacity coupling even for the 2- to 4-mc range because the tubes do not have to produce any amplification. Therefore very low coupling resistances (~200 ohms) can be used. The modulation percentage was checked by measuring the carrier current peaks of the output stage with and without a modulating voltage. Its frequency response characteristic was found to be flat within less than ±0.5 db in the frequency range 50–200,000 cps.

Capacity Calibration of the RTC

The capacity change of the RTC resulting from the application of a certain modulating voltage was calibrated at a low frequency by replacing the microphone with a variable plate condenser driven by an electrodynamic moving coil system. This system was driven by the generator producing the modulating voltage of the RTC. The condenser plate amplitude was controlled by a microscope. The extreme values of its capacity due to this amplitude were measured under static conditions. Because the capacity changes obtained this way were much larger than the capacity changes of the microphone, they were reduced by means of a condenser network.

Determination of the Pressure Equivalent to the Actuator Voltage

The microphone calibration described gives the capacity change of the microphone ΔC_M due to the actuator voltage. To obtain the capacity change resulting from a sound pressure P it is necessary to determine, at one frequency, what sound pressure is equivalent to a certain actuator voltage. In the calibration circuit described above (Fig. 1) with the RTC switched off, the microphone is simultaneously excited by an actuator

[3] C. Travis, "Automatic frequency control," Proc. I.R.E., vol. 23, pp. 1125–1142; October, 1935.
[4] P. Guettinger, "Frequenzmodulation," Zuerich, Leeman, p. 111; 1947.

voltage and by a known low-frequency sound pressure produced in a pressure chamber. The piston of the pressure chamber is driven by an electrodynamic speaker system excited by the actuator voltage generator. By adjusting the amplitude and the phase of the current in the speaker system, the capacity change of the microphone can be made zero. The forces resulting from the actuator voltage and from the sound pressure are equal and opposite in phase in this case.

Results of the Microphone Calibration

With the factor $\Delta C_M/P$ determined, the open circuit voltage E_M resulting from the pressure P in the usual low-frequency circuit with the battery voltage E_B can be calculated when the static capacity C of the microphone is measured:

$$\frac{E_M}{P} = E_B \frac{\Delta C_M}{P} \frac{1}{C}.$$

Fig. 3 shows the results of the calibration of a 640 AA microphone and a small condenser microphone of our own design.[5] To show sensitivity and phase they are presented as polar diagrams. The radial scale indicates the sensitivity of the microphone in decibels, the angles of the radii indicate the phase lag between the pressure acting on the membrane and the output voltage of the microphone. The frequencies are distributed along the curve in a way which depends on the qualities of the microphone. Every loop indicates a resonance in this kind of diagram. The calibration method described represents a pressure calibration. Its results correspond within 1 db with pressure chamber calibrations obtained with the usual methods, however not extended to frequencies that high. They also correspond to free field reciprocity calibrations within 1 db in the low-frequency range, and show the deviation to be expected at high frequencies due to the diffraction on the microphone.

[5] H. F. Koster, H. E. von Gierke, and H. L. Oestreicher, "The calibration of microphones to one hundred kilocycles per second," *Jour. Acous. Soc. Amer.*, vol. 21, p. 58; January, 1949.

Application of the Calibration Method

The calibration method described might be especially useful for the calibration of relatively unsensitive condenser microphones designed for the measurement of high pressures and shock waves. For this calibration it is hardly possible to produce by conventional methods sinusoidal pressures of the necessary magnitude and frequency. By the electrostatic actuator method too only limited forces can be produced on the membrane, how-

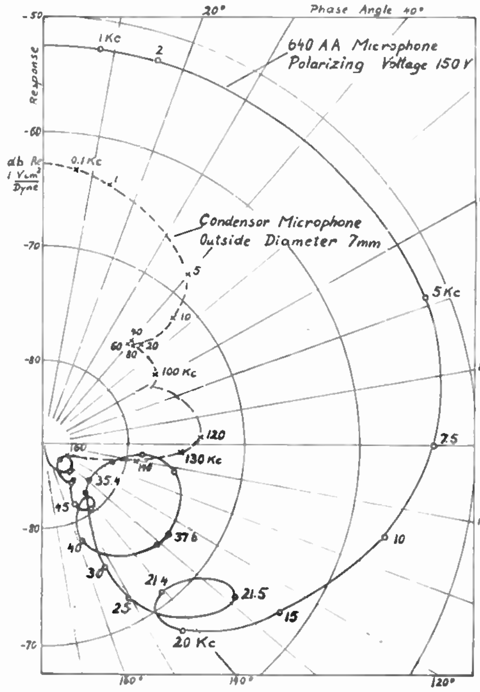


Fig. 3—Sensitivity and phase lag of two different microphones:
radial scale = sensitivity in db ref. 1 volt cm²/dyne
angle of the radii = phase lag in degrees between pressure acting on the membrane
output voltage numbers along the curves = frequencies in cps.

ever not limited in frequency. Even with these limited forces a calibration of insensitive microphones will be possible, because the method described allows diminution of the noise level and a considerable increase in sensitivity by the application of very sharp filters without increasing the difficulties of the measurement.

Grid Current and Grid Emission Studies in Thyratrons—the Trigger-Grid Thyratron*

L. MALTER†, SENIOR MEMBER, IRE, AND M. R. BOYD‡, STUDENT, IRE

Summary—Grid emission leads to unstable control characteristics in thyratrons. Positive ion grid current during the afterglow results in delayed recovery of grid control. Both these drawbacks can be overcome by use of low values of grid resistors, but at the expense of input signal current sensitivity. The requirement for low grid resistance can be eliminated by assigning the firing function to a fine wire electrode known as the trigger grid. The other function of a conventional thyratron grid, which is to promote recovery of control, is assigned to another electrode known as the blocking grid. Since the latter electrode is connected to a bias supply through a low resistance, its emission and positive ion collection do not produce disturbing effects. Comparisons of an experimental trigger grid thyratron with its conventional prototype reveal many advantages of the former.

Introduction

A THYRATRON desideratum is the reduction of the time needed for recovery of grid control following the interruption of a discharge. This time, which is generally referred to as "deionization time," we choose to call "recovery time," T_R, in view of the fact that recovery of grid control can occur when intense ionization is still present in the tube.[1] The period following the interruption of a discharge is generally referred to as the "afterglow."

One of the factors that delays recovery of grid control is the flow of positive-ion current to the grid during the afterglow. This current produces a potential drop in the grid resistor which neutralizes the effect of the grid bias source, thus preventing the early return of the grid potential to its bias value. This can be minimized by a reduction in the effective area of the grid, thus reducing its positive-ion collection. It is the purpose of this paper to describe work in progress along these lines. As will be seen, the step of reducing grid area is effective not only in reducing recovery time, but is accompanied by additional benefits, such as (1) increased sensitivity to input signal current, (2) reduced effects of grid emission with consequent increased stability as regards variation in heater current and plate power, and (3) reduced interelectrode capacitances.

Considerations Regarding Thyratron Recovery

A. Plasma Decay

Fundamental studies of phenomena occurring during the afterglow[1] have led to interesting results that are of value in understanding thyratron behavior. Significant among these results are a conception of a process of plasma decay and the conditions necessary for recovery. The plasma decay has been shown to be due primarily to a diffusion process.

A gas plasma is a region of essentially equal positive ion and electron density. Fundamental studies indicate that when the discharge in a thyratron is interrupted the tube is substantially completely suffused with a plasma, and that as the ionization then diminishes it does so in such a manner that the plasma retains its plasma-like characteristics. This is accomplished by an equal rate of loss of electrons and ions to the walls and electrodes.

The plasma decay can be speeded up by decreasing the molecular weight and pressure of the gas and by shrinking down the geometry of the structure. In general, rapid recovery of grid control requires a rapid decrease of plasma density. Decreasing the pressure or using a lighter gas may lead to difficulties resulting from more rapid cleanup of the gas present, thus resulting in shorter life. Shrinking the geometry can take the form of decreasing interelectrode spacing. This speeds up the diffusion process whereby deionization takes place.

The above described factors which determine T_R may be described as primary since they are inherently properties of the tube geometry or its gas filling, and are effective through the mechanism of altering the rate of plasma decay. T_R can also be varied by changes in operating parameters, such as anode current or voltage, grid voltage, and grid resistance. These are secondary factors which do not materially influence the rate of plasma decay. It will be of interest to examine the role of the grid in promoting recovery. This will throw light on the manner in which recovery is also dependent upon the other operating parameters.

B. The Role of the Grid in Recovery

The grid in a conventional thyratron triode plays a double role. One function is to determine whether or not a discharge is to occur in a previously unfired tube. Thus, it is a firing electrode. An equally important function is to perform in such a fashion as to promote or insure recovery of control, following the interruption of a discharge. This matter, which is fully discussed in the literature[1], is briefly presented here.

As was stated above, during the afterglow the tube is suffused with a plasma. Now if the grid is negative with respect to the plasma, it will surround itself with a positive-ion sheath. As the plasma decays, the sheath thickness increases. The positive-ion current to the grid is determined by the rate at which ions diffuse out of the

* Decimal classification: R337.12. Original manuscript received by the Institute, May 5, 1950; revised manuscript received, November 30, 1950. Presented, 1950 IRE National Convention, New York, N. Y., March 8, 1950.
† RCA Laboratories, Princeton, N. J.
‡ Stanford University, Stanford, Calif.
[1] L. Malter and E. O. Johnson, "Studies of thyratron behavior," *RCA Rev.*, vol. 11, p. 165; June, 1950.

plasma into the sheath. Since the sheath area surrounding the grid varies only slightly with grid potential, one finds the positive-ion current to the grid to be almost independent of grid potential.[2] This in turn implies that the density of ionization in the plasma will be practically uninfluenced by the grid potential.

As the plasma decays, the sheath around the negative grid increases in thickness until it finally extends across the grid opening or openings.[3] When this occurs, the residual plasma in the anode and cathode spaces becomes independent and can assume widely different potentials. Each plasma then assumes a potential very close to that of the most positive electrode with which it can make contact.[3] Thus, if the grid is negative with respect to the cathode and anode, the plasma potential on the cathode side will be substantially that of the cathode and similarly on the anode side the plasma potential will be close to that of the anode. The potential distribution along a section through the cathode-grid aperture is as shown in Fig. 1. The tube now behaves like a thyratron with reduced spacings, the "effective cathode" being at A and the "effective anode" at B. Whether or not the tube fires depends closely on

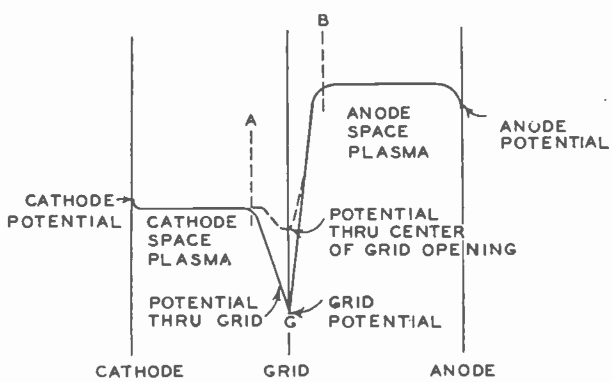


Fig. 1—Potential distribution between thyratron electrodes.

whether or not a thyratron with cathode at A, anode at B, and grid at G, and with the potentials as shown in Fig. 1 will refire. As time progresses, the grid sheath expands, forcing A and B farther away from the grid. Thus, the effective refiring geometry varies in the fashion that one would expect, to explain the observed results that T_R increases with anode refiring voltage. It is also apparent that since the residual plasma density at the discharge interruption is a monotonic function of the discharge current T_R should increase with anode current. As the grid sheath expands, the control ratio gets larger, approaching as a limit the control ratio of the unfired tube.

For the grid to recover control, it must first surround itself with a sheath of sufficient extent to block its own openings. The more negative the grid, the more rapidly does the sheath block the openings and the greater the spacings from grid to the effective anode and cathode positions. This explains why recovery time is reduced by making the grid increasingly negative.

In the conventional triode thyratron the dual functions of firing and of promoting recovery are performed by the same electrode, the control grid. By assigning these functions to separate electrodes one can achieve improved performance in several ways.

Note—There is a common misconception that the decrease of T_R brought about by making the grid increasingly negative arises from a more rapid deionization of the plasma. Evidence for the view that the grid potential has no first-order effects on the rate of deionization is presented in the literature.[1]

C. Grid Resistance and Recovery

We are now in a position to see how the presence of a grid resistor causes an increase in T_R. (We summarize here briefly some of the experimental results and conclusions of the literature.[1])

A three-element thyratron is first fired by simultaneous pulsing of plate and grid. (See Fig. 2.) The grid

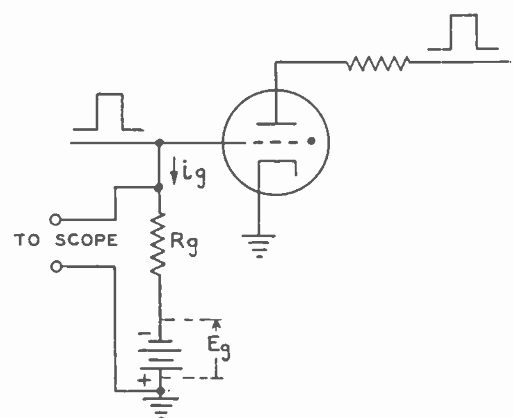


Fig. 2—Thyratron study test circuit.

potential during the afterglow is found to follow a course as shown in Fig. 3. Following the interruption of

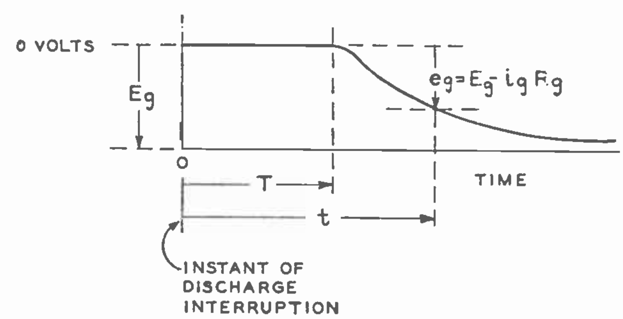


Fig. 3—Thyratron grid voltage during period following interruption of discharge.

the discharge, the grid remains substantially at cathode potential for time T, and then decays along an exponential curve towards the bias value. This is explained in the following manner: Following the discharge inter-

[2] I. Langmuir and H. M. Mott-Smith, "Studies of electric discharges in gases at low pressures," *GE Rev.*, vol. 27, pp. 449–455, 538–548, 616–623, 762–771; 1924.

[3] W. Koch, "Experimental demonstration of the existence of ion sheaths at electrodes of rectifiers during deionization," *Zeit. fur Tech. Phys.*, vol. 17, p. 446; 1936.

ruption, positive-ion current flows to the grid, this current being given by

$$i_g = i_{g0}e^{-t/\tau}, \tag{1}$$

where i_{g0} is positive-ion current to probe immediately following the interruption of the discharge, and τ is the time constant of the plasma decay.

When

$$i_gR_g > E_g \tag{2}$$

(these quantities are defined in Figs. 2 and 3), the drop in the grid resistor exceeds the bias value and the grid tends to go positive. Under this condition, the grid also collects electrons, the electron current being just sufficient to make the grid operate close to cathode potential. This condition persists until

$$i_gR_g = E_g. \tag{3}$$

Beyond this point ($t=T$, in Fig. 3) the grid stops collecting electrons and becomes increasingly negative.

There is no possibility for the recovery of grid control until $t>T$. This follows from the fact that when $t<T$ the anode and cathode plasmas are connected. Under these conditions any potential applied to the anode in excess of the arc drop will cause refiring. From (1) and (3) we obtain

$$T = \tau \ln \frac{i_{g0}R_g}{E_g}. \tag{4}$$

From (4) we see that the greater i_{g0} and R_g, the longer will recovery be delayed. This accounts for the effect of R_g on T_R. Furthermore, we see that if i_{g0} can be decreased, then it becomes possible to get more rapid recovery. Incidentally, if i_{g0} can be decreased sufficiently, then it is possible to increase R_g and still obtain reduced T_R. But increased R_g means increased sensitivity to signal current, a desirable result. A means for decreasing i_{g0} is to decrease the effective area of the grid. We shall see later how this can be effected by a separation of the firing and recontrol functions of the grid.

Considerations Regarding Grid Emission

It has been shown that T_R can be diminished by decreasing the interelectrode spacings. However, when an effort is made to go in this direction, one generally runs into difficulties with increased grid emission due to increased grid temperature, particularly if power levels are not reduced. This grid emission, when flowing through the grid resistor, results in a reduction of the effective grid bias. Since the emission may be nonconstant in time, it can lead to erratic performance, in that the bias may need frequent readjustment in order to prevent a continuous discharge.

Temperature measurements were made on a conventional thyratron triode (see Fig. 4), by means of thermocouples mounted on the grid and anode. Fig. 5 shows the temperature of the electrodes as the tube warmed up and was fired. From the temperature data

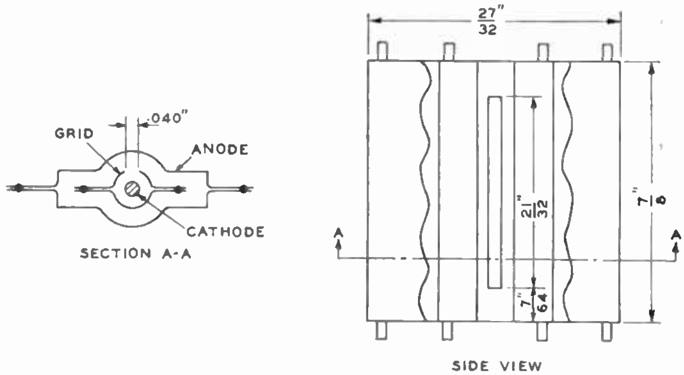


Fig. 4—Sections through conventional thyratron.

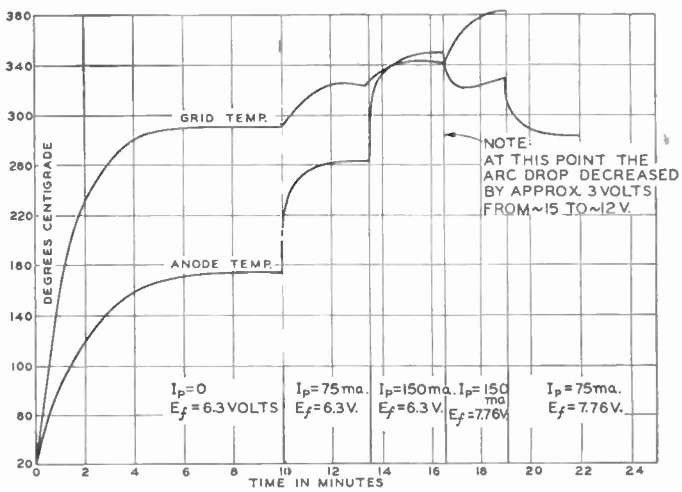


Fig. 5—Thyratron grid and anode temperatures following power turn-on.

and from grid-emission measurements, the work function of the grid surface was found to be approximately 1.5 volts. The constant A of Richardson's equation was 5 amperes per square centimeter per degree squared. During normal operation of the tube, the grid reaches a temperature of 600°K. The close spacings and the presence of an oxide cathode with its accompanying barium evaporation onto the grid make it appear that grid emission would not be unexpected. Fig. 6 shows curves of grid emission from the commercial thyratron (of the form shown in Fig. 4) under various operating conditions. In these tests the grid was held at −25 volts with respect to cathode and anode during the reading of the grid current. This value of potential served to saturate the grid emission without introducing undue ionization in the gas. In the cases where plate current was drawn, the grid emission was measured while the plate was momentarily shorted to ground. The emission variations (particularly the decrease in the case where $E_f=6.3$ v, $I_b=150$ ma) indicate that besides temperature variation of the grid, the barium distribution over its surface must have been changing with consequent change in effective thermionic constants. It is obvious from these figures that if a large grid resistor were employed the bias required to just hold off firing would change with time, thus resulting in unstable operation.

Another factor which makes the effects of grid emission serious is that the electric fields may be sufficient

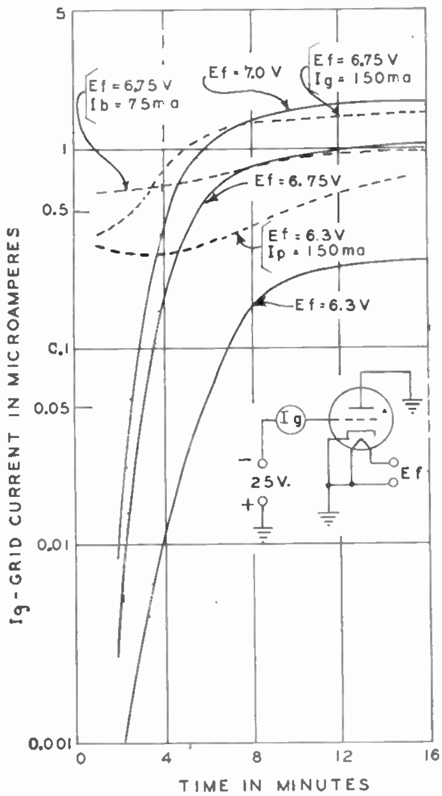


Fig. 6—Thyratron grid current following power turn-on.

to produce field intensified ionization in the gas. Any negative grid bias or positive anode voltage will serve to accelerate emitted electrons and produce ionization which tends to increase the grid current even more. Changes in grid current by factors of over 20 are observed when the grid potential is varied between 0 and −200 volts.

Recapitulation and Problem Formulation

It appears that two obvious methods of securing reduced T_R are:

1. Decreasing electrode spacings
2. Reducing positive-ion current to control grid during the afterglow.

If the first of these methods is adopted, grid emission increases with resultant reduction in stability of the control characteristics. This can be overcome by reduction of effective grid area or of grid resistance. Decreasing the size of the grid resistance is undesirable in that it results in a reduced sensitivity to signal current. But the reduction of effective grid area is just what method 2 calls for in order to secure reduced T_R. Thus the indications are that the direction of attack should be to reduce spacings and to decrease substantially the effective area of the control grid.

The grid in the conventional three-electrode thyratron has a double function: (1) it serves to initiate the discharge, and (2) it serves to produce a sheath of sufficient thickness during the afterglow, to isolate the plasmas on the plate and cathode sides, thus making recovery of grid control possible. A separation of these functions between two electrodes can be of advantage in that the firing electrode, which is the one whose electron emission and whose positive-ion collection during the afterglow are deleterious, can be made small in order to minimize these effects.

The Trigger-Grid Thyratron—Design and Principles

On the basis of the preceding observations and considerations an experimental thyratron of the form shown in Fig. 7 was developed. A typical operating circuit is illustrated in the same figure. It is seen that this modified thyratron differs from that of the standard commercial form shown in Fig. 4, in that an additional electrode in the form of a fine wire has been added and placed within the widened grid opening of the former structure. The fine wire is referred to as a trigger grid because of the fact that its major function is to trigger the discharge. The tube is designated as a "trigger-grid thyratron" (TGT). The recovery function which is

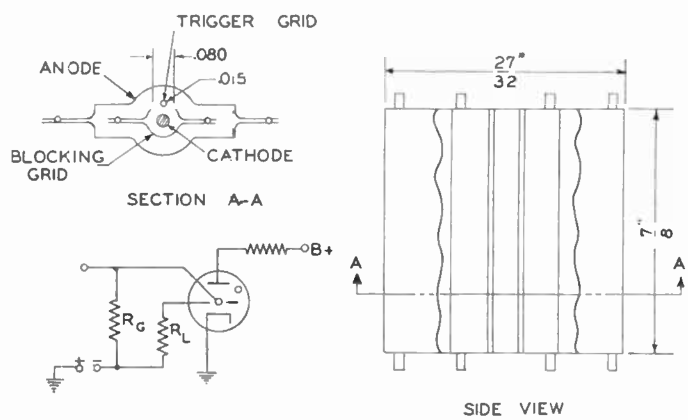


Fig. 7—Sections through trigger-grid thyratron. TGT operating circuit.

normally played by the same grid that initiates the discharge, is now achieved primarily by the blocking grid. The input signal is applied to the trigger grid only. Because of the fineness of the trigger grid, the positive-ion current to it during the afterglow is very small. As a consequence, even though R_g is made large, the trigger grid can return to its bias value quite rapidly, thus giving reduced values of T_R. Concomitantly, one achieves an improved signal current sensitivity due to the permissibly larger value of R_g. Since the blocking grid resistor R_L need be only large enough to prevent the flow of excessive grid currents during the discharge, the blocking grid returns to its bias potential almost instantaneously following the interruption of the discharge. (It was found that a value of R_L of about 5,000 ohms worked satisfactorily in the experimental tubes.) The consequent very rapid sheath buildup around the blocking grid aperture promotes rapid recovery. It is seen thus, that by the incorporation of a trigger grid into the structure of a three-electrode thyratron so as to achieve a separation of functions of firing and sheath formation, one profits by the reduced current flow to the trigger grid during the afterglow.

Another advantage of this design is in the reduced grid-emission effects. Since the blocking grid is tied to the bias source through a low-value resistor, the potential drop across this resistor due to grid emission is negligible compared to the bias potential. Thus the unstable characteristics, which can exist when the triggering and blocking functions are combined in one grid, are obviated. This will be brought out in the results presented in the following section.

The Trigger-Grid Thyratron—Experimental Results

The advantages of the TGT are brought out in comparative studies. These included measurement of the following:

A. Grid emission
B. Control characteristics
C. Critical grid bias
D. Recovery characteristics
E. Interelectrode capacitance
F. Arc drop.

The results will be discussed seriatim.

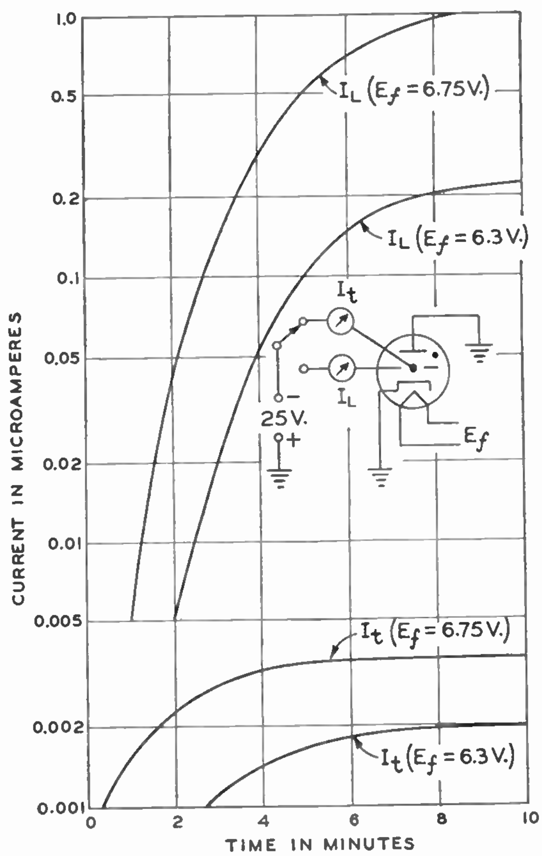


Fig. 8—Trigger-grid and blocking-grid currents following power turn-on.

A. Grid Emission

Grid emission was measured in the circuit shown in Fig. 8. The emission from trigger and blocking grids as a function of time following turning on the heater is shown in Fig. 8. It is seen that assigning the firing action to a small-area electrode has resulted in marked reduction in grid emission from the electrode of importance—the electrode which is connected to the input signal resistor. A comparison of Fig. 8 with Fig. 6 shows that the emission from the blocking grid of the TGT is about the same as that from the control grid of the conventional triode. Thus the grid emission, in so far as its effect on control characteristics is concerned, is seen to have been reduced by a factor well in excess of 100.

Cases for which plate current was drawn were also studied. It was found, as above, that the trigger-grid emission is of the order of 40 to 100 times less than that from the blocking grid of the same tube or that from the control grid of its conventional prototype. (In making these measurements the grid emission was determined while the anode was momentarily shorted to the cathode.)

B. Control Characteristics

Since the TGT contains two grids, a complete presentation of its control characteristics (CC) requires a family of curves. These are presented in Fig. 9. This figure also includes the control characteristic of the conventional thyratron of Fig. 4. A comparison indicates

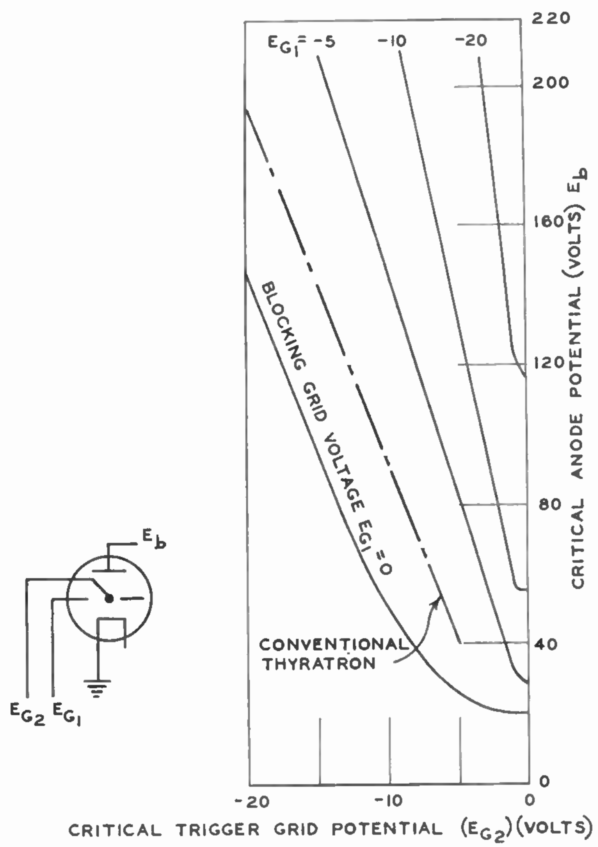


Fig. 9—TGT control characteristics.

that the experimental TGT has a somewhat higher control ratio than the conventional thyratron. Actually, by changes in the size of the blocking-grid aperture and of the position of the trigger grid, the control ratio of the TGT can be varied over a wide range around that of the conventional thyratron of Fig. 4.

C. Critical Grid Bias

The critical grid voltage (CGV) is the grid voltage at which the tube just fires for a particular value of anode voltage. Values of this quantity are given in Fig. 9. If there is no grid emission, then the bias source potential corresponding to that state at which the tube just fires is equal to the CGV. When grid emission is present, then, due to the drop in the grid resistor, the bias potential must be increased so as to prevent firing. We shall call this value of bias potential, the "critical grid bias" (CGB). For dc potentials on the anode, the CGB can be computed from the values of CGV, grid emission, and the magnitude of the grid resistor. If the anode signal is ac in character, then due to capacitance between anode and grid, a portion of the anode signal will appear at the grid (when a grid resistor is present). A still greater bias will then be required to prevent firing.

A set of data to show the relative merits of the various tube types for the case in which the anode voltage is ac in nature is shown in Fig. 10. In this case, instead of operating at fixed values of anode voltage and grid resistance, critical bias values over a range of anode voltage were determined for a number of values of grid resistance. The resultant curves bear a resemblance to the control characteristics and, in fact, coincide with them when $R_g=0$. In taking these data, the tube was permitted to "settle down" to a final steady value of CGB before recording the data. The curves include within themselves not only the effect of primary grid emission, but of gas amplification and of the voltage induced on the grid as a consequence of the interelectrode capacitance. The curves need no comment other than that they, too, clearly show the decidedly greater stability of the TGT, in this case with ac on the plate.

D. Recovery Characteristics

The recovery characteristics of a TGT and of the conventional thyratron of Fig. 4 are shown in Fig. 11.

In these measurements, the tube was fired for 6 microseconds at an anode current of 250 ma. A later 200-volt test pulse was applied to the anode to determine whether the grid had recovered control. The T_R characteristic for the conventional thyratron of Fig. 4 is not very different from the curve obtained by joining the points of the TGT data for which $V_{G1}=V_{G2}$. This is to be expected since, from the point of view of the

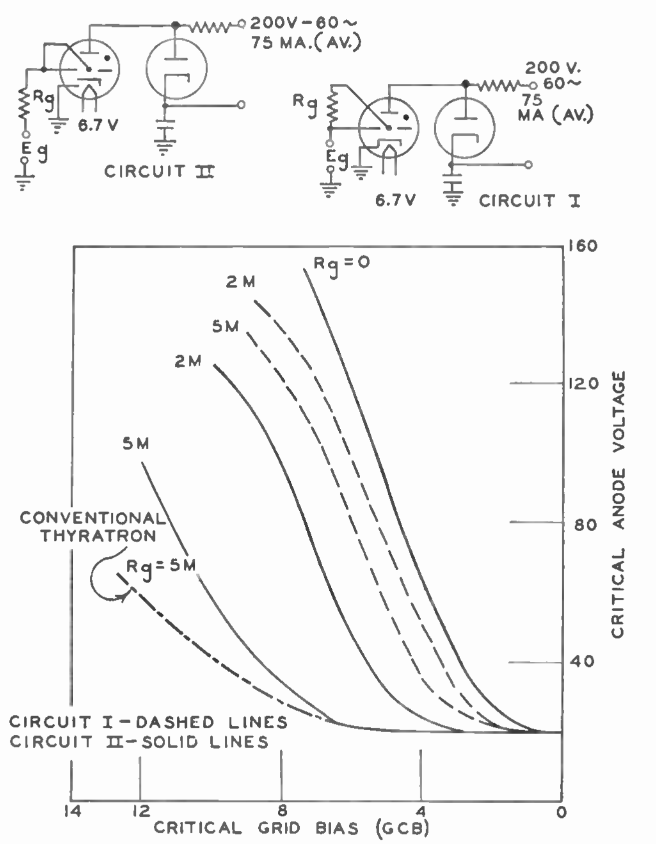


Fig. 10—Control characteristics of TGT and conventional thyratrons for various values of grid resistor.

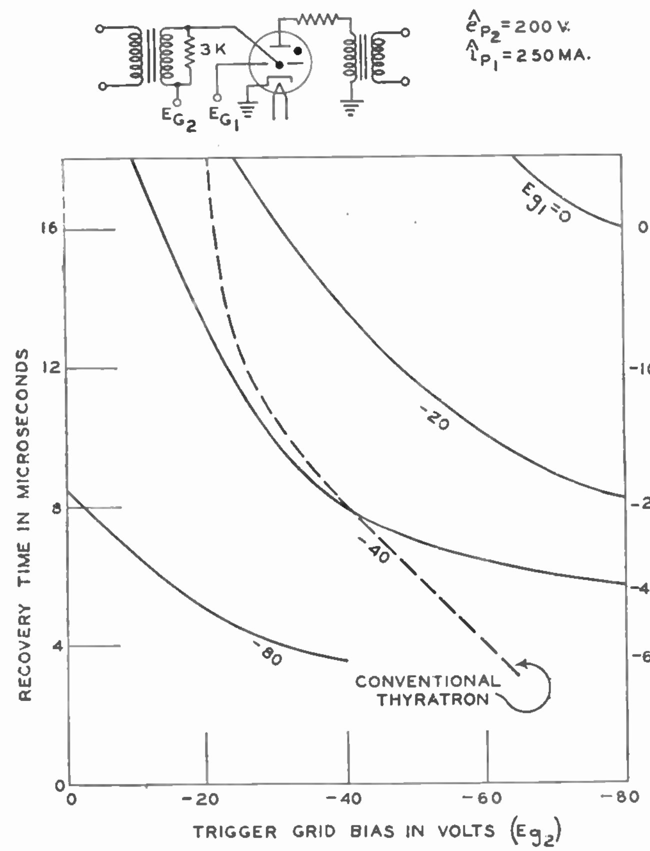


Fig. 11—Recovery times of TGT and conventional thyratron as function of grid voltage.

recovery mechanism, the TGT tube with joined trigger and blocking grids is similar to the prototype thyraton.

As was explained above, the effect of grid resistance is to prevent the applied dc bias from appearing at the grid immediately after the interruption of the discharge because of the voltage drop due to positive-ion current flowing to the grid and thus through the grid resistance. This decreased effective bias results in a longer T_R. The magnitude of the positive-ion current from the decaying plasma to any electrode is almost independent of its potential, but is very much dependent on its area. Thus, if the positive-ion collecting area is made small, the detrimental effect of grid resistance on T_R is reduced. The trigger electrode, designed with a small area to minimize emission, therefore, has the property of collecting less ion current during the decay period. Consequently, it returns to the bias voltage value more rapidly. Fig. 12 brings this out and shows the superiority of the TGT with respect to T_R when series grid resistance is present. The observed behavior would be advantageous in many applications where low T_R is desired.

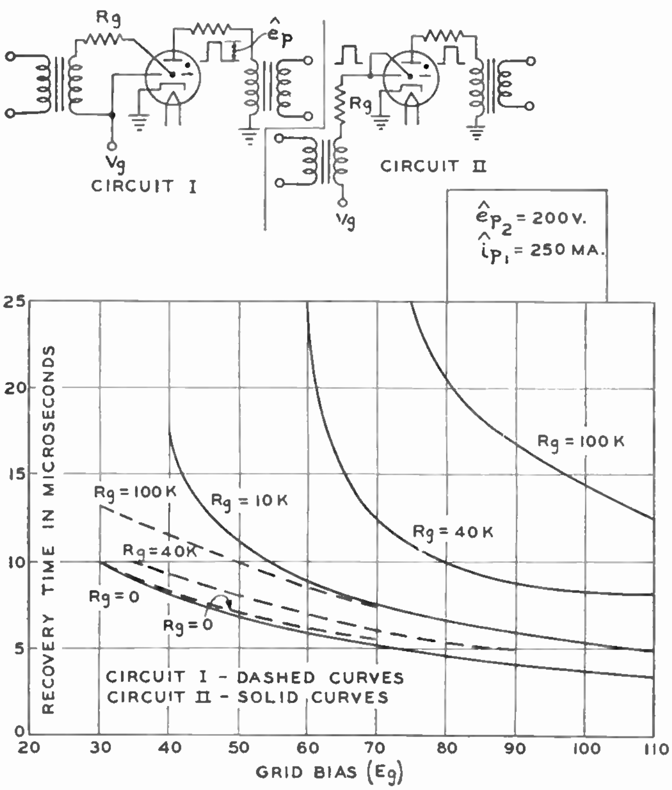


Fig. 12—Recovery times of TGT for various values of grid resistor.

E. Interelectrode Capacitance

In Fig. 12, part of the observed superiority of the TG tube is due to the much lower capacitance between the TG and the anode than between the control grid and anode of the conventional thyratron. The reduced capacitance results in a reduction of the signal induced on the control electrode by varying anode voltage. Thus in the case of the curves of Fig. 12, the instantaneous grid voltage values may be appreciably less than the bias values.

The interelectrode capacitances were measured on a bridge with the results shown in Table I.

TABLE I

Electrodes	Capacitance
Trigger grid to anode	0.83 $\mu\mu f$
Trigger grid to cathode	1.11 $\mu\mu f$
Trigger grid to blocking grid	0.77 $\mu\mu f$
Blocking grid to anode	5.06 $\mu\mu f$
Blocking grid to cathode	1.61 $\mu\mu f$

These results should be compared with those for the conventional thyratron of Fig. 4. The rated values for this tube are: grid to anode capacitance $=6\ \mu\mu f$; grid to cathode capacitance $=2\ \mu\mu f$.

Thus, there is essentially no difference in the input capacitance, but there is a decrease in the feedback capacitance by a factor of about seven. This is very helpful at high-frequency operation, particularly with high grid resistance, since it means that a much smaller disturbing signal appears on the control electrode due to the presence of a varying voltage on the anode. As was seen above, this makes itself felt in smaller effective T_R.

F. Arc Drop

Arc drop readings were taken on the TGT on the thyratron of Fig. 4 and found to be substantially identical.

Summary Comparison of the Trigger-Grid and Conventional Thyratrons

From the experiments described above it is apparent that the TGT has a number of advantages over its conventional prototype particularly in the presence of a large grid resistance.

A summary comparison between permissible operating values of the experimental TGT and its conventional prototype is given in Table II.

TABLE II

Characteristic	Conventional Thyratron	TG Tube
Grid emission R_g max I_p max E_f max	1–10 μamps 0.5 mc* 75 ma* 6.7 v	~0.05 μamps > 10 mc >200 ma >7.7 v
Recovery time		Lower than conventional thyratron when $R_g \neq 0$
C_{g-p}	6 $\mu\mu f$*	0.8 $\mu\mu f$
Control characteristics		Similar to conventional thyratron when $R_g \approx 0$ Improved when $R_g \neq 0$

* Rated values.

Summary and Conclusions

It has been shown that the limitations on a thyratron are set in part by grid emission, and in part by positive-ion-grid current during the afterglow. These factors set an upper limit to the permissible value of grid resistor and plate current. Exceeding these limits results in unstable and time-varying control characteristics and in delayed recovery of grid control.

Efforts to reduce recovery time by reducing electrode spacing lead to increased grid emission with consequent unstable performance.

A promising means for minimizing the limitations involves adding to modified conventional structures an additional small area electrode which is used as the control or firing electrode. Tubes incorporating this extra electrode are known as trigger-grid thyratrons. The grid of a conventional thyratron normally serves a double function, i.e., of firing the tube and of promoting recovery of grid control following the interruption of a discharge. In the TGT these functions are largely separated. The firing function is assigned to the trigger grid only, while the recovery is accomplished mainly by the action of the so-called blocking grid. The benefits of this design are due in part to the reduced emission from the small area of the trigger grid and in part to its smaller collecting area for positive ions during the afterglow. On test it was found that the current rating of an experimental TGT tube was superior to its conventional prototype as regards permissible plate current, grid resistance, and heater voltage. The TGT is also superior as regards recovery characteristics and the effects of grid-plate capacitance.

While the experimental studies and development centered around a low-power thyratron, the principles involved and the methods developed appear applicable to thyratrons of any power rating.

ACKNOWLEDGMENT

The authors wish to express their appreciation for helpful discussions to G. C. Carne and B. Coler of the RCA Victor Division, and to E. O. Johnson of the RCA Laboratories.

A Study of Tropospheric Scattering of Radio Waves*

A. W. STRAITON†, SENIOR MEMBER, IRE, D. F. METCALF†, MEMBER, IRE, AND C. W. TOLBERT†, MEMBER, IRE

Summary—**The theory of tropospheric scattering of radio waves recently advanced by Booker and Gordon[1] makes the prediction that radio-frequency energy may arrive from a ground-based very-high-frequency transmitter at angles well above the horizon, and also that signals received as a result of the scattering process will attenuate less rapidly with distance if the polarization of the source is horizontal than if it is vertical. Experimental observations made at wavelengths of 3 meters and 3 centimeters have verified these deductions.**

I. INTRODUCTION

RADIO FIELD INTENSITIES observed at distances considerably beyond the line of sight have regularly been found to be much higher than would be expected on the basis of the conventional refraction theory of tropospheric waves. This has been noted for both overwater and overland propagation,[2–4] but until this year no satisfactory theoretical treatment has been brought forth.

Booker and Gordon[1] have recently applied the theory of scattering to radio waves in the troposphere. From their results, they conclude that a radio wave beyond the horizon of the transmitting source is made up of a component due to refraction and also a component due to scattering, the latter often being as strong as, or stronger than, the refraction component. At distances not far beyond the horizon, the refracted wave is predominant and the scattered wave is then a cause of fading.

At the greater distances where the result of scattering predominates, the scattering theory predicts that a considerable portion of the energy will arrive from angles well above the horizon. One result of this is an apparent broadening of the acceptance pattern of a high-gain antenna system. The theory further predicts that at great ranges the attenuation of signal with distance is greater for vertical than for horizontal polarization.

A number of experiments performed during the past year at the Electrical Engineering Research Laboratory of The University of Texas have largely verified these predictions. This paper will describe these experiments and present some of the findings.

II. DISTRIBUTION OF SCATTERING FROM A TURBULENT REGION

Booker and Gordon give the scattered power $\sigma(\theta, x)$ per unit solid angle, per unit incident power density, and per unit macroscopic element of volume by

$$\sigma(\theta, x) = \frac{\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2}\left(\frac{2\pi l}{\lambda}\right)^3 \sin^2 x}{\lambda\left[1+\left(\frac{4\pi l}{\lambda}\sin\frac{\theta}{2}\right)^2\right]^2} \tag{1}$$

* Decimal classification: R113.308. Original manuscript received by the Institute, May 16, 1950; revised manuscript received, October 3, 1950.

This work was sponsored in part by the Office of Naval Research and in part by the Central Radio Propagation Laboratory of the National Bureau of Standards.

† University of Texas, Austin, Tex.

[1] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," PROC. I.R.E., vol. 38, pp. 401–412; April, 1950.

[2] J. P. Day and L. G. Trolese, "Propagation of short radio waves over desert terrain," PROC. I.R.E., vol. 38, pp. 165–175; February, 1950.

[3] M. Katzin, R. W. Bauchman, and W. Binnian, "3- and 9-centimeter propagation in low ocean ducts," PROC. I.R.E., vol. 35, pp. 891–905; September, 1947.

[4] K. A. Norton, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. 1, pp. 381–423; 1948.

where

ϵ = the average permittivity

$\Delta\epsilon$ = the departure of the permittivity from its average value

l = the scale of turbulence

λ = the wavelength

x = the angle between the direction of the electric field vector and the direction to the receiver

θ = the angle between a line drawn from the transmitter through the scattering region and a line drawn from the scattering region to the receiver. (θ is small when the scattering region is near the optical horizon).

Referring to Fig. 1, for horizontal polarization the value of angle x will be 90° when the plane formed by the lines from the scattering point to the transmitter and to the receiver is normal to the earth. The sine of x will therefore be approximately unity with the scattering region located anywhere between overhead and the horizon.

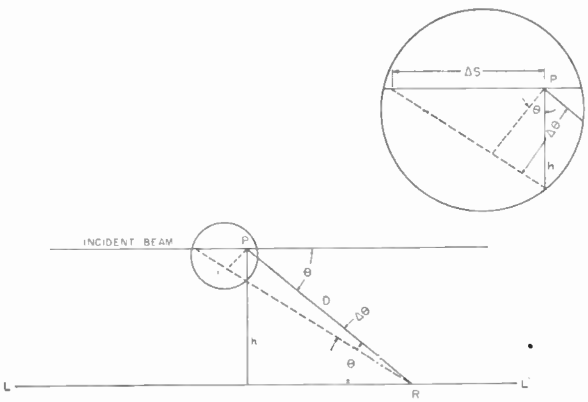


Fig. 1—Geometric quantities in scattering formula.

For vertical polarization, however, the angle x will be zero when the scattering region is overhead and will approach 90° when the scattering region nears the horizon. Thus, scattered signals for vertical polarization will be reduced except when arriving horizontally.

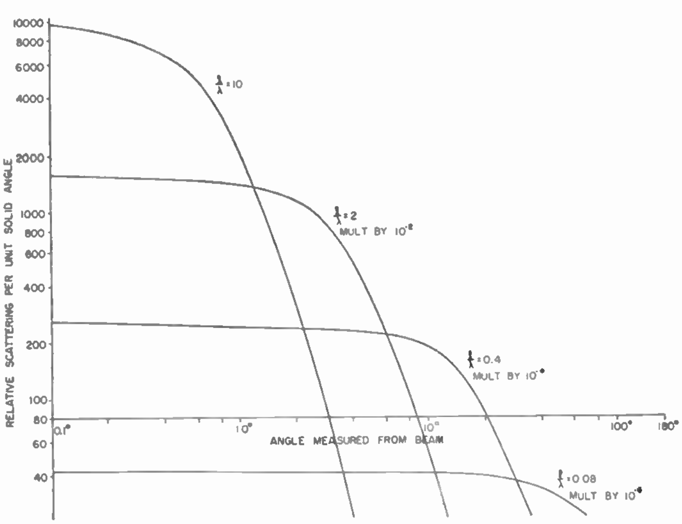


Fig. 2—Relative intensity of scattered signal per unit solid angle, horizontal polarization.

The relative distribution with the angle θ is shown in Fig. 2 for four values of l/λ and for horizontal polarization. If the scale of turbulence is taken as 30 cm, $l/\lambda=0.08$ corresponds to a frequency of 80 mc, and $l/\lambda=10$ corresponds to a frequency of 10,000 mc.

III. Distribution of Signal Scattered from an Isolated Region on a Plane Parallel to the Incident Beam

If the radio scattering occurs from an isolated region P in Fig. 1, the power density received along a line parallel to the incident beam will be given by

$$w(\theta, x) = \sigma(\theta, x)/\frac{4D^2}{\pi}. \tag{2}$$

Replacing D by $h/\sin\theta$ and using the value of $\sigma(\theta, x)$ from (1), we have

$$w(\theta, x) = \frac{\pi\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2}\left(\frac{2\pi l}{\lambda}\right)^3 \sin^2 x \sin^2\theta}{4\lambda h^2\left[1+\left(\frac{4\pi l}{\lambda}\sin\frac{\theta}{2}\right)^2\right]^2}. \tag{3}$$

It is of interest to note that this power density along the line LL' will be maximum at an angle θ_m given by

$$\sin^2\frac{\theta_m}{2} = \frac{1}{2+\left(\frac{4\pi l}{\lambda}\right)^2}. \tag{4}$$

For very large values of l/λ, the power density will be maximum a long distance from the scattering region (i.e., for $\theta\to 0$), while for very small values of l/λ, the maximum power density will be directly below the scattering region.

IV. Power Density from Beam of Uniform Illumination

Let us consider the power density at a receiving point R (Fig. 1) due to scattered radiation originating in an equally illuminated beam. The length of the beam Δs, which will be included in an incremental angle $\Delta\theta$, will be given by $D\Delta\theta/\sin\theta$ or $h\Delta\theta/\sin^2\theta$. The power intensity, $\rho(\theta, x)$ per unit solid angle at R will be given

$$\rho(\theta, x) = w(\theta, x)\frac{\Delta s}{\Delta\theta} = \frac{\pi}{4\lambda h}\;\frac{\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2}\left(\frac{2\pi l}{\lambda}\right)^3\sin^2 x}{\left[1+\left(\frac{4\pi l}{\lambda}\sin\frac{\theta}{2}\right)^2\right]^2}.$$

This distribution of power intensity as a function of the angle θ has the same form except for a constant factor as (1). The graph of Fig. 2 may therefore be thought of as the distribution of the signal received from scattering along a uniformly illuminated beam as a function of angle of elevation.

For large values of l/λ, the scattering would be confined to a small angle near the horizon while for smaller values of l/λ, the scattered radiation will also be observed at higher elevation angles. Evidence of the arrival of radiation at large elevation angles is presented in the following section.

V. The Effect of Antenna Directivity on the Reception of Distant FM Stations

As a test of the arrival of radio energy at large elevation angles, a comparison was made of the signal received from the same commercial FM station on two antennas. One of these antennas was a horizontal dipole and the other consisted of a pair of horizontal dipoles separated horizontally one-half wavelength and fed out of phase. The double-dipole antenna system had a gain of 6.5 db over the simple dipole. This includes transmission-line losses in both systems, which were different. Radiation at right angles to the main lobe of the double-dipole antenna and in a plane normal to the dipole elements was down 15 db from the radiation level in the direction of the main lobe. Two types of antenna comparisons were made as follows:

(1) Directive antenna pointed toward the FM transmitter.

(2) Directive antenna pointed vertically.

A. Directive Antenna Pointed Toward the FM Transmitter.

Simultaneous recording on the two antennas were made from FM stations in Dallas, Houston, and San Antonio, Texas. These stations have frequencies of 97.9, 102.9, and 101.5 mc, respectively, and are at distances of 182, 146, and 73 miles from the receiving site in Austin. The signal levels which were exceeded for various percentages of time were determined from computers or graphs for each hour recorded.

In comparing the signals from Dallas and Houston delivered by the two antennas to the receiving systems, it was noted that the ratio of the signals received from the two antennas was a function of signal level. This trend is illustrated in Fig. 3 for the 50-per cent signal from Dallas for the period December 13–16, 1949. This was a period following the entrance of a cold air mass into the state during which there was a continuous advection of cold air and synoptic conditions were fairly constant.

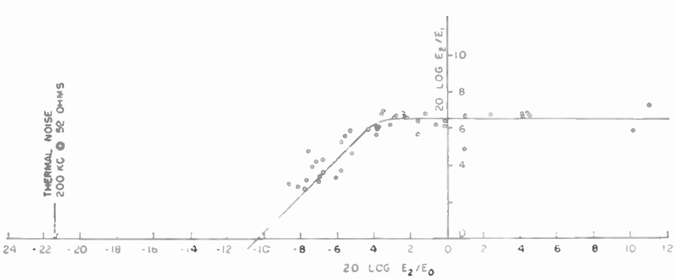


Fig. 3—Comparative signals received with directive and nondirective antennas over 182-mile path, 97.9 mc.
E_1 = open-circuit voltage, dipole antenna
E_2 = open-circuit voltage, directive antenna
$E_0 = 4.8(10^{-6})$ volts.

This same trend was seen in all of the data for the 146- and 182-mile paths although the deviation of the measured points from a smooth curve was greater in some cases. This deviation was attributed to changes in meteorological conditions which occurred during the intervals of measurement.

It is noted in Fig. 3 that, for strong signals, the measured ratio was the nominal gain of the directive antenna over the dipole antenna, indicating that the signals were arriving horizontally. For weaker signals, however, the signal ratio in decibels decreased linearly with the strength of the signal received by the directive antenna. This suggests that the weaker signals came from higher elevation angles. The curve shown was drawn through the points in Fig. 3 by eye. The low-signal asymptote of this curve was a straight line with a slope of 45° which was the condition which would exist if the signal from overhead was statistically independent of time as the signal arriving horizontally became very small. The high-level asymptote was the ratio of response which would exist for a signal arriving horizontally. The values of this scattered signal for a number of four-day periods are shown in Table I. Because of the spread of the measured points, these values are subject to a possible error of ± 0.5 db.

TABLE I

Scattered Signal Strength and Strength of Refracted Wave as Calculated for a Standard Atmosphere

Date	Decibel Relative to 1 Microvolt/Meter /Kilowatt			Microvolt/Meter /Kilowatt		
KPRC-FM, Houston—146 Miles						
Refracted Wave, Calculated for Standard Atmosphere		−45			0.0058	
Scattered Signal	per cent 20	per cent 50	per cent 80	per cent 20	per cent 50	per cent 80
January 7–10, 1950	− 5.7	− 8.8	−10.7	0.52	0.36	0.29
January 3–6, 1950	− 5.2	− 7.3	− 9.7	0.55	0.43	0.33
December 30, 1949–January 2, 1950	− 4.8	− 6.0	− 7.0	0.58	0.50	0.45
WFAA-FM, Dallas—182 Miles						
Refracted Wave, Calculated for Standard Atmosphere		−61			0.00084	
Scattered Signal	per cent 20	per cent 50	per cent 80	per cent 20	per cent 50	per cent 80
December 17–21, 1949	− 8.0	− 9.3	−12.0	0.40	0.34	0.25
December 13–16, 1949	− 8.0	−10.0	−13.6	0.40	0.32	0.21

Table I indicates that the scattered signal was 35 to 50 db stronger than the signal which would have been received by refraction through a standard atmosphere.

The conclusion from these data was that the weaker signals from the distance of 146 and 182 miles came primarily from higher elevation angles and may be attributed to scattering, while the stronger signals were an example of the well-known phenomenon of super-refraction.

For the San Antonio station at 73 miles, the relative signal received on the two antennas indicated that the signal was arriving horizontally. The observed signal was of the same order of magnitude as the refracted signal calculated for a standard atmosphere. The signal from scattering then appeared to be superimposed on the normal refracted signal, but did not change its median value appreciably. For periods of slight fading, the statistical distribution of the signal was Gaussian. However, for periods of deep and rapid fading, the Rayleigh distribution was approximated.

B. Directive Antenna Pointed Vertically

In order to further check the arrival of the FM signals from overhead, the directive antenna was pointed with the minimum in its antenna pattern toward the transmitter and with its main lobe vertical. Simultaneous measurements were made of the signal received from Dallas by the directive antenna in this position and by the dipole antenna. The signal ratio of the antennas was again observed to be a function of the signal strength as shown in Fig. 4. These data were for March 15 and 16, 1950, when the synoptic conditions over Texas were fairly uniform and constant. The ratio of the signal received by the dipole antenna to that received by the

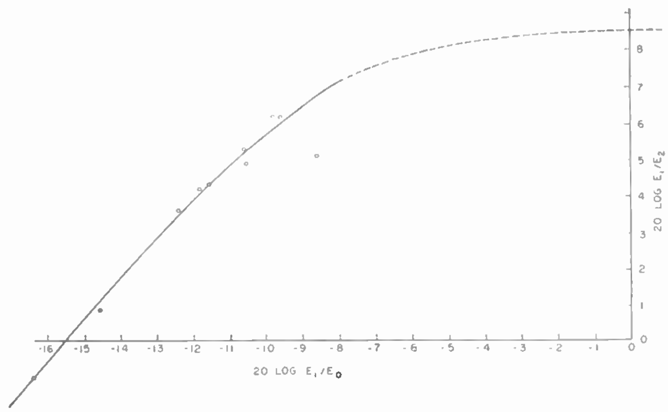


Fig. 4—Comparison of dipole antenna with directive antenna pointed vertically. Station WFAA—FM. Distance, 182 miles. March 15–16, 1950.
E_1 = open-circuit voltage, dipole antenna
E_2 = open-circuit voltage, directive antenna
$E_0 = 4.8(10^{-6})$ volts.

directive antenna pointed vertically was plotted as the ordinate while the signal received by the dipole was plotted as the abscissa. The dotted extension of the curve intersects the ordinate axis at 8.5 db, which was the measured ratio of the response of the two antennas to a nearby source from which waves arrived horizontally. For weaker signals, the antenna pointed vertically shows an improvement in signal relative to the dipole. This again indicated that stronger signals which come from the horizon were due to superrefraction while the weaker signals which came in part from overhead were due to scattering.

VI. Effect of Polarization on Reception of FM Signals

The power density at a given point is seen from (3) to be proportional to the square of the sine of the angle between the electric field vector and the direction from the scattering region to the receiver. This factor will cause the scattered signal arriving from overhead to be less for vertical polarization than for horizontal polarization. In addition, a vertically oriented antenna will have a reception pattern which will discriminate against signals arriving from overhead. These factors indicate that the scattered signal for vertical polarization should drop off more rapidly with distance than that for a horizontally polarized antenna when l/λ is small.

Two tests of the effect of polarization were made for a frequency of 100 mc. These tests are not conclusive but they do show an indication that the vertically polarized signals are attenuated faster at distances beyond the optical horizon than are the horizontally polarized signals.

A. Distance Run on Incidental Vertical Polarization

The signal from station KMHB in Belton, Texas, was observed at a range of distances up to 90 miles by both horizontal and vertical dipole antennas. The antennas were mounted above a small trailer pulled by a truck in which were located the two receivers. Sufficient vertical polarization accrued from the tower to make possible measurement of vertically polarized signals. The KMHB antenna is 300 feet above immediately surrounding ground, but because of its geographical location, the equivalent height over the road followed was very much less.

In order to be able to reduce the results to a comparable radiated power basis, measurements were started within the line of sight of the transmitter. For

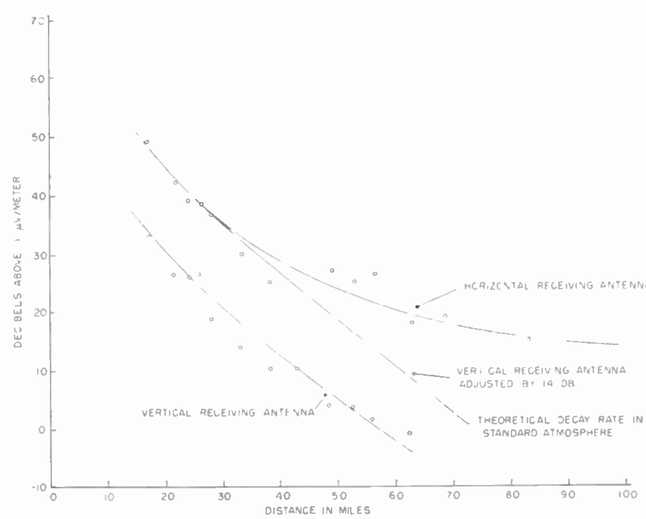


Fig. 5—Comparison of signal attenuation at 97.1 mc for horizontal and vertical polarization. Transmitting source: KMHB, Belton, Texas. Antenna height, 300 feet. Effective radiated power, 12 kw. Receiving antenna height, 10 feet.

distances less than 30 miles, the average signal received by the horizontal antenna was consistently 14 db stronger than the signal received by the vertical antenna. The signal strengths as measured are shown in Fig. 5. In addition, the signal curve for vertical polarization has been redrawn with an increase of 14 db to superimpose it on the curve for horizontal polarization at distances less than 30 miles. The vertical polarization curve was terminated at the distance at which the difference in the signal levels of the two systems approached the known discrimination level of 24 db. From this curve, it is seen that the signal of the vertical antenna decreases at the same rate as the standard refracted wave curve while the signal received by the horizontal antenna decreases more slowly than the theoretical curve. Cross feed between the systems would tend to obscure this difference rather than accentuate it.

B. Distance Run on Antenna Polarized at 45°

As a further test of the polarization, an antenna was polarized at 45° and 100-mc power was radiated from it. The distance run was then repeated. Because of the reduced antenna gain and since the antenna in this case was only 60 feet above surrounding ground, it was not possible to carry this test beyond 30 miles. It was noted that for distances less than 23 miles, a single curve represented both sets of data, while for greater distances, the horizontally polarized component of the signal was somewhat greater than the vertically polarized component.

VII. FM Signals as a Function of Angle of Elevation

In order to further test the variation of scattered signal with elevation angle, a special antenna was constructed which could be tilted through 180°. This antenna was a V-type reflector with planes 20 feet square separated at an angle of 45°. The measured antenna pattern for 100 mc is shown by curve (2) in Fig. 6.

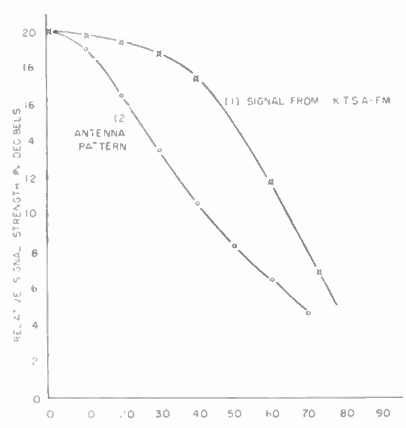


Fig. 6—FM signal versus elevation angle. Distance, 73 miles. Frequency, 101.5 mc. Polarization, horizontal.

Using this antenna at various tilt angles, field-strength recordings were made on the signal received from radio station KTSA-FM in San Antonio, Texas, at a distance of 73 miles. The frequency of this station is 101.5 mc. A 3-minute median level was determined at each tilt angle and the results are plotted as curve (1) in Fig. 6. The effect of diurnal changes in field strength taking place during the experiment was eliminated by means of an auxiliary monitoring receiver using a simple dipole antenna situated near the larger receiving antenna.

It is seen that when scanning the sky for the 100-mc signal, a much flatter curve was obtained than when scanning a nearby point source. This indicates that the radiation is arriving over a range of elevation angles considerably above the horizon.

VIII. Microwave Signals as a Function of Angle of Elevation

It is indicated in Fig. 2 that at wavelengths of a few centimeters and for a scale of turbulence of the order of a meter the energy scattered by the atmosphere will be confined within a few degrees of the horizontal direction. A field test at a wavelength of 3.2 cm was made in the vicinity of Austin. The 20-foot parabola used by the Bell Telephone Laboratories for their original 3-cm angle-of-arrival studies[5] was used for scanning. The center of this parabola was approximately 15 feet above ground, and the transmitting antenna 37 miles away was 12 feet above ground. A pulse signal was used with a peak power of 19 kw. Polarization was horizontal.

Records of the signal received were made in 0.1° steps, and the statistical time distribution of the signal

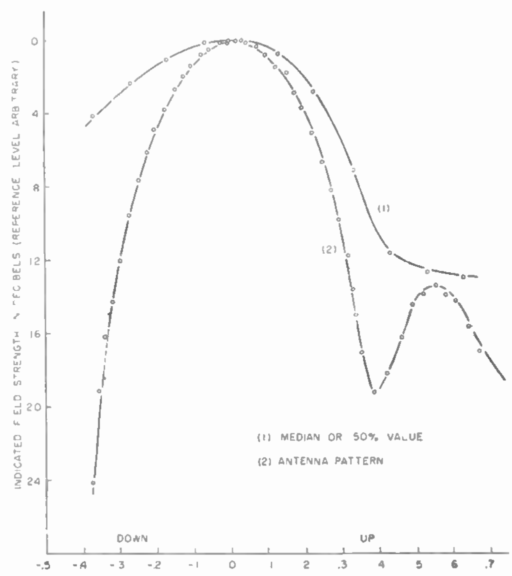


Fig. 7—Microwave signal versus antenna tilt angle. Transmitting antenna: height above ground, 12 feet; aperture, 24λ high×16λ wide. Receiving antenna: height above ground, 15 feet; aperture, 190λ high×5λ wide. Wavelength, 3.2 cm. Polarization, horizontal. Path length, 37 miles.

[5] W. M. Sharpless, "Measurement of the angle of arrival of microwaves," Proc. I.R.E., vol. 34, pp. 837–845; November, 1946.

determined at each angle. From these distributions, the median value was established and this is plotted on Fig. 7 as a function of the angle of tilt. The antenna pattern is also shown in Fig. 7 for comparison. As in the case of the 100-mc signal, the response distribution is seen to be considerably broader than the antenna pattern.

IX. Conclusions

A. Comparison of the reception at large distances with a dipole and a more directional antenna at a frequency of 100 mc indicated that:

1. During periods of stronger signals, the apparent source was near the horizon.
2. During periods of weaker signals the apparent source was more diffuse, and considerable radiation was received at higher elevation angles.

B. Distance runs on a 100-mc source indicated an attenuation with distance as given by the refracted wave theory for vertical polarization, but a smaller attenuation for horizontal polarization.

C. When a 100-mc source was scanned in a vertical plane, a distinct broadening of the antenna response was noted. When a 3.2-cm source was scanned, a similar effect was noted which, however, was confined to a few degrees above the horizon.

D. The collection of results of these experiments on tropospheric scattering appears to substantiate the theory of scattering discussed earlier in this paper.

The Component Theory of Calculating Radio Spectra with Special Reference to Frequency Modulation*

N. L. HARVEY†, SENIOR MEMBER, IRE, M. LEIFER†, SENIOR MEMBER, IRE, AND N. MARCHAND†, SENIOR MEMBER, IRE

Summary—**A new theory for the resolution of a modulated signal into its radio-frequency spectrum is presented. It is shown that this theory is mathematically equivalent to previous methods of analysis. The component theory introduces physical concepts which give an understanding of the process of formation of sidebands. Examples are worked using an analytical method of analysis. The theory also naturally leads to a graphical method which is applicable without harmonic analysis of the modulation. The concepts apply without modification to any type of modulation although special emphasis is given to frequency modulation.**

Introduction

A MAJOR difficulty in the analysis of modulation systems has been the transition from the instantaneous frequency concept to the spectrum solution. The "component" method of analysis, described in this paper, supplies a transition and is shown to give results equivalent to those obtained by the usual methods. It is pointed out that the mathematical procedure for obtaining the spectrum components provides a definition of the carrier frequency of an FM wave.

The Component Method of Analysis

It is of interest to first consider the situation of Fig. 1 which represents a filter with a passband outside the instantaneous frequency range of an FM signal. That energy can be transmitted by the filter is difficult to perceive. Equally difficult to understand is how no energy would be transmitted by a filter which passes

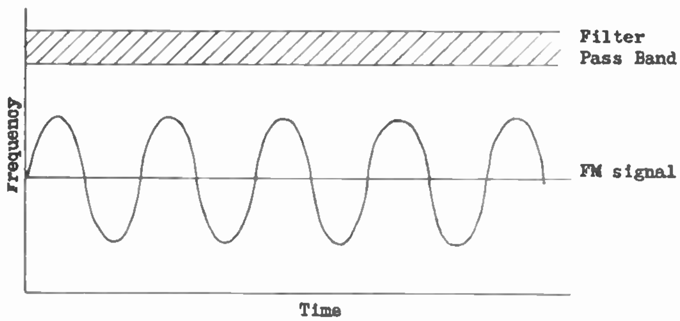


Fig. 1—Example of an FM signal applied to a narrow-band filter.

part of the instantaneous frequency range of the signal. The trouble lies in considering the usual filter steady-state frequency response as being applicable to this situation. Instead, we must treat the experiment from a pseudo-transient point of view. The varying frequency voltage at any instant will cause a current to flow in the filter network, or it will affect the current which is flowing as a result of the voltage applied during the preceding small time interval. If the voltage at any time has a component in phase with the current existing in the filter, then energy will be added to the filter. If the voltage has a component out of phase, energy will be abstracted from the filter. The net energy transfer from the applied signal to the filter is obtained by averaging over a time equal to one period of this process. The existence of a spectrum component is dependent therefore on the nonvanishing of the average energy transfer. This approach is applicable to any type of modulated signal.

* Decimal classification: R148.2. Original manuscript received by the Institute, April 25, 1950; revised manuscript received, October 18, 1950.

† Sylvania Electric Products Inc., Physics Laboratories, Bayside, N. Y.

To develop these ideas, consider a graphical representation of an FM signal as in Fig. 2. The length of the directed line segment is equal to the signal amplitude and the speed of rotation of the directed line segment about the origin is equal to the instantaneous angular frequency of the signal in radians per second. Such a directed line segment is called a phasor. In general, the phase angle of an FM signal is always *monotonically* increasing, but at certain portions of the modulation cycle the rate of increase is greater than at other portions. If the plane of the figure upon which the phasor is drawn is considered to be rotating in a positive sense (i.e., counterclockwise) at the carrier angular frequency (carrying the observer with it), then the phasor representing the FM signal will be seen to swing alternately clockwise and counterclockwise, where the amplitude of the angular swing may include many rotations. Thus if the carrier signal is considered without angular modulation, its speed of rotation is equal to that of the plane upon which it is drawn, and the phasor remains stationary at point A. When angular modulation is applied, the phasor may increase its angular velocity corresponding to the increase in frequency of the signal and will move to B, then reverse its motion to C as the signal frequency decreases, then repeat the cycle. The phasor amplitude is constant, indicating the absence of amplitude modulation.

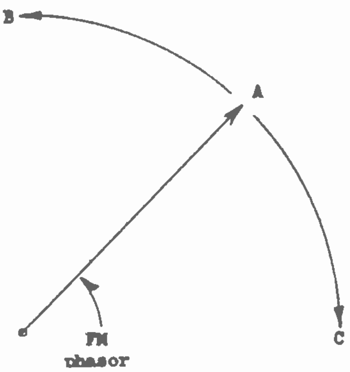


Fig. 2—Graphical representation of an FM signal.

If the speed of rotation of the plane were other than the carrier angular velocity, the signal phasor would have a continuous rotational velocity in one direction and superimposed upon this motion would be the alternating motion due to the frequency or phase modulation. It is seen that this picture presents an approach to the definition of the carrier frequency as that frequency with which it is necessary to consider the plane rotating so that the FM phasor possesses only alternating motion. In other words, the positive and negative phase angle swings shall be equal, corresponding to the mathematical definition of "equal areas."

In the following, there will be no sharp distinction between the abstract picture of the spectrum components of a signal and the experimental procedure which produces these spectrum frequencies, as for example, by means of a narrow bandpass filter. Thus, consider a filter of resonant angular frequency Ω excited by a frequency-modulated signal voltage. Upon the plane of Fig. 3 represent the FM signal and the response voltage at the natural frequency of the filter by phasors. If the plane is rotating at a frequency Ω, then the phasor representing the filter response voltage is stationary, while the FM signal phasor will be in motion as a result of its modulation and since in general its carrier frequency is different from Ω.

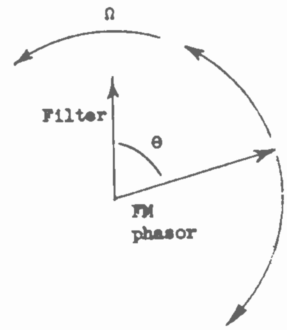


Fig. 3—Graphical representation of signal and filter voltages.

Assume that the component of the FM signal phasor along the direction of the phasor representing the filter response represents the effective voltage to which the filter responds. A current will flow as a result which is proportional to the magnitude of this voltage component, i.e., proportional to the magnitude of the FM signal phasor multiplied by cosine θ, where θ is the instantaneous angle between the signal phasor and the filter response phasor. It is possible to discuss this assumption in such a way as to make it seem quite plausible, but in the final analysis the validity of such an assumption must be justified by the validity of the results which follow. The results will be seen to be correct in all cases. For values of θ less than 90 degrees, the component of the signal voltage along the direction of the voltage response phasor corresponds to energy transfer to the filter. The cosine θ is seen to behave like a power factor. Considering the steady state which will be attained, it appears that the phase of the current in the filter will adjust itself in such a way as to abstract a maximum of energy from the forcing voltage. A useful analogy is to consider how a child's swing adjusts its phase relative to a periodic forcing push so as to reach a maximum height. Hence the phasor representing the filter response may differ from the indicated position in Fig. 3 by some constant angle. In order to locate the exact position of the filter response phasor with respect to the FM signal phasor, the filter voltage can be represented by a pair of orthogonal phasors both rotating at the same frequency Ω, as in Fig. 4. The response to the forcing voltage is obtained by resolving the FM signal phasor upon the two orthogonal filter phasors. The vectorial resultant of the steady-state responses along these two directions gives the resultant response. Thus, the time average of the components along the two orthogonal filter response phasors is calculated by integration over one period of the modulation and the absolute value of the vectorial resultant is the filter response. This procedure will be illustrated below by a series of examples.

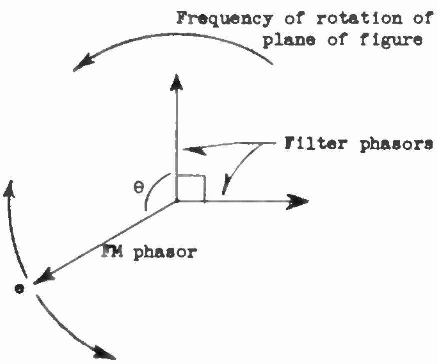


Fig. 4—General representation of signal and filter voltages.

It seems clear at this point that any instantaneous measurement of the filter response (and here a number of pertinent questions as to the meaning of instantaneous measurement are neglected) would indicate a current existing in any filter of any resonant frequency whatsoever. From the view presented here, the forcing voltage component is at one instant in phase with the current, and at a later time opposed in phase to the current. This corresponds to energy alternately being put into and then removed from the filter. In any actual spectrum analysis measurements, the time average of the response of a filter is measured. It must therefore be shown that the time averages of the signal-to-filter energy transfers, computed as explained above, will vanish when the filter does not pass any frequencies corresponding to spectrum lines, and will have the correct nonzero magnitude when the filter passband contains a spectrum component. Several simple examples will be demonstrated, and the general FM spectrum derived.

Example 1: Consider an FM signal with simple sinusoidal modulation, and let this signal be applied to a filter containing the carrier frequency ω_c in its passband.

The input voltage is given by $e = E\cos(\omega_c t + \phi + \beta\sin\omega_m t)$ where

ω_c = the carrier angular frequency
ω_m = the modulation angular frequency
β = the modulation index
ϕ = the rf phase angle.

For the phasor diagram (Fig. 4), let the plane of the figure rotate at frequency ω_c. The phasors representing the filter are stationary and the angle between the signal phasor and one of the filter phasors as a reference is

$$\theta = \phi + \beta \sin \omega_m t.$$

The components of the signal upon the phasors representing the filter response are

$$E\cos(\phi + \beta\sin\omega_m t)$$

and

$$E\sin(\phi + \beta\sin\omega_m t).$$

The time averages are obtained by integration over one cycle of the modulation, giving

$$1/2\pi \int_0^{2\pi} E\cos(\phi + \beta\sin\omega_m t)d(\omega_m t)$$

and

$$1/2\pi \int_0^{2\pi} E\sin(\phi + \beta\sin\omega_m t)d(\omega_m t).$$

In order to perform the integration, use is made of the identities (found in any book on Bessel functions)

$$\cos(\beta\sin\omega_m t) = J_0(\beta) + 2\sum_{n=1}^{\infty} J_{2n}(\beta)\cos 2\omega_m t$$

and

$$\sin(\beta\sin\omega_m t) = 2\sum_{n=1}^{\infty} J_{2n-1}(\beta)\sin(2n-1)\omega_m t.$$

The time average of either the cosine or sine function vanishes, giving as the results of the integrations

$$EJ_0(\beta)\cos\phi \quad \text{and} \quad EJ_0(\beta)\sin\phi.$$

The vector sum of these components has an amplitude of $EJ_0(\beta)$, which is the correct value as derived by the usual mathematical procedure.

Example 2: The amplitude of a spectrum component other than the carrier frequency will now be found.

Let the filter of Example 1 now contain sideband frequency $\omega_c + k\omega_m$ in its pass-band.

If the plane of Fig. 4 is rotating with angular velocity $\omega_c + k\omega_m$, then the phasors representing the filter response are again stationary and the angle θ between the reference phasor and the FM signal phasor is

$$\theta = k\omega_m t - \phi - \beta\sin\omega_m t,$$

obtained by subtraction of the phase angle of the rotating plane from the phase angle of the FM signal. The components of e upon the two phasors are $E\cos\theta$ and $E\sin\theta$.

The time averages are given by the integrals (expanding the integrands)

$$E/2\pi \int_0^{2\pi} [\cos k\omega_m t\cos(\phi + \beta\sin\omega_m t) + \sin(k\omega_m t)\sin(\phi + \beta\sin\omega_m t)]d(\omega_m t)$$

and

$$E/2\pi \int_0^{2\pi} [\sin k\omega_m t\cos(\phi + \beta\sin\omega_m t) - \cos(k\omega_m t)\sin(\phi + \beta\sin\omega_m t)]d(\omega_m t).$$

Using the identities of Example 1 and the orthogonality properties of the trigonometric functions, these time averages reduce to

$$EJ_k(\beta)\cos\phi \quad \text{and} \quad EJ_k(\beta)\sin\phi.$$

The amplitude of the resultant is therefore $EJ_k(\beta)$.

It will now be shown that a filter not containing a spectrum component frequency in its pass-band has a zero average response, even though the filter may pass frequencies within the instantaneous frequency range of the FM signal.

Example 3: Let the filter of Example 1 not contain any sideband frequency $\omega_c \pm k\omega_m$. Thus, let the plane of Fig. 4 rotate with angular velocity $\omega_c+\omega_d$ where ω_d is *not* equal to $\pm k\omega_m$.

The angle θ is equal to $(\omega_d t-\beta\sin\omega_m t)$ and the components of e upon the reference phasors are $E\cos\theta$ and $E\sin\theta$. The time averages are obtained by integration over a period T of the integrand, giving upon expansion of the integrands:

$$\frac{E}{T}\int_0^T \{\cos\omega_d t\}\{J_0(\beta)+2\sum J_{2n}(\beta)\cos\ n\omega_m t\} + \{\sin\omega_d t\}\{2\sum J_{2n-1}(\beta)\sin(2n-1)\omega_m t\}\,dt$$

and

$$\frac{E}{T}\int_0^T \{\sin\omega_d t\}\{J_0(\beta)+2\sum J_{2n}(\beta)\cos 2\ n\omega_m t\} - \{\cos\omega_d t\}\{2\sum J_{2n-1}(\beta)\sin(2n-1)\omega_m t\}\,dt.$$

Since ω_d is assumed not equal to any multiple of ω_m, each term of these integrations vanishes by reason of the orthogonality of the trigonometric functions, or the time-averaged response is zero.

The general expression for any spectrum component of an angularly modulated signal can be similarly derived, as is shown in the next example.

Example 4: Let the general FM signal be $e=E\cos[\omega_c t+\beta\int_0^{\omega_m t}F(\omega_m t)]d(\omega_m t)$, where β is the usual modulation index and $F(\omega_m t)$ is the function describing the way in which the frequency varies.

For simplicity, write $e=E\cos[\omega_c t+g(t)]$ where $g(t)$ describes the angular modulation.

Consider a filter which accepts frequency $\omega_c+\omega_d$ in its pass-band.

Then, just as in Example 3, the angle θ between e and the reference phasor is $\theta=\omega_d t-g(t)$ and the components of the signal phasor upon the phasors representing the filter response are $E\cos\theta$ and $E\sin\theta$.

The time averages over one period of the angle θ are:

$$E/T\int_0^T \cos\{\omega_d t-g(t)\}\,dt \quad \text{and} \quad E/T\int_0^T \sin\{\omega_d t-g(t)\}\,dt.$$

Inasmuch as $g(t)$ is a periodic function describing the angular modulation, we may expand $\cos g(t)$ and $\sin g(t)$ into Fourier series giving:

$$\cos g(t) = \sum_{n=0}^{\infty}(b_n\cos n\omega_m t + a_n\sin n\omega_m t)$$

$$\sin g(t) = \sum_{n=0}^{\infty}(d_n\cos n\omega_m t + c_n\sin n\omega_m t).$$

Expanding the integrands and using these Fourier expansions, we note from the orthogonality of the trigonometric functions that the integrals are zero except when $\omega_d=\pm k\omega_m$ with $k=0, 1, 2\cdots$.

In the latter case, the time averages become

$$E/2(b_k \pm c_k) \quad \text{and} \quad E/2(a_k \mp d_k),$$

whereupon the amplitude of the component of frequency $\omega_c \pm k\omega_m$ is

$$E/2[(b_k \pm c_k)^2 + (a_k \mp d_k)^2]^{1/2}.$$

This result is a perfectly general one and requires only a Fourier analysis of the cosine and sine of the modulation function.

The pair of orthogonal phasors used to represent the filter response may be replaced by a single phasor at any arbitrary angle in the diagrams as shown in Fig. 5. The time average of the component of the FM signal phasor upon this single filter phasor is obtained in the usual fashion, and then is maximized with respect to the arbitrary angle. The correct spectrum amplitudes are obtained in this way verifying the physical concept of the filter response adjusting its phase in such a way as to extract a maximum of energy from the signal. For illustration, Example 1 is repeated by this method. It is required to obtain the carrier amplitude of a sinusoidally modulated FM signal.

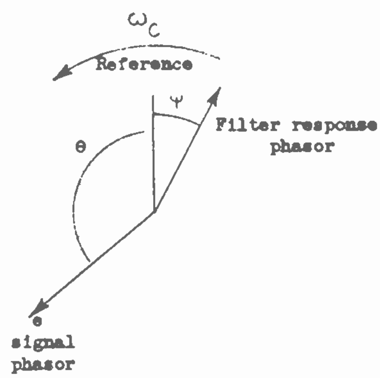


Fig. 5—Representation of filter voltage by a single phasor.

Suppose the phasor representing the filter response is at an angle Ψ. Let the FM signal be $e=E\cos(\omega_c t+\phi+\beta\sin\omega_m t)$, and let the plane of the figure be rotating at angular velocity ω_c. Then the component of e upon the filter phasor is $E\cos(\phi+\Psi)$, which gives an integrated time average of $EJ_0(\beta)(\cos\phi\cos\Psi-\sin\phi\sin\Psi)$. This is maximized with respect to Ψ by setting the derivative equal to zero, giving $\tan\Psi=-\tan\phi$, whereupon the amplitude reduces to $EJ_0(\beta)$.

The component view may be extended without difficulty to amplitude-modulated signals, and to signals which are modulated both in amplitude and angle. For example, Fig. 6 shows an amplitude-modulated signal phasor of angular frequency ω_c and two phasors for the filter response at this frequency. Both the signal phasor and the filter phasors are stationary on the plane rotating at angular velocity ω_c. The components of the AM signal phasor are obtained by projection upon the axes, and the time average gives the usual result.

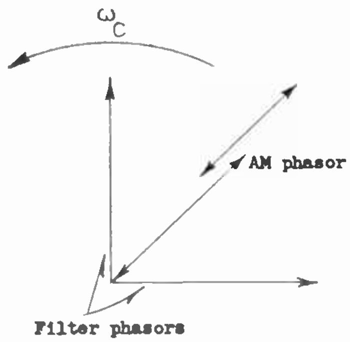


Fig. 6—Representation of an AM signal.

The component method lends itself to graphical analysis of the spectrum of modulated signals, and in fact a good deal of use has been made of this approach in treating certain problems where the modulation waveform has been given in graphical form. The projections of the signal phasor can be measured for any required number of positions of the phasor over a modulation cycle and the results averaged without recourse to integrations or Fourier analysis.

A simple example is presented in Fig. 7, where one seeks the amplitude of the carrier frequency component of a hypothetical signal which is modulated both in phase and amplitude, and where the modulation waveforms for one period are given in graphical form. Fig. 7(a) shows the phase variation of the signal and Fig. 7(b) shows the normalized amplitude variation. (For simplicity the periods have been chosen as equal although generally this will of course not be the case.) The time interval of one period is divided into any convenient number of parts, and for each value of time so chosen the amplitude and phase of the signal are replotted as a phasor, as shown in Fig. 7(c) for the point marked E. The components of this phasor upon the axes are easily read and tabulated. The numerical averages of the tabulated components are obtained, squared, added and the square root of the sum of the squares of the averages gives the carrier amplitude.

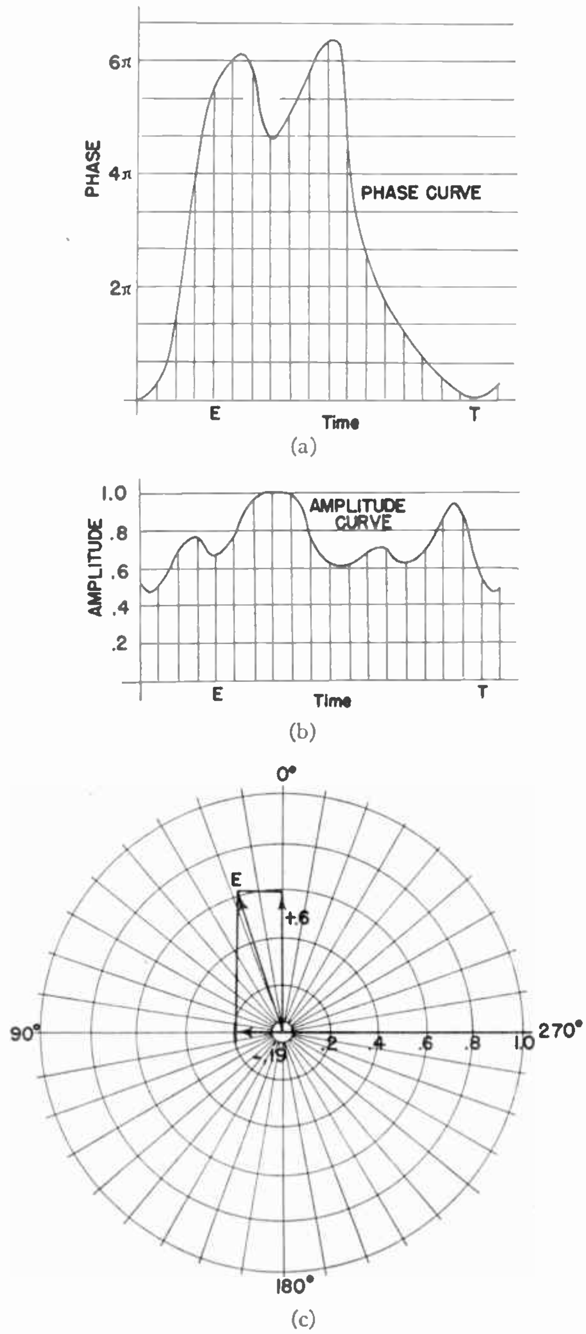


Fig. 7—Graphical procedure for calculation of the radio spectrum of an amplitude- and frequency-modulated signal.

Conclusion

In the 1922 Proceedings of the I.R.E., Carson[1] first presented the spectrum solution for a sine wave angularly-modulated signal. He then stated that, "The foregoing solutions, though unquestionably mathematically correct, are somewhat difficult to reconcile with our physical intuitions and our physical concepts of such 'variable frequency' mechanisms as for example the siren."

It is just this difficulty which the component theory has attempted to resolve. In brief review, this theory considers the time average of the response of a band selective filter to a modulated signal. The existence of a spectrum component is indicated by the presence of a finite output signal caused by a net energy transfer from the signal to the selective network. This approach yields results identical with those given by the usual mathematical techniques. Its usefulness lies not in its verification of the mathematics, but in furnishing concepts which aid in attacking new problems and which treat all types of modulation from the same viewpoint.

[1] John R. Carson, "Notes on the theory of modulation," Proc. I.R.E., vol. 22, pp. 57–64; February, 1922.

Acknowledgment

The authors wish to express their appreciation for the many helpful suggestions and criticisms contributed by the Circuits Section of Sylvania Physics Laboratories.

Methods of Measuring Adjacent-Band Radiation from Radio Transmitters*

NEAN LUND†, ASSOCIATE, IRE

Summary—A review of three possible methods of measuring or estimating adjacent-band radiation characteristics of a radio transmitter is given. These three methods differ in the type of signal applied to the transmitter and may be termed the two-tone, normal signal, and thermal noise methods. Measurements on a multichannel single-sideband transmitter using each of these methods are presented to show that there is a good correlation between the normal signal and thermal noise methods.

An empirical method for calculating the slope of the adjacent-band radiation as a function of frequency from the measured two-tone distortion values is given, and the measured and calculated slopes are shown to be in fairly good agreement.

Introduction

GENERALLY, a radio transmitter has some radiation outside its necessary transmission band, and this radiation may be sufficient to interfere with services occupying other channels. This radiation includes many modulation products generated because of nonlinearity in the transmitter circuits. Usually, the radiation which appears as harmonics of the modulated frequencies can be reduced to satisfactory values by adding selective circuits. Similarly, the very-low-frequency modulation products are sufficiently suppressed by the normal selectivity of the system. However, the radiation which appears just outside the transmitted band cannot be easily reduced by using ordinarily selective circuits.[1] In this paper we will be concerned only with the radiation which falls just outside the transmitted band, and our purpose will be to describe methods of measuring this adjacent-band radiation and to give some results obtained when using these methods on a radio-telephone transmitter.

Three possible methods of measuring or estimating adjacent-band radiation will be described. These are:

1. Measure certain distortion products when two tones of equal amplitude are modulating the transmitter.

2. Measure the radiation in the regions adjacent to the transmitted band when the normal modulating signals are applied to the transmitter.

3. Measure the radiation in the regions adjacent to the transmitted band when thermal noise of equivalent bandwidth and amplitude is substituted for the normal modulating signals.

These three methods of measuring adjacent-band radiation have been applied to a newly designed Western Electric Type LD-T2 single-sideband reduced-carrier transmitter. This transmitter normally operates in the frequency range of 4 to 23 mc and is capable of carrying four telephone channels when used with the proper terminal equipment. The nominal over-all bandwidth of the transmitter is 12 kc consisting of two nominal 6-kc channel groups. These two channel groups are located on either side of the reduced carrier. The reduced-carrier amplitude is normally transmitted 26 db below the rated envelope peak power of 4 kilowatts, and is placed at midband frequency. This transmitter is designed primarily for telephone service, and therefore we are interested in the adjacent-band radiation that results when four talkers are loading the transmitter.

Two-Tone Distortion Measurements

The first or two-tone method for estimating adjacent-band radiation may be carried out by applying two tones of equal amplitude to the transmitter and measuring with a selective analyzer the amplitude of the modulation products which results. A block diagram of the circuit used in making the two-tone tests is shown in Fig. 1. Distortion products of any order of modulation may be measured by using this method. However, in the study of distortion products which lie just outside the transmitted band only odd-order products need be considered.

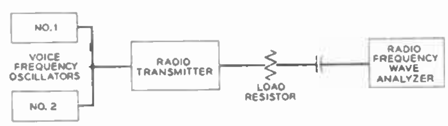


Fig. 1—Block diagram of circuit used in making two-tone distortion measurements.

The results of making the usual two-tone distortion measurements on the LD-T2 transmitter are shown in Fig. 2. The solid curves show the measured ratio of the

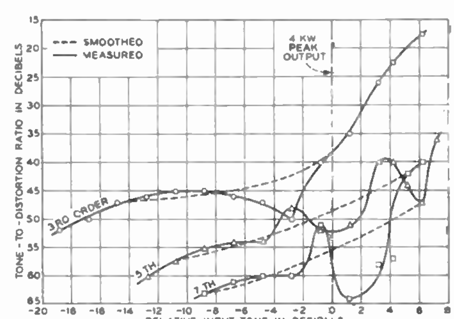


Fig. 2—Tone-to-distortion ratios plotted against input tone amplitude for the LD-T2 radio transmitter. The solid curves show measured values and the dotted curves show smoothed values.

* Decimal classification: R254.1×R148.13. Original manuscript received by the Institute, June 2, 1950; revised manuscript received, September 15, 1950.

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

[1] I. J. Karr, "Some notes on adjacent channel interference," Proc. I.R.E., vol. 22, pp. 295–313; March, 1934.

power in one of the two output tones to the power in a single distortion product as a function of the transmitter input tone amplitude. Data are plotted for third-, fifth-, and seventh-order products. When the transmitter is operating at its envelope peak-power rating of 4 kw, these curves show that the third-order distortion ratio is 38 db, the fifth-order distortion is 52 db, and the seventh-order distortion is 59 db. Some very general conclusions might be drawn from these two-tone curves. Since the distortion decreases with the order of modulation product, it might be expected that the adjacent-band radiation would decrease with frequency from the center of the transmitted band. Also, since the average amplitude of the distortion decreases as the signal input decreases, it would be expected that the adjacent-band radiation would decrease as the signal input is decreased. These conclusions are qualitative and of small value in determining the actual amplitude of adjacent-band radiation. However, such distortion measurements on one transmitter can be compared to similar measurements on another transmitter to obtain a reasonably accurate idea of the relative adjacent-band radiation performance.

Measurements Made with Normal Signal Loading

In order to carry out the second or normal signal method it was necessary to provide four sources of speech signals which could be used to modulate all four channels. Also, in order to be able to use these four speech sources it was necessary to have channel shifter equipment which allowed two 2.5-kc speech channels to be placed in the channel group frequency range from 100 to 6,000 cycles per second. A block diagram of the circuit used in the normal signal loading tests is shown in Fig. 3. Since the transmitter is to be used in two-way

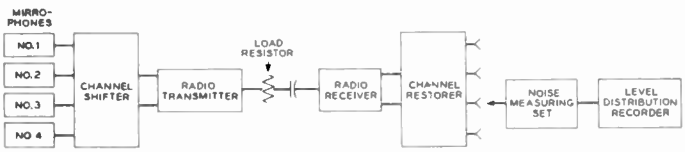


Fig. 3—Block diagram of circuit used in making the normal signal loading tests.

telephone communication, a subscriber would be expected to be talking on the circuit about 50 per cent of the time his call is in progress. Therefore, to represent this condition, recordings were made of conversational type speech, that is, the speech recorded was one side of a two-way conversation. Each record was made from a different conversation. These records were made on magnetic tape recorders which were capable of continuously playing a one-minute recording.

The adjacent-band radiation was measured with the Western Electric type LD-R1 single-sideband receiver, tuned to definite spot frequencies outside the band of the transmitter. Measurements were made at 3-kc intervals out to 40 kc from the transmitted carrier using a receiving bandwidth of 2.5 kc.

In order to actually measure fluctuating adjacent-band radiation, it is necessary to use some instrument which is capable of giving readings which can be interpreted. This is particularly important when measuring widely fluctuating modulation noise amplitudes such as are produced by speech modulation. An instrument which is quite satisfactory for this purpose is a Level Distribution Recorder.

The results of these adjacent-band radiation measurements have been plotted as ratios of the speech amplitude in one channel to the adjacent-band radiation amplitude in a band of the same width. These results are shown by the dashed curves plotted in Fig. 4 for three different input volumes which correspond to normal input, 5 vu above and 5 vu below normal. These three loading conditions give transmitter output volumes of 50, 55, and 59.5 vu per channel.

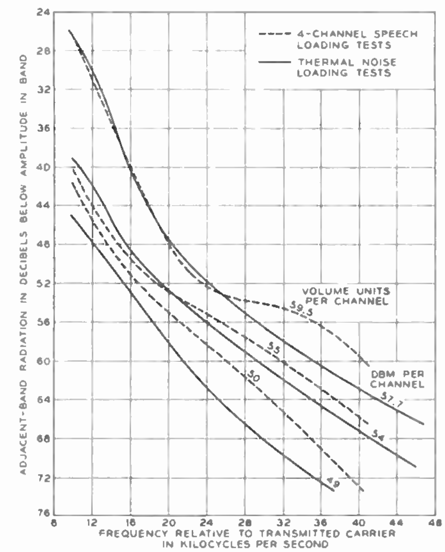


Fig. 4—Comparison of results of four-channel speech loading tests and thermal noise loading tests on the LD-T2 radio transmitter.

Measurements Made with Thermal Noise Loading

In the third or thermal noise method, it is necessary to have a source of thermal noise which has about the same band of frequencies as the speech band. This is necessary in order that the intermodulation between the noise frequencies will produce the same range of adjacent-band components as are produced by the speech modulation. Since the LD-T2 transmitter has two 5.9-kc channel groups, it was necessary to have two independent sources of thermal noise each producing a band from 100 to 6,000 cycles. Another requirement is that the noise energy per unit of bandwidth shall be constant with frequency. A convenient source of such thermal noise was the 5.9-kc channel group outputs of an LD-R1 single-sideband receiver. A block diagram of the circuit used in the thermal noise tests is shown in Fig. 5.

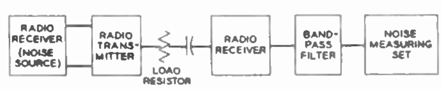


Fig. 5—Block diagram of circuit used in making the thermal noise loading tests.

The adjacent-band noise was measured by using a Western Electric type 2B Noise Measuring Set. There was very little fluctuation in the reading so that simply recording the reading of this instrument was sufficient. Filters were used in the analyzer receiver output to limit the received bandwidth to 2.5 kc. Measurements were made with the analyzer receiver tuned to the transmitted band so as to obtain the amplitude of the transmitted thermal noise in a 2.5-kc band. The measured adjacent-band modulation noise was then compared to this amplitude in order to obtain a ratio of single channel power to adjacent-band radiation power. The results are shown by the solid curves in Fig. 4. Measurements were made in 3-kc steps out to about 40 kc from the carrier for three values of noise input to the transmitter. As will be explained later, these three loading conditions approximate normal loading and conditions 5 db above and 5 db below normal. The adjacent-band radiation for about normal loading is shown by the curve for 54 dbm power output per channel.

Correlation of Adjacent-Band Radiation Measurements and Two-Tone Distortion Measurements

Since the adjacent-band radiation measurements require a fairly elaborate testing arrangement, it would be desirable to be able to predict the adjacent-band radiation for any transmitter on the basis of two-tone tests. Some calculations have been made for the LD-T2 transmitter to determine a correlation between two-tone distortion and adjacent-band radiation. These calculations have been made using the following assumptions:

1. It is assumed that a uniform 12-kc band of noise is applied to the transmitter amplifier and that this noise can be analyzed as being equal to a large number of equally spaced sine wave signals having equal amplitudes and random phase in the 12-kc band.
2. The intermodulation between these sine wave signals gives a distribution of modulation products which can be determined for each order of modulation product.
3. The contribution of each distribution to the total distribution is found by multiplying each distribution by the two-tone distortion ratio for that particular order of modulation product.
4. The two-tone distortion ratios used in step 3 are the values measured for an envelope peak tone input which is 6 db greater than the total average noise power input.
5. The modulation noise at any frequency can be found by summing up the contributions found in step 3 for each order of modulation product.
6. The addition of the distributions in step 5 is made on the basis of power addition of the individual distributions.

In order to carry out the procedure which has been outlined, it is necessary to determine the frequency distribution of modulation products for each order of modulation product. A method for determining a distribution of modulation products is to assume a definite number of modulating tones and calculate the number of modulation products which fall into a bandwidth equal to the spacing of the modulating tones. Such a process has been carried out by W. R. Young of the Bell Telephone Laboratories so as to determine formulas for third-order modulation products. A distribution of third-order products, calculated from third-order formulas, is shown in Fig. 6. The curve plotted in this figure shows the relative number of modulation products as a function of the relative frequency from the center of the modulating band. It has been assumed that this curve can be used to determine the distributions for higher orders of modulation, as well as for third-order, by stretching the abscissa values in proportion to the order of modulation. This is a simplifying assumption and no estimate of the error that results is given here.

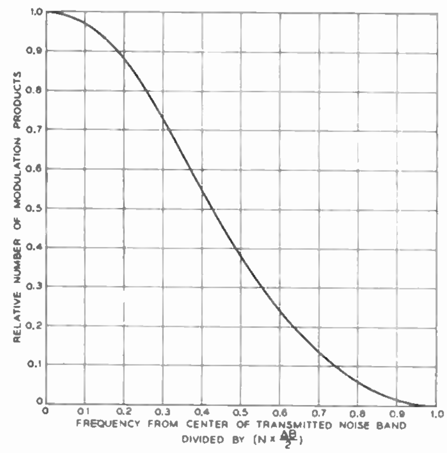


Fig. 6—Distribution of modulation products used to calculate adjacent-band radiation. N is the order of modulation product and B is the width of the transmitted noise band.

It is desired, then, to multiply the frequency distributions by the corresponding two-tone distortion ratio shown in Fig. 1. The data actually used in this figure were the smoothed or dashed curves shown. The reason for using the smoothed curves is to minimize the effects of the dips or cancellation points in the two-tone curves. The cancellation effect takes place over a small range of amplitudes which can be observed on a two-tone test. However, for noise inputs the amplitudes have a large range of values and the cancellation effect takes place over only a small portion of this range. For noise amplitudes slightly greater or less than the critical amplitude, the modulation products are much stronger so that products generated when the noise envelope is at the critical amplitude contribute very little to the total distortion.

The results of both measured and calculated adjacent-band radiation are shown in Fig. 7, and show that the calculated values are in fairly good agreement with the measured values for all three values of loading. It may be noted that the procedure for calculating the adjacent-band radiation has been developed through the use of empirical methods, and therefore it would be desirable to obtain some further checks by application of the method to other types of transmitters.

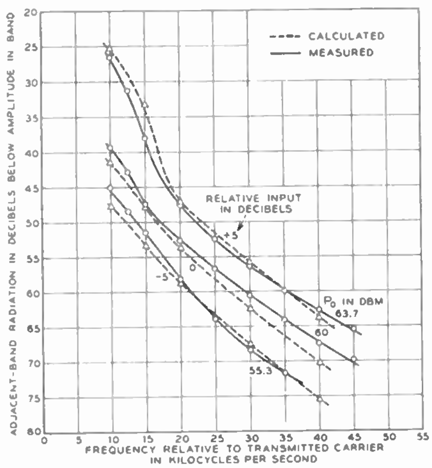


Fig. 7—Comparison of measured and calculated adjacent-band radiation for the LD-T2 radio transmitter. The solid curves show measured values for the thermal noise loading tests. The dashed curves show calculated values.

Comparison of Normal Signal and Thermal Noise Tests

A direct comparison of the measured results obtained by the speech and thermal noise methods is shown in Fig. 4. The comparison data show that in general the slopes of the adjacent-band radiation for both tests are the same, and that the absolute values of the adjacent-band radiation for both tests are about the same. In Fig. 4 the three curves for speech loading are shown in terms of vu output per channel and the three curves for thermal noise loading are shown in terms of dbm output per channel. These results indicate that on the average a speech loading which is 1 vu greater than the noise loading in dbm will produce about the same amount of adjacent-band radiation.

Conclusions

It is concluded that the thermal noise method of making measurements gives results which are essentially the same as those obtained with normal signal loading. The thermal noise method is much superior to the normal signal method in that it requires considerably less equipment for generating the modulating signals and for making the measurements. Also in the thermal noise method there is very little fluctuation in the amplitudes to be measured, and therefore the readings may be obtained quite rapidly and accurately.

Acknowledgment

The author wishes to express his indebtedness to the Overseas Radio Group at Deal, N. J., for their assistance in making the measurements, and to F. B. Llewellyn for his valuable suggestions in connection with the thermal noise method.

A Very-High-Frequency—Ultra-High-Frequency Tail-Cap Antenna*

LOUIS E. RABURN†, SENIOR MEMBER, IRE

Summary—**A description is given of a broad-band vhf-uhf antenna for a small, single-engine type of aircraft consisting of the vertical stabilizer that is tip-fed by the metal tail cap. The resulting antenna covers the frequency range of 100 to 400 mc, is vertically polarized, and has radiation patterns that are essentially omnidirectional. It is a zero-drag type of antenna that is very adaptable to aircraft production techniques.**

The tail cap, together with the rest of the vertical stabilizer, performs as a vertical sleeve-fed stub on an unsymmetrical ground plane. For this reason, the antenna radiation fields are not perfectly omnidirectional, but they are adequate for two-way communication in the frequency range 100 to 400 mc, and in this range the standing wave ratio on RG-8/U 50-ohm coaxial cable is less than 2 to 1.

The predicted radiation field-strength patterns for a specified transmitter power are calculated from measured radiation field patterns of the scale-model aircraft and full-scale standing-wave ratio measurements.

Introduction

The possibilities of exciting aircraft wings, fuselage, and empennage for use as low-drag or zero-drag antennas at frequencies above the medium frequency band has been investigated over a period of years by a number of people, both in this country and abroad.[1–6] An extensive study of tip-

* Decimal classification: R326.21×R326.7. Original manuscript received by the Institute, May 29, 1950; revised manuscript received, September 5, 1950.

† Electronics Research, Inc., Evansville, Ind.

[1] M. P. Hanson, "HF Collecting and Radiating Structure," U. S. Patent No. 2,044,779; June 23, 1936.

[2] S. Zisler, "Theory and Technique of Antennas, Part I," pp. 24–30. This is a German document which has been translated and declassified. It is available at the Central Air Documents Office, Wright Field, Ohio.

[3] W. A. Johnson, "Recent developments of aircraft communication aerials," *Jour. IEE*, Part IIIA, p. 452; 1947.

[4] F. H. Behrens, "Drag cut with plastic antenna housings," summarized by McLarren, *Aviation Week*, vol. 49, pp. 18–24; November 29, 1948.

[5] G. E. Beck, "Suppressed aircraft aerials," *Wireless World*, vol. 26, p. 468; December, 1949.

[6] J. V. N. Granger, "Shunt-excited flat-plate antennas with applications to aircraft structures," Proc. I.R.E., vol. 38, p. 280; March, 1950.

excited wings and stabilizers was sponsored several years ago by the Army Air Force and was performed at the Rocky Point Laboratories of RCA.[7]

This study showed that it was possible to excite a wing by insulating and feeding the outer one-fifth of its length. The resulting antenna bears some physical resemblance to a sleeve stub and subsequent work has confirmed the fact that most of the electrical characteristics are also similar.[8] It has been found that sleeve-type stubs with large-diameter elements of optimum proportions have very desirable impedance characteristics over broad frequency bands.[9]

When it was required to develop a zero-drag vertically polarized omnidirectional antenna for use with vhf and uhf communications equipment in a small, single-engine aircraft of relatively conventional type, the tip-fed vertical stabilizer configuration seemed most feasible. The development of this antenna will be described, and because the feeding tip is the vertical stabilizer cap, the antenna will be referred to as the tail-cap antenna throughout the rest of this paper.

Radiation Pattern Study

A tail-cap antenna in the vertical stabilizer of a conventional-type aircraft has similar electrical characteristics to a sleeve stub near the side of a horizontal ground plane. For this reason, it isn't considered possible to change the radiation patterns very much by changing the shape or size of the tail-cap alone. It is, therefore, desirable to perform a preliminary pattern study using scale-model techniques to determine whether or not the radiation characteristics of the tip-excited tail antenna would be acceptable before the full-scale prototype antenna is developed.

This preliminary pattern study was performed using the measuring range at the roof laboratory shown in Fig. 1. A one-eighth scale, precision model of the aircraft was made of wood and sprayed with copper to provide a highly conductive coating. A scale model of the capacity-hat tail-cap antenna with the proposed feeding and matching configuration was attached to the vertical stabilizer of the model in Fig. 1. The model antenna had an SWR of less than 4 to 1 throughout the model frequency range of 800 to 3,200 mc, and did not need any special matching system for the pattern measurements.

A complete set of principal plane patterns and cone-angle patterns was recorded for both E_θ (level-flight vertical polarization), and E_ϕ (level-flight horizontal polarization) at simulated frequencies of 130, 250, 300, and 362 mc. The E_θ and E_ϕ polarizations are illustrated in the aircraft co-ordinate system of Fig. 2(a). The set of principal-plane patterns for 250 mc is shown in Fig. 3 to best indicate the general shape of the patterns for both E_θ and E_ϕ polarization. The cone-angle patterns, plotted on a basis of relative directivity only, were then integrated and computations were made[10] to convert the relative patterns into field-strength contour patterns at a distance of one mile from the aircraft for a transmitting power of one watt.

The complete set of patterns is very bulky, and the important data within the cone angles of $+30°$ and $-30°$ can be consolidated in the more convenient form

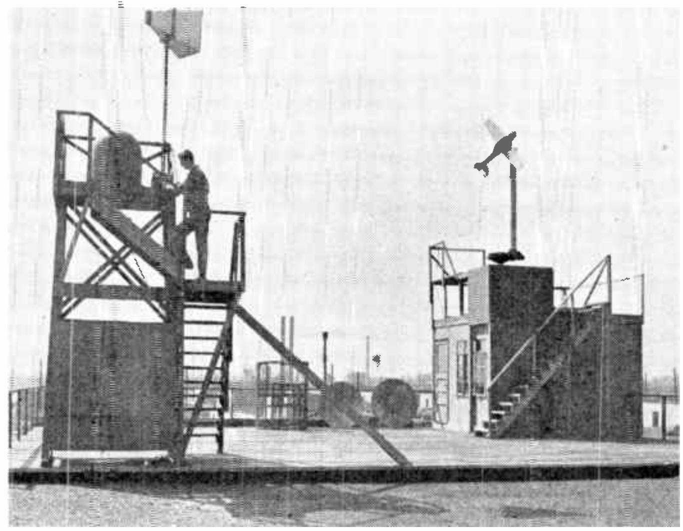


Fig. 1—Scale-model aircraft and radiation pattern-measuring range.

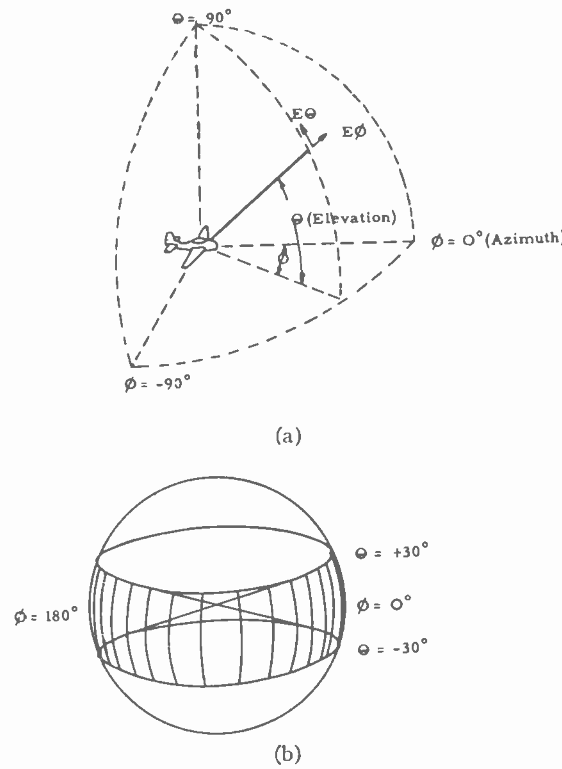


Fig. 2—(a) Radiation pattern co-ordinate system. (b) Equatorial belts for region $\theta = +30°$ to $-30°$.

[7] P. S. Carter, H. C. Lawrence, R. S. Wehner, and J. D. Woodward, "HF communications development program, Item 1," RCA Report, pp. 7–48; June, 1947.

[8] E. L. Bock, J. A. Nelson, and A. Dorne, "Very High Frequency Techniques," pp. 119–123, McGraw-Hill Book Co., New York, N. Y.; June, 1947.

[9] N. E. Lindenblad, "Television transmitting antenna for Empire State Bldg.," *RCA Rev.*, vol. 3, p. 387; April, 1939.

[10] F. E. Terman, "Radio Engineer's Handbook," McGraw-Hill Book Co., New York, N. Y., pp. 782–784; 1943.

of equatorial belt diagrams of the type shown in Figs. 2(b) and 4. In these diagrams a rectangular co-ordinate system is used with azimuth plotted as abscissas and cone angle as ordinates. Contours of equal field strength are formed on the diagram by joining points of constant field strength obtained from the complete set of radiation patterns for each test frequency. The radiation patterns of the tail cap are practically symmetrical about the axis of the aircraft, and therefore, only the port half of each spherical radiation pattern (for vertical polarization) is shown. Those areas in the equitorial belt where the field strength is less than one-half millivolt per meter at one mile for one-watt input are indicated in solid black and represent the prominent nulls which might impair successful communication. Nulls are therefore immediately obvious from an inspection of the equatorial belt diagrams, and as can be seen there are only a few null areas, and these are so small they would not seriously affect the communications during flight.

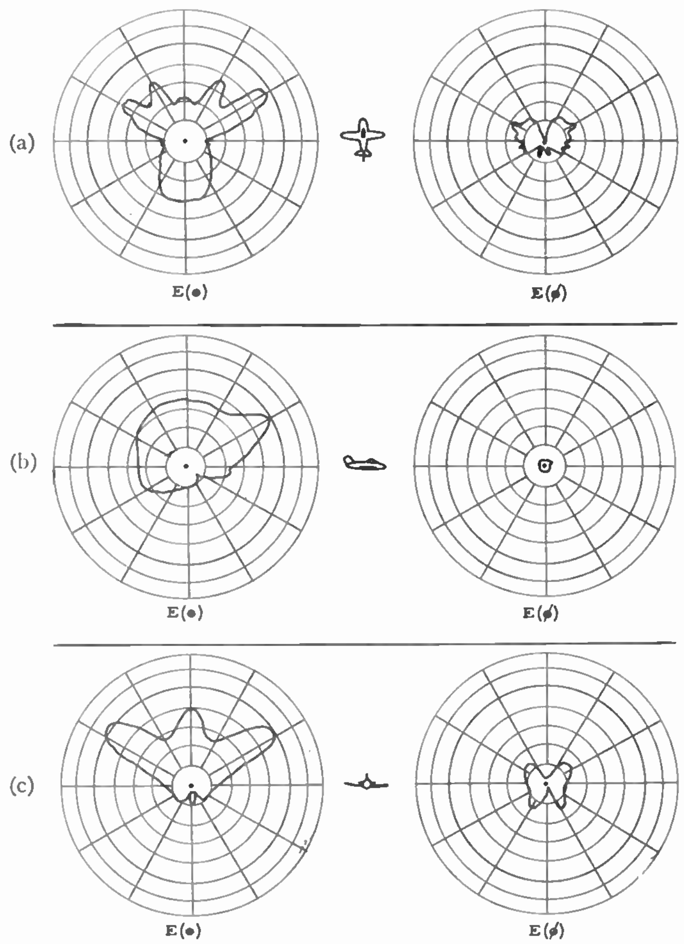


Fig. 3—Principal-plane patterns at 250 mc. (a) Horizontal plane. (b) Fore-and-aft plane. (c) Transverse plane.

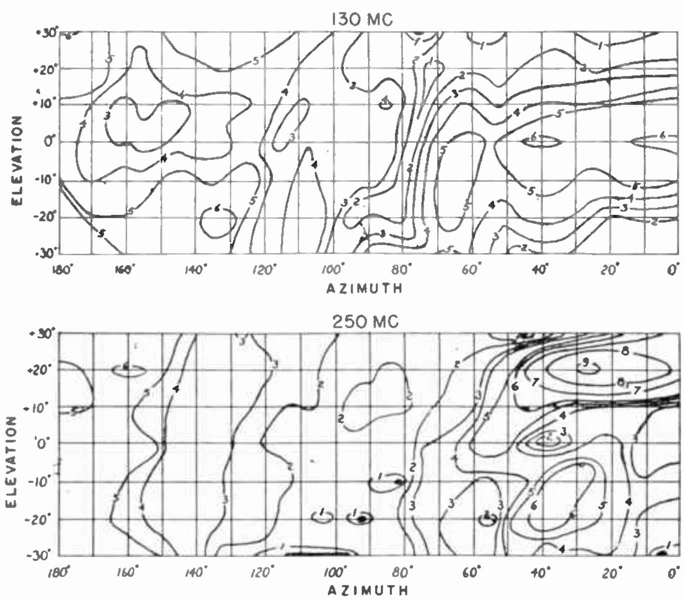


Fig. 4—Absolute field strength in equatorial belt. (Field-strength units—millivolts per meter for one watt at a distance of one mile.) Black areas represent field strength less than 0.5 millivolt per meter for one watt at a distance of one mile. Polarization—E_θ vertical polarization when aircraft is in level flight.

The ultimate purpose of the absolute field-strength patterns is to predict what the maximum ranges for acceptable operation will be when the antenna is used with specified radio equipments. These predictions are based on operational characteristics of the pertinent equipment, propagation data such as given in the literature,[11] assumed cable attenuation, and so forth. The predicted ranges depend upon the altitude of the aircraft, and are usually plotted as curves of maximum range versus altitude. The predictions from this model study indicate that the tail-cap antenna has satisfactory directivity characteristics, and it was concluded that a properly matched tail-cap antenna would give acceptable results when installed on the actual aircraft.

Development of Electrical Prototype

After the radiation pattern study was completed with acceptable results, the development of the full-scale electrical prototype tail-cap was begun. This phase was performed using a full-scale mock-up of the vertical portion of the tail above the horizontal stabilizer. The mock-up was constructed of plywood covered with copper and hardware cloth as shown in Fig. 5.

The mock-up was set on a 15-foot-square ground plane to simulate the empennage and fuselage for purposes of VSWR and matching measurements. An eight-foot length of RG-8/U coaxial cable was connected between the antenna and a precision slotted line located in the hut below.

The tail cap of the aircraft is 16 inches tall and provides the best solution for meeting all the design requirements. The all-metal stabilizer tip is isolated from the remainder of the stabilizer by a small strip of dielectric material, thus eliminating the need for a fiberglas stabilizer cap made by special production techniques.

The copper prototype tail cap was mounted on the stabilizer rib by two fiberglas pillars as shown in Fig. 5. The tail-cap face of the feeding gap connects to the

[11] C. R. Burrows, et al., "Radio Wave Propagation," Academic Press, New York, N. Y.; 1949.

center conductor of a coaxial cable receptacle, and the stabilizer face of the gap connects to the outer shield of the coaxial receptacle by a metal pyramid having a base $4\frac{1}{2}$ inches by 9 inches across. The pyramid provides a smooth transition between the coaxial receptacle and the feeding gap, and the slope of the pyramid was chosen for best broad-band performance.

The impedance-matching problem was solved in part by the use of an optimum pyrámid configuration, but

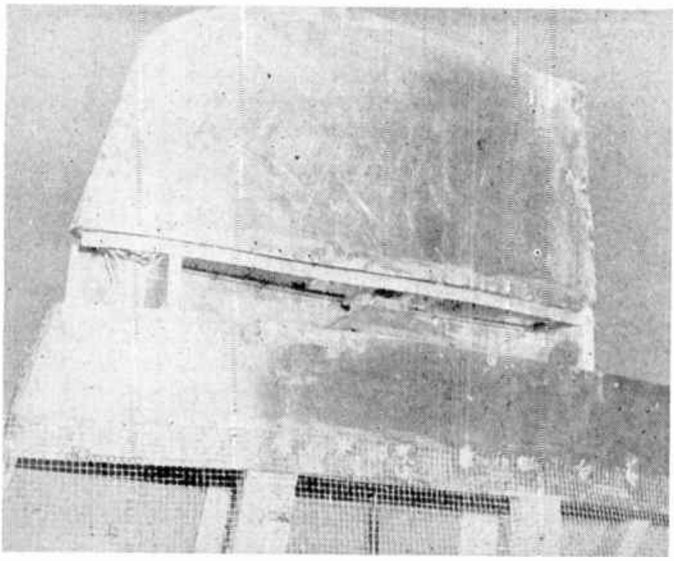


Fig. 5—Full-scale mock-up, prototype antenna.

in order to meet the SWR requirements at the low end of the frequency band, it was found necessary to employ a shunt inductance across the gap. The final configuration of the shunt consists of two copper wires 0.125-inch diameter and $3\frac{1}{2}$ inches long placed in the gap near the leading edge of the stabilizer.

Fig. 6 is a plot both of the SWR of the antenna as measured with 22 feet of RG-8/U cable as lead-in and also with the SWR at the antenna terminal corrected

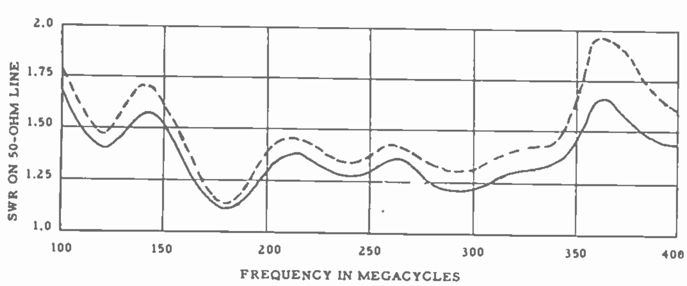


Fig. 6—Standing-wave ratio for tail-cap antenna. (———SWR with 22 feet RG-8/U. - - - - - SWR calculated at antenna.)

for the attenuation of the cable. The resultant SWR in both cases is less than 2 to 1. All final impedance and SWR measurements were made with a fiberglas rudder cap in place and with a fiberglas belt around the capacity hat.

Mechanical Design and Assembly

The mechanical design of the antenna incorporates a $\frac{1}{8}$-inch thick fiberglas belt 3 inches wide and two fiberglas structural spars $4\frac{1}{2}$ inches wide by $\frac{1}{2}$ inch thick shown in Fig. 5. The spars provide structural strength in addition to that of the fiberglas belt around the gap. The section of the rudder tip above the feeding gap in the stabilizer is made of fiberglas to avoid shielding and shunting effects on the tail cap. All fiberglas used is Formica FF-55 which has a dielectric constant of about 7. In order to control the current distribution in the metal part of the rudder, flexible bonding straps are run between the rudder and vertical stabilizer.

For convenience in production assembly, the pyramid and shunt wires are designed so that they may be installed after the tip has been mounted, or, if preferred, the pyramid may be assembled to the tip before the tip is mounted. The top of the pyramid is a small box that can be adjusted to change the over-all height of the pyramid a sufficient amount to compensate for production variations in the width of the feeding gap.

Conclusions

The tail-cap antenna described in this paper has acceptable electrical characteristics; is easy to construct, install, and service; does not cause any increase in aerodynamic drag; and last, but not least, it does not cause an increase in the weight in the tail of the aircraft. For these reasons, it is considered to be a good antenna for this application.

There are two shortcomings that this type of antenna may have. Some aircraft fuselage types may be so designed as to cause poor radiation patterns and also narrow-band characteristics. In these cases, the performance can not be made acceptable regardless of the tail cap configuration. The second shortcoming is that when this type of antenna is installed on a high-speed aircraft, moisture and ice particles errode the leading edge of the fiberglas belt and destroy its smooth contour. Perhaps this problem, which also exists with all types of fiberglas radomes and covers, will be solved to a great degree by continuing development of plastic materials.

Acknowledgments

Much credit is due to Richard Anderson and William Barrick of Electronics Research, Inc., for the measurements and development given in this paper, and to Eugene H. Barnard also of Electronics Research, Inc., who contributed considerably to the final configuration of the electrical prototype.

The Patterns of Antennas Located Near Cylinders of Elliptical Cross Section*

GEORGE SINCLAIR†, SENIOR MEMBER, IRE

Summary—**The patterns of small dipole and loop antennas mounted on or near a perfectly conducting cylinder of elliptical cross section are calculated by determining the open-circuit terminal voltages of the antennas when receiving plane waves. The antennas are assumed to be such that they cause negligible distortion of the field when they are open-circuited, thus reducing the problem to that of calculating the diffraction of a plane wave around the cylinder. The results have applications in the design of aircraft antennas and slotted-cylinder antennas.**

Introduction

IN THE DESIGN of antenna systems the problem of determining the effect of a nearby obstacle on the performance of the system frequently arises. An antenna may be mounted on or near a metallic structure, such as a building or an aircraft, with resultant distortion of the pattern. In some cases this distortion is undesirable and must be minimized while in others the distortion may be utilized to obtain a more desirable pattern. In either case, a knowledge of the distortion is required in carrying out the design of the system.

An antenna system in which the mounting structure plays an important part is that of an airborne antenna since the aircraft structure itself is often an essential portion of the radiating system. The effects of the aircraft structure on the pattern may be calculated approximately by replacing the complicated shape of the aircraft by a simpler geometrical shape more amenable to calculation. For example, if the antenna is mounted on the fuselage, it is known that an infinitely long circular cylinder is often a good approximation. When the antenna is mounted on a wing or on the empennage, a cylinder of elliptical cross section forms an adequate approximation in most cases.

An analysis of the effect of a nearby structure on the pattern of an antenna has applications also in the design of slotted-cylinder antennas since a slot antenna can be analyzed in terms of an equivalent distribution of magnetic dipoles. A slot antenna located on the surface of a cylindrical metal structure has a directional pattern, and a knowledge of the directive properties is usually required in designing the system. Since the pattern is dependent to some extent on the cross-sectional shape of the cylindrical surface, it is possible to obtain some control of the pattern by properly shaping the cylinder.

The patterns of simple antennas mounted on or near cylinders of circular cross section have been investigated by P. S. Carter[1,2] by a method based on diffraction theory and the reciprocity theorem. In the following, the analysis is extended to antennas located near cylinders of elliptical cross section. The method yields information only on the distant field radiated by the antenna, and hence produces no information on the reactive component of the terminal impedance.

Relationship Between Antenna Patterns and Diffraction Theory

The application of diffraction theory to the calculation of antenna patterns follows naturally from a consideration of the fields existing around an antenna receiving plane waves from a distant transmitter. Consider, for example, a dipole antenna whose dimensions are small compared with the wavelength and which is located near a given object of arbitrary shape, receiving waves from a given direction. When the antenna is open-circuited at its terminals, it causes negligible distortion of the field which would exist with the antenna absent. The dipole is merely a small probe which samples the component of the electric field parallel to its axis. If this component of the field can be calculated by diffraction theory, for different directions of incidence of the plane wave, the pattern of the dipole as a receiving antenna can be obtained. The distant field radiated by the antenna when transmitting is then determined from the reciprocity theorem. Similarly, if the antenna is a loop whose dimensions are small compared with the wavelength and with its terminals open-circuited, the receiving pattern can be determined from a knowledge of the component of the magnetic field intensity parallel to the axis of the loop.

The diffraction of plane waves by an elliptic cylinder has been considered by several investigators,[3,4] but their solutions are not sufficiently general since they considered only waves incident in directions normal to the axis of the cylinder. In the Appendix, expressions are obtained for the field produced near an infinitely long, perfectly conducting elliptic cylinder when illuminated by a plane wave of arbitrary polarization and traveling in an arbitrary direction.

* Decimal classification: R120×R326.21. Original manuscript received by the Institute, June 16, 1949; revised manuscript received, August 7, 1950.

The work described in this paper was carried out, in part, under contract between the Air Materiel Command, Wright-Patterson Air Force Base, and The Ohio State University Research Foundation, Columbus, Ohio.

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

[1] P. S. Carter, "Antenna arrays around cylinders," Proc. I.R.E., vol. 31, pp. 671–693; December, 1943.

[2] P. S. Carter, "Antennas and Cylindrical Fuselage," Report No. 895-11, RCA Laboratories, Rocky Point, N. Y.; December 24, 1943.

[3] P. M. Morse and P. J. Rubenstein, "The diffraction of waves by ribbons and slits," *Phys. Rev.*, vol. 54, pp. 895–898; December, 1938.

[4] B. Seiger, "Die beugung einer ebenen elektrischen wellen an einem schirm von elliptischen querschnitt," *Ann. der Phys.*, vol. 27, pp. 626–664; November, 1908.

Patterns of Antennas Near an Elliptic Cylinder

In specifying the location of the antenna near the elliptic cylinder, it will be convenient to employ elliptic cylinder co-ordinates (u, v, z) which may be defined by the following transformation from Cartesian co-ordinates:

$$x = d \cosh u \cos v \tag{1}$$

$$y = d \sinh u \sin v \tag{2}$$

$$z = z, \tag{3}$$

where $2d$ is the distance between the foci of the ellipses. The foci are located at the points $(x = \pm d, y = 0)$, as shown in Fig. 1. The surfaces u = constant represent a family of confocal cylinders of elliptical cross section. The semimajor and semiminor axes of an ellipse $u = u_0$ are given by

$$a' = d \cosh u_0 \tag{4}$$

$$b' = d \sinh u_0. \tag{5}$$

An open-circuited antenna of infinitesimal dimensions (a dipole or a loop antenna) is situated at the point $P_1(u_1, v_1, 0)$ outside (or on) the conducting surface of the cylinder $u = u_0$. No loss in generality results from assuming $z_1 = 0$ since the cylinder extends to infinity in both directions along the z axis from P_1. The pattern of the antenna is to be calculated by determining the open-circuit terminal voltage as a function of the direction of incidence of the plane wave.

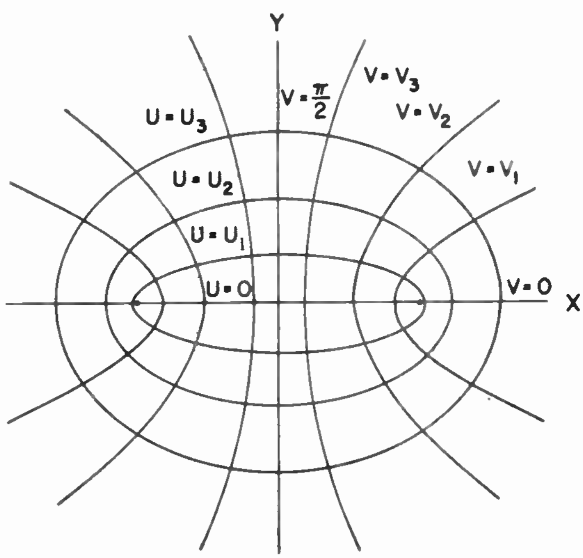


Fig. 1—The elliptic co-ordinate system.

When the antenna is being used for transmission, it produces a distant field which is essentially a spherical wave, so that in specifying the pattern of the antenna it is more suitable to employ a spherical co-ordinate system. The distant field can be resolved into the two spherical components E_θ (theta-polarized component)[5] and E_ϕ (phi-polarized component),[5] as shown in Fig. 2.

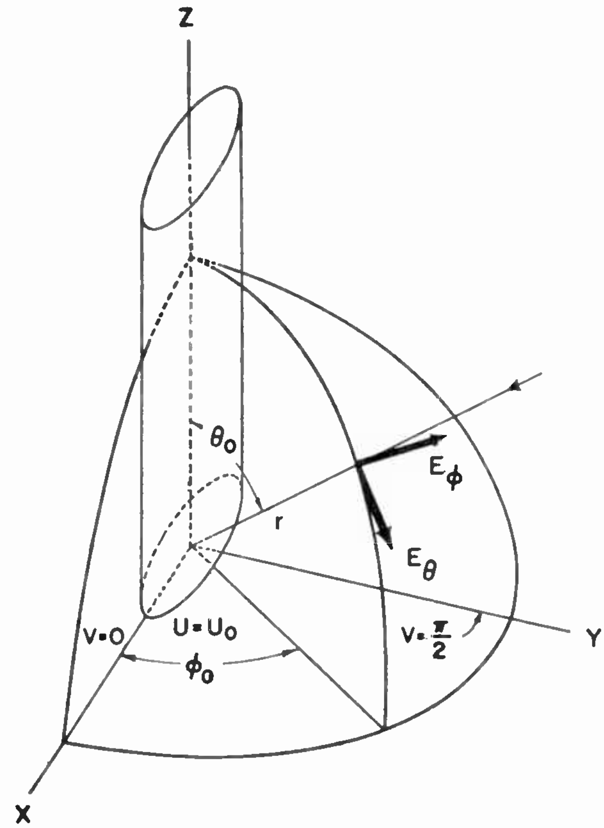


Fig. 2—Co-ordinate system, showing the resolution of the field into components.

By the reciprocity theorem[6] it can be shown that the relative pattern for either component can be obtained by determining the open-circuit terminal voltage of the antenna when receiving a plane wave polarized parallel to the component under consideration, the direction of propagation of the plane wave being varied to obtain the pattern.

While it is possible to calculate the patterns of loop and dipole antennas of any arbitrary orientation with respect to the cylinder, only the cases of dipole and loop antennas whose axes are tangent to co-ordinate curves will be considered. Consider an infinitesimal dipole of effective length l, receiving a plane linearly polarized wave, polarized in the direction for measuring the E_θ component of the transmitting pattern. Let the open-circuit voltage produced be V_θ. This voltage is equal to the product of the effective length and the component of the electric field intensity parallel to the axis of the antenna. Similarly, for a phi-polarized wave, the open-circuit voltage is the effective length times the component of electric intensity parallel to the axis. For a loop antenna of area S, the open-circuit voltage is equal to the product of its effective length $l = j\omega\mu S$ and the component of the magnetic intensity parallel to the axis of the loop. Using the expressions for the field com-

[5] I.R.E. Standards on Antennas, Definitions of Terms, 1948.

[6] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., p. 476; 1943.

ponents derived in the Appendix, the following are the expressions for calculating the open-circuit terminal voltages for the different cases:

Radial Dipole

$$V_\theta = \frac{l}{h_1} \frac{\partial^2 \Pi_z}{\partial u \partial z}\bigg|_{P_1} \quad (6)$$

$$V_\phi = -j \frac{\omega\mu l}{h_1} \frac{\partial \Pi_z^*}{\partial v}\bigg|_{P_1}, \quad (7)$$

where Π_z and Π_z^* are the Hertzian potentials[7] used in calculating the field (see Appendix).

Circumferential Dipole

$$V_\theta = \frac{l}{h_1} \frac{\partial^2 \Pi_z}{\partial v \partial z}\bigg|_{P_1} \quad (8)$$

$$V_\phi = j \frac{\omega\mu l}{h_1} \frac{\partial \Pi_z^*}{\partial u}\bigg|_{P_1}. \quad (9)$$

Dipole Parallel to the Axis

$$V_\theta = lk^2 \sin^2 \theta_0 \Pi_z \big|_{P_1} \quad (10)$$

$$V_\phi = 0. \quad (11)$$

Loop with Axis Radial

$$V_\theta = -\frac{k^2 S}{h_1} \frac{\partial \Pi_z}{\partial v}\bigg|_{P_1} \quad (12)$$

$$V_\phi = j \frac{\omega\mu S}{h_1} \frac{\partial^2 \Pi_z^*}{\partial u \partial z}\bigg|_{P_1}. \quad (13)$$

Loop with Axis Circumferential

$$V_\theta = \frac{k^2 S}{h_1} \frac{\partial \Pi_z}{\partial u}\bigg|_{P_1} \quad (14)$$

$$V_\phi = j \frac{\omega\mu S}{h_1} \frac{\partial^2 \Pi_z^*}{\partial v \partial z}\bigg|_{P_1}. \quad (15)$$

Loop with Axis Parallel to Axis of Cylinder

$$V_\theta = 0 \quad (16)$$

$$V_\phi = j\omega\mu k^2 S \sin^2 \theta_0 \Pi_z^* \big|_{P_1}. \quad (17)$$

In these expressions, θ_0 defines the direction of incidence with respect to the axis of the cylinder (see Fig. 2), h_1 is the metrical coefficient, and $k^2 = \omega^2\mu\epsilon$.

In each of the equations (6) through (17), the derivatives are to be evaluated at the point P_1 where the antenna is located. The total terminal voltage when waves of both theta and phi polarizations are present is the sum of V_θ and V_ϕ, taking due account of phase differences between the components of the incident wave.

The effective length of the antenna is easily obtained from the expressions for the received voltages. The vector effective length $\mathbf{h}$ of the antenna[8] can be resolved into two components whose values are easily seen to be

$$h_\theta = V_\theta / E_0. \quad (18)$$

$$h_\phi = V_\phi / E_1. \quad (19)$$

Thus the transmitted field, when the antenna is used for transmission is[8]

$$E = j \frac{Z_0 I h}{2\lambda r} e^{-jkr}. \quad (20)$$

A careful distinction must be made between the vector effective length $\mathbf{h}$, and the effective length l used in (6) through (17). The effective length l is the value of the effective length for the antenna when the conducting cylinder is absent, whereas the vector effective length $\mathbf{h}$ is its value when the conducting cylinder is present.

When the actual expressions for the potentials Π_z and Π_z^* are inserted in (6) through (17), and use is made of (18) and (19), it is found that the components of the transmitted field can be written in the forms

$$h_\theta = lK_1 \sum_{n=0}^{\infty} j^n \{A_n Se_n(s, \cos\phi_0) + B_n So_n(s, \cos\phi_0)\} \quad (21)$$

$$h_\phi = lK_2 \sum_{n=0}^{\infty} j^n \{C_n Se_n(s, \cos\phi_0) + D_n So_n(s, \cos\phi_0)\}. \quad (22)$$

The coefficients A_n, B_n, C_n, D_n, K_1, and K_2 are listed in Table I. Se_n and So_n are the even and odd angular Mathieu functions and the angle ϕ_0 is shown in Fig. 2; and $s = k^2 d^2 \sin^2\theta_0$.

The procedure in calculating the pattern of a given antenna near an elliptic cylinder is as follows: First, the distance $2d$ between the foci of the elliptical surface is determined and the parameter s calculated for the desired direction θ_0. The separation constants of Mathieu's equation[9] be_n and bo_n, and the Fourier coefficients De_m and Do_m are obtained from tables[10,11] or calculated.[12] The radial Mathieu functions listed in Table I are then calculated for the value $u = u_0$ defining the cylindrical surface, and for $u = u_1$ defining the location of the antenna. The values of the coefficients K_1, K_2, A_n, B_n, C_n, and D_n are then calculated as listed in Table I. The patterns, as functions of the azimuthal angle ϕ_0 can then be calculated using the series (21) and (22).

Antennas Mounted on a Circular Cylinder

A circular cylinder can be considered to be a limiting form of an elliptical cylinder, so the patterns of dipole

[7] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., p. 32; 1941.

[8] G. Sinclair, "The transmission and reception of elliptically polarized waves," Proc. I.R.E., vol. 38, pp. 148–151; February, 1950. Correction, Proc. I.R.E., vol. 38, p. 1216; October, 1950.

[9] The notation for the Fourier coefficients and the separation constants was suggested by G. Blanch of the National Bureau of Standards, and conforms to that used in a forthcoming set of tables of these quantities. In the notation of footnote reference 10, $De_m = D_m^n$, $Do_m = F_m^n$, $be_n = b_n$, $bo_n = b_n'$, $s = c^2$.

[10] J. A. Stratton, P. M. Morse, L. J. Chu, and R. A. Hutner, "Elliptic Cylinder and Spheroidal Wave Functions," John Wiley and Sons, Inc., New York, N. Y.; 1941.

[11] "Table of Characteristic Values of Mathieu's Differential Equation; NDRC Applied Mathematics Panel Report No. 165-1R;" September, 1945.

[12] G. Blanch, "On the computation of mathieu functions," *Jour. Math. Phys.*, vol. 25, February 1946.

TABLE I

ANTENNAS ON AN ELLIPTIC CYLINDER

Antenna	K_1	K_2	A_n
Radial dipole	$\frac{-j\sqrt{8\pi}\cos\theta_0}{kh_1\sin\theta_0}$	$\frac{j\sqrt{8\pi}}{kh_1\sin\theta_0}$	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_nHe_n^{(2)\prime}(s,\xi_1)]$
Circumferential dipole	$\frac{-j\sqrt{8\pi}\cos\theta_0}{kh_1\sin\theta_0}$	$\frac{-j\sqrt{8\pi}}{kh_1\sin\theta_0}$	$\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$
Axial dipole	$-\sqrt{8\pi}\sin\theta_0$	0	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$
Loop with axis radial	$\frac{-j\sqrt{8\pi}}{Z_0kh_1\sin\theta_0}$	$\frac{-j\sqrt{8\pi}\cos\theta_0}{Z_0kh_1\sin\theta_0}$	$\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$
Loop with axis circumferential	$\frac{j\sqrt{8\pi}}{Z_0kh_1\sin\theta_0}$	$\frac{-j\sqrt{8\pi}\cos\theta_0}{Z_0kh_1\sin\theta_0}$	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_nHe_n^{(2)\prime}(s,\xi_1)]$
Loop with axis axial	0	$\frac{-\sqrt{8\pi}\sin\theta_0}{Z_0}$	0

$\xi_1=\cosh u_1$ $\quad c_n=-\frac{Je_n(s,\cosh u_0)}{He_n^{(2)}(s,\cosh u_0)}$ $\quad d_n=-\frac{Jo_n(s,\cosh u_0)}{Ho_n^{(2)}(s,\cosh u_0)}$ $\quad s=k^2d^2\sin^2\theta_0$

Antenna	B_n	C_n	D_n
Radial dipole	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_nHo_n^{(2)\prime}(s,\xi_1)]$	$\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$	$\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$
Circumferential dipole	$\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_n'He_n^{(2)\prime}(s,\xi_1)]$	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_n'Ho_n^{(2)\prime}(s,\xi_1)]$
Axial dipole	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$	0	0
Loop with axis radial	$\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_n'He_n^{(2)\prime}(s,\xi_1)]$	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_n'Ho_n^{(2)\prime}(s,\xi_1)]$
Loop with axis circumferential	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_nHo_n^{(2)\prime}(s,\xi_1)]$	$\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$	$\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$
Loop with axis axial	0	$\frac{Se_n(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$	$\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$

$h_1=d\sqrt{\cosh^2 u_1-\cos^2 v_1}$ $\quad c_n'=-\frac{Je_n'(s,\cosh u_0)}{He_n^{(2)\prime}(s,\cosh u_0)}$ $\quad d_n'=-\frac{Jo_n'(s,\cosh u_0)}{Ho_n^{(2)\prime}(s,\cosh u_0)}$

and loop antennas mounted near a circular cylinder can be obtained from the above results by taking the appropriate limits. The transition to co-ordinates of a circular cylinder is given by the limits

$$d\rightarrow 0 \tag{23}$$

$$u\rightarrow\infty \tag{24}$$

in such a manner that the product de^u remains finite and equal to the radius ρ of the circular cylinder. The angular variable v becomes the azimuthal angle ϕ. It is, however, more convenient to obtain the expressions directly as shown by Carter.[1,2] The resulting expressions for the components of the vector effective length are Fourier series of the forms

$$h_\theta = lK_1\sum_{n=0}^{\infty} \epsilon_n j^n\{A_n \cos n\phi_0 + nB_n \sin n\phi_0\} \qquad (25)$$

$$h_\phi = lK_2\sum_{n=0}^{\infty} \epsilon_n j^n\{C_n \cos n\phi_0 + nD_n \sin n\phi_0\}, \qquad (26)$$

where ϵ_n is Neumann's number ($\epsilon_0=1$, otherwise $\epsilon_n=2$), and the coefficients K_1, K_2, A_n, B_n, C_n and D_n are listed in Table II. The results given in Table II are due to Carter[1,2] and are repeated here for convenience and completeness. The antenna is located at the point $P_1(b, \phi_1, z_1)$, and there is no loss in generality in assuming that $\phi_1=z_1=0$. The radius of the perfectly conducting infinite circular cylinder is $\rho_0=a$ (see Fig. 3).

Experimental Verification of the Calculated Patterns

A number of measurements have been made to test the accuracy of the calculated patterns with the results shown in Figs. 4 and 5. In all cases the antenna was mounted directly on the surface of the cylinder so that $u_1=u_0$. Fig. 4 shows the patterns for a very short radial stub antenna for two different locations, while Fig. 5 shows the patterns for a loop antenna. The poorer agreement between measurements and calculations in the case of the loop antenna is attributed to the finite dimensions of the loop used in the measurements. Measurements verifying the accuracy of the calculations for antennas mounted near a circular cylinder have been described previously.[1,2,13]

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

Antennas of Arbitrary Configurations

The results derived above for the patterns of antennas located near circular and elliptic cylinders are actually the expressions for the distant fields radiated by

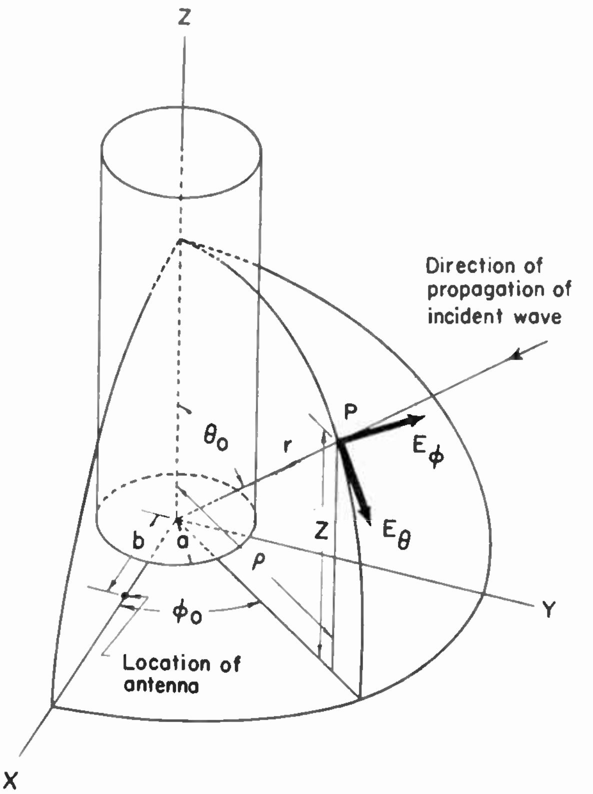


Fig. 3—Co-ordinate system for the circular cylinder.

TABLE II

Antennas on a Circular Cylinder

Antennas	K_1	K_2	A_n	B_n	C_n	D_n
Radial dipole	$-j\cos\theta_0$	$-j\dfrac{1}{w}$	$J_n'(w)+c_nH_n^{(2)\prime}(w)$	O	O	$J_n(w)+c_n'H_n^{(2)}(w)$
Circumferential dipole	$j\dfrac{\cos\theta_0}{w}$	$-j$	O	$J_n(w)+c_nH_n^{(2)}(w)$	$J_n'(w)+c_n'H_n^{(2)\prime}(w)$	O
Axial dipole	$-\sin\theta_0$	O	$J_n(w)+c_nH_n^{(2)}(w)$	O	O	O
Loop with axis radial	$-\dfrac{1}{jZ_0w}$	$-j\dfrac{\cos\theta_0}{Z_0}$	O	$J_n(w)+c_nH_n^{(2)}(u)$	$J_n'(w)+c_n'H^{(2)\prime}(w)$	O
Loop with axis circumferential	$-\dfrac{1}{jZ_0}$	$j\dfrac{\cos\theta_0}{Z_0w}$	$J_n'(w)+c_nH_n^{(2)\prime}(w)$	O	O	$J_n(w)+c_n'H_n^{(2)}(w)$
Loop with axis axial	O	$-\dfrac{\sin\theta_0}{Z_0}$	O	O	$J_n(w)+c_n'H_n^{(2)}(w)$	O

$w=kb\sin\theta_0$ $\qquad c_n=-\dfrac{J_n(ka\sin\theta_0)}{H_n^{(2)}(ka\sin\theta_0)}$ $\qquad c_n'=-\dfrac{J_n'(ka\sin\theta_0)}{H_n^{(2)\prime}(ka\sin\theta_0)}$

infinitesimal elements of electric and magnetic currents. Therefore, it is possible to use these results to determine the patterns of more complicated antennas when the distributions of current or voltage along the antenna are known since it is merely necessary to sum the contributions to the total field from each part of the antenna. It is possible to carry out the integrations involved in this calculation in certain cases, as, for example, when the current or voltage can be assumed distributed sinusoidally.

Conclusions

Expressions have been obtained for the distant field radiated by small dipole and loop antennas located on or near cylinders of elliptical cross section. Arrays of such antennas can be treated by the superposition of the fields due to the individual radiators.[1]

Acknowledgements

The measurements shown in Figs. 4 and 5 were made by E. A. Fouty; the calculations, by Miss Frances M. Nichols. Some assistance in calculating the Mathieu functions was received from G. Blanch of the National Bureau of Standards.

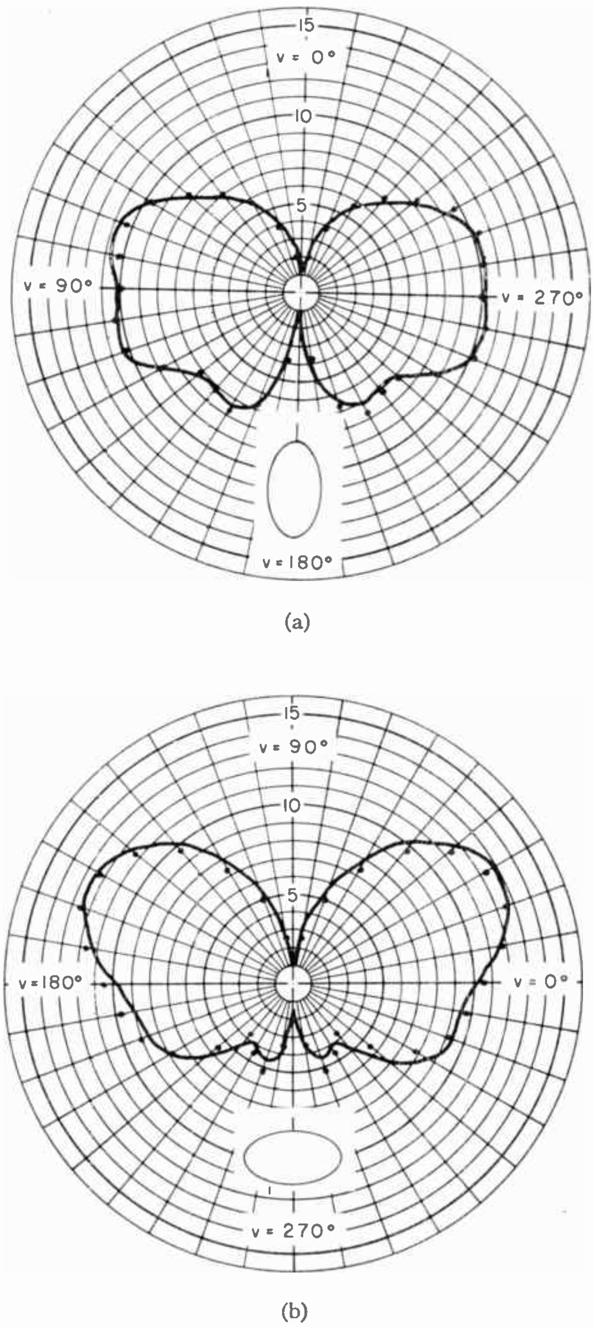


Fig. 4—Comparison of measured and calculated patterns for a short radial stub antenna mounted on an elliptic cylinder of semimajor axis $a' = 0.637\lambda$, and semiminor axis $b' = 0.390\lambda$. Curves measured at 1,500 mc and points calculated. (a) Stub located at $v_1 = 0$, $u_1 = u_0$. (b) Stub located at $v_1 = \pi/2$, $u_1 = u_0$.

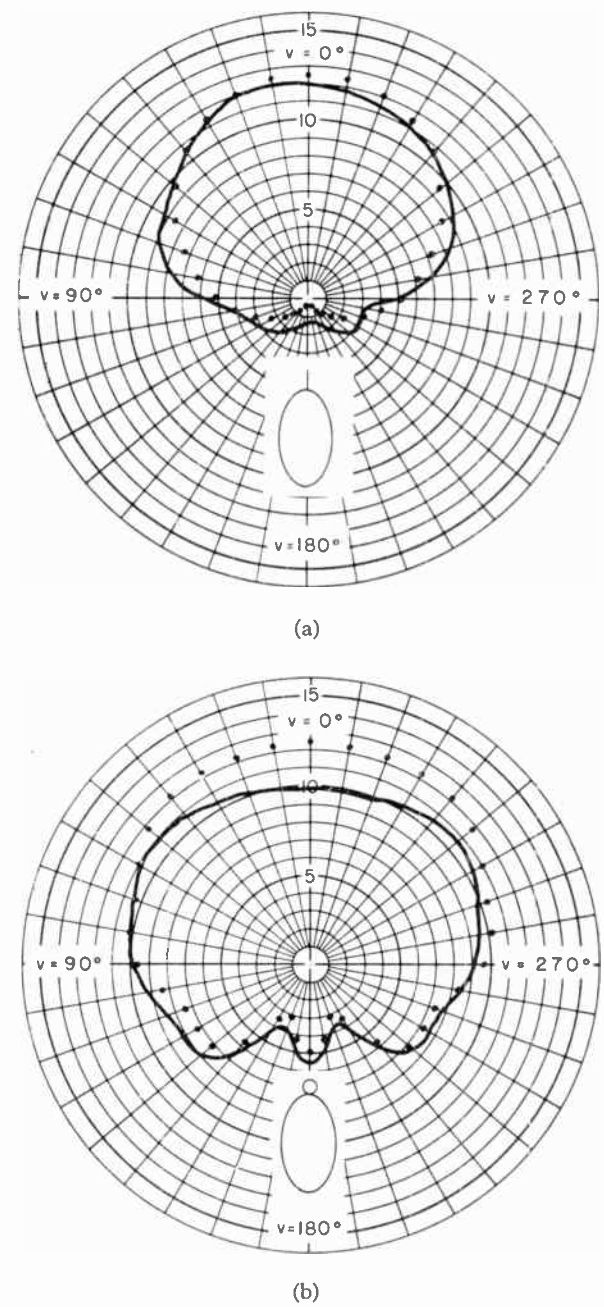


Fig. 5—Similar to Fig. 4, but for a loop antenna $u_1 = u_0$, $v_1 = 0$. (a) Loop axis directed circumferentially. (b) Loop axis directed parallel to axis of cylinder.

Appendix

Diffraction of a Plane Wave Around an Elliptic Cylinder

The diffraction of a plane wave around an elliptic cylinder is readily solved by expressing the vector wave equation in elliptic cylinder co-ordinates.[14] The equation for the z component of the Hertzian potential $\mathbf{\Pi}$ is

$$\frac{1}{h_1^2}\left\{\frac{\partial^2\Pi_z}{\partial u^2}+\frac{\partial^2\Pi_z}{\partial v^2}\right\}+\frac{\partial^2\Pi_z}{\partial z^2}+k^2\Pi_z=0, \quad (27)$$

where the metrical coefficient is

$$h_1^2 = d^2(\cosh^2 u - \cos^2 v). \quad (28)$$

This equation is readily solved by separation, that is, by assuming a solution of the form

$$\Pi_z = f_1(u)f_2(v)f_3(z), \quad (29)$$

where the functions f_i are functions only of the single variable shown for each.

The function $f_3(z)$ is the exponential function

$$f_3(z) = e^{\pm i\gamma z}, \quad (30)$$

where γ is a constant resulting from the separation of the differential equation.

For the function $f_2(v)$, the periodic even and odd angular Mathieu functions are chosen

$$Se_n(s, \cos v) = \sum_{m=0,1}^{\infty}{}' De_m \cos mv \quad (31)$$

$$So_n(s, \cos v) = \sum_{m=1,2}^{\infty}{}' Do_m \sin mv, \quad (32)$$

where

$$s = d^2(k^2 - \gamma^2) \quad (33)$$

and the prime on the summation sign indicates a summation over even values of m if n is even, and over odd values of m if n is odd.

The radial Mathieu functions $f_1(u)$ associated with the even and odd angular Mathieu functions are[15]

$$Ze_n(s, \cosh u) = \sqrt{\frac{\pi}{2}}\sum_{m=0,1}^{\infty}{}' j^{n-m}De_m Z_m(s\cosh u) \quad (34)$$

$$Zo_n(s, \cosh u) = \sqrt{\frac{\pi}{2}}\sum_{m=1,2}^{\infty}{}' j^{n-m}mDo_m Z_m(s\cosh u), \quad (35)$$

where $Z_m(s)$ is any cylinder function.

Since any linear combination of solutions is still a solution of (27), a complete solution is therefore

$$\Pi_z = \sum_{n=0}^{\infty}\{c_n Se_n(s, \cos v)Ze_n(s, \cosh u) + d_n So_n(s, \cos v)Zo_n(s, \cosh u)\} \times e^{\pm\gamma z}. \quad (36)$$

The constant γ and the coefficients c_n and d_n are to be determined by the physical conditions of the boundary value problem.

Consider a plane wave polarized parallel to the theta component of the radiated field of the antenna shown in Fig. 2. The incident plane wave is expressed by the equation

$$E^i = E_0 e^{ikr+j\omega t}, \quad (37)$$

where r is distance measured from the origin of the co-ordinate system along the direction of propagation (see Fig. 2). In elliptical cylinder co-ordinates,

$$r = d\sin\theta_0(\cosh u\cos v\cos\phi_0 + \sinh u\sin v\sin\phi_0) + z\cos\theta_0, \quad (38)$$

where the angles ϕ_0 and θ_0 define the direction of incidence. It is readily shown that this plane wave can be derived from the following Hertzian potential having only a z component:

$$\Pi_z{}^i = \frac{-E_0 e^{ikr}}{k^2\sin\theta_0} = \frac{-\sqrt{8\pi}\,E_0}{k^2\sin\theta_0}\sum_{n=0}^{\infty} j^n\left\{\frac{1}{N_n}Je_n(s, \cosh u)\cdot Se_n(s, \cos v)Se_n(s, \cos\phi_0) + \frac{1}{N_n'}Jo_n(s, \cosh u)So_n(s, \cos v)\cdot So_n(s, \cos\phi_0)\right\}e^{ikz\cos\theta_0}, \quad (39)$$

where

$$N_n = \pi\left\{2(De_0)^2 + \sum_{m=2}^{\infty}{}'(De_m)^2\right\} \text{ when } n \text{ is even} \quad (40)$$

$$N_n = \pi\sum_{m=1}^{\infty}{}'(De_m)^2 \text{ when } n \text{ is odd} \quad (41)$$

$$N_n' = \pi\sum_{m=1,2}^{\infty}{}'(Do_m)^2 \quad (42)$$

$$s = k^2d^2\sin^2\theta_0 \quad (43)$$

$$\gamma = k\cos\theta_0. \quad (44)$$

It is assumed that the boundary conditions can be satisfied by choosing for the reflected field the following potential:

$$\Pi_z{}^r = \frac{-\sqrt{8\pi}\,E_0}{k^2\sin\theta_0}\sum_{n=0}^{\infty} j^n\left\{\frac{c_n}{N_n}He_n^{(2)}(s, \cosh u)\cdot Se_n(s, \cos v)Se_n(s, \cos\phi_0) + \frac{d_n}{N_n'}Ho_n^{(2)}(s, \cosh u)So_n(s, \cos v)\cdot So_n(s, \cos\phi_0)\right\}e^{ikz\cos\theta_0}, \quad (45)$$

where $He_n^{(2)}(s, \cosh u)$ and $Ho_n^{(2)}(s, \cosh u)$ are radial

[14] See pp. 375–386 of footnote reference 7.

[15] The notation employed for the general Mathieu function is an obvious extension of the notation used in footnote reference 10.

Mathieu functions of the fourth kind, and are the appropriate functions for ensuring that the reflected field consists only of outward travelling waves.[16]

The total field can be derived from the potential

$$\Pi_z = \Pi_z^i + \Pi_z^r. \tag{46}$$

The components of the total field are found from the expressions

$$\mathbf{E} = \text{curl curl } \mathbf{\Pi} \tag{47}$$

$$\mathbf{H} = j\omega\epsilon \text{ curl } \mathbf{\Pi}. \tag{48}$$

Thus the tangential components of the electric vector are

$$E_v = \frac{1}{h_1} \frac{\partial^2 \Pi_z}{\partial v \partial z} \tag{49}$$

$$E_z = k^2 \sin^2 \theta_0 \Pi_z. \tag{50}$$

Since the boundary conditions require that the tangential components of the electric vector must vanish on the conducting surface $u=u_0$, it is apparent that c_n and d_n must have the values

$$c_n = -\frac{Je_n(s, \cosh u_0)}{He_n^{(2)}(s, \cosh u_0)} \tag{51}$$

$$d_n = -\frac{Jo_n(s, \cosh u_0)}{Ho_n^{(2)}(s, \cosh u_0)} \cdot \tag{52}$$

Thus the components of the total field can be found.

For an incident plane wave polarized parallel to the phi component of the pattern, the expression for the plane wave is

$$E^i = E_1 e^{jkr+j\omega t}. \tag{53}$$

This field can be derived from the potential

$$\pi_z^{*i} = \frac{-E_1 e^{jkr}}{Z_0 k^2 \sin \theta_0} \cdot \tag{54}$$

Expressing this potential in terms of the wave functions and assuming a suitable form for the potential of the reflected field, it is found that the total field can be derived from the potential

$$\begin{aligned}\Pi_z^* &= \Pi_z^{*i} + \Pi_z^{*r} \\ &= \frac{-\sqrt{8\pi}\, E_1}{Z_0 k^2 \sin \theta_0} \sum_{n=0}^{\infty} j^n \Bigg\{ \frac{1}{N_n} [Je_n(s, \cosh u) \\ &\quad + c_n' He_n^{(2)}(s, \cosh u)] Se_n(s, \cos v) Se_n(s, \cos \phi_0) \\ &\quad + \frac{1}{N_n'} [Jo_n(s, \cosh u) + d_n' Ho_n^{(2)}(s, \cosh u)] \\ &\quad \cdot So_n(s, \cos v) So_n(s, \cos \phi_0) \Bigg\} e^{jkz \cos \theta_0},\end{aligned} \tag{55}$$

where c_n' and d_n' are to be determined by the boundary conditions, namely, the vanishing of the tangential components of the electric vector. The components of the field due to this potential are found from

$$\mathbf{E} = -j\omega\mu \text{ curl } \mathbf{\Pi}^* \tag{56}$$

$$\mathbf{H} = \text{curl curl } \mathbf{\Pi}^*. \tag{57}$$

The tangential components of the electric vector are

$$E_v = j\frac{\omega\mu}{h_1} \frac{\partial \Pi_z^*}{\partial u} \tag{58}$$

$$E_z = 0. \tag{59}$$

The boundary conditions are satisfied by choosing

$$c_n' = -\frac{Je_n'(s, \cosh u_0)}{He_n^{(2)\prime}(s, \cosh u_0)} \tag{60}$$

$$d_n' = -\frac{Jo_n'(s, \cosh u_0)}{Ho_n^{(2)\prime}(s, \cosh u_0)} \tag{61}$$

where the primed Mathieu functions mean the derivatives with respect to the argument u, that is, for example,

$$Je_n'(s, \cosh u_0) = \left[\frac{d}{du} Je_n(s, \cosh u)\right]_{u=u_0}. \tag{62}$$

In Table I, the prime on the angular Mathieu functions means, for example,

$$Se_n'(s, \cos v_1) = \left[\frac{d}{dv} Se_n(s, \cos v)\right]_{v=v_1}. \tag{63}$$

List of Symbols

A_n = series coefficient (see Tables I and II)
B_n = series coefficient (see Tables I and II)
C_n = series coefficient (see Tables I and II)
D_n = series coefficient (see Tables I and II)
De_m = Fourier coefficient for even Mathieu functions
Do_m = Fourier coefficient for odd Mathieu functions
$\mathbf{E}$ = electric vector
$\mathbf{H}$ = magnetic vector
$He_n^{(2)}(s, \cosh u)$ = even radial Mathieu function of fourth kind
$Ho_n^{(2)}(s, \cosh u)$ = odd radial Mathieu function of fourth kind
$Je_n(s, \cosh u)$ = even radial Mathieu function of first kind
$Jo_n(s, \cosh u)$ = odd radial Mathieu function of first kind
K_1 = see Tables I and II
K_2 = see Tables I and II
N_n = normalization factor for $Se_n(s, \cos v)$
N_n' = normalization factor for $So_n(s, \cos v)$
S = area of loop antenna in square meters
$Se_n(s, \cos v)$ = even angular Mathieu function
$So_n(s, \cos v)$ = odd angular Mathieu function
V = open-circuit terminal voltage
Z_0 = intrinsic impedance of free space (120π ohms)

[16] Note that $e^{+j\omega t}$ has been assumed for the time dependence of the field.

a = radius of perfectly conducting circular cylinder
a' = semimajor axis of an ellipse
b = radial distance to location of antenna in circular cylindrical co-ordinates
b' = semiminor axis of an ellipse
c_n = series coefficient (see (51))
c_n' = series coefficient (see (60))
d = half the distance between foci of the ellipses
d_n = series coefficient (see (52))
d_n' = series coefficient (see (61))
h_1 = metrical coefficient
$h_1 = d\sqrt{\cosh^2 u - \cos^2 v}$
k = propagation constant, $k = 2\pi/\lambda$
l = effective length of dipole or loop antenna
r = radial distance from origin
$s = k^2d^2 \sin^2\theta_0$
u = radial co-ordinate in elliptic cylinder co-ordinates
u_0 = radial co-ordinate for surface of perfectly conducting elliptical cylinder
u_1 = radial co-ordinate of location of antenna
v = angular co-ordinate in elliptic cylinder co-ordinates
v_1 = angular co-ordinate of location of antenna
$w = kb \sin \theta_0$ (see Table II)
z = axial co-ordinate
$\mathbf{\Pi}$ = Hertzian vector potential
$\mathbf{\Pi}^*$ = Hertzian vector potential
γ = separation constant, $\gamma = k \cos \theta_0$
ϵ = dielectric constant of free space, $\epsilon = 10^{-9}/36\pi$
ϵ_n = Neumann's number ($\epsilon_0 = 1$; $\epsilon_n = 2$ if $n \neq 0$)
θ_0 = angle between z axis and direction of propagation
λ = wavelength in free space
μ = permeability of free space, $\mu = 4\pi \times 10^{-7}$
$\xi = \xi_1 = \cosh u_1$ (see Table I)
ρ = radial co-ordinate in circular cylindrical co-ordinates
ϕ = azimuthal angular co-ordinate
ϕ_0 = azimuthal angle of direction of propagation.

Current Distributions on Helical Antennas*

JAMES A. MARSH†, ASSOCIATE, IRE

Summary—**The current distribution on a uniform circular helix has been measured over a frequency band extending from 600 to 1,700 mc. Although the measured distributions are highly complex functions of distance and also change in an anomalous manner with frequency, it is possible to analyze the distributions in terms of traveling waves associated with three different transmission modes on the helix. The relative amplitude functions of these traveling waves, as well as their associated phase velocities, are approximated. Current distributions which have been calculated by superposing three or more traveling waves are in good agreement with the measured data.**

I. Introduction

In recent years considerable effort has been spent in analyzing the propagation of electrical energy along a helical conductor. Many applications of the helix have been found. One of these, discovered by Kraus,[1-5] employs the helix as an end-fire antenna. In this application the pitch angle may lie between 10 and 20 degrees while the circumference of the imaginary cylinder having diameter D (see Fig. 1) is approximately 0.80 to 1.3 wavelengths. In this range of frequencies a helix, of at least a few turns, radiates approximately circularly polarized waves in the direction of its axis. The radiation pattern of this antenna has a beamwidth which may be made quite narrow by employing a sufficiently large number of turns. The helix, when used in this manner, may be called a "helical beam antenna" and is said to be radiating in the "axial mode."

Although the current distributions of many types of antennas are well known, only a few distributions have been reported for the helical antenna.[2] The current

* Decimal classification: R326.7×R242. Original manuscript received by the Institute, February 3, 1950; revised manuscript received, August 23, 1950.
Work described in this paper was carried out, in part, under a contract between the Wright-Patterson Air Force Base of the Air Materiel Command and the Ohio State University Research Foundation. This is a portion of a doctorate dissertation, entitled, "A study of phase velocity on long cylindrical conductors," presented, Graduate School, Ohio State University, 1949. Presented, 1950 National IRE Convention, New York, N. Y., March 7, 1950. Published as a special report, The Ohio State University Research Foundation, February 28, 1950.
† Formerly, The Ohio State University Research Foundation, Columbus 10, Ohio; now, Aerophysics Laboratory, North American Aviation, Inc., Downey, Calif.

[1] J. D. Kraus, "Helical beam antenna," *Electronics*, vol. 20, pp. 109–111; April, 1947.
[2] J. D. Kraus and J. C. Williamson, "Characteristics of helical antennas radiating in the axial mode," *Jour. Appl. Phys.*, vol. 19, pp. 87–96; January; 1948.
[3] O. J. Glasser and J. D. Kraus, "Measured impedances of helical beam antennas," *Jour. Appl. Phys.*, vol. 19, pp. 193–197; February; 1948.
[4] J. D. Kraus, "Helical beam antennas for wide-band applications," Proc. I.R.E., vol. 36, pp. 1236–1242; October, 1948.
[5] J. D. Kraus, "The helical antenna," Proc. I.R.E., vol. 37, pp. 263–272; March, 1949.

distribution on the helix is complex, and varies in an anomalous manner with changes in frequency. Accordingly, a systematic set of measurements was undertaken in order to observe the current distribution both as a function of distance along the helix and also as a function of the frequency. Examples of these measured distributions and their analysis in terms of traveling waves on the helix are presented in this paper.

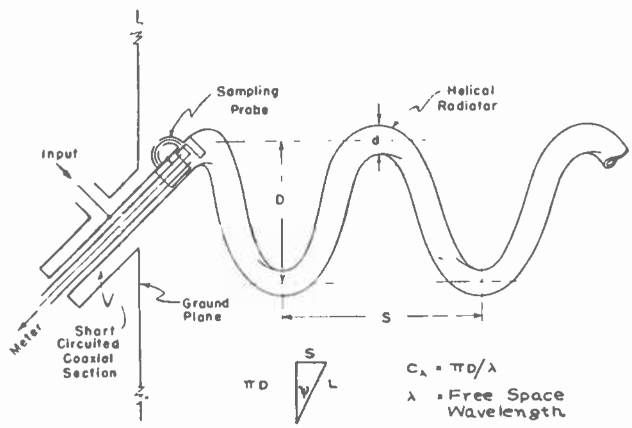


Fig. 1—Sketch of feed and sampling system used in the investigation of the helix.

All of the measurements were made in the uhf range on helices made from copper cylindrical tubing having small outer diameters in terms of a wavelength. Therefore, the assumption is made that the current on the helix is a one-dimensional surface current flowing only in the direction of the conductor axis. This assumption permits the use of the method of feeding and sampling shown in Figs. 1 and 2.[6] The hollow-center conductor

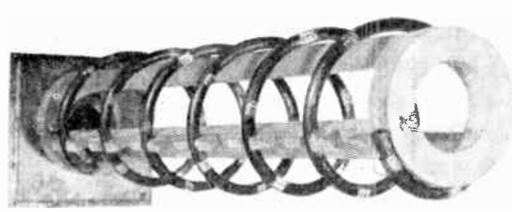


Fig. 2—Six-turn uniform circular helix, pitch angle $\psi = 12.6°$, mean diameter $D = 8.61$ cm.

of a short-circuited coaxial section is extended and wound in the shape of a uniform circular helix. The helix is excited by feeding energy through the side of the coaxial section. The current on the helix is sampled by means of a balanced loop which is positioned in an axial slot in the radiator. The energy sampled by the loop is fed back through the inside of the helix on a coaxial line to the measuring equipment located behind the ground plane.

Since the currents on the helix vary harmonically with time, must satisfy Maxwell's equations, and are assumed to be one-dimensional, it is possible to obtain expressions for the currents by the superposition of one-dimensional traveling waves.

The analysis of a measured current distribution in terms of traveling waves consists of assigning a relative amplitude function and a relative phase velocity to each of the component waves. The analysis becomes difficult for the present case, since there are at least three traveling waves of significant magnitude which, in general, have different phase velocities. For simplicity, the amplitude of the component waves are assumed to be either constant or exponentially attenuated. Therefore, each component wave can be expressed in the form:

$$I = I_0 e^{-\alpha x + j(\omega t \mp \beta x)},$$

where α is the attenuation constant, ω is the angular frequency, and β is the phase constant. The phase velocity of such a wave is usually defined[7] as $v = \omega/\beta$ and is independent of the attenuation constant α. In practical systems, however, the existence of a single traveling wave of the form above is rarely encountered. The current distributions that are usually measured consist of two or more traveling waves, and only the sum total of these currents is measurable.

It is common practice to measure the relative phase and amplitude of the total current as functions of distance and plot these data, choosing a convenient reference point in the system for the co-ordinate, in a manner as shown in Figs. 3 to 5. The relative amplitude curves are labeled $|I(x)|^2$ and the relative phase curves are labeled ϕ. The angle ϕ, essentially a lag angle for waves traveling to the right, is plotted positively for convenience. Since the angular frequency ω is readily determined in any practical measurement, it is only necessary to obtain the wave number β to find the phase velocity. If there were but one wave traveling with a constant velocity in the system under consideration, the $\phi(x)$ curve would be linear with a slope equal to β. The determination of the phase velocity for this case is straightforward. However, where more than one wave is present, the curve is not linear and the determination of β must be made indirectly. That is, an average slope of the ϕ curve, or some other method which uses the distance between successive points which are 360° apart must be employed to obtain the phase velocity.

A generalized phase velocity can be defined for the case where more than one traveling wave is present in the system by letting the slope $d\phi/dx = \beta(x)$ and defining $\hat{v} = -\omega/\beta(x)$, where $\hat{v}$ is called the generalized

[6] This method has been used by a number of investigators; see for example, G. Barzilai, "Experimental determination of the distribution of current and charge along cylindrical antennas," Proc. I.R.E., vol. 37, pp. 825–829; July, 1949. Also, M. Aronoff, "Measured phase velocity and current distribution characteristics of helical antennas radiating in the beam mode," an unpublished thesis for the degree of M.Sc., Ohio State University, 1948.

[7] J. A. Stratton, "Electromagnetic Theory," 1st ed., McGraw-Hill Book Co., Inc., New York, N. Y.; 1941.

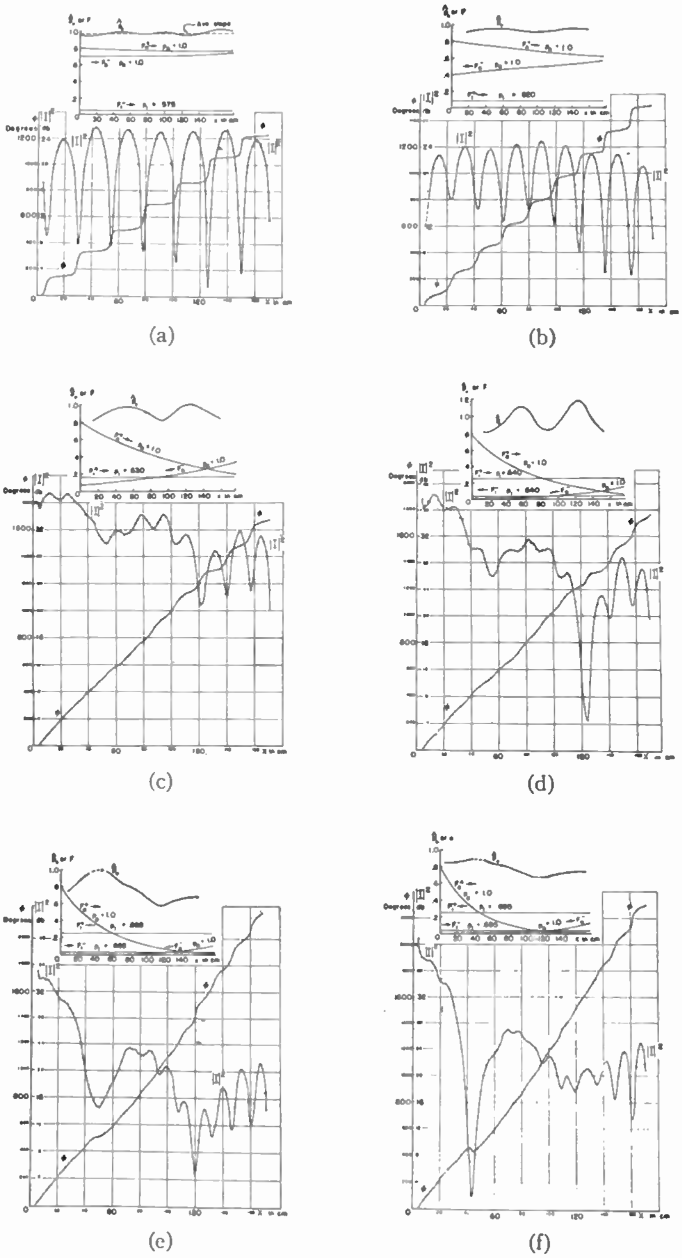


Fig. 3—Measured current distributions on the helix as a function of C_λ, the circumference in wavelengths:
(a) $C_\lambda=0.542$ (b) $C_\lambda=0.665$
(c) $C_\lambda=0.691$ (d) $C_\lambda=0.715$
(e) $C_\lambda=0.774$ (f) $C_\lambda=0.855$.

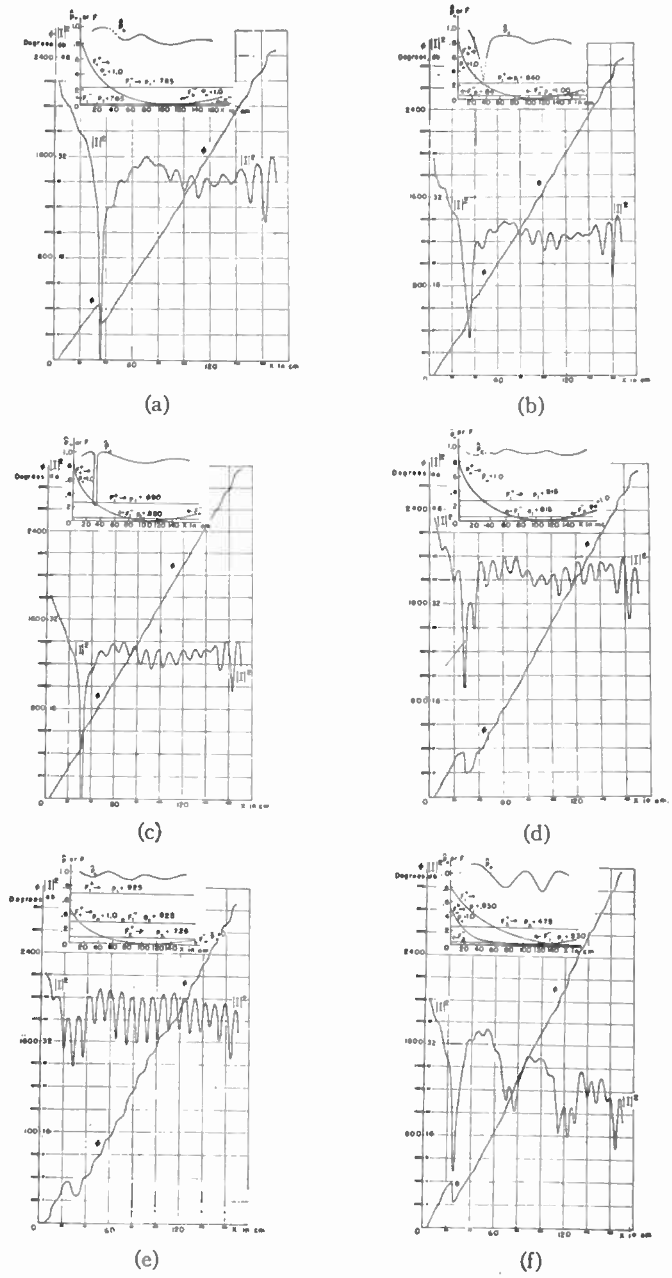


Fig. 4—Measured current distributions on the helix as a function of C_λ, the circumference in wavelengths:
(a) $C_\lambda=1.01$ (b) $C_\lambda=1.09$
(c) $C_\lambda=1.175$ (d) $C_\lambda=1.26$
(e) $C_\lambda=1.36$ (f) $C_\lambda=1.50$.

phase velocity, and the negative sign is necessary because of the convention adopted in plotting ϕ positive for a wave traveling in the positive x direction. Such a generalization leads to many interesting interpretations of phenomena often observed.[8] If ω and $\phi(x)$ are measured or known mathematically, the quantity $\hat{v}$ may be found graphically or calculated directly. Since the phase velocity relative to the velocity of light is usually used to describe the traveling waves in the system, the data are presented in the form $\hat{p}=\hat{v}/c$, where c is the velocity of light and $\hat{p}$ is called the generalized relative phase velocity. Plots of $\hat{p}_+$, the generalized relative phase velocity for those waves traveling in the positive x direction only, may be found in Figs. 3 and 4.

II. Measured Distributions

The antenna investigated (see Fig. 2) is a six-turn helix which has a pitch angle $\Psi=12.6°$, and a mean diameter $D=8.61$ cm. This antenna is mounted on a three-foot-square ground plane which is supported so that the helix is horizontal and about six wavelengths above ground. The current distribution was obtained as a function of distance measured along the slot in the

[8] For a more complete discussion, see J. A. Marsh, "A study of phase velocity on long cylindrical conductors," presented, Graduate School, Ohio State University, 1949. Presented, 1950 National IRE Convention, New York, N. Y., March 7, 1950. Published as a special report, The Ohio State University Research Foundation, February 28, 1950.

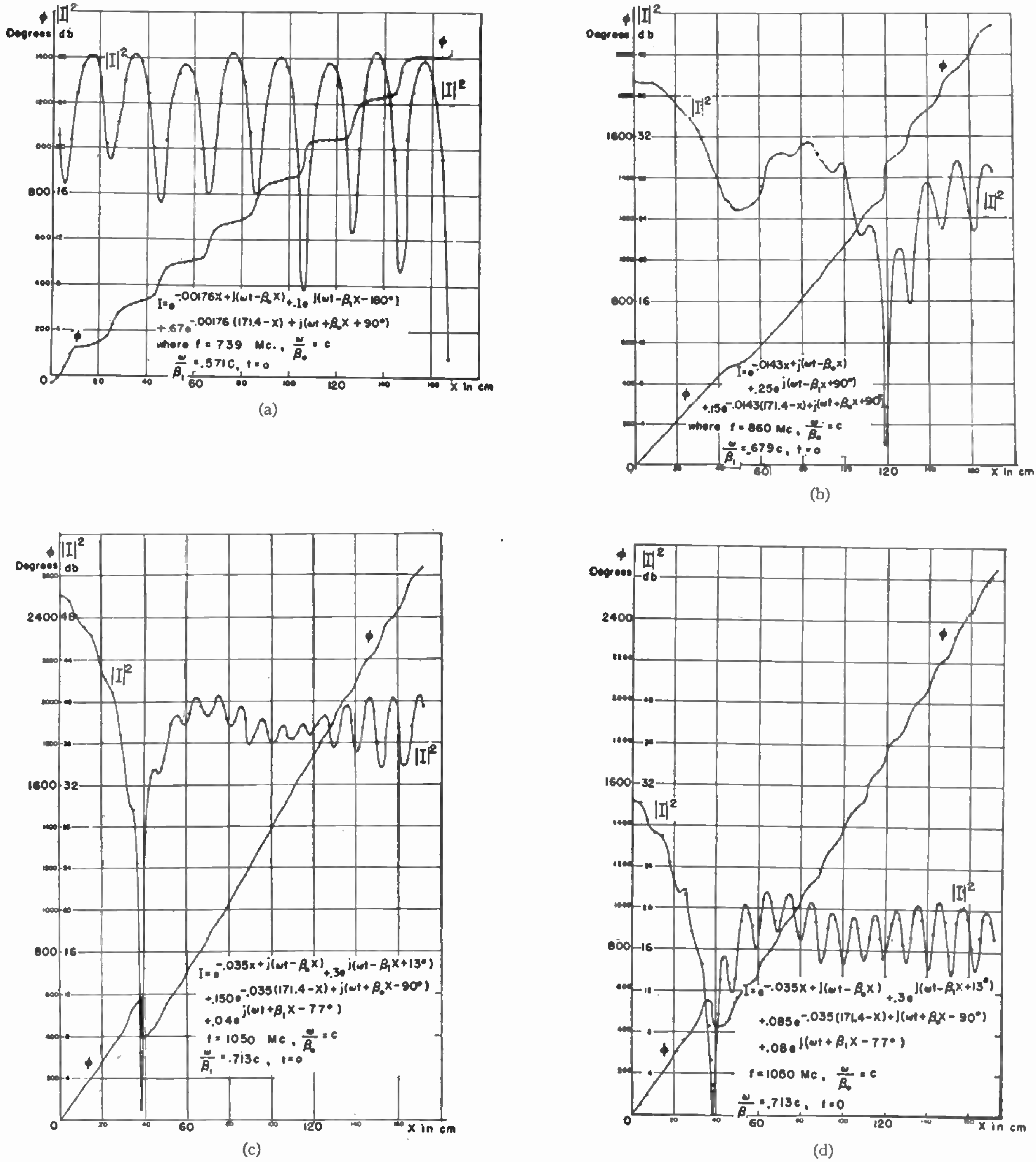


Fig. 5—Calculated current distributions on the helix:
(a) To mode region (compare with Fig. 3(b)).
(b) Transition region from T_0 to T_1 modes (compare with Fig. 3(d)).
(c) T_1 mode region (compare with Fig. 4(a)).
(d) Transition region from T_1 to T_2 modes (Compare with Fig. 4(d) and 4(e)).

hollow tubing using the ground plane as a reference.

The measured current distribution on the helix for the range of frequencies 602 to 1,667 mc may be seen in Figs. 3 and 4. The distributions are labeled in terms of the parameter C_λ, which is defined in Fig. 1 as the circumference in free-space wavelengths of the cylinder with diameter D. The data[9] are arranged in order of increasing C_λ. In each of the distributions shown in

[9] The magnitude of the current I is expressed in decibels relative to an arbitrary reference current I_0 by db $= 10 \log (I^2/I_0^2)$.

Figs. 3 and 4 there is an insert which contains a portion of the analysis made of the data for that distribution. These inserts show an approximate plot of the generalized relative phase velocity, $\hat{p}_+$ for the waves which travel only in the positive x-direction away from the ground plane. The other curves, shown in the inserts, are plots of the amplitude or F functions for the number of one-dimensional traveling waves which may be used to obtain an approximate expression for the current distribution shown. The plus and minus signs are used to show that the associated waves are traveling in the positive or negative x-direction, respectively. It should be stated that these curves are only initial assumptions needed to "fit" the associated distribution. Because of the number of variables involved, the actual task of fitting any one of these curves closely is tremendous. Thus, no attempt was made to "fit" a particular distribution exactly. Four distributions (see Fig. 5) were calculated to show how distributions in four different regions of C_λ could be obtained.

III. Analysis

In the analysis of the data shown in Figs. 3 and 4, it is found that the current distributions can be interpreted in terms of traveling waves associated with three different modes of transmission on the helix. These modes[5] are distinguished by means of the symbols T_0, T_1, and T_2. The T_0 mode of transmission on the helix is found to predominate when $C_\lambda < 0.675$. This is the mode of transmission which is encountered in the traveling wave tube. The T_1 mode of transmission predominates when C_λ lies in the region from 0.80 to 1.30. Since the T_1 mode of transmission on the helix makes possible the axial or beam mode of radiation, it is the center of interest in this investigation. The T_2 mode of propagation exists when $C_\lambda > 1.25$. The presence of energy in this mode is first detected at the upper frequency limit of the helical beam antenna. The subscripts 0, 1, and 2 are used in the inserts of Figs. 3 and 4 to connect the relative amplitude functions and phase velocities of the traveling waves with these transmission modes.

A. The T_0 Mode Distribution

The current distribution shown in Fig. 3(a) is a typical example of the distributions found in the range of frequencies where C_λ varies from 0.348 to 0.600. Traveling waves of current, which may be associated with the T_0 mode of transmission, predominate in this range of frequencies. The amplitudes of the T_0 traveling waves encounter only a slight attenuation and the waves are assumed to travel with the velocity of light. This results in an amplitude characteristic which has a large standing-wave ratio and a phase characteristic which is nearly a step function.

The presence of T_1 mode currents may be detected in the current distributions when $C_\lambda = 0.542$ and 0.665 (see Fig. 3(a), (b)) by observing the undulating nature of the $\hat{p}_+$ curves in the inserts. However, the relative importance of the T_1 mode currents in these two distributions is very slight.

A calculated T_0-mode-region current distribution is presented in Fig. 5(a). This distribution is a first attempt to approximate the measured current distribution shown in Fig. 3(b), when $C_\lambda = 0.665$. Although it is obvious from a comparison of the two distributions that refinements on the first attempt could be made, it is felt that the calculated distribution shows clearly that, by minor manipulations of the available parameters of the three traveling waves, one could obtain calculated and measured distributions which agree within experimental accuracy. The assumed current expression consists of two T_0-mode traveling waves which travel with the velocity of light in opposite directions and have a slight exponential amplitude attenuation. The third, T_1 mode, wave is assumed to be constant in amplitude and to travel with a relative velocity $p_1 = 0.571$.

B. The Transition Region from the T_0 to the T_1 Mode

The transition region from the range of frequencies where the T_0 mode predominates to the range of frequencies where the T_1 mode predominates, occurs in the range $C_\lambda = 0.675$ to $C_\lambda = 0.800$. A study of the measured current distributions in Fig. 3 shows how this transition occurs. A calculated current distribution for this region of transition is given in Fig. 5(b) and is seen to approximate the measured distribution when $C_\lambda = 0.715$.

A comparison of the expressions used to obtain Figs. 5(a) and 5(b) shows that the amplitude of the two waves associated with the T_0 mode become more sharply attenuated as C_λ is increased. This results in less total energy reaching the open end of the helix, and therefore, less energy reflected to create standing waves. The traveling waves associated with the T_1 mode begin to take on more and more importance in this region of transition, not only because of the increased attenuation encountered by the T_0 waves, but also because of the increase in the relative value of the constant amplitude of the T_1 waves compared to the amplitude of the T_0 waves at $x = 0$.

C. The T_1 Mode Distribution

The T_1 or beam mode of radiation from the circular helix may be said to occur in the range of frequencies where C_λ takes on the values 0.8 to 1.3. Figs. 3 and 4 show measured current distributions obtained in this region. A study of these data reveals that although the attenuated waves associated with the T_0 mode are always in evidence they attenuate rapidly so that the T_1 waves predominate a few turns from the feed point. Most of the energy which reaches the open end of the helix is reflected in the form of a T_0 wave, but this also attenuates rapidly leaving only the F_1^+ wave to predominate over most of the helix. This results in only a small reflected wave reaching the feed end of the helix and accounts for the relatively constant input impedance as a function of frequency. Fig. 5(c) shows a calculated current distribution which should be compared with the measured distributions when $C_\lambda = 1.01$ (see Fig. 4(a)).

The amplitude characteristic of the current distribution in the T_1 mode region can best be explained by observing a trend which may be seen in Figs. 3 and 4. When $C_\lambda = 0.691$ (see Fig. 3(c)), there are two minima which occur at $x = 53$ and 123 cm. As C_λ takes on higher values, the minimum, which occurs in the vicinity of $x = 120$ cm, first becomes more pronounced and then disappears when C_λ has values greater than 1.0. The minimum in the vicinity of $x = 50$ cm goes through the same variation and happens to be very pronounced in the distributions which occur in the range of frequencies where the T_1 mode is said to predominate. Similar minima occur again in the distributions when C_λ is greater than 1.4.

These minima can be interpreted as the points where the F_0^+ and F_1^+ waves interfere since the two waves are traveling in the same direction with different phase velocities. The depths of the minima are controlled by the relative magnitudes of the two waves when they are in phase opposition. The increase in attenuation which the T_0 mode currents encounter, as C_λ is increased, causes the minima first to become more pronounced as the initially larger F_0^+ wave attenuates to a value almost equal to the smaller, but constant F_1^+ wave. Then as the attenuation of the F_0^+ wave becomes large, the magnitude of this wave becomes so small compared to the F_1^+ wave at a point of interference that the minima cannot be observed.

The position of these minima, created by the outward traveling waves, is sometimes obscured and other times accentuated by the presence of reflected energy from the open end of the helix. The reflected waves also propagate in at least two modes. When C_λ has values between 1.0 and 1.3, the SWR created by the reflected wave varies in a beatlike fashion. The points of small SWR can be interpreted as points of interference between the F_0^- and F_1^- waves.

Plots are presented in Fig. 6 of the component traveling waves which combine to produce the distribution shown in Fig. 5(c). These plots show how the F_0^+ and F_1^+ waves add to give the total outward traveling current distribution as well as the manner in which the reflected F_0^- and F_1^- waves combine to give the total reflected current distribution. Fig. 6 reveals how the pronounced minimum in the amplitude characteristic is obtained by the interference of the F_0^+ and F_1^+ waves. In addition, one observes in the absence of any reflected waves that there is a very rapid phase advance in the vicinity of the point of maximum interference ($x \cong 1.25\lambda_0$). However, since the outward traveling currents are at such a low level at this point, the reflected energy predominates and causes a reversal of the direction of energy flow over this short segment of the helix. Therefore, the total phase characteristic encounters a dip. It is observed in Figs. 3 and 4 that this effect is encountered many times in measurements; however, occasionally the reflected energy is not large enough to create the phase reversal, and the inherent phase jump is measured. (See Figs. 4(b), 4(c)).

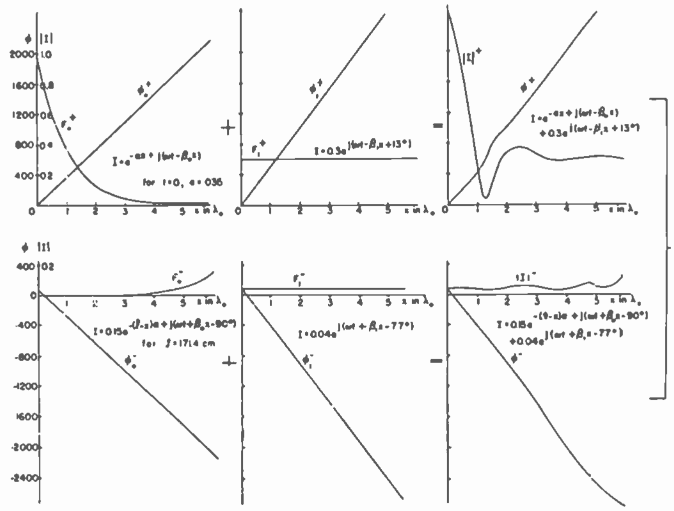


Fig. 5(c)

Fig. 6—Component traveling waves and the manner in which they add to produce a T_1 mode current distribution. (a) Outward traveling waves. (b) Reflected traveling waves.

D. The Transition from the T_1 to the T_2 Mode

The effect of the T_2 mode currents upon the total current distribution may be observed in Fig. 4(e), 4(f), when $C_\lambda = 1.36$ and 1.50. The transition phenomena observed in this range of values for C_λ is quite similar to those observed in the transition region between the T_0 and T_1 modes discussed above; therefore, only a few general observations will be made.

The presence of a T_2 mode on the helix has been detected by earlier investigators.[5,10] However, the radiation pattern of the helix in the range of frequencies where the presence of this mode has been detected[10] has broken up into a large number of lobes. Therefore, the investigation of the helix in this range of frequencies has not been stressed.

A calculated current distribution for the transition region from the T_1 to T_2 modes is illustrated in Fig. 5(d), and should be compared with the measured distributions shown in Figs. 4(d), 4(e), when $C_\lambda = 1.26$ and 1.36.

In the insert of Fig. 4(f), one will observe that the F_1^+ wave has an exponential attenuation. When C_λ is made larger than 1.4, the T_1 mode also develops an

[10] C. K. Bagby, Jr., "A theoretical investigation of electromagnetic wave propagation on the helical beam antenna," an unpublished thesis for the degree M.Sc., The Ohio State University, 1948.

exponential attenuation similar to that of the T_0 mode discussed above.

IV. Phase Velocity

It is apparent from the above analysis that there are always two modes of propagation present on the helix in the range of frequencies where a detailed investigation has been made. This means that there is no single phase velocity associated with the helix, except in certain regions of the helix at certain frequencies. At any given frequency there are always traveling waves which can be associated with two or more different modes, each of which has its own characteristic phase velocity. In analyzing the measured current distributions of Figs. 3 and 4 it is possible to distinguish the different modes present and their relative phase velocities. The results of this analysis may be seen in Fig. 7, where the p versus C_λ curves for the individual modes are given.

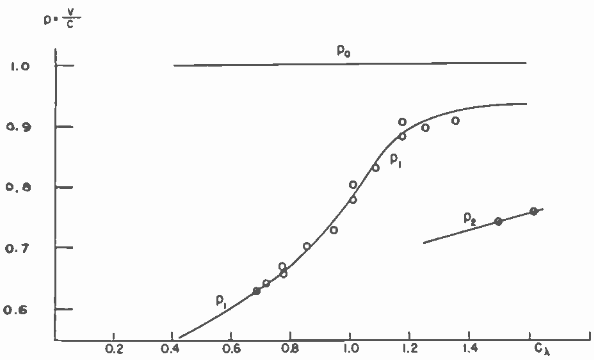


Fig. 7—p versus C_λ curves for the T_0, T_1, and T_2 modes on the helix.

It has been shown by Pierce,[11] and by Chu and Jackson,[12] that propagation along the helix in the T_0 mode has a characteristic phase velocity greater than the velocity of light. This velocity decreases as C_λ increases and approaches closely to the velocity of light when $C_\lambda > 0.7$. Therefore, the relative phase velocity p_0 of the T_0 mode is assumed to be 1.0 at all larger values of C_λ. The maximum error introduced by this assumption occurs for small values of C_λ, and amounts to about 2 per cent when $C_\lambda = 0.7$.

In Fig. 7, one observes that p_1 and p_2 are functions of frequency and are always less than 1.0 in the range of the measurements. The quantities p_1 and p_2 are the phase velocities of the traveling waves associated with the T_1 and T_2 modes, respectively, relative to the velocity of light. The relative phase velocities assigned to each of the traveling waves in the inserts of Figs. 3 and 4 were obtained from Fig. 7. The information available in the inserts of Figs. 3 and 4, therefore, should enable one to calculate, to a good approximation, the far field-radiation pattern of the helix. The radiation pattern of the helix operating in the beam mode has been calculated by Kraus,[2,5] assuming that the current distribution is essentially a single unattenuated outward-traveling wave. This is a satisfactory approximation for a long helix. Kraus uses a phase velocity measured in the central portion of a seven-turn helix and good agreement is obtained between calculated and measured patterns.

The inserts of Figs. 3 and 4 show that the current distribution on the helix when operating in the T_1 mode is essentially the F_1^+ wave. Therefore, in the T_1 mode the p_1 versus C_λ curve should agree fairly closely with a measured p versus C_λ curve obtained by Kraus.[13] The two curves are plotted in Fig. 8 for comparison. The maximum directivity line also shown in Fig. 8 was obtained from an expression derived by Kraus.[5] This expression

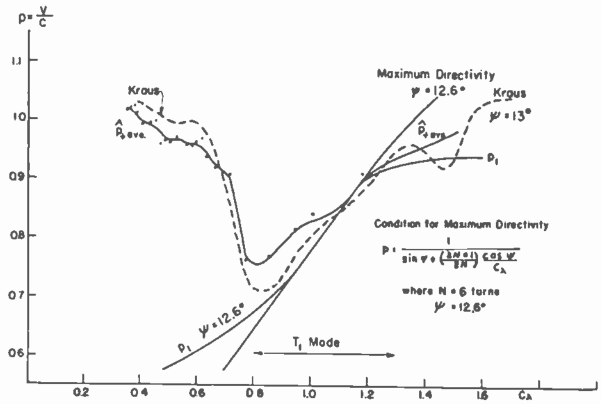


Fig. 8—p curves for the uniform circular helix.

gives the value p should have for the radiation from an array of helical turns, considered as point sources, to add up along the axis of the helix in such a manner as to produce maximum directivity in the axial direction. In Fig. 8, it is observed that the p_1 curve, the maximum directivity curve, and Kraus' curve are all in close agreement within the region of the T_1 mode.

Since the task of calculating the radiation pattern for the helix is laborious at best, it would be desirable to reduce to at least two the number of traveling waves considered, even outside the region of the T_1 mode. In the T_1 mode, the F_1^+ wave alone may be used to calculate the patterns as has been done by Kraus. However, at both ends of the T_1 mode region and beyond, the traveling waves associated with the other modes cannot be neglected.

To facilitate calculations, it is suggested that all of the energy which flows in the positive x direction be considered as one traveling wave, which has a phase velocity that is weighted in terms of the relative magnitudes of the component traveling waves which it is approximating. The energy which flows in the negative x direction can be treated in the same manner. The

[11] J. R. Pierce, "Theory of the beam-type traveling wave tube," Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.

[12] L. J. Chu and D. Jackson, "Field theory of traveling wave tubes," Proc. I.R.E., vol. 36, pp. 853–863; July, 1948.

[13] J. D. Kraus, "Antennas," McGraw-Hill Book Co., Inc., New York N. Y., in press; see Figs. 7–19.

generalized phase velocity, defined above, is such a weighted phase velocity. In the inserts of Figs. 3 and 4, one will notice curves which are labeled p_+. These curves are plots of the generalized phase velocity for the energy traveling in the positive x direction only. These curves were obtained from calculations made on the associated measured current distribution. In drawing the p_+ curves, the regions where the associated ϕ versus x curve dips were omitted for two reasons: first, the peculiar behavior in the region of the phase dips is almost always created by the reflected wave and, therefore, should not be allowed to affect p_+; second, the current in this region is always down in amplitude from the maximum by a factor of 20 db and the effect this small segment of the helix can have on the over-all radiation must be negligible.

The shape and trends shown by these p_+ curves are interesting. Each of the four regions discussed have p_+ curves which may be considered characteristic of that region. In the region of the T_0 mode, the p_+ curve is fairly constant and has a value between 0.95 and 1.0. The p_+ curve has an oscillatory nature in the regions of transition from one mode to another. The oscillations are similar to those observed when two waves are traveling at different velocities in the same direction. The center of oscillation, of course, shifts down as the T_1 mode currents tend to predominate. The p_+ curves in the T_1 mode region are characteristically high for low values of x where the T_0 mode currents are still large. As the value of x increases, the p_+ curve settles down to the value p_1 for most of the remaining length of the helix. The p_+ curves are often more sensitive indications of the presence of other modes than are the amplitude and phase curves themselves.

Each of the generalized p_+ curves was integrated graphically and the average value plotted in Fig. 8. The curve drawn through these points is labeled $p_{+\text{ave}}$. The values of p, obtained from this curve, can be used as a weighted phase velocity for a single traveling wave approximation of the energy traveling in the positive x direction. It is observed that this curve follows roughly Kraus' curve in Fig. 8. Differences between these two curves should be expected since the $p_{+\text{ave}}$ curve is an average taken over the entire helix, whereas Kraus' curve was measured in the central region of a seven-turn helix.

V. Charge and Current on a Long Helix

The above analysis has been made almost entirely on the basis of data obtained from the helix shown in Fig. 2. As a check on this analysis, a second helix was constructed similar to the antenna of Fig. 2 but with 3.75 additional turns.

From the analysis made on the shorter helix, one should be able to predict the effect of extending the helix in length. For example, when $C_\lambda = 0.774$ (see Fig. 3(e)), the total current distribution is relatively free of any reflected energy when $x < 80$ cm. Extending the length of the helix at this frequency, therefore, should not alter the current distribution appreciably in the region $x < 80$ cm. However, by making the helix longer one would expect the amplitude curve in the vicinity of $80 \leq x \leq 160$ cm to be smoothed out since the attenuated reflections would occur at some point beyond $x = 172$ cm.

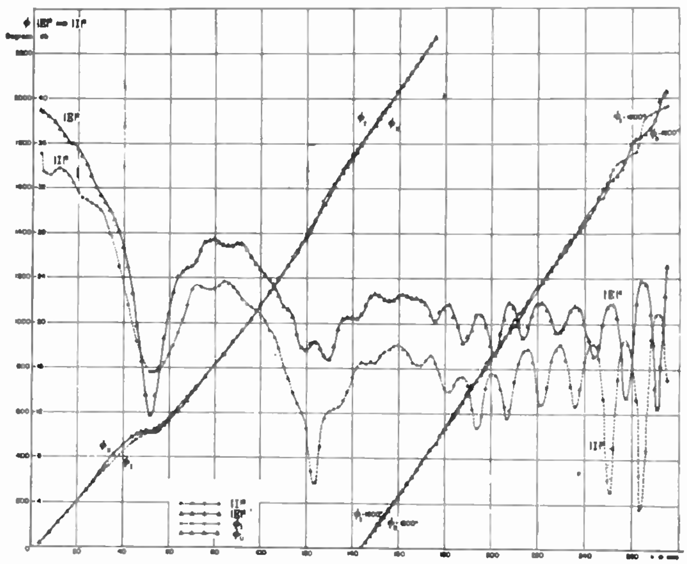


Fig. 9—Measured current and charge distribution on a 9.75-turn helix, pitch angle $\psi = 12.6°$, mean diameter $D = 8.61$ cm for $C_\lambda = 0.774$.

The current distribution shown in Fig. 9 was measured on the long helix, 9.75 turns, for $C_\lambda = 0.774$. Comparing Fig. 9 with Fig. 3(e) for the six-turn helix at the same value of C_λ, one finds that all of the predictions are fulfilled.

As a further check on the analysis, a charge or E field measurement as a function of distance along the long helix was made and plotted in Fig. 9. Comparing the E and H field measurements shown in Fig. 9, one observes that for low values of x, where the outward traveling waves predominate, the E and H fields are in phase and have the same general shape. However, as x takes on larger values, the effect of the reflected energy causes the E and H fields to approach a relative phase quadrature. That is, the two phase characteristics take on an interlaced appearance, a maximum E field amplitude being found where a minimum H field is observed, and vice versa.

VI. Acknowledgment

The advice and encouragement of J. D. Kraus, and the assistance received from the staff of the Ohio State University Research Foundation Antenna Laboratory, are gratefully acknowledged.

An Atmospherics Waveform Receiver*

WILLIAM J. KESSLER†, ASSOCIATE, IRE, AND SYDNEY E. SMITH†

A KNOWLEDGE of the instantaneous field-intensity variations, or waveforms, of atmospherics noise provides a useful starting point for studying the characteristics and mechanism of lightning discharges, as well as many radio propagational phenomena.

The present summary constitutes a brief description of a driven-sweep oscillograph of special design for the observation of atmospherics waveforms. A 15-foot vertical antenna coupled to a combination signal-power dual coaxial transmission line through a cascaded cathode-follower impedance transformer is provided to permit locating the antenna in a remote interference-free area.

The signal amplifier employs resistance-capacitance parallel-T filter sections exhibiting transmission nulls at one megacycle and at 60 cycles per second to eliminate interfering signals due to radio-broadcast transmitters and further reduce the interference from stray power-line fields. An 18-microsecond delay line permits delineation of the entire leading edge of the atmospherics waveform. A sweep control circuit is arranged to trigger the horizontal sweep generator, regardless of the polarity of the initial maximum of the atmospherics. Limiting of the display-rate to about eight waveforms per second prevents overlapping of the waveforms on the photographic record.

A marker generator permits the measurement of time intervals and provides a continuous check upon the stability and linearity of the sweep generator. The display of the markers, spaced either 50 or 100 microseconds, is unique in that they appear on a second sweep displayed 2 inches above, rather than superimposed upon the atmospherics waveforms. Confusion of the markers with the characteristic variations of the waveform is thus prevented. This display is accomplished with a cathode-ray tube containing a single electron gun by blocking the signal channel during the application of the marker sweep which is delayed approximately 1,500 microseconds after the end of the signal sweep.

The atmospherics waveform receiver was developed to implement a research program to investigate the possibilities of thunderstorm location through waveform analysis of the associated atmospherics. Further applications may be found in other geophysical areas, such as the mapping of subterranean strata with the aid of artificially injected earth shocks or the investigation of the lower ionospheric regions employing atmospherics as natural pulses. The application of the atmospherics waveform receiver for ionospherics research is of particular interest because of the information provided on the ionosphere in the low-frequency region down to 10 or 20 kc.

Fig. 1 is an atmospherics waveform observed during the night. Successive pulses which are approximate replicas of the parent pulse that have experienced multiple reflections between the lower regions of the E layer and the surface of the earth are clearly visible. Processing of the time-intervals between successive reflections observed in Fig. 1 through a graphical procedure based on a method of least squares yields a great circle distance of approximately 1,175 kilometers to the discharge source and a mean reflection height of about 90 kilometers.

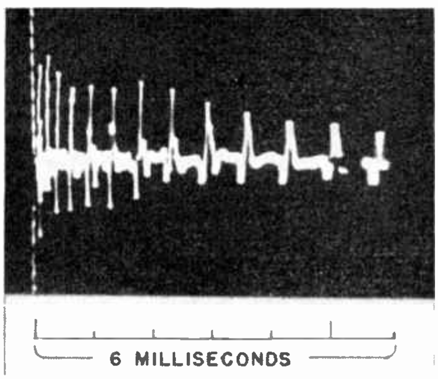
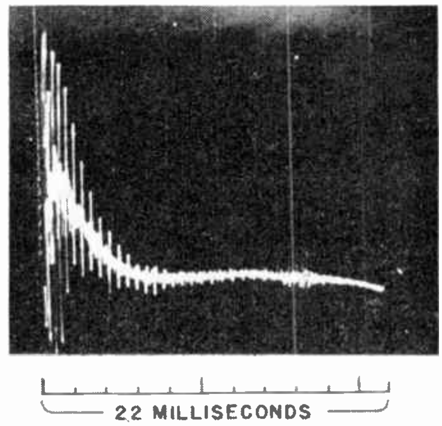
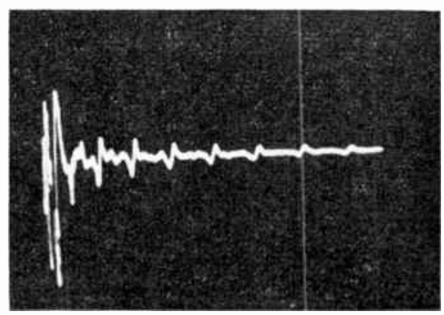


Fig. 1—Characteristic atmospherics waveform recorded during night.

Examples of multiple returns from lightning discharges associated with an overhead thunderstorm in Gainesville are shown in Fig. 2(a). Approximately 20 vertically-incident returns are shown which exhibit a delay of about 610 microseconds between consecutive returns corresponding to an effective reflection height of about 91 km. Fig. 2(b) shows another waveform recorded during similar conditions with an expanded time scale. The effects of energy absorption and the small degree of dispersion are evident.



(a)



(b)

Fig. 2—Multiple returns from overhead thunderstorm; (a) 22-millisecond sweep; (b) 5-millisecond sweep.

* Decimal classification: R272.1. Original manuscript received by the Institute. May 10, 1950; abstract received, October 11, 1950. Presented, IRE National Convention, New York, N. Y., March 9, 1950.

The development described in this paper was carried out under an Atmospherics Research Contract No. W28-003-sc-1306 between the Evans Signal Laboratory, Belmar, N. J., and the Engineering and Industrial Experiment Station, University of Florida.

† University of Florida, Gainesville, Fla.

On the Response of a Directive Antenna to Incoherent Radiation*

FREDERICK W. SCHOTT†, MEMBER, IRE

Summary—The relative response of directive antennas is dependent on the mechanism whereby the energy is conveyed. This fact, which often seems to be neglected, is of particular importance when the energy arriving at a receiving antenna can be represented in terms of a spectrum of incoherent plane-wave components. Such may be true frequently in the extraoptical propagation of microwaves.

A study of an idealized situation is presented which compares the behavior of several pencil-beam antennas of varying aperture. A brief comparison with some recent observations is made.

Introduction

THE CONCEPT of antenna gain is a useful tool for computing the power received by an antenna when the incident radiation consists of a single, substantially plane, wave. However, when the incoming radiation is of a more complex nature as, for example, when due to an aggregate of scattering elements randomly located, the gain concept must be applied with some caution.

The purpose of this paper is to indicate the manner in which the power received by a directive antenna depends upon the distribution of the incident radiation and to illustrate the result with an application to a pencil-beam antenna.

The Weighted Antenna Gain

A directive antenna whose axis is chosen as the reference direction is shown in Fig. 1. A parabolic antenna is shown; however, the form of the antenna is immaterial

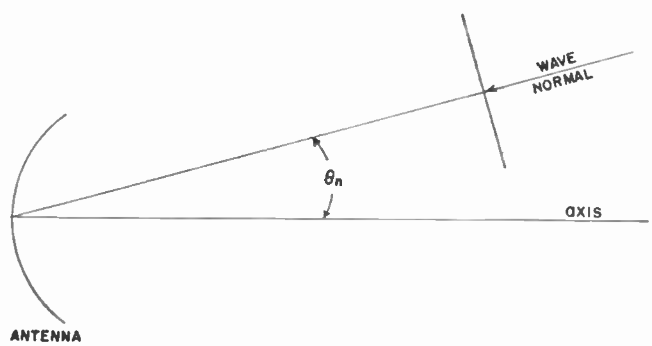


Fig. 1—Directive antenna and associated notation.

at this point. It is supposed that the incident radiation consists of many plane-wave components, and it is assumed that the nth component is arriving from the direction θ_n, ϕ_n, where θ_n represents the colatitude and ϕ_n the azimuth. Because of the directive characteristics of the antenna, this wave produces a component of voltage of relative amplitude a_n at the antenna output, i.e.,

$$a_n = a(\theta_n, \phi_n).$$

The resultant amplitude (voltage) at the antenna output due to the many plane waves is then

$$E = \sum_{n=1}^{N} a_n e^{j\alpha_n}, \tag{1}$$

where α_n represents the phase of the nth component relative to some arbitrary reference and is assumed to be *random*.

The power associated with this resultant voltage (except for a constant of proportonality) is then

$$S = |E|^2 = \left| \sum_{n=1}^{N} a_n e^{j\alpha_n} \right|^2 \tag{2}$$

and the average power

$$\overline{S} = \sum_{n=1}^{N} a_n^2 \tag{3}$$

because of the random nature of the α's. If the plane-wave components are sufficiently numerous, the above summation over n can be represented by a corresponding integration over the appropriate directions. Since the element of area is proportional to $\sin \theta d\theta d\phi$, the average is given by[1]

$$\overline{S} = C \int_0^{2\pi} \int_0^{\pi} a^2(\theta, \phi) \sin \theta d\theta d\phi, \tag{4}$$

where C is a constant and $a(\theta, \phi)$ now represents the probability per unit area that the component of amplitude a is associated with the direction θ, ϕ.

Because $a(\theta, \phi)$ is a measure of the relative amplitude at the antenna output, it represents the effect of antenna directivity as well as the directional character of the radiation itself. Hence

$$a^2(\theta, \phi) = a_1^2(\theta, \phi)G(\theta, \phi),$$

where $a_1(\theta, \phi)$ is the relative amplitude of the plane-wave component in the direction of polarization of the antenna arriving from the direction θ, ϕ, and $G(\theta, \phi)$ is the gain function[2] of the antenna in a homogeneous medium.

If an isotropic antenna were used, the average received power would be given by

* Decimal classification: R125×R326.8. Original manuscript received by the Institute, May 9, 1950; revised manuscript received, October 2, 1950. Presented, IRE-URSI Symposium on Antennas and Propagation, San Diego, Calif., April 3, 1950.

† University of California, Los Angeles, Calif.

[1] See P. M. Morse, "Vibration and Sound," McGraw-Hill Book Co., New York, N. Y., p. 383; 1948.

[2] As defined, for example, by S. Silver, "Microwave Antenna Theory and Design," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 12, p. 3; 1949.

$$\bar{S}_I = C \int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) \sin \theta d\theta d\phi. \qquad (5)$$

Thus, the power received by the directive antenna relative to that received by an isotropic antenna is

$$\frac{\bar{S}}{\bar{S}_I} = \frac{\int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) G(\theta, \phi) \sin \theta d\theta d\phi}{\int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) \sin \theta d\theta d\phi}, \qquad (6)$$

which might be termed the "weighted" gain. This rather general result must include, as a special case, that of radiation arriving from a single direction, θ_1, ϕ_1. In the latter case the directivity of the incident radiation may be represented by the delta function, i.e., $a_1(\theta, \phi) = \delta(\theta - \theta_1, \phi - \phi_1)$, giving

$$\frac{\bar{S}}{\bar{S}_I} = G(\theta_1, \phi_1),$$

as required.

It should be noted that the result using (6) is determined by the method used to illuminate the scattering elements as well as their own directivity characteristics, since this determines $a_1(\theta, \phi)$. An analysis by Booker and Gordon,[3] based on the work of Pekeris,[4] has given explicit results for certain cases of forward scattering due to atmospheric inhomogeneities.

Examples

As an illustration of the use of (6), the following case will be considered: It will be assumed that the antenna is viewing the "center" of a scattering region such that $a_1(\theta, \phi) = a_1(\theta)$ only, and furthermore that the incident plane-wave components arrive uniformly from all directions within the cone semiangle θ_0, and that no radiation is arriving from outside this region. This is indicated in Fig. 2. In addition, it will be assumed that

$$G(\theta, \phi) = G(\theta)$$

only. Thus (6) becomes

$$\frac{\bar{S}}{\bar{S}_I} = \frac{\int_0^{\theta_0} G(\theta) \sin \theta d\theta}{\int_0^{\theta_0} \sin \theta d\theta}. \qquad (7)$$

The assumptions leading to (7) would be, in general quite severe; however, there are instances where highly directive antennas are used in which such restricted results may be of value.

To obtain specific results, four parabolic antennas having apertures of 12, 18, 30, and 48 inches, respectively, have been studied, using (7). For each antenna the gain function in the E and H planes was measured at 9,375 mc and the average used as $G(\theta)$. This procedure is appropriate[5] inasmuch as the significant portion of the pattern is nearly identical in the two planes.

The results, expressed as a function of θ_0, are shown in Fig. 3.

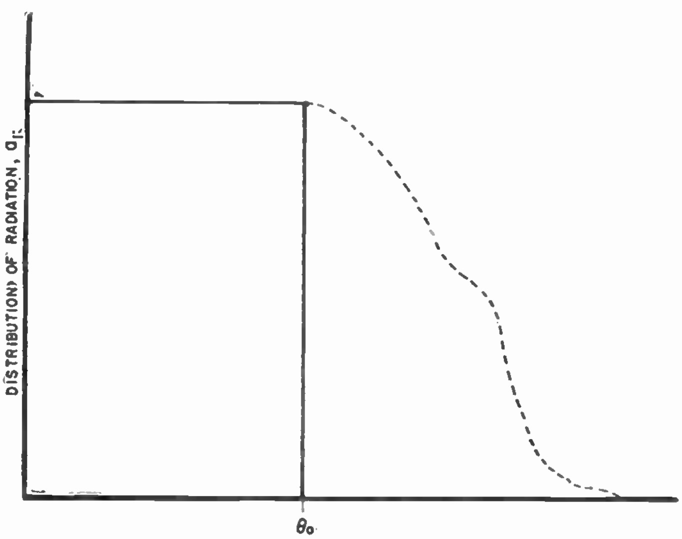


Fig. 2—A theoretical distribution of incident radiation.

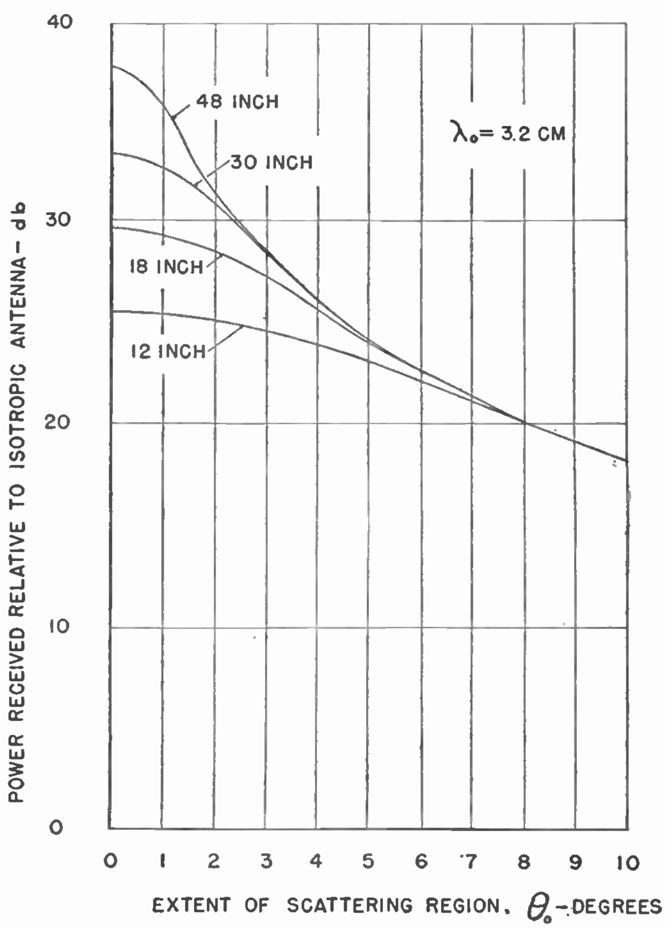


Fig. 3—The response of parabolic antennas to scattered radiation.

[3] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," Proc. I.R.E., vol. 38, pp. 401–413; April, 1950.

[4] C. L. Pekeris, "Note on scattering in an inhomogeneous medium," Phys. Rev., vol. 71, p. 268; February, 1947.

[5] See page 413 of footnote reference 2.

It may be seen that if the radiation is arriving from a region which is sufficiently large, i.e., if θ_0 is large relative to the beam width, the received power is independent of the antenna aperture. Furthermore, in a case such as this, where the minor lobe pattern is insignificant, the incident radiation could vary in any manner outside the main portion of the beam, such as indicated by the dotted curve of Fig. 2, for example, without invalidating the above observation.

Microwave propagation experiments in the Arizona Desert have indicated that in many cases scattering may have been the dominant mechanism involved, and recent studies have been made to learn more of the nature of these scattered fields. Although a complete analysis of these data will follow in a later paper, two preliminary observations may be noted:

1. Only during the evening, under conditions of a strong temperature inversion, or during the day when high in the diffracted region, was the average received power proportional to the gain of the receiving antenna. At these times the fields did not fluctuate greatly.

2. During the daytime with low antenna heights, when rapid scintillations of large amplitude were noted, the average received power would usually increase when a higher gain antenna was used, but never in proportion to the gain increase. In a typical situation, changing the aperture from 18 to 30 and from 30 to 48 inches corresponded to measured gain increments of 3.7 and 4.5 db. whereas the observed power increased in steps of 2.1 and 2.4 db, respectively. This is illustrated in the lower portion of Fig. 4. The upper data represent a less common but more striking example of this phenomenon. (The markers at the left indicate 10-db intervals.)

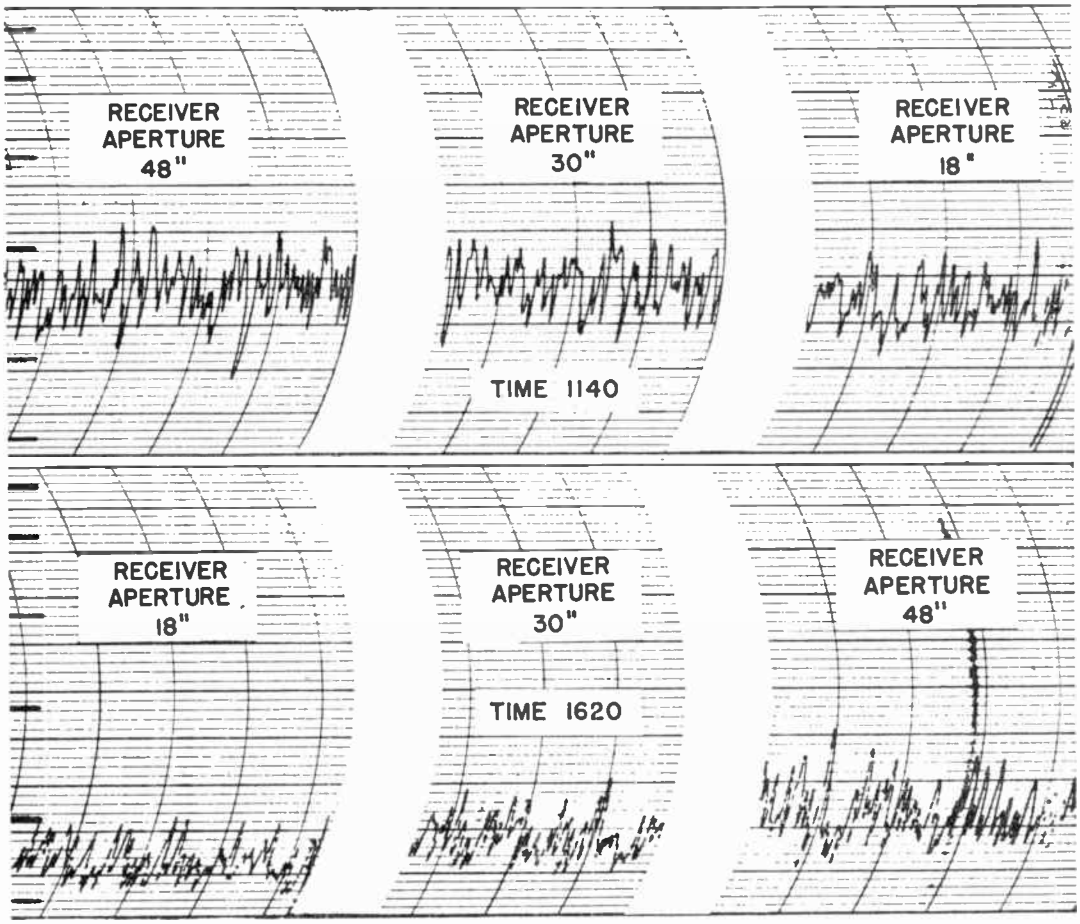


Fig. 4—Relative response of directive antennas. (March 6, 1950; 9,340 mc; transmitter aperture, 18 inches; transmitter and receiver cab elevation, 0 foot; separation, 46.2 miles.)

Conclusions

There are many situations where an arriving wave can be represented only by a spatial spectrum of essentially incoherent plane wave components. This seems to be true often in the extraoptical propagation of microwaves,[6] and it is felt to be of significance in the optical range in the presence of appreciable rain attenuation.

[6] Recent experimental results presented by A. W. Straiton and D. F. Metcalf at the IRE-URSI Symposium on Antennas and Propagation held in San Diego, during April 3–5, 1950, suggest that extraoptical propagation at FM frequencies may also be due to atmospheric scattering.

When it is desired to predict the manner in which a particular receiving antenna will perform under such circumstances, considerations of the type indicated may be appropriate.

In this connection it must be emphasized that the frequently used engineering procedure for expressing the power associated with an incident field may be misleading and erroneous. It is quite common to find fields expressed in decibels relative to a particular microvolt-per-meter level.[7] This is accomplished by the use of a calibration which is appropriate when the wave front is essentially plane; however, this same calibration is then used *regardless of whether or not the radiation is known to be of this type.*

If the distribution of the incident radiation is unknown, one can only state the power received by the particular antenna configuration used.

[7] E. W. Allen, W. C. Boese, and H. Fine, "Summary of Tropospheric Propagation Measurements and the Development of Empirical VHF Propagation Charts (Revised)," Federal Communications Commission, T.I.D. Report No. 2.4.6.; May 26, 1949.

Optimum Operation of Echo Boxes*

W. M. HALL†, SENIOR MEMBER, IRE, AND W. L. PRITCHARD†, MEMBER, IRE

Summary—The parameters determining ring time for an echo box in a radar system are examined, and the conditions for optimum ring time established. Two specific arrangements, namely, the echo box coupled directly to the guide and the echo box coupled through a directional coupler, are treated in detail.

Representative curves of ring time as functions of various parameters, and curves of power extracted by the echo box are included. It will be seen that simply increasing the loaded Q of a cavity will not increase the ringing time past a certain maximum, and that there is an optimum value of loaded Q which is a function of the radar power, receiver sensitivity, pulse length, and the method of coupling the cavity to the transmitter.

I. Introduction

The use of echo boxes in checking over-all radar system performance is well known.[1,2] Although the design of the cavity per se has been treated in considerable detail,[1,3] and the general operation of echo boxes and other resonant cavities discussed,[2,4,5] the full significance of the factors involved in obtaining maximum ring time from a resonant cavity used in conjunction with a particular radar is not too readily apparent. It is the purpose of this paper to consider in detail these factors. Two methods of coupling the echo box to the waveguide are considered. In one, the echo box cavity is mounted directly on the side of the guide. In the other, it is mounted on the end of a secondary guide, which in turn is coupled to the main guide through a directional coupler.

The behavior of the echo box is calculated for the following assumed conditions. The transmitted pulse is assumed to be of constant amplitude and frequency throughout its duration; i.e., the rise and decay times are assumed negligible, compared to the pulse duration, and the pulse is assumed to be free from frequency modulation. The transmission line is assumed to be terminated in its characteristic impedance, both at the antenna and at the transmitter-receiver. If a particular system is known to depart significantly from the above assumed conditions, corrections may be made in the calculations, but the same general principles should apply. For example, if the transmitter is known to be frequency modulated, the amount of energy available to an echo box tuned to a specific frequency will be reduced, and so forth.

As in the case of any other measuring device, it is desirable that the effect of the measuring device on the performance of the system be negligible. Hence, it is advisable that the introduction of the echo box into the system have no appreciable effect, either on the impedance seen by transmitter or on the power delivered to the antenna. This requirement may preclude the adjustment for maximum ring time. The limitations are discussed in some detail below.

II. List of Symbols

R = equivalent cavity resistance
C = equivalent cavity capacitance
L = equivalent cavity inductance
Z_0 = characteristic impedance of transmission lines
e = voltage across cavity
e_{min} = minimum detectable value of e
E = transmitter voltage across matched line
R_g = generator impedance in equivalent circuit
E_g = generator voltage in equivalent circuit
e_r = voltage across receiver
K = voltage coupling factor of directional coupler
Q_L = loaded Q of cavity
Q_0 = unloaded Q of cavity
a = length of exciting pulse
ω_0 = resonant angular frequency
S_{-1} = unit step function

* Decimal classification: R211.112×R374.112. Original manuscript received by the Institute, April 12, 1950; revised manuscript received, August 28, 1950.

† Raytheon Manufacturing Company, Newton, Mass.

[1] I. G. Wilson, C. W. Schramm, and J. P. Kinzer, "High Q resonant cavities for microwave testing," *Bell Sys. Tech. Jour.*, vol. 25, pp. 408–434; July, 1946.

[2] E. I. Green, H. J. Fisher, and J. G. Ferguson, "Techniques and facilities for microwave radar testing," *Bell Sys. Tech. Jour.*, vol. 25, pp. 435–482; July, 1946.

[3] J. P. Kinzer and I. G. Wilson, "Some results on cylindrical cavity resonators," *Bell Sys. Tech. Jour.*, vol. 26, pp. 410–445; July, 1947.

[4] C. G. Montgomery, "Technique of Microwave Measurements," vol. 11, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., p. 341; 1947.

[5] "Notes on Lectures of W. W. Hansen," Radiation Laboratory, (limited classified printing), chap. 12; 1943.

p = complex variable of Laplace transform
t = time
$j = \sqrt{-1}$
T = ring time of cavity measured from end of pulse
F = voltage "dynamic range" constant of transmitter-receiver combination
M = ratio of power taken by echo box to power delivered to load
A = ratio of average power into echo box, short pulse, to average power, long pulse.

III. Basic Coupling Systems

Figs. 1 and 2 are schematic representations of the two basic types of coupling systems herein considered. Fig. 1 shows a cavity coupled directly to the line, and Fig. 2 shows a directional-coupler type of connection. The lumped circuit equivalent of both of these two circuits is shown in Fig. 3.

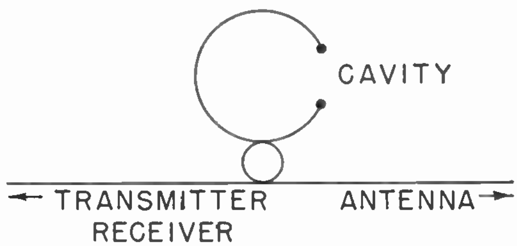


Fig. 1—Echo box coupled directly on line.

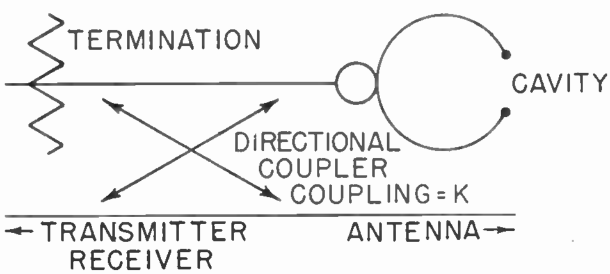


Fig. 2—Echo-box system using directional coupler.

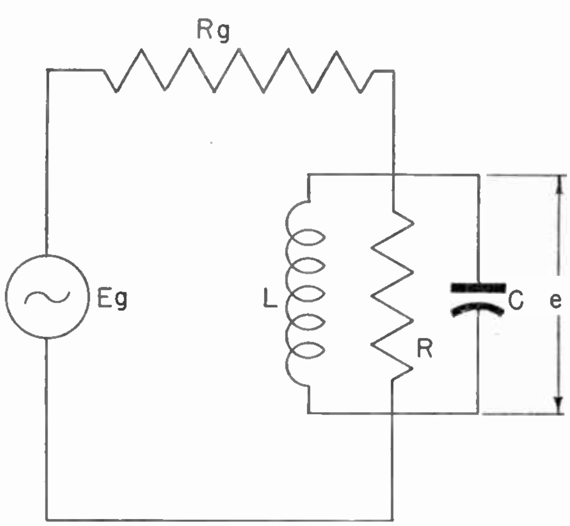
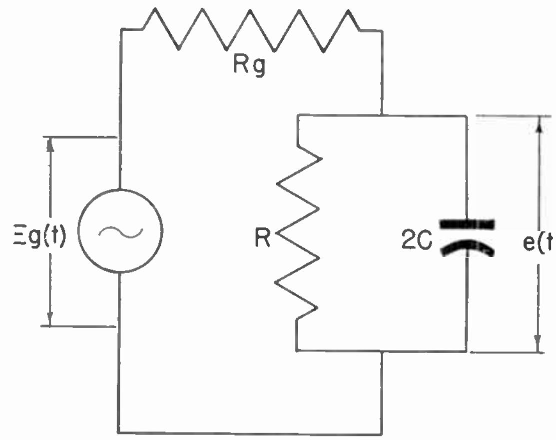


Fig. 3—General equivalent circuit.



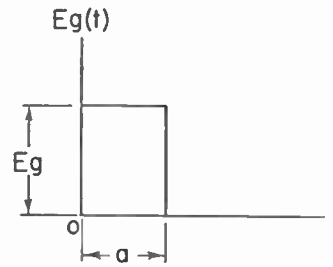


Fig. 4—Low-pass analog.

The transmitter can be treated as an open-circuit voltage $2E$ in series with an internal impedance Z_0; the voltage across the matched line to the antenna is then E, and the equivalent generator voltage in the auxiliary line of Fig. 2 is E/K. If the voltage across the cavity at any instant is e, the voltage across the receiver, during the decay of the "echo" is $e_r = e/2$ in Fig. 1 and $e_r = e/K$ in Fig. 2.[5]

The following table gives the equivalent circuit values for each case:

TABLE I

Equivalent Circuit Fig. 3	Cavity on Line Fig. 1	Directional Coupler Fig. 2
R_g	$2Z_0$	Z_0
E_g	$2E$	E/K
e	$2e_r$	Ke_r

The basic relations derived from elementary resonant circuit theory, between the parameters of the circuit shown in Fig. 3 and the resonant angular frequency ω_0, loaded ϕ_L, and unloaded ϕ_0 of the cavity are

$$Q_0 = R\sqrt{\frac{C}{L}} \tag{1}$$

[5] If the echo box is coupled to the radar system by means of an auxiliary antenna, instead of being coupled to the transmission line, the analysis can be made the same as for the directional coupler case by letting the equivalent generator voltage in the transmission line from the auxiliary antenna to the echo box equal E/K.

$$Q_L = \frac{R_g R}{R_g + R}\sqrt{\frac{C}{L}} \tag{2}$$

$$\omega_0 = \frac{1}{\sqrt{LC}} \tag{3}$$

$$\frac{R}{R + R_g} = 1 - \frac{Q_L}{Q_0} \tag{4}$$

$$\frac{R_g + R}{2RR_gC} = \frac{\omega_0}{2Q_L}. \tag{5}$$

In order to analyze the response of the above circuit to a pulse-modulated sine wave, we make a conventional band-pass to low-pass transformation to arrive at the circuit of Fig. 4.[7] This transformation is permissible because of the small cavity bandwidth compared to its center frequency and the long pulse length compared to the period of the radio-frequency carrier. This circuit is now analyzed for its response to a rectangular pulse.

IV. Calculation of Ringing Time

This analysis is performed by conventional Laplace transform methods. The transform of a rectangular pulse of width a and height E_g, using the "real translation" theorem, is given as

$$E_g(p) = \frac{E_g}{p}(1 - \epsilon^{-ap}). \tag{6}$$

Standard network theory yields as the transform of the output voltage

$$e(p) = \frac{\dfrac{E_g}{2R_gC}(1 - \epsilon^{-ap})}{p\left(p + \dfrac{R + R_g}{2RR_gC}\right)}. \tag{7}$$

The inverse transform is found from tables of Laplace transforms, and (4) and (5) enable us to write

$$e(t) = E_g\left[1 - \frac{Q_L}{Q_0}\right][1 - \epsilon^{-\omega_0/2Q_L} - S_{-1}(t - a)(1 - \epsilon^{-\omega_0/2Q_L(t-a)})]. \tag{8}$$

Fig. 5 is a graphical picture of (8) and shows the effects of pulse length and coupling on the maximum amplitude of the energy in the cavity and on the ring time. The factor $S_{-1}(t-a)$ becomes equal to one for $t>a$. We define $t'=t-a$ and ring time T as the value of t' when $e(t')$ is the minimum detectable, $e_{\min}$. Equation (8) is then solved for T.

$$T = \frac{2Q_L}{\omega_0}\ln\left[F\left(1 - \frac{Q_L}{Q_0}\right)(1 - \epsilon^{-(\omega_0/2Q_L)a})\right], \tag{9}$$

where $F = E_g/e_{\min}$.

[7] Ernst A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. II; 1935.

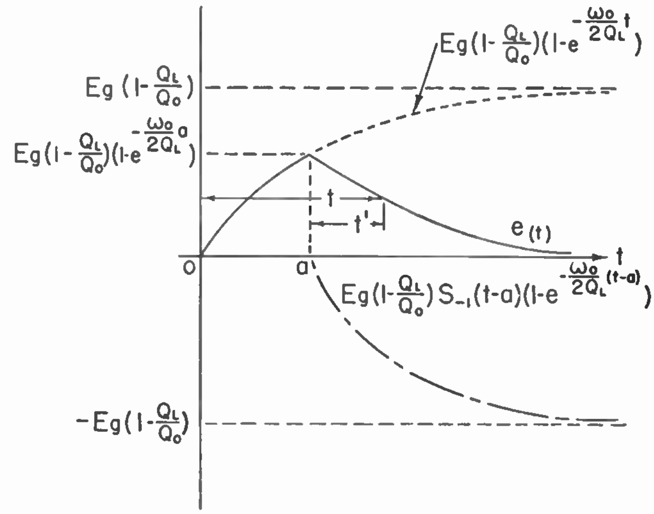


Fig. 5—Graphical interpretation of (8).

Equation (9) should be differentiated with respect to Q_L/Q_0 and the derivative set equal to zero to determine the optimum ratio, but this is an awkward procedure because of the unwieldy transcendental equations. Hence families of curves have been plotted using values of the parameters representative of a wide range of typical radar systems. These curves are shown in Figs. 6 through 8.

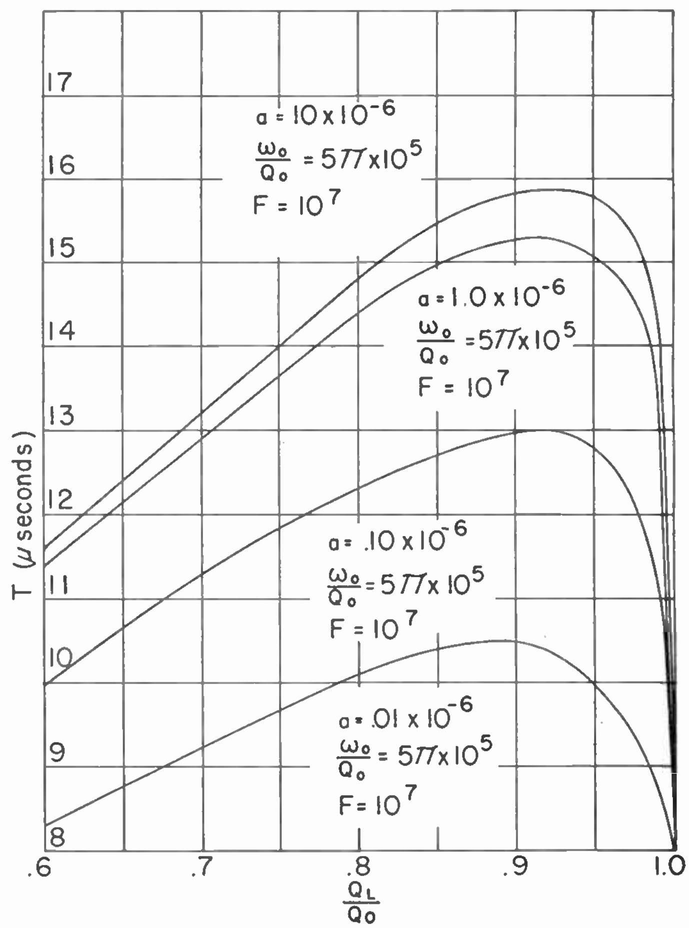


Fig. 6—Ring time versus Q_L/Q_0 for different pulse lengths.

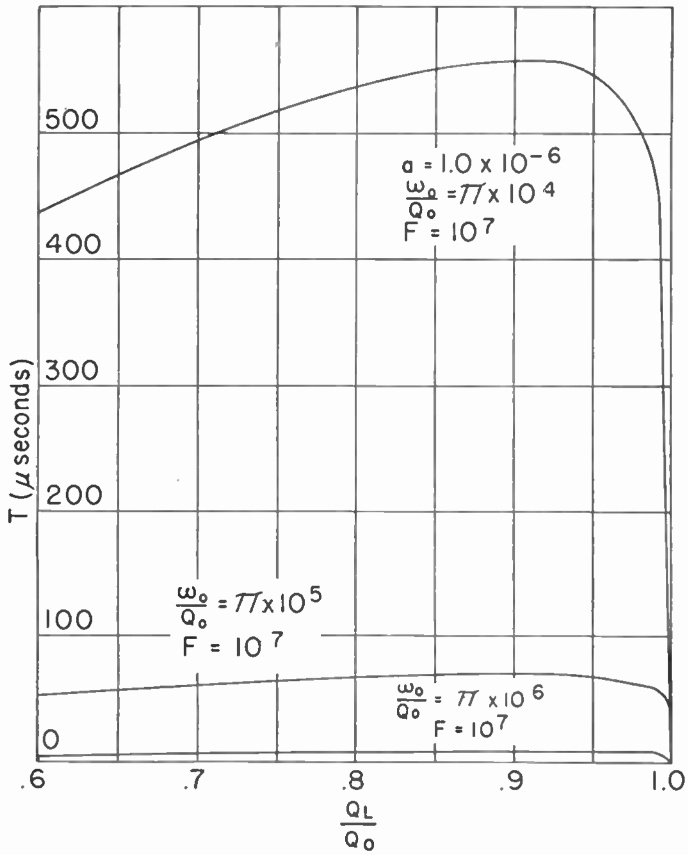


Fig. 7—Ring time versus Q_L/Q_0 for various unloaded Q's.

We also note that for all system parameters constant, and for a constant ratio of loaded to unloaded Q, the loaded Q is proportional to frequency for a given ring time.

Fig. 6, which is a plot of ring time versus Q_L/Q_0 for several different pulse lengths, clearly shows that there is an optimum loaded Q which will yield maximum ring time. The ratio of Q_L/Q_0 is about 0.9 for a 0.01-microsecond pulse and gets progressively higher as the pulses get longer. This is to be expected physically since the shorter pulses do not give a high-Q cavity sufficient time to store its maximum energy. It is important to note that if the cavity loaded Q is too low, the loss in ring time will not be serious, but if it is too high the ring time will degenerate very rapidly.

Fig. 7 is similar to Fig. 6 except that all these curves are for the same pulse length (1 microsecond) but different ratios of frequency to unloaded Q. It is apparent from these curves that the ringing time is very sensitive to unloaded Q and that every effort should be made to obtain as high an *unloaded* Q as possible.

Fig. 8 has as a parameter F, the dynamic range in voltage between the transmitted signal and minimum discernible received signal. These curves show the optimum Q_L/Q_0 decreasing as F decreases. A general conclusion to be drawn from these curves, along with those of Fig. 6, is that low-power short-pulse radars demand a lower optimum loaded Q where the higher-power long-pulse systems can use a higher loaded Q to achieve maximum ring time.

If a directional coupler is used to couple the cavity, then the factor K^2 appears in F in accordance with Table I. Otherwise the computation for ring time with or without a directional coupler is identical.

The above curves cover the range of parameters normally encountered in microwave radars.

V. Power Loss in Echo Box

For the case of the echo box coupled through a directional coupler, the power extracted from the main guide is proportional to the square of the voltage coefficient of coupling of the directional coupler

$$M = \frac{P_{\text{extracted}}}{P_{\text{load}}} = K^2. \tag{10}$$

Thus, for example, if K is less than 0.1, the maximum power extracted by the measuring system will be less than one per cent of the total power.

For the case of the cavity mounted directly on the guide, the average power extracted during the transmitted pulse is determined by the coupling and also by the pulse length.

For very long pulses, a steady-state condition may be assumed, the energy *stored* in the echo box may be neg-

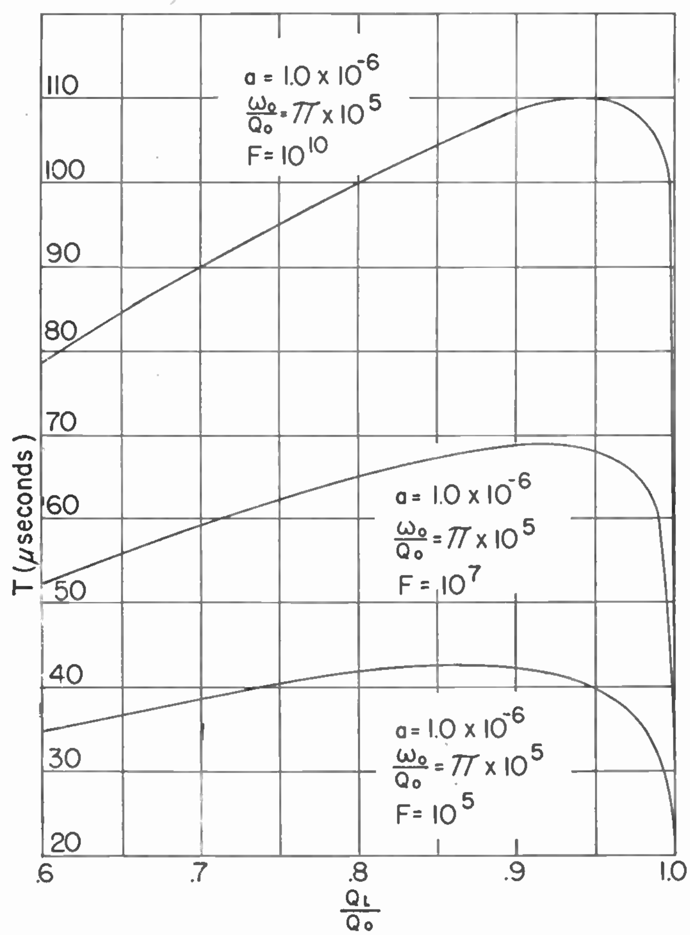


Fig. 8—Ring time versus Q_L/Q_0 for various dynamic ranges.

lected in comparison with the energy *dissipated* therein, and the ratio M of power into the echo box to power into the load is accordingly given by

$$M = \frac{2R}{R_g}, \tag{11}$$

wherein R and R_g are shown in Fig. 3. Relating R and R_g to Q_L and Q_0 by (4), (11) becomes

$$M = 2\left(\frac{Q_0}{Q_L} - 1\right). \tag{12}$$

This is plotted in Fig. 9 as the curve for infinite pulse length ($\omega_0 a/2Q_L = \infty$). For short pulses, the average

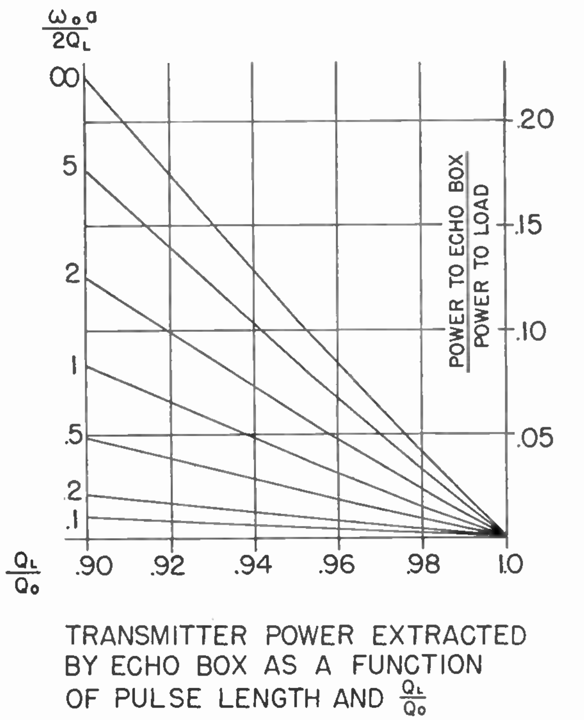


Fig. 9—Ratio of cavity power to load power versus Q_L/Q_0.

power into the echo box will be less than indicated in (12) because of the effect of the initial transient condition. The ratio of power extracted during a short pulse to power delivered to the load can be approximated by multiplying the above factor M by A, where A is the ratio of average power into the echo box, for the short pulse, to the power into the echo box for the steady-state condition.

The energy delivered to the echo box in time a, in the steady-state condition, would be

$$\frac{\left[E_g\left(1 - \frac{Q_L}{Q_0}\right)\right]^2 a}{R}.$$

The energy dissipated in the echo box, during a pulse of length a, is

$$\frac{1}{R}\int_0^a e^2 dt.$$

The energy *stored* in the echo box at the end of the pulse is

$$e^2(a)\,\frac{Q_0}{\omega_0 R}.$$

Hence

$$A = \frac{\frac{1}{R}\int_0^a e^2 dt + e^2(a)\,\frac{Q_0}{\omega_0 R}}{\left[E_g\left(1 - \frac{Q_L}{Q_0}\right)\right]^2 a/R}. \tag{13}$$

As

$$e = E_g\left(1 - \frac{Q_L}{Q_0}\right)\left(1 - \epsilon^{-\omega_0 t/2Q_L}\right), \tag{14}$$

it can be shown that A is given approximately by the equation

$$A = 1 - \frac{2Q_L}{\omega_0 a}\left(1 - \epsilon^{-\omega_0 a/2Q_L}\right). \tag{15}$$

Multiplying the curve given in Fig. 9 for very long pulses by this factor gives the other curves shown in that figure. These curves would indicate that radar systems using very short pulses can tolerate much tighter coupling to an echo box than would normally be used.

VI. Conclusions

The dependence of the optimum Q and maximum ring time of an echo box system on coupling method, pulse length, unloaded cavity Q, and transmitter-receiver properties has been developed. The power taken by a direct-coupled echo box has been plotted as a function of the ratio of loaded to unloaded Q.

A Network Theorem and Its Application*

F. W. BUBB, JR.†, ASSOCIATE MEMBER, IRE

Summary—The proof of a theorem useful in analyzing the response of certain types of circuits to modulated signals and examples of the use of the theorem are presented.

Introduction

It is known that the response of a high-frequency circuit to an amplitude-modulated sinusoidal input signal may sometimes be obtained by analyzing an "equivalent" low-frequency circuit whose input is the envelope only. Those who have used such analyses in the literature[1,2] probably are aware of the proof of the theorem presented here. However, the writer has not seen such a proof, and it appears to be in order to show in detail the basis for these excellent works. Therefore a theorem will be stated, its proof given, and examples of its use will be shown. The Laplace transformation and its inverse will be used as the most convenient means of presenting the material.

Theorem: Let $Z_H(s)$ be the transfer function of a high-frequency circuit normally operated at an angular frequency in the vicinity of w_0. Let $\zeta = s - jw_0$, where s is the usual "complex frequency" and, of course, ζ is also complex.

Then, if

$$Z_H(s) = Z_L(s - jw_0) = Z_L(\zeta) \tag{1}$$

is the transfer function of some low-frequency circuit, the response of this low-frequency circuit to a driving force $m_i(t)$ is identical to the envelope of the response of the high-frequency circuit to a sinusoid of angular frequency w_0 whose envelope is $m_i(t)$.

To enlarge on these statements, let the input to $Z_H(s)$ be

$$i(t) = m_i(t)e^{jw_0 t},$$

where it is understood as in steady-state alternating-current circuit analysis that the real, or imaginary, part of $i(t)$ is the physical driving force, and let its response to $i(t)$ be

$$e(t) = m_0(t)e^{jw_0 t}. \tag{2}$$

These are shown in Fig. 1(a). The theorem postulates that if (1) holds, the output of $Z_L(\zeta)$ will be $m_0(t)$ for an input $m_i(t)$, as shown in Fig. 1(b).

Proof: Suppose that the Laplace transform of $m_i(t)$ is $M_i(s)$. The transform of $i(t)$ is then

$$\begin{aligned} I(s) &= \int_0^\infty e^{-st} m_i(t) e^{jw_0 t} dt \\ &= \int_0^\infty e^{-(s-jw_0)t} m_i(t) dt = M_i(s - jw_0). \end{aligned} \tag{3}$$

The Laplace transform of the response of the high-frequency circuit to $i(t)$ is given by the product of the transform of its input and its transfer function:

$$E(s) = I(s)Z_H(s), \tag{4}$$

while the time-response may be found by means of the inversion integral, provided it exists. That is,

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}E(s)ds, \tag{5}$$

where $\gamma > Re(s)$ for all poles of the integrand.

Using (3) and (4) in (5) we have

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}M_i(s - jw_0)Z_H(s)ds$$

or in view of (1)

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}M_i(s - jw_0)Z_L(s - jw_0)ds.$$

Changing the variable of integration to $s - jw_0$,

$$e(t) = \frac{e^{jw_0 t}}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t(s-jw_0)}M_i(s-jw_0)Z_L(s-jw_0)d(s-jw_0).$$

Now using the fact that $\zeta = s - jw_0$, we have finally

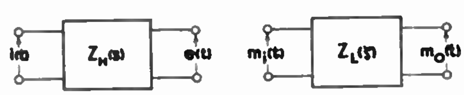
$$e(t) = e^{jw_0 t}\left[\frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t\zeta}M_i(\zeta)Z_L(\zeta)d\zeta\right].$$

On comparing this last expression to (2), it is obvious that the coefficient of $e^{jw_0 t}$ is the envelope of the response, namely

$$m_0(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t\zeta}M_i(\zeta)Z_L(\zeta)d\zeta. \tag{6}$$

We see that the right-hand side of (6) is also the response of the circuit characterized by $Z_L(\zeta)$ to $m_i(t)$, which proves the theorem.

Example 1. A Simple Tuned Circuit.[3] The simplest example of the use of the theorem is the analysis of an



(a) (b)

Fig. 1—(a) Showing the input and output of the high-frequency circuit. (b) Showing the input and output of the low-frequency circuit.

* Decimal classification: R143. Original manuscript received by the Institute, May 18, 1950; revised manuscript received, March 5, 1951.

† St. Louis University, St. Louis, Mo.

[1] G. E. Valley, Jr., and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., New York, N. Y., chap. 7; 1948.

[2] Eugene F. Grant, "Time response of an amplifier of N identical stages," Proc. I.R.E., vol. 36, pp. 870–871; July, 1948.

[3] Grant's paper finds the response for n stages of amplification, all stages tuned to the same frequency.

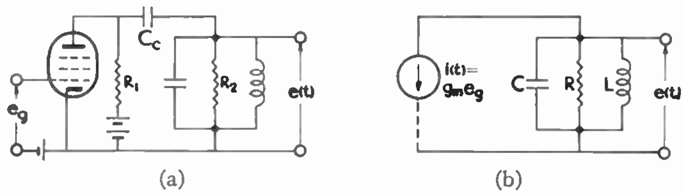


Fig. 2—(a) The tube-amplifier circuit. (b) The equivalent circuit.

amplifier whose plate load impedance is a simple tuned circuit. (See Fig. 2.) The operational impedance of the tuned circuit is

$$Z_H(s) = \frac{R}{1 + Q_0\left(\dfrac{s}{w_0} + \dfrac{w_0}{s}\right)},$$

where

$$\frac{1}{R} = \frac{1}{r_p} + \frac{1}{R_1} + \frac{1}{R_2},$$

and if $|\zeta| \ll w_0$, this becomes

$$Z_H(s) \approx \frac{R\pi B}{\zeta + \pi B},$$

where

$$Q_0 = w_0 RC$$

$$w_0 = \frac{1}{\sqrt{LC}}$$

and

$$B = \frac{1}{R(2C)},$$

where B is the bandwidth between the 3-db points of $|Z_H(s)|$. Thus

$$Z_H(s) \approx \frac{1/(2C)}{\zeta + \dfrac{1}{R(2C)}} = Z_L(\zeta). \qquad (7)$$

We see that (7) gives the operational impedance of the circuit shown in Fig. 3. It is such a simple matter to analyze this equivalent circuit for any reasonable $m_i(t)$, that no details will be put down here. It should be noted,

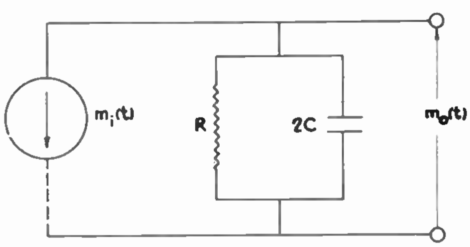


Fig. 3—The low-frequency equivalent of a simple tuned circuit.

however, that it is the approximation to $Z_H(s)$ in (7) which is equal to $Z_L(\zeta)$. We expect the results of an analysis of the equivalent circuit to be strictly accurate,

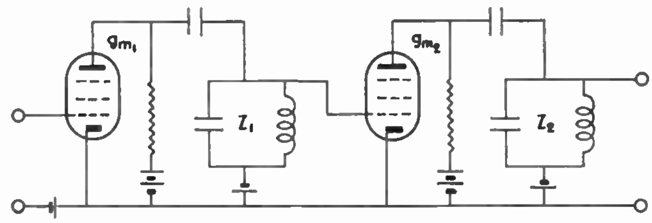


Fig. 4—A two-stage staggered-tuned amplifier.

then, only when $|\zeta|$ is indeed much less than w_0, but this is not a fault of the theorem.

Having obtained the equivalent low-frequency circuit, the reader might be tempted to include the components R_1 and C_c (see Fig. 2(a)) ahead of those in Fig. 3, since these components may not be neglected at the frequencies contained in $m_i(t)$. However, when it is remembered that the frequencies dealt with in the original problem are close to w_0, at which the reactance of C_c may be truly negligible, this pitfall is avoided.

Example 2. Two Staggered-Tuned Stages of Amplification. Consider the circuit of Fig. 4. In view of the previous section, the high-frequency equivalent circuit of each stage of amplification is like that of Fig. 2(b). If we denote the plate-load impedance of the first stage by $Z_1(s)$ and that of the second by $Z_2(s)$, then it is apparent that the gain of the two-stage amplifier is

$$Z_H(s) = g_{m_1} g_{m_2} Z_1(s) Z_2(s). \qquad (8)$$

Supposing that the two tuned circuits have equal bandwidths and are tuned respectively to angular frequencies w_{0_1} and w_{0_2}, they have impedances which are approximately

$$Z_1(s) \approx \frac{\pi B R_1}{(s - jw_{0_1}) + \pi B}$$

and

$$Z_2(s) \approx \frac{\pi B R_2}{(s - jw_{0_2}) + \pi B},$$

where it has been assumed that $|s - jw_{0_1}| \ll w_{0_1}$ and $|s - jw_{0_2}| \ll w_{0_2}$. The magnitudes of these impedances (for $Re(s) = 0$, $Im(s) = w$) plotted versus w will look something like the curves shown in Fig. 5.

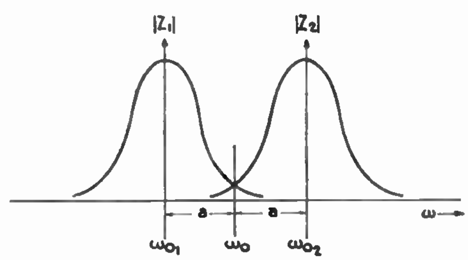


Fig. 5—The response curves of the staggered-tuned circuits.

Letting

$$w_{0_1} = w_0 - a$$

and

$$w_{0_2} = w_0 + a$$

the impedances may be written

$$Z_1(s) = \frac{\pi BR_1}{s - j(w_0 - a) + \pi B} = \frac{\pi BR_1}{\zeta + \pi B + ja} \quad (9)$$

$$Z_2(s) = \frac{\pi BR_2}{s - j(w_0 + a) + \pi B} = \frac{\pi BR_2}{\zeta + \pi B - ja} \cdot (10)$$

On inserting (9) and (10) in (8), we have

$$Z_H(s) = \frac{\pi^2 B^2 g_{m_1} R_1 g_{m_2} R_2}{(\zeta + \pi B + ja)(\zeta + \pi B - ja)} = Z_L(\zeta). \quad (11)$$

Equation 11 suggests the low-frequency equivalent circuit of Fig. 6, where

$$R = \frac{\pi^2 B^2 g_{m_1} R_1 g_{m_2} R_2}{\pi^2 B^2 + a^2}$$
$$L = \frac{\pi B g_{m_1} R_1 g_{m_2} R_2}{2(\pi^2 B^2 + a^2)^2} \quad (12)$$
$$C = \frac{2(\pi^2 B^2 + a^2)}{\pi B g_{m_1} R_1 g_{m_2} R_2}.$$

For experimentally analyzing the response of odd-shaped modulations, this circuit might be of some value;

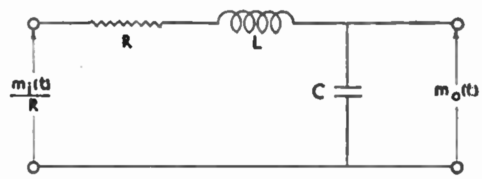


Fig. 6—The low-frequency equivalent circuit of a staggered-tuned amplifier.

however, to find the response of the amplifier to a unit step-function envelope is an easy matter. The inversion integral may be used to find $m_0(t)$ from (11):

$$m_0(t) = \frac{g_{m_1} R_1 g_{m_2} R_2}{2\pi j} \int_{\gamma - j\infty}^{\gamma + j\infty} \frac{\pi^2 B^2 e^{t\zeta} d\zeta}{(\zeta + B + ja)(\zeta + B - ja)}.$$

We see from (11) that for $\zeta = 0$ (which means the circuit is excited by a sinusoid of angular frequency w_0) the gain is $g_{m_1} R_1 g_{m_2} R_2 / 1 + (a/\pi B)$.[2] Letting $m_0'(t)$ be the normalized response, then,

$$m_0'(t') = \frac{m_0(t)[1 + \alpha^2]}{g_{m_1} R_1 g_{m_2} R_2}$$
$$= \frac{1 + \alpha^2}{2\pi j} \int_{\gamma - j\infty}^{\gamma + j\infty} \frac{e^{t'\xi} d\xi}{(\xi + 1 + j\alpha)(\xi + 1 - j\alpha)}, \quad (13)$$

where, by a variable change, $\xi = \zeta/\pi B$ and $\alpha = a/\pi B$; that is, ξ is the complex angular frequency in units of half-bandwidths of either tuned circuit, and α is the difference between the resonant angular frequency of either tuned circuit and the band center, also in half-bandwidths. Also $t' = \pi Bt$ is in units of half-bandwidth-seconds, or in radians.

We easily find $m_0'(t)$ by summing the residues of the poles of the integrand of (13). These poles occur at $\xi_1 = 0$, $\xi_2 = -(1 + j\alpha)$, and $\xi_3 = -(1 - j\alpha)$.

$$R_{\xi_1} = \frac{1}{1 + \alpha^2}$$
$$R_{\xi_2} = \frac{(1 - j\alpha)e^{-(1+j\alpha)t}}{2j\alpha(1 + \alpha^2)}$$
$$R_{\xi_3} = -\frac{(1 + j\alpha)e^{-(1-j\alpha)t'}}{2j\alpha(1 + \alpha^2)}$$

or

$$m_0'(t) = (1 + \alpha^2) \sum R$$
$$= 1 - e^{-t'} \left(\frac{1}{\alpha} \sin \alpha t' + \cos \alpha t' \right), \quad (14)$$

showing that the envelope of response of the staggered-tuned amplifier to a step-function modulated input first rises above and then oscillates about its final value. The damping of the oscillations depends on the bandwidth of either tuned circuit; this certainly is expected since the bandwidth itself is determined by the circuit time constant. The angular frequency of the oscillations of the envelope is equal to the separation of the resonant frequency of either circuit from the band center.

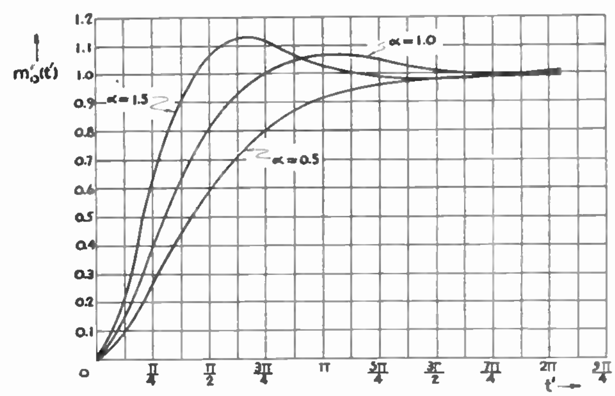


Fig. 7—The response of a staggered-tuned amplifier to a step-modulated signal.

Curves of $m_0'(t)$ versus t' for several values of α are shown in Fig. 7. The points where $m_0'(t)$ reaches its maxima and minima are $t' = n\pi/\alpha$, the maxima being

$$m_0'(n\pi/\alpha) = 1 + e^{-(n\pi/\alpha)}, \; n \text{ odd}$$

and the minima

$$m_0'(n\pi/\alpha) = 1 - e^{-(n\pi/\alpha)}, \; n \text{ even}.$$

These are shown in Fig. 7.

The magnitude of the first "overshoot" is then $e^{-\pi/\alpha}$; obviously this is zero only when $\alpha = 0$, and in general it is smaller as α becomes small. For the maximally flat response curve,[4] $\alpha = 1$, the overshoot is $e^{-\pi} = 0.044$. If a small value of the overshoot is to be a criterion for choosing α, then this should, of course, be as small as is consistent with a usefully large bandwidth, the maximally flat response curve providing a good comparison in this respect.

[4] The response curve, that is, $|Z_H(s)|$ plotted versus $Im(s) = w$ has only one maximum, and no minimum for $\alpha = 1$.

From the curves we also see that the time for the response first to reach its final level is larger as α becomes smaller. This also follows from the fact that for $m_0'(t)$ to be equal to unity,

$$t' = \frac{1}{\alpha} \tan^{-1}(-\alpha).$$

If quick response is to be the criterion for choosing α, then α should be as large as is allowed by overshoot considerations and steady-state response-curve flatness, α greater than unity resulting in the familiar double-humped response curve.

Conclusions

The theorem provides a powerful means of analyzing the response of certain types of circuits to amplitude modulated signals. The step function has been used as a representative type of modulation, but the response to sinusoids, ramp functions, and the like, are easily investigated using the same methods.

The writer is at present attempting to extend the theorem to include modulated signals which are periodic, but not necessarily sinusoidal, and some success in this direction has been achieved. It is hoped that more general methods of analyzing circuits using modulated signals will be the subject of another paper.

Constant-Resistance Networks of the Linear Varying-Parameter Type*

L. A. ZADEH†, MEMBER, IRE

Summary—**The term "constant-resistance network" is generally used to describe a fixed network whose input impedance is a constant independent of frequency. The present paper generalizes the concept of constant-resistance networks by extending this concept to linear varying-parameter systems. It is shown that any self-dual variable network is a constant-resistance network and this result is used to obtain such constant-resistance structures as the lattice and the bridged-tee. Explicit expressions for the transfer functions of these structures are derived and an illustrative example involving a constant-resistance variant of Cowan's modulator is considered. In addition to self-dual structures, a non-self-dual type of constant-resistance network is described and the possibility of using this structure as a constant-resistance capacitive modulator is indicated.**

Introduction

The term "constant-resistance network" is generally used to describe a fixed network whose input impedance is a constant independent of frequency. Such networks play an important role in the synthesis of four-terminal networks and, in particular, are widely used in the design of equalizers.[1]

The present paper generalizes the concept of constant-resistance networks by extending this concept to linear varying-parameter networks. This class of networks constitutes the most general form of linear systems; it includes as special cases the various kinds of amplitude and phase modulators, as well as the ordinary fixed linear networks.

As a preliminary to the consideration of variable[2] constant-resistance networks it is necessary to become acquainted with certain definitions and relations which form the basis for the circuit analysis[3] of variable networks in the frequency domain. These are as follows:

A. System Function

The fundamental concept on which the circuit analysis of variable networks is based, is that of the system function[4] $H(j\omega;\, t)$. Briefly, the system function of a variable network N is defined as a function $H(j\omega; t)$ such that $H(j\omega;\, t)e^{j\omega t}$ represents the response of N to $e^{j\omega t}$. Thus, let $u(t)$ denote the input to N and let $v(t)$ be the response to $u(t)$. Then, the definition of $H(j\omega;\, t)$ reads

$$H(j\omega;\, t) = \left. \frac{v(t)}{u(t)} \right|_{u(t)=e^{j\omega t}}. \tag{1}$$

The salient property of the system function is that it relates the output and input of N in much the same manner as the conventional system function $H(j\omega)$ relates the output and input of a fixed network. For example, the response of N (at rest) to an input $u(t)$ ($u(t) = 0$ for $t<0$) may be expressed in terms of $H(j\omega; t)$ as

$$v(t) = L^{-1}\{H(s;\, t)U(s)\}, \tag{2}$$

where $s=\sigma+j\omega$ is the complex frequency, $U(s)$ is the Laplace transform of $u(t)$, and L^{-1} represents the operation of inverse Laplace transformation (treating t in $H(s;\, t)$ as if it were a constant parameter). Similarly, in the operational form, one may write

$$v(t) = H(p;\, t)u(t), \tag{3}$$

where $H(p;\, t)$ should be treated as a usual Heaviside operator, i.e., t in $H(p;\, t)$ should be regarded as a con-

* Decimal classification: R143. Original manuscript received by the Institute, August 8, 1950; revised manuscript received, December 19, 1950. Presented, 1951 IRE National Convention, New York, N. Y., March 20, 1951.

† Columbia University, New York, N. Y.

[1] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., chap. 12; 1945.

[2] The term "variable," as used here and elsewhere in this paper, is synonymous with "linear varying-parameter."

[3] L. A. Zadeh, "Circuit analysis of linear varying-parameter networks," *Jour. Appl. Phys.*, vol. 21, pp. 1171–1177; November, 1950.

[4] L. A. Zadeh, "Frequency analysis of variable networks," Proc. I.R.E., vol. 38, pp. 291–299; March, 1950.

stant parameter. When written as $H(p;\ t)$, the system function will be referred to as a time-dependent Heaviside operator, or simply, as an operator.

In dealing with electrical networks it is convenient to refer to a system function by a distinctive name, such as impedance, admittance, gain, etc., which places in evidence the physical significance of $u(t)$ and $v(t)$. For instance, the *impedance* of a two-terminal network N is in effect the system function relating the voltage across N, $v(t)$, to the current flowing through N, $i(t)$. Thus, denoting the impedance of N by the symbol $Z(j\omega;\ t)$, (1) becomes

$$Z(j\omega;\ t) = \left.\frac{v(t)}{i(t)}\right|_{i(t)=e^{j\omega t}}. \tag{4}$$

Similarly, the *admittance* of N is defined by the relation

$$Y(j\omega;\ t) = \left.\frac{i(t)}{v(t)}\right|_{v(t)=e^{j\omega t}}. \tag{5}$$

As a simple illustration of these definitions consider a variable capacitance $C=C(t)$, in which case the relation connecting $v(t)$ and $i(t)$ is

$$i = \frac{d(Cv)}{dt}. \tag{6}$$

Applying (4) and (5), and using s in place of $j\omega$, the expressions for the impedance and admittance of a variable capacitance are found to be

$$Z(s;\ t) = \frac{1}{C(t)s} \tag{7}$$

and

$$Y(s;\ t) = C(t)s + \dot{C}(t), \tag{8}$$

where the dot represents differentiation with respect to time.

B. Product Relation

The product relation to be given below expresses essentially the rule of composition of successive operations on a function of time. Thus, let $v(t)$ be the result of successive operations by two operators, say $H_1(p;\ t)$ and $H_2(p;\ t)$, on $u(t)$. In other words,

$$v(t) = H_2(p;\ t)\{H_1(p;\ t)u(t)\}. \tag{9}$$

The two successive operations involved in (9) may be replaced by a single operation with an operator $H_3(p;\ t)$,

$$v(t) = H_3(p;\ t)u(t), \tag{10}$$

where $H_3(p;\ t)$ is called the *product* of $H_2(p;\ t)$ and $H_1(p;\ t)$. It can readily be shown that $H_3(p;\ t)$ is related to $H_2(p;\ t)$ and $H_1(p;\ t)$ through an operational expression

$$H_3(s;\ t) = H_2(p + s;\ t)H_1(s;\ t), \tag{11}$$

in which $H_2(p+s;\ t)$ is the operator and $H_1(s;\ t)$ plays the role of a function of time involving s as a parameter. Equation (11) may conveniently be expressed in a symbolic form

$$H_3 = H_2 * H_1, \tag{12}$$

where the symbol $*$ has all the usual properties of an algebraic product except for commutativity. Thus, one can write

$$H_a * (H_b + H_c) \equiv H_a * H_b + H_a * H_c \tag{13}$$

and

$$H_a * (H_b * H_c) \equiv (H_a * H_b) * H_c, \tag{14}$$

where H_a, H_b, and H_c represent arbitrary operators. However, in general, $H_a * H_b$ is not the same as $H_b * H_a$.

C. Inverse of a System Function

The inverse of an operator $H(p;\ t)$, or simply H, is defined as an operator H^{-1} such that the product of H and H^{-1} is equal to unity. Thus,

$$H * H^{-1} = H^{-1} * H = 1, \tag{15}$$

where the product symbol $*$ should be interpreted in the sense of (12). It should be noted that in general H^{-1} *is not* the algebraic reciprocal of H. The determination of the inverse of a given time-dependent operator is generally a difficult problem. There is no need, however, to be concerned with this problem in the present paper.

Combining the definitions of the impedance and admittance of a variable network ((4) and (5)) with that of the inverse of a system function, it is readily seen that $Y(p;\ t)$ is the inverse of $Z(p;\ t)$, and vice versa. From this it follows that the impedance and admittance of a variable network are related to each other by the operational relation

$$Z(p + s;\ t)Y(s;\ t) = Y(p + s;\ t)Z(s;\ t) = 1 \tag{16}$$

or more simply

$$Z * Y = Y * Z = 1. \tag{17}$$

This relation represents a generalization of the familiar algebraic relation

$$Z(s)Y(s) = 1, \tag{18}$$

which holds in the case of fixed networks.

There are three important identities involving system functions which will be needed later in this paper. These are:

$$(H_a * H_b * H_c)^{-1} \equiv H_c^{-1} * H_b^{-1} * H_a^{-1} \tag{19}$$

$$\begin{aligned} H_a^{-1} + H_b^{-1} &\equiv H_a^{-1} * (H_a + H_b) * H_b^{-1} \\ &\equiv H_b^{-1} * (H_a + H_b) * H_a^{-1} \end{aligned} \tag{20}$$

$$\begin{aligned} (H_a^{-1} + H_b^{-1})^{-1} &\equiv H_b * (H_a + H_b)^{-1} * H_a \\ &\equiv H_a * (H_a + H_b)^{-1} * H_b. \end{aligned} \tag{21}$$

The properties of system functions as outlined above will now be used to extend the concept of constant-resistance networks to linear varying-parameter systems.

Constant-Resistance Networks

A variable network N will be said to be a constant-resistance network if its input impedance $Z(s; t)$ is independent of both s and t—that is, if $Z(s; t)$ is equal to a constant k. It is evident that if $Z(s; t)=k$ then $Y(s; t)=1/k$, and hence a constant-resistance network can be so characterized on the basis of either its impedance or its admittance. In what follows, the constant k will be assumed to be equal to unity. This simplifies the discussion without entailing any loss in generality.

An important class of constant-resistance networks can be obtained by making use of the properties of dual variable networks. Duality in the case of variable networks involves the same relationships as in the case of fixed networks. The only difference is that in the case of variable networks the relations between the dual elements assume a more general form, which is as follows:

The dual of an inductance $L=f(t)$ is a capacitance $C=f(t)$, and vice versa.

The dual of a resistance $R=h(t)$ is a conductance $G=h(t)$, and vice versa.

It will be recalled that if N and N' are two fixed networks which are the duals of each other, then the input impedance of N is equal to the input admittance of N', and vice versa. The same relation holds in the case of variable networks. Thus, let $Z(s; t)$ and $Y(s; t)$ denote the input impedance and admittance of N, and let those of N' be denoted by the corresponding primed symbols. Then

$$Z(s; t) = Y'(s; t) \tag{22}$$

and

$$Y(s; t) = Z'(s; t). \tag{23}$$

From the above relations it follows at once that any self-dual network—that is, any network which is identical with its dual—is a constant-resistance network. For, if N is identical with N', then (22) and (23) give

$$Z(s; t) = Y(s; t). \tag{24}$$

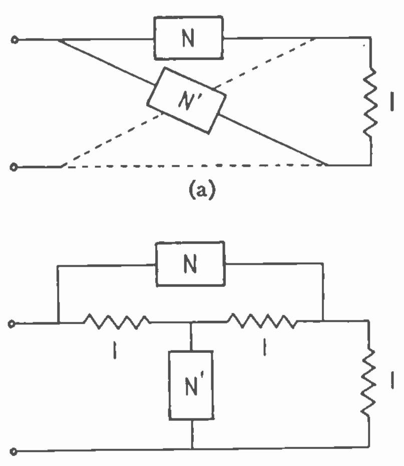


Fig. 1—Self-dual variable networks. (a) Constant-resistance lattice; (b) constant-resistance bridged-tee.

But, $Z(p; t)$ is the inverse of $Y(p; t)$, and therefore (24) implies that $Z(p; t)$ is equal to its inverse, which is impossible unless $Z(p; t)$ is equal to unity. Thus, it may be concluded that: *Any self-dual variable network is a constant-resistance network.*

This result may be used as a means for obtaining many forms of constant-resistance structures. For example, consider the lattice and the bridged-tee shown in Figs. 1(a) and 1(b), in which N and N' are dual variable networks. It can readily be verified that both of these structures are self-dual. Hence, we may conclude that:

The lattice and the bridged-tee are constant-resistance networks whenever the branches N and N' consist of dual variable networks.

In view of the importance of the constant-resistance lattice and the constant-resistance bridged-tee in the design of conventional equalizers, it is of interest to examine in somewhat greater detail the characteristics of these structures for the case where N and N' are variable networks.

Let $G(s; t)$, or simply G, denote the gain of these structures and let $Z(s; t)$ and $Z'(s; t)$, or simply Z and Z', represent the input impedances of N and N'. Then, by using (14), (19), (20), and (21), and noting that Z' is equal to Z^{-1}, it is readily found that the gain of the lattice is given by

$$G = (1 - Z) * (1 + Z)^{-1}, \tag{25}$$

where $(1+Z)^{-1}$ is the inverse of $(1+Z)$. This relation is a generalization of the well-known expression

$$G = \frac{1 - Z}{1 + Z}, \tag{26}$$

which represents the gain of a *fixed* constant-resistance lattice.

From (26), Z may be expressed in terms of G as follows:

$$Z = (1 - G) * (1 + G)^{-1}, \tag{27}$$

which is a generalization of

$$Z = \frac{1 - G}{1 + G} \cdot \tag{28}$$

For a constant-resistance bridged-tee the corresponding relations are

$$G = (1 + Z)^{-1} \tag{29}$$

and

$$Z = G^{-1} - 1. \tag{30}$$

As a simple illustration of the use of the above relations, consider the case where it is desired that G be a function of time of the form shown in Fig. 2(a). Applying (27) and noting that Z' (the impedance of N') is equal to Z^{-1}, it is easily found that Z and Z' should be variable resistances varying with time in the manner shown in Figs. 2(b) and 2(c). In this way one arrives at the constant-resistance variant of Cowan's modulator which was suggested by Tucker[5] on the basis of a different line of reasoning. It will be recognized, of course, that (27) and (30) make it possible to realize any desired form of modulating characteristic by using either a constant-resistance lattice or a constant-resistance bridged-tee.

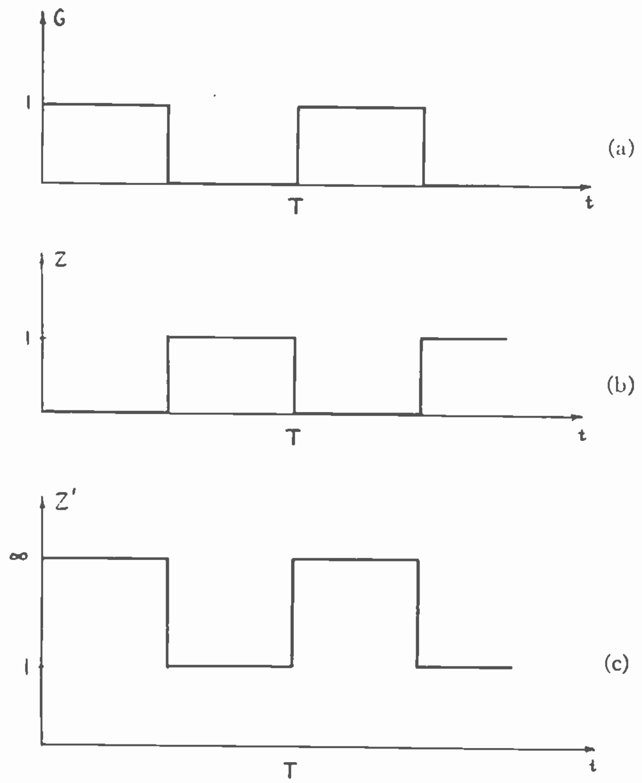


Fig. 2—(a) The desired variation of gain versus time; (b) variation of the impedance of series branches versus time; (c) variation of the impedance of crossed branches versus time.

It is evident that there are many forms of constant-resistance networks which are not of the self-dual type. The simplest of these is shown in Fig. 3. Here, as before, N' is the dual of N. It will be noted that this structure was obtained simply by taking a well-known fixed constant-resistance network and replacing N and N' by variable networks. The same procedure can be used for obtaining variable constant-resistance networks of a more complicated form.

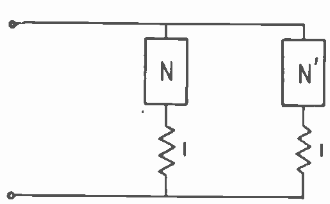


Fig. 3—One of the simplest non-self-dual constant-resistance networks.

[5] D. G. Tucker, "Rectifier resistance laws," *Wireless Eng.*, vol. 25, pp. 117–128; April, 1948. British Provisional Patent, 35540; 1946.

To prove that the network shown in Fig. 3 is indeed a constant-resistance network, it is sufficient to show that its input admittance is equal to unity. Using the same notation as before, the input admittance in question may be written as

$$Y_{in} = (1+Z)^{-1} + (1+Z^{-1})^{-1}. \tag{31}$$

By using (21), (31) becomes

$$Y_{in} = (1+Z)^{-1} + Z * (1+Z)^{-1} \tag{32}$$

or

$$Y_{in} = (1+Z)^{-1} * (1+Z), \tag{33}$$

and since the second factor in this expression is the inverse of the first factor, therefore Y_{in} is equal to unity, which is the desired result.

For the particular case where N is a capacitance $C=f(t)$, the network shown in Fig. 3 reduces to that shown in Fig. 4, in which the inductance L' is the dual of C and hence is equal to $f(t)$. The branch of this network comprising C and $R(R=1)$ may be regarded as a capacitive modulator of the type frequently used in practice. Thus, the network as a whole constitutes a constant-resistance capacitive modulator.

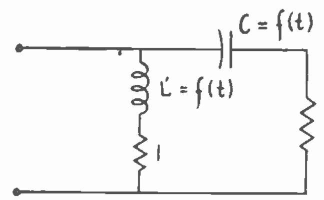


Fig. 4—A constant-resistance capacitive modulator.

Only a few basic types of constant-resistance networks have been considered in the preceding discussion. However, the outlined principles and methods are quite general and can be used both in the analysis and synthesis of various constant-resistance structures. In designing such structures it is useful to remember that the dual of any constant-resistance network is also a constant-resistance network, and that a complex constant-resistance structure may be obtained through the interconnection of a number of simpler constant-resistance networks.

Conclusion

In conclusion, it may be stated that there is no essential difference between constant-resistance networks of the fixed and variable type. The potentialities of variable constant-resistance networks remain largely unexplored. They might prove useful in the design of constant-resistance modulators; in the simulation of time-dependent operators by means of electrical networks; in the synthesis of variable networks; and, as electromechanical analogs, in the design of constant-torque mechanical drives.

Distortion in Linear Passive Networks*

LEON RIEBMAN†, ASSOCIATE, IRE

Summary—It is well known that distortion is suffered by a modulated wave in passing through a practical transmission network.

The increasing use of frequency modulation in television relays and other uses where high fidelity is important has made the problem of FM distortion of current interest. One of the earliest analytical approaches to the problem was by Carson and Fry[1] and later applied by Jaffe[2] to single- and double-tuned circuits. The Carson and Fry method depends on a series solution to obtain numerical results.

The theory to be developed is based on a series solution of the superposition integral. A recent approach along this line is by Bradley.[3] He obtained a rapidly convergent series that is useful when the network transfer function is in a form which permits separation of attenuation and phase of the network and can be differentiated conveniently.

It is the purpose of this paper to provide a method of calculating distortion, when the transfer function is known in pole and zero form, that is rapidly convergent under most cases considered in practice.

Theory

GIVEN A general input function $f_1(t)$, applied to a linear passive network $Y(\omega)$, the Fourier integral expression for the output is

$$f_2(t) = \int_{-\infty}^{\infty} f_1(\tau)\left[\frac{1}{2\pi}\int_{-\infty}^{\infty} Y(\omega)e^{j\omega(t-\tau)}d\omega\right]d\tau. \qquad (1)$$

The integral in the brackets will be recognized as the network response to a unit spike impressed at time t. This is usually called the characteristic transient[3] and denoted by the symbol $\mathrm{T}\!\!\!\!/\,(t-\tau)$.

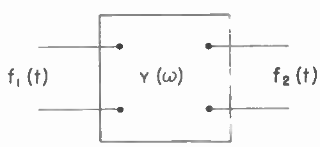


Fig. 1—General linear passive network.

Noting that $\mathrm{T}\!\!\!\!/\,(t-\tau)=0$ for $\tau>t$ in any physically realizable case, the output function $f_2(t)$ can be written as

$$f_2(t) = \int_{-\infty}^{t} \mathrm{T}\!\!\!\!/\,(t-\tau)f_1(\tau)d\tau. \qquad (2)$$

Now assume an input waveform

$$f_1(t) = Me^{j\omega_0 t}, \qquad (3)$$

where M is the modulating function and is a function of time. The output waveform may be considered to be

$$f_2(t) = e^{\alpha+j(\omega_0 t+\beta)}, \qquad (4)$$

where α and β are functions of M.

The general expression for the characteristic transient of a system of poles and zeros can be written as follows:[4]

$$\mathrm{T}\!\!\!\!/\,_t = \sum_n \sum_r \left[A_{nr}t^{(r-1)}e^{-(P_{nr}-j\omega_0)t} + A_{nr}^{*}t^{(r-1)}e^{-(P_{nr}^{*}+j\omega_0)t}\right], \qquad (5)$$

where

A_{nr}^{*} is the conjugate of A_{nr}, P_{nr}^{*} is the conjugate of P_{nr}, and

$$P_{nr} = \alpha_{nr} - j(\beta_{nr}-\omega_0) = \alpha_{nr} - j\sigma_{nr}.$$

From pole and zero theory, the summation over n denotes n different poles and the summation over r denotes the nth pole is raised to the power r.

Combining (2), (3), (4), and (5),

$$e^{\alpha+j\beta} = \sum_n \sum_r t^{(r-1)}A_{nr}e^{-P_{nr}t} \int_{-\infty}^{t} M(\tau)e^{P_{nr}\tau}\left(1-\frac{\tau}{t}\right)^{(r-1)}d\tau + A_{nr}^{*}e^{-(P_{nr}^{*}-j2\omega_0)t} \int_{-\infty}^{t} M(\tau)e^{(P_{nr}^{*}-j2\omega_0)\tau}\left(1-\frac{\tau}{t}\right)^{(r-1)}d\tau. \qquad (6)$$

If the percentage bandwidth is small the second term in (6) becomes very small compared to the first term, and can be neglected. If wide percentage bandwidth is used (i.e., >12 per cent), the second term should be included and the following method is equally applicable. Equation (6) can now be rewritten for the narrow-band case,

$$e^{\alpha+j\beta} = \sum_n \sum_r A_{nr}t^r \int_0^{\infty} M(y)e^{-P_{nr}ty}y^{(r-1)}dy,$$

$$y = 1 - \frac{\tau}{t}. \qquad (7)$$

At this point it is necessary to know the form of the modulating function $M(y)$ in order to apply this equation

1. Input FM Wave

For an input FM wave, the modulating function takes the form

$$M(\tau) = e^{j\phi(\tau)}, \quad \phi(\tau) = \phi[t(1-y)] = \theta(y).$$

Expanding $e^{j\theta(y)}$ in a power series in y,

* Decimal classification: R148.11×R143. Original manuscript received by the Institute, January 19, 1950; revised manuscript received, August 22, 1950.

† Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Pa.

[1] J. Carson and T. Fry, "Variable frequency electric circuit theory with application to the theory of frequency-modulation," *Bell Sys. Tech. Jour.*, vol. 16, p. 513; October, 1937.

[2] D. Jaffe, "Tuned circuit F-M distortion," Proc. I.R.E., vol. 33, pp. 318–333; May, 1945.

[3] W. Bradley, "A Theory of FM Distortion," Philco Research Report #114.

[4] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y., vol. I, pp. 159–164; 1945.

$$e^{j\theta(y)} = e^{j\theta(o)} \sum_{s=0}^{\infty} A_s y^s, \quad \text{where}$$

$$A_s = \frac{1}{s!}\left[\frac{d^s}{dy^s}\left(e^{j\theta(y)}\right)\right]_{y=0} e^{-j\theta(o)}. \tag{8}$$

Combining (7) and (8) and performing the integration,[5] then

$$e^{\alpha+j\beta} = (B_0 + B)e^{j\phi(t)}, \tag{9}$$

where,

$$B_0 = \sum_n \sum_r A_{nr} A_0 \frac{(r-1)!}{P_{nr}^r},$$

$$B = \sum_n \sum_r \sum_{s=1} A_{nr} A_s \frac{t^{-s}(r-1+s)!}{P_{nr}^{r+s}}.$$

Although this form is frequently used to demonstrate distortion, it is not very useful in practice for application to distortion measurements. Taking the logarithm of both sides and letting $\log B_0 = \alpha_0 + j\beta_0$,

$$\alpha + j\beta = \alpha_0 + j[\beta_0 + \phi(t)] + \log(1 + B/B_0). \tag{10}$$

α_0 and β_0 are independent of ϕ, and therefore represent the attenuation and phase shift of the carrier in the absence of modulation. As they are constants, these two factors do not cause distortion. The last term is a function of the derivatives of ϕ and an expansion of this term will yield the distortion. From Appendix A,

$$B_0 + B = \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{(r-1)!}{(1-R)^r}\left[1 + j\frac{\phi''}{2P_{nr}{}^2} \cdot \frac{r(r+1)}{(1-R)^2} - j\frac{\phi'''}{6P_{nr}{}^3} \cdot \frac{r(r+1)(r+2)}{(1-R)^3} - \frac{\phi''^2}{8P_{nr}{}^4} \cdot \frac{r(r+1)(r+2)(r+3)}{(1-R)^4} + \cdots\right]. \tag{11}$$

For deviation ratios greater than one, this series converges very rapidly, the first term alone usually giving sufficient accuracy in practice. Although only the first term will be used in the remainder of this paper, the second term will provide a criterion for the accuracy of this method. Then,

$$B_0 + B \cong \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{(r-1)!}{(1-R)^r} \tag{12a}$$

$$B' = \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{r!R'}{(1-R)^{r+1}}$$

$$B' = \frac{dB}{dt} \quad \text{and} \quad R' = \frac{dR}{dt}. \tag{12b}$$

Returning to (10), the output frequency modulation is,

$$\beta' = \phi' + \Omega_D = \phi' + \text{imag.}\left[\frac{B'}{B_0 + B}\right], \tag{13}$$

[5] H. B. Dwight, "Tables of Integrals," 861.2.

where Ω_D is the output frequency-modulation distortion. The output amplitude-modulation distortion is,

$$M_D = e^{\alpha} - 1 = |B_0 + B| - 1. \tag{14}$$

The output wave can then be written,

$$f_2(t) = (1 + M_D)e^{j(\phi + \int \Omega_D dt)}. \tag{15}$$

The form of the above expression is convenient for application. If the above waveform is impressed on an ideal linear AM detector, the output is

$$f_{AM} = 1 + M_D. \tag{16}$$

The output of an ideal FM detector is,

$$f_{FM} = \phi' + \Omega_D. \tag{17}$$

It is profitable at this point to examine a typical FM receiver using a discriminator type detector circuit as illustrated by Fig. 2 in block diagram.

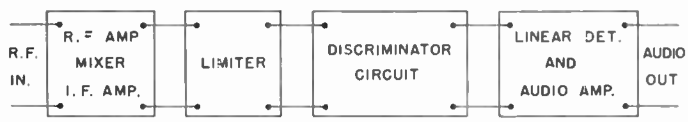


Fig. 2—Block diagram of typical FM receiver.

If an undistorted FM wave is applied to the receiver, this wave will suffer both AM and FM distortion in passing through the amplifier stages before the limiter. Assuming an ideal limiter, the AM distortion will be removed before the signal is applied to the discriminator. Thus, the FM distortion is of primary importance when considering the radio-frequency and intermediate-frequency amplifiers.

If a pure FM wave is applied to the discriminator, both FM and AM distortion are again produced by the tuned circuits. The output of the discriminator is applied to a linear detector that is assumed to be independent of frequency. Therefore, the AM distortion is of primary interest for this case. After detection, the fundamental frequency term in the AM distortion provides the useful output of a discriminator, and the higher harmonic terms provide the audio harmonic distortion.

In general, discriminator circuits are either one- or two-pole circuits (when the narrow-band approximation is used). The amplifier part of the receiver usually contains a larger number of poles. It will, therefore, be desirable to be able to calculate the amplitude distortion for one- or two-pole circuits and the FM distortion for any reasonable number of poles.

At this point it is convenient to consider special cases.

A. Synchronous Single-Tuned Amplifier.

For this case all stages have equal damping and are tuned to the same frequency (thus, $n=1$ and $P_{1r}=P$). It is general practice to design the over-all rf and IF bandwidth to be equal to or greater than twice the maximum frequency deviation. For cases where this holds and the deviation ratio is sensibly greater than one, (12a) and (12b) can be used without appreciable error. If the percentage bandwidth is small, the term of major importance in the expression for the characteristic transient is the rth germ. For this case, using (12a), (12b), and (13), the general expression for the frequency distortion becomes,

$$\Omega_{Dr} = \text{imag.}\left[\frac{rR'}{1-R}\right] = \text{imag.}\left[\frac{-jr\phi''}{P+j\phi'}\right]. \qquad (18)$$

It should be noted that the above form of the expression for FM distortion does not require the amplifier to be tuned to the carrier frequency of the incoming signal. The only requirement is that the r poles all be tuned to the same frequency and have the same damping.

Assume the modulating wave is a sinusoid, $\phi' = \Delta\omega \cos \omega_a t$. Rewriting (18) for this case, expanding into a Fourier series (see Appendix B), and with some trigonometric manipulation,

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \left(\frac{m_b}{m_T}\right)^m \sin m(\theta_1 - \theta_3) \sin m\omega_a t, \qquad (19)$$

where

$$m_b = \frac{\Delta\omega}{\sqrt{\alpha^2+\sigma^2}}, \qquad \theta_1 = \tan^{-1}\frac{\alpha}{\sigma}$$

$$\theta_2 = \tfrac{1}{2}\tan^{-1}\frac{m_b^2 \sin 2\theta_1}{1 - m_b^2 \cos 2\theta_1}$$

$$\theta_3 = \tan^{-1}\left(\frac{\sin\theta_2 \sqrt[4]{1 m_b^4 - 2m_b \cos 2\theta_1}}{1 + \cos\theta_2 \sqrt[4]{1 m_b^4 - 2m_b \cos 2\theta_1}}\right)$$

$$m_T e^{j\theta_3} = 1 + \sqrt{1 - X^2}.$$

The above equation is in a convenient form for calculating purposes. It should be noted that considerable simplification occurs when $\sigma = 0$ (i.e., the circuit is tuned to the incoming carrier), and when $\sigma = \alpha$.

Several interesting general conclusions can be made from inspection of the above equation.

1. The distortion amplitude is linearly related to the modulation frequency.
2. For a given bandwidth per stage, the distortion increases linearly with the number of stages.
3. As there is no dc term, the incoming carrier is not shifted in frequency.
4. Only sine wave terms are present. The fundamental term represents "phase" distortion of the FM modulation, and the higher terms represent harmonic distortion.

Items 3 and 4 are surprising at first glance, as the carrier is not necessarily at the center frequency of the amplifier. The dc and cosine terms are, of course, present in the more rigorous general solution.

A special case of (19) will be considered:

1. Circuits tuned to incoming carrier, $\sigma = 0$. For this case

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \left(\frac{m_b}{1+\sqrt{1+m_b^2}}\right)^{2m-1} \cdot \sin(2m-1)\omega_a t, \qquad (20)$$

where $m_b = \Delta\omega/\alpha$.

As might be expected, only odd harmonics are produced.

If the over-all bandwidth of the amplifier is maintained constant, the value of m_b increases with the number of stages. From synchronously tuned amplifier circuit theory,[6] it can be shown that

$$\omega_B = 2\alpha_1 = 2\alpha_0\sqrt{2^{1/r}-1}, \qquad (21)$$

where α_1 is the α for a one-stage single-tuned amplifier of bandwidth ω_B. α_0 is the α for an r-stage single-tuned amplifier of over-all bandwidth ω_B.

Combining (20) and (21),

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \frac{\dfrac{2\Delta\omega}{\omega_B}\sqrt{2^{1/r}-1}}{1 + \sqrt{\left(\dfrac{2\Delta\omega}{\omega_B}\right)^2 (2^{1/r}-1) + 1}} \cdot \sin(2m-1)\omega_a t. \qquad (22)$$

Usually, in practice, the over-all bandwidth ω_B is equal to twice the maximum frequency of deviation. For this case, setting $\omega_B = 2\Delta\omega$, the percentage harmonic distortion becomes

$$\frac{\Omega_{Dr}}{\Delta\omega} = \frac{2r}{m_f}\sum_{m=1}^{\infty}(-1)^m\left(\frac{2^{1/2r}-1}{2^{1/2r}+1}\right)^{m-1/2} \sin(2m-1)\omega_a t, \qquad (23)$$

where $m_f = \Delta\omega/\omega_a$ is the frequency deviation ratio as usually defined. The absolute values of the amplitudes of the percentage harmonic distortion are plotted in Fig. 3 for several values of r. It should be noted that the fundamental distortion term can have a larger amplitude than the desired signal for m_f near one. This term contributes both amplitude and phase distortion. The third harmonic distortion is very small for large values of m_f, and for values of m_f near one, the third harmonic is usually outside the pass band of the audio amplifier. Thus, the fundamental distortion term is the most troublesome for this case.

NO. OF STAGES (r)	% FUNDAMENTAL DISTORTION	% 3RD HARMONIC DISTORTION
1	$.849/m_f$	$.142/m_f$
2	$1.200/m_f$	$.095/m_f$
3	$1.441/m_f$	$.083/m_f$
4	$1.819/m_f$	$.094/m_f$
5	$2.256/m_f$	$.114/m_f$

Fig. 3—Absolute values of amplitudes of the percentage harmonic distortion.

Using (20) or (22), the percentage harmonic distortion at any harmonic can be easily computed for a multiple stage synchronous single-tuned amplifier.

[6] S. N. Van Voorhis, "Microwave Receivers," McGraw-Hill Book Co., New York, N. Y., pp. 160–162; 1948.

The amplitude modulation in the output can be found from (12a) and (14), and using the value of A given in Appendix C,

$$M_D + 1 = \left(\frac{g_m}{g}\right)^r \cdot \frac{1}{\left[\left(1+\frac{1}{m_\sigma^2}\right)\left(1+\frac{m_b^2}{2}\right)\right]^{r/2}} \cdot \frac{1}{[1 + C\cos\omega_a t + D\cos 2\omega_a t]^{r/2}}, \quad (24)$$

where

$$C = \frac{4m_b\sigma^2}{(2+m_b^2)\Delta\omega}, \qquad D = \frac{m_b^2}{2+m_b^2}.$$

The writer was not able to obtain a Fourier series expansion for (24) with coefficients in closed form. Fortunately, most applications only require $r=1$, and for this case the coefficients can be obtained in the form of rapidly converging power series. This is illustrated in the next example.

B. Stagger-Tuned Discriminator.

The transfer function between the input current $i(t)$ and either output voltage, e_A or e_B, in Fig. 3, can be represented by a single pole transfer function, $n=1$ and

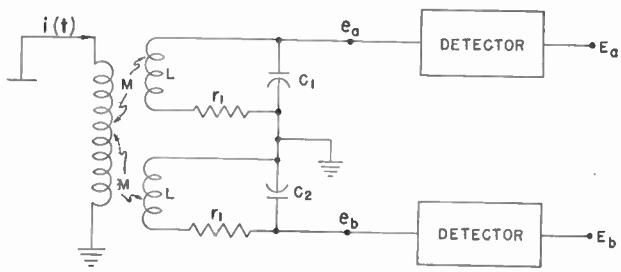


Fig. 4—Stagger-tuned discriminator.

$r=1$.[7] When used as a stagger-tuned discriminator circuit, one pole is tuned slightly higher, $+\sigma$, than the carrier and the other tuned slightly lower, $-\sigma$. If the secondary Q's are equal, the detected voltages (see (16)) at terminals A and B are

$$E_A = M_D(+\sigma) + 1 \quad (25a)$$

$$E_B = M_D(-\sigma) + 1. \quad (25b)$$

The detected discriminator outputs are then subtracted in the usual manner to give the audio output E_0.

$$E_0 = E_A - E_B = M_D(+\sigma) - M_D(-\sigma). \quad (26)$$

From (24) for $r=1$,

$$M_D(+\sigma) + 1 = \left(\frac{g_m}{g}\right)^r \cdot \frac{1}{\sqrt{\left(1+\frac{1}{m_\sigma^2}\right)\left(1+\frac{m_b^2}{2}\right)}} \sum_{n=0}^{\infty} a_n \cos n\omega_a t, \quad (27)$$

where

$$a_0 = 1 + \frac{3}{16}(C^2+D^2) - \frac{15}{64}C^2D + \cdots$$

$$a_1 = \frac{1}{2}C + CD - \frac{15}{64}C^3 - \frac{15}{32}CD^2 + \cdots$$

$$a_2 = \frac{1}{2}D + \frac{3}{16}(C^2+D^2) - \frac{15}{32}C^2D + \frac{3}{4}D^3$$

$$a_3 = CD - \frac{5}{64}C^3 - \frac{15}{64}C^2D.$$

It is clear that changing the sign of σ changes the sign of C. Therefore, combining (25a), (25b), and (26),

$$E_0 = \left(\frac{g_m}{g}\right)^r \frac{1}{\sqrt{\left(1+\frac{1}{m_\sigma^2}\right)\left(1+\frac{m_b^2}{2}\right)}} \cdot \sum_{n=1}^{\infty} b_{(2n-1)}\cos(2n-1)\omega_a t, \quad (28)$$

where, $b_1 = 2a_1$, $b_3 = 2a_3$, and so forth.

Two interesting facts are apparent:

1. All even harmonics are cancelled leaving only odd harmonics.
2. Amplitude distortion is present in the fundamental.

The amplitude of the fundamental would be proportional to m_b if there were no distortion. In other words, the amplitude of the fundamental should be proportional to the frequency deviation. An examination of the fundamental Fourier coefficient b_1, shows that the term C is the undistorted output, while the remaining terms give amplitude distortion. To minimize amplitude distortion, it is desirable to make C much larger than D, but smaller than one. This is done most conveniently by choosing the value of σ properly (i.e., $\sigma \geqq \Delta\omega$).

2. Input Amplitude-Modulated Wave

For an amplitude-modulated input, the modulating function takes the form,

$$M(\tau) = 1 + \cos\omega_a\tau, \quad M(y) = 1 + \cos\omega_a t(1-y). \quad (29)$$

Inserting this expression for $M(y)$ into equation (7), and performing the integration[8]

$$\alpha + j\beta = \log[H_0 + H\cos(\omega_a t - \phi)] \quad (30)$$

where,

$$H_0 = \sum_n \sum_r \frac{A_{nr}}{P_{nr}^r}(r-1)!$$

$$H_s = \sum_n \sum_r A_{nr}(r-1)!\left(\frac{1}{P_{nr}^2 + \omega_a^2}\right)^{r/2} \cdot \sin\left(r \tan^{-1}\frac{\omega_a}{P_{nr}}\right)$$

[7] L. Arguimbau, "Vacuum Tube Circuits," John Wiley and Sons, Inc., New York, N. Y., pp. 487–494; 1948.

[8] Bierens De Haan, "Nouvelles Tables D'Integrales Definis," G. E. Stechert, New York, N. Y., Tables 363, nos. 1 and 2, p. 507; 1939.

$$H_c = \sum_n \sum_r A_{nr}(r-1)!\left(\frac{1}{P_{nr}^2+\omega_a^2}\right)^{r/2} \cdot \cos\left(r \tan^{-1}\frac{\omega_a}{P_{nr}}\right)$$

$$H = \sqrt{H_c^2 + H_s^2}, \qquad \phi = \tan^{-1}\frac{H_s}{H_c} = \theta_\alpha + j\theta_\beta.$$

The output frequency modulation becomes

$$\beta' = \text{imag.}\left[\frac{-\omega_a R \sin(\omega_a t - \phi)}{1 + R\cos(\omega_a t - \phi)}\right], \text{ where } R = \frac{H}{H_0}. \quad (31)$$

This is similar in form to (20), using the Fourier series expansion

$$\beta' = 2\omega_a \sum_{m=1}^{\infty}(-1)^m\left(\frac{R}{1+\sqrt{1-R^2}}\right)^m \cdot [\sin m(\theta_R - \theta_s)\sinh m\theta_\beta \cos m(\omega_a t - \theta_\alpha) - \cos m(\theta_R - \theta_s)\cosh m\theta_\beta \sin m(\omega_a t - \theta_\alpha)], \quad (32)$$

where

$$m_r\epsilon^{j\theta_r} = R$$
$$m_s\epsilon^{j\theta_r} = 1 + \sqrt{1-R^2}.$$

It should be noted that if all the $\sigma_{nr}=0$, then $\beta'=0$. Thus β' exists only when some poles are not tuned to the carrier frequency. This means that, in general, when an amplifier is not tuned to the carrier of the input AM wave, FM appears in the output wave.

The amplitude modulation for the general case is

$$M_A = \epsilon^\alpha = |H_0 + H\cos(\omega_a t - \phi)|. \quad (33)$$

If $\sigma_{nr}=0$, H and H_0 are real, and there is no harmonic amplitude distortion; only phase distortion is present.

It is convenient at this point to consider a special case:

1. Synchronous Single-Tuned Amplifier: If the percentage bandwidth is small, the term of major importance is the rth term. Assuming further that the amplifier is tuned to the incoming carrier, it is clear from (32) that $\beta'=0$. The amplitude modulation becomes,

$$M_A = \frac{A}{\alpha^r}(r-1)!\left[1 + \left(\frac{\alpha^2}{\alpha^2+\omega_a^2}\right)^{r/2} \cdot \cos\left(\omega_a t - r\tan^{-1}\frac{\omega_a}{\alpha}\right)\right]. \quad (34)$$

Using this equation it is not difficult to calculate the output amplitude modulation for any value of r.

Conclusion

The theory presented in this paper illustrates a method by which a wide variety of distortion problems can be analyzed. The multiple-stage synchronous single-tuned amplifier has been analyzed and the method can be extended to synchronous double-tuned amplifiers with a little added labor. Stagger-tuned amplifier problems can be handled without much difficulty, if the number of different pole positions is only two or three and symmetry exists around the midfrequency. Further application of the method to nonsinusoidal periodic modulating functions can be made by evaluation of (7) by methods similar to steady-state operational analysis.

In all examples worked out in this paper, the resultant distortion is found in terms of quantities that can be easily checked experimentally.

Acknowledgment

The writer wishes to acknowledge the assistance of B. Haimowitz, Research Associate at the Moore School of Electrical Engineering, in obtaining several Fourier series expansions.

Appendix A

Expansion of B_0+B

In order to simplify notation, p will be used to denote differentiation with respect to y and $(p_n x)^m$ will signify the nth derivative of x with respect to y raised to the mth power. It will be understood that when n or m is negative, the expression is equal to zero. Then, from (8),

$$s!A_s\epsilon^{j\theta(o)} = [p^s\epsilon^{j\theta(y)}]_{y=0}$$
$$= \left[(pj\theta)^s + \frac{s(s-1)}{2}(pj\theta)^{s-2}(p^2j\theta) + \frac{s(s-1)(s-2)}{1\cdot2\cdot3}(pj\theta)^{s-3}(p^3j\theta) + \frac{s(s-1)(s-2)(s-3)}{2\cdot4}(pj\theta)^{s-4}(p^2j\theta)^2\cdots\right]_{y=0}\epsilon^{j\theta(o)}. \quad (35)$$

Denoting differentiation with respect to t by primes and noting that $p\theta(y) = -t\phi'(t)$, (35) can be combined with (8) and (9) to give

$$B_0 + B = \sum_n\sum_r\sum_{s=0}\frac{A_{nr}}{P_{nr}^r}\frac{(r-1+s)!}{s!} \cdot \left[R^s + j\frac{s(s-1)}{2}R^{s-2}\frac{\phi''}{P_{nr}^2} - j\frac{s(s-1)(s-2)}{1\cdot2\cdot3}R^{s-3}\frac{\phi'''}{P_{nr}^3} - \frac{s(s-1)(s-2)(s-3)}{2}R^{s-4}\frac{\phi''^2}{P_{nr}^4} + \cdots\right], \quad (36)$$

where $R = -j(\phi'/P_{nr})$.

For $|R|<1$, the summations over s can be performed and yield (11).

Appendix B

The Fourier series expansion of

$$\frac{r\omega_a X \sin \omega_a t}{1 + X \cos \omega_a t}, \qquad X = j\frac{\Delta\omega}{p} \tag{37}$$

is obtained as follows:

$$\frac{1}{1 + X \cos \omega_a t} = \frac{a_0}{2} + \sum_{m=1}^{\infty} a_m \cos m\omega_a t, \ |X| < 1, \tag{38}$$

where

$$a_m = \frac{1}{\pi}\int_0^{2\pi} \frac{\cos mx dx}{1 + X \cos x} = \frac{1}{\pi}\int_0^{2\pi} \frac{\epsilon^{imx} dx}{1 + X \cos x}$$

let $Z = \epsilon^{ix}$, $dZ = iZdx$

$$a_m = \frac{1}{\pi X i}\oint \frac{Z^m}{Z^2 + \frac{2}{X}Z + 1}$$

$$= \frac{2(-1)^m}{\sqrt{1 - X^2}}\left(\frac{X}{1 + \sqrt{1 - X^2}}\right)^m. \tag{39}$$

The integral around the unit circle being evaluated at the two poles.

Combining (37), 39), and (20),

$$\Omega_{Dr} = \text{imag.}\left[2r\omega_a \sum_{m=1}^{\infty} (-1)^m \cdot\left(\frac{X}{1 + \sqrt{1 - X^2}}\right)^m \sin m\omega_a t\right]. \tag{40}$$

Appendix C

Characteristic Transient for an r-Stage Synchronously Tuned Amplifier.

Using the narrow-band form for the Laplace transform of the admittance of a single-tuned circuit, i.e., $Y(p) = 2C(p - p_1)$, the characteristic transient for a synchronously tuned amplifier of identical stages is

$$\mathcal{T}(t) = \mathcal{L}^{-1}\left[\left(\frac{g_m}{2C}\right)^r \cdot \frac{1}{(p - p_1)^r}\right], \tag{41}$$

where

$$p_1 = \alpha_1 - j\sigma_1.$$

$\mathcal{L}^{-1}$ indicates the inverse Laplace transform. Then,

$$\mathcal{T}(t) = \left(\frac{g_m}{2C}\right)^r \frac{t^{r-1}}{(r - 1)!}\epsilon^{-p_1 t} \tag{42}$$

or,

$$A_r = \left(\frac{g_m}{2C}\right)^r \cdot \frac{1}{(r - 1)!}. \tag{43}$$

Correction

John Ruze, author of the paper, "Wide-Angle Metal-Plate Optics," which appeared on pages 53–59 of the January, 1950, issue of the Proceedings of the I.R.E., has advised the editorial department that Fig. 6 was in error.

Mr. Ruze therefore prepared the following illustration presented in the right-hand column, for the benefit of Proceedings readers who may have been confused. The editors are glad to co-operate in publishing this revised figure.

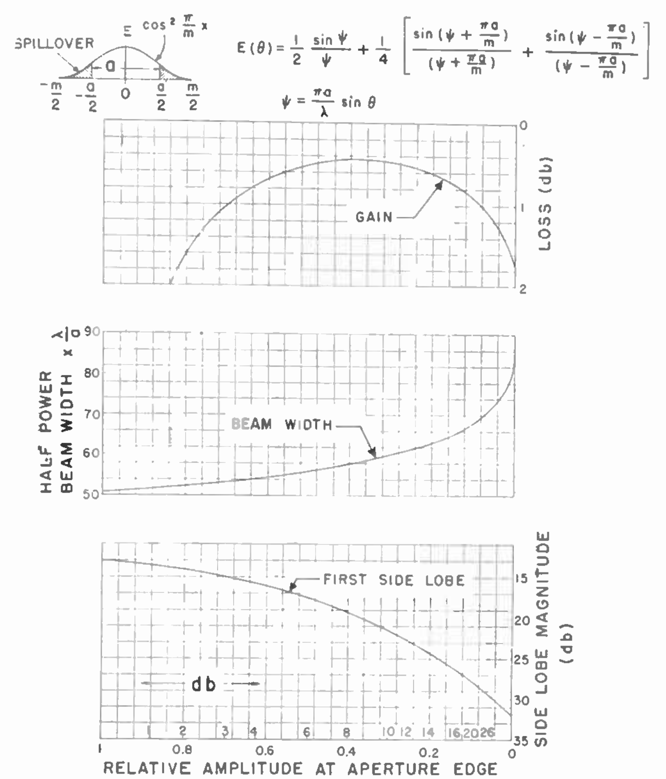


Fig. 6—Radiation characteristics of tapered rectangular aperture.

Contributors to the Proceedings of the I.R.E.

Richard C. Borden was born in Wilkes Barre, Pa., on April 9, 1919. After being graduated from Tulane University with the B.S. degree in physics in 1941, he was employed in the Division of Geomagnetism and Seismology of the United States Coast and Geodetic Survey. While on active duty in the Navy, he attended the Naval Officers' Radar Schools at Cruft Laboratory, Harvard University, and the Massachusetts Institute of Technology, and also served as radar officer aboard the aircraft carrier U.S.S. *Belleau Wood*, and as instructor at the Naval Officers' Pre-Radar School, Bowdoin College, Brunswick, Maine.

R. C. Borden

After separation from the Naval Service in 1946, Mr. Borden joined the engineering staff of the Civil Aeronautic Authority's Technical Development and Evaluation Center at Indianapolis, Ind. During this time he has been actively engaged in the development and evaluation of radar aids to air navigation and traffic control, particularly distance-measuring equipment. He is presently serving as Chief of the Radar Branch, Electronics Division.

❖

Malcolm R. Boyd (S'50) was born on January 23, 1924, in Boulder, Colo. He received the B.S. degree in electrical engineering from the University of Colorado in 1947. In addition, he attended the University of Pittsburgh and Penn State College under the ASTP. From 1947 to 1949 Mr. Boyd was employed at the RCA Laboratories in Princeton, N. J., where he was engaged primarily in gaseous electronic studies. Since September, 1950, he has been a graduate student and research assistant in the electrical engineering department at Stanford University.

M. R. Boyd

Mr. Boyd is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

For a photograph and biography of H. J. Reich, see page 1451 of the December, 1950, issue of the Proceedings of the I.R.E.

F. W. Bubb, Jr. (A'44) was born on July 12, 1921, at St. Louis, Mo. He received the B.S. degree in electrical engineering with honors in 1943, and the M.S. degree in applied mechanics in 1949, both from Washington University. From 1943 to to 1948 he was successively instructor at the Massachusetts Institute of Technology Radar School in Boston; chief product engineer at the Navy division of McQuay-Norris Manufacturing Company, St. Louis, Mo., a VT fuze assembly plant; senior design engineer at the Naval ordnance division of Eastman Kodak Company, Rochester, N. Y.; and research engineer engaged in work on automatic control devices at Washington University, St. Louis, Mo. During 1948, Mr. Bubb invented and has applied for patent on an electronic instrument for measuring and controlling fluid phases and extremely low humidities. In 1948, he joined the staff of St. Louis University as an assistant professor of engineering.

F. W. Bubb, Jr.

Mr. Bubb is a member of Tau Beta Pi, Sigma Xi, and Pi Mu Epsilon, and is a registered professional engineer in the fields of applied mechanics and electrical engineering.

❖

Georg Goubau (A'49) was born in Munich, Germany, on November 29, 1906. He received the Dipl. Phys. degree in 1930, and the Dr. Ing. degree in 1931, both from the Munich Technical University. From 1931 to 1939 he was employed in research and teaching in the physics department of the same university, under Professor Zenneck. During this time he was principally concerned with ionospheric investigations. He established the first German Ionospheric Research Station (Herzogstand/Kochel), and was in charge of the research work carried on at this station.

Georg Goubau

In 1939 Dr. Goubau was appointed professor and director of the department of applied physics at the Friedrich-Schiller University, in Jena, Germany. Before he arrived in this country, he was the senior author of the volumes on electronics of the FIAT Review of German Science, published by the Military Government for Germany. Dr. Goubau is now a consultant at the Signal Corps Engineering Laboratories, in Fort Monmouth, N. J.

William M. Hall (A'33–M'44–SM'45) was born in Burlington, Vt., on July 10, 1906. He attended the University of Vermont and received the B.S. degree in electrical communications from the Massachusetts Institute of Technology in 1928.

William M. Hall

From 1928 until 1930 he was with the Sound Picture Industry in New York, N. Y. Following this, he was on the staff of the electrical engineering department of MIT until the beginning of World War II. In 1934 he received the Sc.D. degree from that institution. During 1940 and 1941, Dr. Hall was on the staff of the MIT Radiation Laboratory.

Since 1941 he has been associated with the Raytheon Manufacturing Company, where he was a project engineer on various radar and countermeasures systems during the war and where he now holds the position of technical assistant to the manager of the radar department.

Dr. Hall is a member of Sigma Xi and a Fellow of the Acoustical Society of America.

❖

Norman L. Harvey (A'41–M'46–SM'48), was born in Ossian, Iowa in 1911. He was graduated from the Iowa State College in 1934 with the B.S. degree in electrical engineering. In 1941, Mr. Harvey joined the research department of Sylvania Electric Products, Inc., at Emporium, Pa., and was actively engaged in a number of developments for the armed services. In 1943 he was transferred to the Central Engineering department at Flushing, L. I., N. Y., as supervisor of the circuits section. Here his duties comprised responsibility for a number of developments including a special radar unit, a part of which was later used in the "moon radar" experiments, as well as in ultra-high-frequency communication equipment.

Norman L. Harvey

At present Mr. Harvey is director of engineering of the Colonial Radio and Television Division of Sylvania, in Buffalo, N. Y

❖

Meyer Leifer (A'46–M'48–SM'50) was born in New York, N. Y., in 1914. He received the B.S. and M.S. degrees from

Brooklyn College and Columbia University, respectively, the former in mathematics and the latter in physics. After teaching for several years in the New York City secondary schools, he joined the United States Navy and served as a radio-radar officer during the war. Mr. Leifer became a member of the physics laboratories staff of Sylvania Electric Products Inc. upon his release from active duty, and has been engaged principally in the application of the theory of communication to navigation and television systems.

MEYER LEIFER

Mr. Leifer is now the leader of the theoretical group of the circuits section, and is teaching in the Graduate School of the Brooklyn Polytechnic Institute. He is a member of the American Physical Society.

❖

Donald Gordon Lindsay was born on May 2, 1909, in Australia, and was educated at North Sydney High School and Sydney University, where he completed a two-year Science course. He joined the Amalgamated Wireless (A/sia) Ltd. in 1933 as a design engineer, after previous experience with Stromberg-Carlson (A/sia) Ltd.

DONALD G. LINDSAY

Mr. Lindsay's association with the Institution of Radio Engineers, Australia, as a foundation member dates from 1932. He was elected a Fellow of this Society in 1949.

❖

Nean Lund (A'40) was born in Madison, Wis., on May 13, 1912. He was graduated in 1935 from the University of Wisconsin, where he received the B.Sc. degree in electrical engineering. After a year of postgraduate study at that University, and a brief period as a radio consultant, he joined the staff of the Bell Telephone Laboratories, Inc. He first worked on the development and testing of single-sideband radio receivers. From 1941 to 1943 he was concerned with circuit development of loran, radar, and other equipment used by the military.

NEAN LUND

In recent years, Mr. Lund has worked on the measurement of interchannel modulation noise from multiplex single-sideband transmitters, and on the measurement of extraband radiation from multiplex single-sideband radio transmitters. He is a member of the American Institute of Electrical Engineers, Phi Eta Sigma, and Eta Kappa Nu.

❖

Louis Malter (A'37-SM'45) was born on April 28, 1907, in New York, N. Y. He received the B.S. degree from the College of the City of New York in 1926 and the M.A. and Ph.D. degrees in physics from Cornell University in 1931 and 1936, respectively. From 1926 to 1928 he taught physics at the College of the City of New York. He was with the Acoustic Research and Photophone Divisions of Radio Corporation of America from 1928 to 1930 and during the summer of 1931. From 1933 to 1942 Dr. Malter was associated with RCA Manufacturing Company at Camden, N. J., and from 1942 to 1943 with the RCA Laboratories Division at Princeton, N. J. During the period from 1943 to 1946, he was in charge of the Special Development Division of RCA Manufacturing Company at Lancaster, Pa. From May, 1946, to December, 1947, he was in charge of the Vacuum Tube Research Section of the Naval Research Laboratory in Washington, D. C. In December, 1947, he returned to the RCA Laboratories where he is now engaged in research on gaseous electronics.

L. MALTER

Dr. Malter is a fellow of the American Physical Society, and a member of Sigma Xi.

❖

Nathan Marchand (A'39-M'44-SM'50) was born on June 20, 1916, in Shawinigan Falls, Canada. He received the B.S. degree from the College of the City of New York and the M.S. degree from Columbia University, both in electrical engineering.

NATHAN MARCHAND

Mr. Marchand has been a lecturer in electrical engineering and has served as a consultant in the fields of cardiology and antenna and circuit design. He is at present the head of the circuits section of the physics laboratories, at Bayside, L. I., N. Y., as well as a consultant at the New York University-Bellevue Medical Center. He is the author of the books "Frequency Modulation," and "Ultrahigh Frequency Transmission and Radiation."

Mr. Marchand is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

James A. Marsh (S'49-A'50) was born in Youngstown, Ohio, on January 7, 1922. He received the degree of A.B. from Ohio Wesleyan University in 1946, and both the M.Sc. and Ph.D. degrees from The Ohio State University in 1947 and 1949, respectively, where his field of specialization was electrical engineering.

J. A. MARSH

During the war, Dr. Marsh was an electronics officer in the Army Air Force, working in the field of radar. He taught for the year 1947–1948 on the electrical engineering department staff at The Ohio State University. From 1948 to 1950 he was associated with the Antenna Laboratory of The Ohio State University, where he did research on flush-mounted airborne antennas. In February, 1950, Dr. Marsh joined the staff of the Aerophysics Laboratory of North American Aviation, Inc. He is now directing research in the antenna and microwave section. Dr. Marsh is a member of Phi Beta Kappa, Eta Kappa Nu, Pi Mu Epsilon, and Phi Mu Alpha.

❖

De Forrest Metcalf (M'48) was born in El Paso, Tex., on November 25, 1916. After receiving the B.S. degree from the University of Texas in 1937 he was employed by the Humble Oil and Refining Company. In 1942 he entered the employ of the Raytheon Manufacturing Company and participated in its microwave development program. He joined the staff of the Electrical Engineering Research Laboratory of The University of Texas in 1948, where he has been in charge of the wave propagation research conducted under contract with the National Bureau of Standards.

D. F. METCALF

Mr. Metcalf is a member of Tau Beta Pi.

❖

Wilbur L. Pritchard (A'45-M'48) was born in New York, N. Y., on May 31, 1923. He received the B.E.E. degree from the College of the City of New York in 1943, and has since done graduate work at the Massachusetts Institute of Technology.

W. L. PRITCHARD

During the period from 1943 to 1946, Mr. Pritchard worked as an engineer for the Philco Radio and Television Corporation, where he was engaged in the development of airborne radar systems and later with the home radio and phonograph division. Since 1946 he has been associated with Raytheon Manufacturing Company, first as a development engineer in the microwave communications department and then as section head in charge of the radio-frequency and antenna section, his present position.

❖

Louis E. Raburn (S'41–A'44–M'46–SM'50) was born in Manhattan, Kan. on June 4, 1919, and received the B.S. degree in electrical engineering from Kansas State College in 1941. From 1935 to 1941 he was employed as an engineer at radio station KSAC. From August, 1941, until March, 1942, he was employed as a carrier engineer with the long lines department of the American Telephone and Telegraph Company.

Louis E. Raburn

From March, 1942, until October, 1945, Mr. Raburn was a research associate at Radio Research Laboratory, Cambridge, Mass., and attended Harvard University graduate school.

Mr. Raburn has been with Electronics Research, Inc., since November, 1945. He has been engaged in the development of low-drag and zero-drag aircraft antennas, FM broadcast antennas, and antenna test equipment. In addition, as an antenna engineer, he has supplied consultant services to several aircraft companies on the application of zero-drag antennas. At present, Mr. Raburn is the senior engineer of the antenna division of Electronics Research, Inc.

❖

Leon Riebman (S'43–A'45) was born on April 22, 1920, in Coatesville, Pa. He received the B.S. degree in electrical engineering in 1943, and the M.S. degree in 1947, both from the Moore School of Electrical Engineering, University of Pennsylvania.

Leon Riebman

After attending the United States Navy Midshipman School, he was assigned to duty in radar development at the Navy Research Laboratory in 1944. In March, 1946, he joined the research and development staff of Philco Corporation as a senior engineer. From October 1, 1948, until June, 1949, he attended the Moore School of Electrical Engineering, University of Pennsylvania, under an Atomic Energy Commission pre-doctoral fellowship. Since June, 1949, he has been associated with the Moore School as a Research Associate and part-time Instructor in electrical engineering.

Mr. Riebman is a member of Sigma Xi and Tau Beta Pi.

❖

Peter M. Schultheiss (S'50–A'51) was born in Munich, Germany, on October 18, 1924. He received his education in Germany, England, and the United States, where, in 1945, he received the B.E. degree from Yale University.

Credit C. T. Alburtis

P. M. Schultheiss

During the period from 1944 to 1946, Mr. Schultheiss served with the U. S. Army Signal Corps. Following this he returned to Yale for graduate work and received the M.E. degree in 1948. In that year he was appointed an instructor in electrical engineering at Yale, where, since 1950, he has been an assistant professor.

Mr. Schultheiss has specialized in the field of servomechanisms and has also been active in research on transistor circuits.

❖

George Sinclair (A'37–SM'46) was born in Hamilton, Ont., Canada, on November 5, 1912. He received the B.Sc. degree in electrical engineering in 1933 and M.Sc. degree in 1935 from the University of Alberta, and the Ph.D. degree in 1946 from the Ohio State University. Dr. Sinclair was an instructor in electrical engineering at the University of Alberta for one year, and an engineer for the Northern Broadcasting Corporation for two years.

George Sinclair

From 1941 to 1947 Dr. Sinclair was a research associate in the department of electrical engineering of the Ohio State University, supervising the research program of the Antenna Laboratory. He is now an associate professor of electrical engineering at the University of Toronto, and is also a consultant for the Antenna Research Laboratory Inc., Columbus, Ohio.

❖

Frederick William Schott (S'41–A'45–M'50) was born in Phoenix, Ariz., on October 2, 1919. In 1940, he received the degree of Bachelor of Arts (Honors) from San Diego State College. In 1941, he was awarded a university scholarship and began graduate work at Stanford University. The following year he continued under a Westinghouse Fellowship, and received the degree of electrical engineer in 1943. He resumed graduate study at Stanford University in 1947 and received the Ph.D. degree in 1948.

F. W. Schott

He has held the following positions: junior engineer, San Diego Gas and Electric Co., San Diego, Calif., 1943 to 1944; ensign, United States Navy, 1944 to 1946; instructor in engineering and physics, San Diego State College, 1946 to 1947; acting instructor, Stanford University, summer, 1947; research associate, Stanford University, 1947 to 1948; and assistant professor, University of California at Los Angeles, 1948 to date. During 1949 and 1950 he was on leave of absence with the U. S. Navy Electronics Laboratory.

Dr. Schott is a member of Tau Beta Pi, Sigma Xi, Sigma Pi Sigma, the American Association for the Advancement of Science, and the American Institute of Electrical Engineers, and is a registered electrical engineer in the State of California.

❖

Robert C. Sprague was born on August 3, 1900, in New York, N. Y. He attended Hotchkiss, the United States Naval Academy and Post-Graduate School, and the Massachusetts Institute of Technology, where he also took postgraduate work during 1923–1924. He continued his naval career as a naval architect and later was a member of the staff which superintended the design and building of the aircraft carrier U.S.S. *Lexington*.

R. C. Sprague

Mr. Sprague founded the Sprague Electric Company (originally the Sprague Specialties Co.), in 1926. In addition to being president and director of this company, he is a director of the Sprague Products Company of North Adams, Mass., and also of the Herlec Corporation, of Milwaukee, Wis. He actively participated in co-ordinating industry and government activities during World War II as a member of the WPS Advisory Committee on Electric Condensers, 1942–1945; chairman of the OPA Industry Advisory Committee for the Radio Parts Industry, 1944; and member of the Executive Committee of the Massachusetts Committee on Postwar Reconversion, 1942. He has been a Director of the Radio Manufacturers' Association since 1943, and has been chairman of the RMA Parts Division for two terms, 1944–1945 and 1945–1946. More recently he has been chairman of the RMA "Town Meetings" committee, which directed numerous activities in the interest of radio and television dealers and service technicians. He is now chairman of the board of Radio-Television Manufacturers' Association.

Mr. Sprague is a Fellow of the American Institute of Electrical Engineers, and also vice-president and member of the executive committee of the Associated Industries of Massachusetts.

❖

A. W. Straiton (M'47–SM'49) was born in Tarrant County, Tex., on August 27, 1907. He received the degrees of B.S. in

Institute News and Radio Notes

Technical Committee Notes

A Task Group on Pulse Definitions was set up under the **Standards Committee** recently, for the purpose of obtaining a set of definitions which would satisfy the needs of all IRE Technical Committees. This Task Group, headed by C. J. Hirsch of Hazeltine, has prepared the **Standards on Pulses: Definitions of Terms, 1951,** which is published in this issue of the PROCEEDINGS. J. G. Brainerd announced that arrangements have been made by the IRE Executive Committee to effect co-operation with the Institution of Radio Engineers in Australia on standardization work.

The Institute wishes to express its gratitude to Professor Brainerd for his leadership and guidance during the past two years as Chairman of the Standards Committee, and for his sincere interest and co-operation as IRE Representative on the ASA Standards Council. A. G. Jensen, the new Standards Committee Chairman, will replace Professor Brainerd on the ASA Council.

A meeting of the **Receivers Committee** was held on March 28 under the Chairmanship of R. F. Shea. This Committee has prepared two standards: namely, a proposed Standards on Radio Receivers: Open Field Methods of Measurement of Spurious Radiation, Frequency Modulation and Television Broadcast Receivers; and a proposed Standards on Receivers: Definitions of Terms. Both of these were submitted to the Standards Committee for approval on April 27.

The **Committee on Sound Recording and Reproducing** held a meeting on April 3, H. E. Roys, Chairman, at which the chairmen of the various subcommittees announced their tentative work schedules for the year. This Committee is concerned primarily with definitions of terms and measurement procedure in the fields of magnetic, mechanical, and optical recording.

A. Gardner Fox, Chairman of the **Antennas and Waveguides Committee,** presided at a meeting of that group on April 10. The Committee is drafting definitions of terms relating to waveguides and transmission lines.

R. E. Shelby and R. M. Bowie have been appointed Chairman and Vice-Chairman, respectively of the **Television Systems Committee.**

A meeting of the **Circuits Committee** was held on April 13, W. N. Tuttle, Chairman. Work on the preparation of definitions is progressing.

The **IRE Co-ordination Committee with International Technical Organizations (CCITO),** under the chairmanship of Axel G. Jensen, held a meeting on April 9 and proposed a question regarding a standard frequency-band designation to be put before the CCIR at its next convention in Geneva.

Atomic Standards of Length Now Available

The availability to science and industry of an ultimate standard of length was announced today by the National Bureau of Standards and the Atomic Energy Commission. The standards consist of spectroscopic lamps containing a single pure isotope of mercury. These lamps enable any research organization which has the auxiliary optical equipment to have for the first time an ultimate primary standard of length in its own laboratories.

Distribution of the lamps will be handled by the National Bureau of Standards. They will be available to qualified government, industrial, and education laboratories, both in this country and abroad, engaged in precision length measurements and related research. All requests for information and applications should be addressed to the Co-ordinator of Atomic Energy Commission Projects at the National Bureau of Standards, Washington 25, D. C.

Professional Group Notes

Two new Professional Groups were formed during the month of May: the **IRE Professional Group on Industrial Electronics,** and the **IRE Professional Group on Information Theory.** Two petitions for the formation of the **Industrial Electronics Group** were received; one from Eugene Mittelmann of Chicago, Ill., and one from Carl E. Smith of the United Broadcasting Co., Cleveland, Ohio. On May 22, this Group held its first conference in Cleveland, Ohio, sponsored jointly with the Cleveland Sections of the IRE and the AIEE. Eight technical papers were presented.

At least four Professional Groups have already decided to assess their members in order to raise money for the purposes of Group publications, or for conducting Group symposia. Up to **$1,000** of each Group's funds will be matched by Institute funds, and the Groups will be well on their way toward financial independence.

> The following newly organized Professional Groups have not yet had an opportunity to conduct a national membership drive:
>
> **Airborne Electronics**
> **Industrial Electronics**
> **Information Theory**
> **Radio Telemetry and Remote Control**
>
> Membership application forms for these Groups are available upon request from IRE Headquarters.

The Editorial Department of the IRE has offered to serve the Professional Groups in the following ways: (1) it will endeavor to procure papers for Groups; (2) it will notify a Group when a paper of particular interest to it is received; (3) it will publish a boxed notice preceding a paper procured by a Group, stating that the paper originated with that Group, (4) upon submission of a Group paper for publication, it will give it expedited handling; (5) it will furnish suggested formats for Professional Group publications for both mimeographed or offset publications, and for the more formal Group TRANSACTIONS; and (6) it will place at the disposal of the Groups the experience of the Editorial Department in editing or making arrangements with printers.

A meeting of the Administrative Committee of the **IRE Professional Group on Airborne Electronics** was held in Dayton, Ohio, on April 5. Plans were made for the Group's participation in the Airborne Electronics Conference, which was held from May 23 to May 25 at the Biltmore Hotel in Dayton. The Group sponsored an evening symposium and a luncheon at the Conference, and held a business meeting of its members. A constitution and bylaws were drawn up.

Abstracts of papers presented at the URSI/IRE meeting held in Washington, D. C., on April 16–19 appear in this issue of the PROCEEDINGS. The Institute's participation was conducted by the **IRE Professional Group on Antennas and Propagation.** The Group hopes that these papers will later appear in complete form in a publication of the Group's TRANSACTIONS. A meeting of the Administrative Committee of the Group was held on April 18 in Washington, D. C., to discuss plans for participation in the IRE West Coast Convention which will take place in San Francisco, Calif., August 22–24. There will be three sessions sponsored jointly by the Group and the West Coast Convention.

The **IRE Professional Group on Audio** is currently making plans to publish two papers for its members.

A constitution and bylaws are being drawn up by the **Professional Group on Broadcast Transmission Systems.** The local Group in Boston held a meeting on March 8 in the studios of WCOP. Forty-five people attended, and Roger W. Hodgkins, Chairman, presided. A paper entitled, "Use and Interpretation of Noise and Distortion Meters," was presented by Charles Cady. A meeting has been held, also, in Troy, N. Y., to initiate a local Group in Northern New York State.

The **Professional Group on Circuit Theory** is drawing up a constitution and bylaws. Several IRE Sections have shown interest in the formation of local Circuit Theory Groups.

The **IRE Professional Group on Instrumentation** has a financial balance resulting from co-sponsorship of the High-Frequency Measurements Conference in January of this year. This will be matched by Institute funds in accordance with the new IRE policy. The following people have been elected as new members of the Group's Administrative Committee to serve a three-year term commencing July 1: John F. Byrne, Rudolf Feldt, W. D. Hershberger, and Hugh S. Knowles. Several members of the Group, under the leadership of W. M. Rust, Jr., Regional Director of IRE Region No. 6, will participate in the Sixth National Instrument Conference to be held in Houston, Texas, next fall, by organizing an IRE session on Instrumentation. The Group will organize a session on High-Frequency Measurements for the 1951 National Electronics Conference in Chicago, Ill., next October. The local Detroit Group, with 32 members under the Chairmanship of Seymour Sterling, has already scheduled several meetings, and its activities are well under way. Interest in the formation of additional local Groups exists in Chicago, Los Angeles, and in other areas.

The money earned by the **Professional Group on Nuclear Science** as a result of its Conference on Electronic Instrumentation in Nucleonics and Medicine will be retained by the Group and augmented by a similar amount from Institute funds in accordance with the new financial policy for Professional Groups. The Group is planning another conference, to be sponsored jointly with the AIEE, which will probably be held next fall. Urner Liddell will serve as Chairman of the Planning Committee for the conference. Ballots for the election of two new members of the Group's Administrative Committee have been mailed to the membership.

The **IRE Professional Group on Quality Control** has amended its constitution and bylaws, and has submitted them for approval. Election of six new members to the Group's Administrative Committee is being conducted. The Committee held a meeting on April 23 in Chicago, Ill., details of which will be reported in these Notes next month.

A membership drive is currently being conducted by the **Professional Group on Radio Telemetry and Remote Control.** The Group plans to publish two newsletters each year, to sponsor papers for the IRE West Coast Convention, and to hold a symposium on Telemetering on the West Coast in the spring of 1952.

Ballots have been mailed to members of the **Professional Group on Vehicular Communications** to elect five new members of the Administrative Committee. Plans are under way for the Group's technical meeting to be held in the fall. Reprints of five technical papers which have appeared in the PROCEEDINGS were furnished through the courtesy of Bell Telephone Laboratories, and mailed by IRE Headquarters to the Group's entire membership.

IRE Members, Eligible for Draft—Attention!

The Signal Corps has recently established the following procedures which will assist in the proper assignment of technically skilled IRE members upon entry into the Army:

The Institute of Radio Engineers will provide, upon request, to members who are about to be called into military service in enlisted grades, a Statement of Experience indicating the most applicable military specialties in the radio-electronic field for which a member's experience and background qualify him. This Statement of Experience should be presented by the member upon reporting for military service for reference in assignment to duty.

Interested members should write to George W. Bailey, Chairman, Liaison Committee, 1 East 79 St., New York 21, N. Y., giving full particulars as to education and experience in the radio-electronic field.

British IRE Plans Summer Convention

The British Institution of Radio Engineers is planning to hold its third convention this summer, during the Festival of Britain. The program will include the following sessions: "Electronic Instrumentation in Nucleonics," University College, London, July 3–4; "Valve Technology and Manufacture," University College, London, July 5–6; "Radio Communication and Broadcasting," University College, Southampton, July 24–25; "Radio Aids to Navigation," University College, Southampton, July 26–27; "Television Engineering," King's College, Cambridge, August 21–24; and "Audio-Frequency Engineering," The Richmond Hall, Earls Court, September 4–6.

Oak Ridge Institute to Give Additional Courses

Three additional basic courses in radioisotope technique of four-week duration, and a three-week autoradiography course will be offered this summer by the Special Training Division of the Oak Ridge Institute of Nuclear Studies.

The basic courses will be the 21st, 22nd, and 23rd to be given by the Institute since the program was initiated in June, 1948, and are scheduled to begin on June 11, July 9, and August 15. They are to combine lectures, demonstrations, and laboratory work. Sufficient space is available to permit individual laboratory work. Thirty-two participants will be accepted for each course.

The autoradiography course beginning on July 2 will be the first of its kind to be offered by the Institute. It is intended for personnel who will direct medical or biological research utilizing the autoradiographic process.

George A. Boyd, formerly of the University of Rochester, well-known for his work in autoradiography, will direct the course. Subjects to be covered include the theory of photographic process, reaction of ionizing particles with photographic emulsions and the interpretation of results, and techniques of making gross and microscopic autoradiograms. Twenty participants will be accepted for the course.

Lecturers and laboratory leaders of the course will include the following: Julian Webb and John Spence, Eastman Kodak Co.; Robert Dudley, MIT; Margaret Holt, New England Deaconess Hospital; C. P. Leblond and Rita Bogorach, McGill University; L. F Belanger, University of Ottawa S. R. Pelc, Hammersmith Hospital, London; Agnes Williams, University of New Mexico; and Harvey Blank, University of Pennsylvania.

Registration is $25 per course. Additional information and application forms are available from Ralph T. Overman, Chairman, Special Training Division, Oak Ridge Institute of Nuclear Studies, P.O. Box 117, Oak Ridge, Tenn.

Engineers Wanted!

Electrical and electronics engineers are urgently needed for work in research and development with the Armed Forces Security Agency. All positions are permanent insofar as this agency is concerned, and are located in the metropolitan area of Washington, D. C. Neither Civil Service examinations nor Civil Service status is required. Salaries range from GS-7, $3825.00, to GS-12, $6400.00 per annum, depending upon applicants' education and experience. For further information, write to the Director, Armed Forces Security Agency, Washington 25, D. C., Att'n: AFSA-153C.

Engineering Research Fellowships Offered

Research fellowships in electrical engineering will be awarded by the Institute of Industrial Research, University of Denver. The stipend is $3,300 for 21 months starting September 1, 1951. The recipients will work approximately 25 hours per week in the Institute of Industrial Research on industrial research projects leading to the fulfillment of the thesis requirement, while engaged in graduate study leading to the M.S. degree in electrical engineering. Interested applicants should send transcript of college record, personal data, and names of three persons acquainted with their work to the Director, Institute of Industrial Research, University of Denver, Denver 10, Col.

Attention!

Firms in New York Area

The New York Section of the IRE is interested in arranging field trips of technical interest to its members to places conveniently located to New York City. It is contemplated that such trips be scheduled on Saturdays, and that about 100 or more persons participate.

Organizations in the communications, electronic, or allied fields able to accommodate an IRE field trip should contact A. L. Stillwell, Chairman of Field Trip Committee, IRE New York Section, Bell Telephone Laboratories, Inc., Murray Hill, N. J.

Over 600 Attend New England Radio Engineering Meeting

On April 21 the Copley Plaza Hotel in Boston, Mass., was the scene of a very successful New England Radio Engineering Meeting, sponsored by the North Atlantic Region of the IRE. The all-day program of papers and exhibits attracted over 600 engineers, scientists, and educators from the six-state area.

The morning technical session was under the chairmanship of Herbert L. Krauss, Chairman of the Connecticut Valley Section. The following papers were presented: "Radio Frequency Problems in the Design of a Linear Accelerator," by H. L. Schultz and W. G. Wadley, Yale University, New Haven, Conn.; "Considerations in the Design of a Line of Inexpensive Test Equipment," by D. B. Sinclair, General Radio Co., Cambridge, Mass.; and "Spectrum Utilization in Color Television," by R. B. Dome, General Electric Co., Syracuse, N. Y.

Willard H. Hansen, Chairman of the Boston Section, presided over the afternoon session at which three more papers were read: "Physiological Effects of Radiation," by W. A. Meissner, New England Deaconess and Baptist Hospitals, Boston, Mass.; "Civilian Defense Against Atomic Attack," by J. W. M. Bunker, Massachusetts Institute of Technology, Cambridge, Mass.; and "Instrumentation in the Field of Radioactivity," by W. A. Higinbotham, Brookhaven National Lab., Upton, L. I., N. Y.

Following the afternoon session, a Regional meeting was held for IRE members with Regional Director H. J. Reich presiding.

Sharing the spotlight with the sessions was the informative exhibit of recent radio-engineering and electronic developments. The products and services of 101 firms were shown in an outstanding representation of the radio-electronic industry.

Social activities included a luncheon at which IRE President I. S. Coggeshall gave an address of welcome.

Calendar of COMING EVENTS

54th Annual Meeting of the American Society for Testing Materials, Chalfonte-Haddon Hall, Atlantic City, N. J., June 18–22

IRE 7th Regional Conference, University of Washington, Seattle, Wash., June 20–22

Electron Devices Conference, University of New Hampshire, Durham, N. H., June 21–22

1951 AIEE Summer General Meeting, Royal York Hotel, Toronto, Ont., Canada, June 25–29

1951 Annual IAS Summer Meeting, 7660 Beverly Blvd., Los Angeles, Calif., June 27–28

Institute of Navigation National Meeting, Hotel New Yorker, New York, N. Y., June 28–30

International Isotope Conference, Oxford, England, July 16–21

1951 IRE West Coast Convention, San Francisco, Calif., Aug. 22–24

1951 National Electronics Conference, Edgewater Beach Hotel, Chicago, Ill., October 22–24

Radio Fall Meeting, King Edward Hotel, Toronto, Ont., Canada, October 29–31

1952 IRE National Convention, Waldorf-Astoria Hotel, and Grand Central Palace, New York, N. Y., March 3–6

Pacific Coast Regional Conference Slates Program

The Seventh IRE Regional Conference, to be held at the University of Washington, Seattle, Wash., on June 20, 21, and 22, has announced the following program:

On Wednesday morning, June 20, registration will take place in the Student Union Building, after which a paper on color television will be read.

In the afternoon the following subjects will be considered: transformer design, new coupling circuits, loudspeaker design, optimum servo design, and computers.

On Thursday morning, June 21, papers will be read on such topics as telemetering in 12 mev, a low-frequency generator, a micro-second pulser, and impedance measurement to 1,000 mc.

The afternoon session will be devoted to multiple operation of transmitters and receivers, linear rf amplifier circuits, spectrophotometers, facsimile, cutoff waveguides, and a miniature hf amplifier.

The topics to be dealt with on Friday morning, June 22, are as follows: lf ionosphere measurement, ground-wave phase shift, aircraft antenna design, VOR cavity antenna, a slot aircraft antenna, and signal distribution from broadcast DA.

In the afternoon theater television equipment, a 30-inch tv receiver, uhf television systems, a 5-kw uhf tv amplifier, and a micro-wave relay for television discussed.

In addition to these technical sessions, field trips are being planned to the Puget Sound Navy Yard on a destroyer escort, to the Boeing Aircraft Co., to the University of Washington's new cryclotron, to a television station, and other places of interest.

Scientists Wanted For Civil Service Positions

The United States Civil Service Commission has called attention to the continuing need in the Federal Service for chemists, metallurgists, physicists, mathematicians, and electronic scientists. Applications are being accepted for positions in and around Washington, D. C., paying from $3,825 to $10,000 a year.

No written test will be given to applicants for these positions. To qualify, they must meet a basic requirement of appropriate college education and/or experience. In addition, they must show that they have had professional experience in the field for which they apply. These and other requirements are explained in detail in the examination announcement.

Information and applications may be obtained at most first- and second-class post offices, from Civil Service regional offices, or from the United States Civil Service Commission, Washington 25, D. C. Applications are to be sent to the Executive Secretary, U. S. Civil Service Committee of Expert Examiners, National Bureau of Standards, Washington 25, D. C., where they will be accepted until further notice.

Industrial Engineering Notes[1]

Controls

Defense Production Administrator William H. Harrison has announced the **establishment of an Electronics Production Board** which will be responsible for over-all coordination of electronics production under the mobilization program. DPA officials stated that the board "will consist of a Chairman to be appointed by the Administrator, the Director of the Electronics Division of the National Production Authority, the General Manager of the Atomic Energy Commission, representatives of the Defense Production Administration and the Department of Defense, and such additional representatives of the Munitions Board and the Armed Services as may be agreed upon." Harry Ehle, International Resistance Co., and Don G. Mitchell, Sylvania Electric Products Inc., have been designated to represent the Army and Air Force, respectively, on the Board. The Navy will be represented by Vice Admiral A. G. Noble, Chief, Office of Naval Material, and the Munitions Board by Chairman John D. Small. W. W. Watts, assistant to DPA Administrator Harrison, has been serving as Acting Chairman of the Board pending the appointment of a permanent chairman. . . . The National Production Authority is considering a **method to maintain production of receiving tubes** at the highest possible level until defense requirements are known. This disclosure was made at a recent meeting of the Receiving Tube Industry Advisory Committee with NPA officials. Committee members advised NPA that the industry's biggest problem is obtaining adequate nickel supplies. NPA pointed out that because of the industry's importance to the defense effort, the agency plans to reserve a supply of nickel which would be made available to the industry during the second quarter of this year. . . . All manufacturers of radio and television receivers have received a **letter from RTMA General Manager James D. Secrest** informing them that a critical situation may exist on the availability of selenium suitable for use in selenium rectifiers. A questionnaire accompanying the letter called for each company's anticipated requirements for se-

[1] The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of March 16, March 23, March 30, April 6, and April 13, published by the Radio-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

lenium rectifiers. When returned to RTMA Washington headquarters, the data will be compiled on an industry-wide basis and presented to the National Production Authority so that agency will know the over-all industry requirements for selenium. Data submitted by individual companies will be confidential, and will be identified by code number rather than by name. . . . The National Production Authority was warned that the electronic industry is running into trouble in obtaining a specific type of copper used in magnetrons, resistors, and other equipments. **The shortage of OFHC copper** is said to be due to a shortage of facilities to process and draw the metal as well as to the copper shortage. NPA has moved to aid the industry in obtaining the metal. At the same time it has been learned that the Electronic Production Board is acting now to prevent an expected acute shortage of tungsten in the radio-television industry.

Military News

The Armed Services plan to issue a **new policy on reservists** which may assist industry in planning with regard to valuable trained technical personnel who are members of the Reserve. One of the major objectives of the policy is to give reservists not yet called definite information as to whether they may be summoned to active duty, and when. Further, it is expected to provide for release of some reservists now in the service. . . . **Brig. Gen. George Irving Back, General MacArthur's Signal Officer in Tokyo since 1947, has been nominated by the President to be Chief Signal Officer of the U. S. Army.** If confirmed by the Senate, he will succeed Maj. Gen. S. B. Akin, who was Chief Signal Officer from April 1, 1947, until he retired from the Army on March 31 of this year. During World War II, from September, 1944, to November, 1945, General Back served in the Mediterranean Theater of Operations, as Deputy Chief Signal Officer of the Allied Force Headquarters, and as Chief Signal Officer of the Mediterranean Theater. He also held several important wartime positions in the Office of the Chief Signal Officer in Washington. . . . **Maj. Gen. Francis H. Lanahan, Commanding General of the Signal Corps Center, Fort Monmouth, N. J., has been assigned to Supreme Headquarters Allied Powers in Europe** and is expected to report in the near future, according to the Department of the Army. His successor will be announced at a later date.

Television News

Initial steps toward **lifting the "freeze" on new television station construction** were taken recently by the Federal Communications Commission with the issuance of a "Third Notice of Further Proposed Rule Making." Public hearings begin on June 11. Principal provisions of the FCC's new allocation plan are: Sixty-five or 70 uhf channels are to be made available for television in addition to present 12 vhf channels, laying ground work for about 2,000 television stations ultimately; 52 uhf channels will be made available throughout the United States on an intermixture basis with vhf channels; increased power for existing vhf stations is provided; adjacent channel separation for vhf and uhf stations is reduced to 70 and 65 miles, respectively, while the minimum separation of stations on adjacent channels is 60 miles for vhf and 55 for uhf; 31 vhf stations now operating will be required to shift channels if the plan is adopted; provision is made for earmarking about 10 per cent of the channels, chiefly in the uhf band, or about 200 stations for noncommercial educational use; earlier lifting of vhf "freeze" in Alaska, the Hawaiian Islands, Puerto Rico and the Virgin Islands is proposed. While the Commission definitely has "broken the ice" of the "freeze" and has moved towards a partial lifting in the near future, it is most unlikely that the construction ban can be lifted before autumn. The allocation of either 65 or 70 uhf channels will depend on how the FCC rules on a common carrier request for five channels in the 470–500 mc band. If this proposal is rejected by the FCC, Channel No. 14 would begin at 470 mc; but, if it is granted, the uhf broadcasting channels would commence at 500 mc. The FCC gave no indication when this issue would be decided. In rejecting industry proposals to keep uhf and vhf stations separate, the FCC said it had decided "that the adoption of an assignment table based on nonintermixture constitutes a short-term view of the problem and is inadvisable." In proposing a reduction in adjacent channel separation, the FCC stated that problems which have been encountered have not been serious and can be solved to a very considerable extent by improvements in receiver design which are neither difficult nor costly. The Commission added that it is of "the opinion that these separations should be based upon receiver performance which may reasonably be expected of manufacturers and not on the characteristics of the poorer receivers." Recognized as a victory for Commissioner Frieda Hennock, the reservation of some 200 assignments for noncommercial education use drew supplemental opinions from Chairman Coy and Commissioner Hennock. Commissioners E. M. Webster, George E. Sterling, and Robert F. Jones dissented in part from the decision. . . . **The standard intermediate frequency of 41.25 mc** recommended by the RTMA engineering department was "taken into account" by the FCC in making the uhf allocations, the Commission said in an appendix to its proposed order, and added that the use of an intermediate frequency of 111 mc in television sets "is not feasible at this time." . . . **The frequencies of 31 vhf stations now broadcasting will be shifted** if the FCC's new allocation plan is adopted as proposed. Only three such shifts were proposed in the Commission's earlier allocation plan. "In preparing the 'table' of television allocations," the Commission stated it "proposes to alter existing television authorizations in 31 instances. The alterations with respect to Channel 9 in the city of Cleveland, Ohio, Channel 5 in the city of Syracuse, N. Y., and Channel 6 in the city of Rochester, N. Y., resulted from the Commission's efforts to arrive at an equitable distribution of television channels between the United States and the Dominion of Canada. The remaining 28 channel substitutions resulted from the Commission's efforts to reduce interference, make available a reasonable number of channels, and to effect the maximum utilization of vhf television channels in the United States." The stations which would be required to change frequencies are listed in Table I.

TABLE I

Licensee or Permittee	City Affected	Present Channel Assignment	Proposed Channel Assignment
WOI-TV	Ames, Iowa	4	5
WSB-TV	Atlanta, Ga.	8	11
WBRC-TV	Birmingham, Ala.	4	6
WTTV	Bloomington, Ind.	10	4
WBKB	Chicago, Ill.	4	2
WLWT	Cincinnati, Ohio	4	5
WKRC-TV	Cincinnati, Ohio	11	12
WCPO-TV	Cincinnati, Ohio	7	9
WXEL	Cleveland, Ohio	9	8
WNBK	Cleveland, Ohio	4	3
WLWC	Columbus, Ohio	3	4
WLWD	Dayton, Ohio	5	2
WHIO-TV	Dayton, Ohio	13	7
WOC-TV	Davenport, Iowa	5	6
WLAV-TV	Grand Rapids, Mich.	7	8
WSAZ-TV	Huntington, W. Va.	5	8
WJAC-TV	Johnstown, Pa.	13	6
WGAL-TV	Lancaster, Pa.	4	8
WAVE-TV	Louisville, Ky.	5	3
WHAS-TV	Louisville, Ky.	9	11
WMCT	Memphis, Tenn.	4	5
WTMJ-TV	Milwaukee, Wis.	3	4
WNHC-TV	New Haven, Conn.	6	8
WTAR-TV	Norfolk, Va.	4	10
WKY-TV	Oklahoma City, Okla.	4	7
WDTV	Pittsburgh, Pa.	3	2
WJAR-TV	Providence, R.I.	11	10
WHAM-TV	Rochester, N.Y.	6	5
WRGB-TV	Schenectady, N.Y.	4	6
WSYR-TV	Syracuse, N.Y.	5	3
WDEL-TV	Wilmington, Del.	7	12

IRE People

E. U. Condon (M'42–SM'43), Director of the National Bureau of Standards, was recently elected honorary member of the Société Française de Physique, the French Physical Society. This is an honor which has been accorded only nine other scientists in the history of the society.

E. U. CONDON

Born in Alamogordo, N. M., in 1902, Dr. Condon received the B.A. degree in 1924, and the Ph.D. in 1926, both from the University of California. The recipient of one of the Rockefeller Foundation's National Research Council Fellowships, he continued his studies on the postdoctoral level in Germany from 1926 to 1927. Upon his return to America, he began his teaching career as a lecturer in physics at Columbia University, a career which included appointments to the University of Minnesota, to Princeton, and to the University of Pittsburgh.

In 1937 Dr. Condon left the classroom for the industrial laboratory, having been appointed associate director of the Westinghouse Research Laboratories in East Pittsburgh, where he was instrumental in working on the only large atom smasher to be operated by industry.

Beginning with the fall of 1940, Dr. Condon devoted all of his time to various phases of research for military purposes, both with Westinghouse and with many government projects. He became a consultant to the National Defense Research Committee when that group was first established by the late President Roosevelt. In this capacity he helped organize the Radiation Laboratory, for which he wrote the basic text on the subject of microwaves, introduced the radar program to the Westinghouse Corporation, and organized the microwave electronics research divisions of the company. In 1941 he became a member of the National Defense Research Committee's Rocket Program group.

In the autumn of 1941, he was appointed to a subcommittee of the NDRC, charged with studying the feasibility of producing atomic bombs. The investigations of this committee resulted in the recommendation that the United States proceed with a large-scale atomic bomb project immediately.

In 1945 he resigned from Westinghouse to become Director of the National Bureau of Standards by Presidential appointment. At the same time, by a second Presidential appointment, he became a member of the National Advisory Committee for Aeronautics. He also served, during 1945 and 1946, as scientific adviser to the Special Committee on Atomic Energy of the United States Senate which, under the chairmanship of Senator Brien McMahon, reported on the bill establishing the Atomic Energy Commission. In the summer of 1946, Dr. Condon was appointed to the President's Evaluation Board for "Operation Crossroads." He also belongs to the scientific advisory board of the Brookhaven National Laboratory of the AEC.

Dr. Condon has written more than 60 research papers, most of which are "required reading" in nuclear physics, as well as a large number of less rigorously technical articles. He is the co-author of two books, "Quantum Mechanics," published in 1929, and the "Theory of Atomic Spectrum," which appeared in 1935.

In 1944 he was elected to the National Academy of Sciences, and in 1947 to the American Academy of Arts and Sciences. He was president of the American Physical Society during 1946, and is a member of the governing board of the American Institute of Physics. Among the other scientific organizations to which he belongs are the American Institute of Electrical Engineers, the Institute of Aeronautical Sciences, and the American Association for the Advancement of Science. He is also a member of Sigma Xi, Phi Beta Kappa, and Sigma Pi Sigma. For his scientific achievements, particularly his contributions during the war, the Junior Chamber of Commerce of Pittsburgh elected him "Man of the Year" for 1945.

❖

Joseph C. Ferguson (A'41–SM'46), a former Chairman of the Fort Wayne IRE Subsection, died suddenly some time ago, it has been learned.

Born in Beaumont, Texas, in 1900, Mr. Ferguson received the B.S. degree from Louisiana State College in 1930, and did graduate work at Harvard University. He began his career as a junior radio engineer at Westinghouse Electric; shortly thereafter he joined RCA as a radio-design engineer. Four years later he became a senior radio engineer at the Farnsworth Television and Radio Corporation, where he was placed in charge of television-broadcasting equipment and design. In 1942 he was promoted to the position of chief engineer of the electronics apparatus division, in charge of the development and design of electronic equipment for the armed services.

Mr. Ferguson was a member of the RTMA Television Transmitter Committee, and served as chairman of their subcommittee on television studio facilities from 1944 to 1947.

W. B. Whalley (A'37), engineering specialist for Sylvania Electric Products Inc., Bayside, L. I., N. Y., was appointed adjunct professor in electrical engineering at the Polytechnic Institute of Brooklyn. The appointment was made in recognition of his professional standing and of his distinguished experience in the electrical and electronic fields.

W. B. WHALLEY

Professor Whalley received the B.A.Sc. degree from the University of Toronto in 1932; he served on the University's department of electrical engineering for the next four years, during which time he received the M.A. Sc. degree. Shortly thereafter he became associated with the Radio Valve Company of Toronto, and with the RCA Manufacturing Company. In 1940 he began his work on radar development for the Canadian government which included the technical organization of government-owned cathode-ray tube plant in Toronto for the production of radar screen tubes.

From 1943 to 1947 he was engaged in research at the RCA Laboratories in Princeton, N. J., where he continued work in the radar field and in the development of transmitting tubes and circuits for television. He also had done notable prewar work to improve television techniques, particularly with camera pickup tubes, television-receiver picture tubes, advanced circuits, and tubes for television systems.

His most recent work, at the Sylvania Physics Laboratories, has been in the field of simplified television receivers. He has consistently predicted a trend toward receiver simplification to reduce the number of required parts, increase reliability, reduce size of cabinets, and lower cost to the consumer. To prove his contentions, he has been demonstrating a 16-tube television receiver in direct comparison with commercial sets that require from 20 to 30 tubes for comparable results.

Professor Whalley is a member of Sigma Xi, the American Association of University Professors, and the American Physical Society.

❖

Edward G. Hall (A'43–M'48) has been appointed engineering representative for the sales and engineering service department of the Lenkurt Electric Company in San Carlos, Calif. Prior to joining Lenkurt, Mr. Hall had been for several years general transmission engineer for the Indiana Associated Telephone Corporation in Lafayette, Ind. His engineering experience includes also several years' affiliation with the Wisconsin Bell Telephone Company, and the Wilcox Electric Company of Kansas City, Mo.

W. A. Weiss (S'40–A'41–M'44) has been appointed manager of the new Sylvania radio receiving tube plant in Burlington, Iowa, it was announced recently. It is expected that he will take up his new duties early next autumn, at which time the construction of the plant will have been completed.

W. A. Weiss

Mr. Weiss, who has been manager of the Sylvania receiving tube plant in Emporium, Pa., since 1947, joined the company in 1940 as a student engineer in Emporium. In 1942 he was named supervisor of quality control and served in this capacity until 1947, when he became quality control manager for the entire Sylvania radio tube division.

Born in Alton, Ill., in 1913, Mr. Weiss was graduated from Pennsylvania State College in 1941 with the B.S. in electrical engineering. Before attending college, he had worked for a time with the Philco Corporation in Philadelphia. He is a member of the American Institute of Electrical Engineers and of The American Management Association.

❖

A. M. Zarem (S'42–A'46), director of Stanford Research Institute's Los Angeles division, and chairman of its applied physics research department, has been named one of "America's Ten Outstanding Young Men of 1950" on the United States Junior Chamber of Commerce list. His nomination was sponsored by the Alhambra, Calif., Junior Chamber, and was supported by numerous West Coast and Midwest educators, scientists, and newspaper and business executives.

The citation was made "for new photographic techniques making possible the taking of one million photos a second for the purposes of research in ballistics and similar fields."

Dr. Zarem, a native of Chicago, has won several awards based on photographic techniques he developed. His most spectacular achievement is the Zarem camera, so named by the Navy. With the speed of 100,000,000 frames per second, and an effective exposure time as fast as one-hundredth of a millionth of a second, it is the fastest known to science. This work opened a whole new field in physics research, termed "microtime," which has become vital in research on explosions, ballistics, shaped charges, and cavitation.

In 1948, Dr. Zarem was selected "The Outstanding Young Electrical Engineer in the United States" by Eta Kappa Nu.

In his capacity of chief administrative and scientific representative of Stanford Research Institute in Southern California, Dr. Zarem has concerned himself directly with the major problems facing the future development of the region, including mass transportation, water resources, and air pollution. He suggested, organized, and directed the First National Air Pollution Symposium sponsored by SRI and held in Pasadena in November, 1949.

In the past he has supervised groups of physicists and engineers whose work was directly used in the development of the atomic bomb. His department at Stanford Research Institute is presently engaged in studies dealing with fundamental explosion phenomena and with certain aspects of the effects of atomic weapons.

❖

Eginhard Dietze, a retired electrical engineer, died at Bradenton, Fla., according to a recent announcement. He had been in failing health for several months.

Mr. Dietze had had wide experience in many phases of telephone, and in the past few years had been associated with the development of the new and improved telephone set recently announced by Bell Telephone Laboratories.

Born on August 27, 1891, in Berlin, Mr. Dietze received his early education in Dresden, and moved to the United States in 1915. He enrolled at the University of Michigan where, in 1917, he received the B.S. degree in electrical engineering.

He then became associated with the American Telephone and Telegraph Company, where he engaged in telephone circuit transmission studies. Two years later he joined the company's department of development and research, which became a part of Bell Telephone Laboratories in 1934.

During World War II, Mr. Dietze was on leave of absence to the National Defense Research Committee, where he served as technical director of a group working on sonar.

A former member of the Institute, he served on the Annual Review Committee, the Papers Procurement Committee, and the Standards Committee; he was Chairman of the Electroacoustics Committee from 1946 to 1948.

❖

Correction

An error was made in the Fellow citation of A. G. Kandoian on page 572 of the May, 1951, issue of the Proceedings. The correct version of the citation is given below:

"For his important contributions to the development of antennas and navigation systems."

Emanuel R. Piore (A'38, M'42, SM'43, F'50) was appointed Deputy and Chief Scientist of the Office of Naval Research, it was announced recently. Up to now he has held the title of Deputy for Natural Sciences.

E. R. Piore

Dr. Piore was graduated from the University of Wisconsin in 1930, and in 1936 received the Ph.D. from that university, where he remained until 1938 as an assistant instructor in physics.

From 1938 to 1942 he was engineer in charge of the television laboratories of the Columbia Broadcasting System, and was instrumental in the development of color television. In 1944, from a position as senior physicist in the Navy's Bureau of Ships, he was called to active duty in the Navy Reserve as a lieutenant commander.

Dr. Piore joined the Office of Research and Inventions (which later became the Office of Naval Research) in 1946 as head of the electronics branch, and his service to ONR has been continuous except for the year 1948–1949, when he was a guest scientist in electronics at the research laboratory of the Massachusetts Institute of Technology.

A member of the Institute's Board of Directors, Dr. Piore belongs to Sigma Xi, the American Society of Naval Engineers, the Association for Computing Machinery, and is a Fellow of the American Physical Society.

❖

Eugene F. Grant (A'44), formerly chief of the computer branch, Air Force Cambridge Research Laboratories, has been appointed co-ordinator of electronics research and development at the W. L. Maxson Corporation in New York, N. Y.

Eugene F. Grant

Mr. Grant was born in Baker, Ore., and received the M.S. degree in electrical engineering from Oregon State College in 1942. He was then employed by the Westinghouse Research Laboratories, where he worked on the design of airborne radar. At the Sperry Gyroscope Company from 1945 to 1946, he was concerned with the design of Doppler radar systems. At the Air Force Cambridge Research Laboratories he was in charge of the design and construction of specialized digital computing machines.

Mr. Grant is a member of Phi Kappa Phi, Eta Kappa Nu, Sigma Pi Sigma, Pi Mu Epsilon, Sigma Tau, and Sigma Xi.

Report of the Secretary—1950

TABLE I.—TOTAL MEMBERSHIP DISTRIBUTION BY GRADES

Grade	As of Dec 31, 1950 Number	% of Total	As of Dec. 31, 1949 Number	% of Total	As of Dec. 31, 1948 Number	% of Total
Fellow	323	1.1	284	1.1	259	1.1
Senior Member	2,726	9.4	2,421	9.0	2,192	9.4
Member	3,803	13.1	3,559	13.3	3,334	14.2
Associate	14,590‡	50.3	12,309†	46.0	11,713*	50.0
Student	7,560	26.1	8,196	30.6	5,939	25.3
Totals	29,002		26,769		23,437	

* Includes 1,260 Voting Associates.
† Includes 1,159 Voting Associates.
‡ Includes 1,072 Voting Associates.

TABLE II.—FIVE-YEAR ANALYSIS OF U. S. FOREIGN MEMBERSHIP

	1950	1949	1948	1947	1946
TOTAL	29,002	26,769	23,437	21,037	18,154
U. S. & Possessions	26,708	24,434	21,048	18,723	15,898
Foreign (including Canada)	2,294	2,335	2,389	2,314	2,256
Per Cent Foreign	7.9	8.72	10.2	11.0	12.4

TO THE BOARD OF DIRECTORS,
THE INSTITUTE OF RADIO ENGINEERS

Gentlemen:

The tremendously expanding technical field which the Institute represents is reflected in the activities of the organization as portrayed in the following pages of this Report of the Secretary for the year 1950. Not only is the factor of growth shown by the continued increases in almost every item which could represent an index, but also by the gradually evolving pattern of dealing with diversified membership interests by technical fields which the Professional Group method represents. This latter approach seems to be progressing successfully in the direction of providing enhanced value of membership to the individual in the Institute.

Attention is particularly directed to the figures concerning membership, Technical Committees and their meetings, and the financial statement.

The 1950 National Convention scored a record registration of 17,689 persons as compared with 15,710 in 1949, held 36 technical sessions with 170 papers as compared with 33 and 170, respectively, in 1949, and had 253 exhibits as compared with 225 in 1949. Other meetings were: Fourth Annual Spring Technical Conference (Sponsored by Cincinnati Section), 1950 IRE Technical Conference (Sponsored by Dayton Section), Radio Fall Meeting (Syracuse, N. Y.), IRE/URSI Meeting, Commissions 1, 6, 7, 4 (Washington, D. C.), Conference on Improved Quality Electronic Components (Washington, D. C.), IRE West Coast Convention (Long Beach, Calif.), New England Radio Engineering Meeting (Boston, Mass.) IRE Conference on Electron Devices (University of Michigan, Ann Arbor, Mich.), National Electronics Conference (Chicago, Ill.), IRE/AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine (New York, N. Y.), National Conference of the IRE Professional Group on Vehicular Communications (Detroit, Mich.), IRE/AIEE Conference on Electron Tubes for Computers (Washington, D. C.), and UHF-Microwave Conference (Sponsored by Kansas City Section).

The problem of the growing deficit because the revenues from Student dues have, in recent years, been insufficient to cover the basic cost of providing students with the PROCEEDINGS has been overcome by increasing them to $5.00 per year, which amount will become effective in 1951. There has been no increase in Student dues since the establishment of the Student grade in 1932.

The year 1951 brings new problems due to changed world conditions with the inevitable shiftings of industrial and economic emphasis, and technical personnel, that always occur during a National emergency. The management of the Institute has geared itself to keep track of these trends better to organize for providing a maximum of service to its members as changes take place.

Respectfully submitted,

Haraden Pratt

HARADEN PRATT, *Secretary*

February 23, 1951

Membership

At the end of the year 1950, the membership of the Institute, including all grades, was 29,002, an increase of 2,233, or 8 per cent over the previous year. The 2,233-member increase in 1950 was less than 2,400 and 3,332, the increases for 1948 and 1949, respectively. The percentage increase was 11 per cent in 1948, and 14 per cent in 1949. The membership trend from 1912 to date is shown graphically in Fig. 2.

The distribution of members in the various grades for the years 1948, 1949, and 1950 is shown in the accompanying plot, Fig. 3. Actual figures for 1948, 1949, and 1950 are shown in Table I. Of the 13,518 nonvoting Associates, 3,919 have been in that grade for more than five years. 1,786 Students transferred to the Associate grade in 1950, compared to 951 in 1949. The membership ratio of Associates to higher grades was 6 to 1 in 1944, 4 to 1 in 1945, less than 3 to 1 in 1946, and about 2 to 1 in 1947, 1948, 1949, and 1950, a very satisfactory trend. Monetary exchange difficulties and devaluation of the pound have contributed to the reduced increase in foreign membership during the past three years, shown in Table II.

It is with deep regret that this office records the death of the following members of the Institute during the year 1950:

Fellows

Edwin H. Colpitts (A'14–F 26)
Lawrence C. F. Horle (A'14–M'23–F'25)
Karl G. Jansky (A'28–M'34–SM'43–F'47)
Fred Kolster (A'12–M'13–F'16)
William D. Loughlin (A'29–M'29–F'35)
Roger M. Wise (A'26–M'30–F'37)

Senior Members

John E. Allen (A'23–SM'45)
Francis E. Bash (SM'45)
John F. Bates (A'36–SM'45)
Norman R. Beers (SM'50)
Loyd A. Briggs (M'29–SM'43)
Roy W. Chesnut (M'42–SM'43)
Frederick A. Cobb (A'26–M'28–SM'43)
B. Ray Cummings (A'18–M'20–SM'43)
Hammond H. Hollis (A'28–M'38–SM'43)
James H. Ludwig (A'37–VA'39–SM'48)
George O. Milne (M'43–SM'43)
John V. Murphy (SM'47)
Nicholas M. Oboukhoff (M'23–SM 43)
Otis W. Pike (A'26–M'29–SM'43)
Earl H. Schoenfeld (A'35–SM'45)

Members

M. S. Alexander (A'43–M'44)
A. B. Ellis (A'35–M'45)
Margaret S. Heagy (M'46)
Leonard D. Scalise (M'49)
Paul E. Waples (M'46)
Robert C. Woodhead (A'37–M'46)

Voting Associates

George W. Bain (A'31–VA'39)
John C. Bain (A'36–VA'39)
Robert J. Hare (A'32–VA'39)
Aristote Mavrogenis (A'38–VA'39)
Albert J. Muchow (A'35–VA'39)
Walter E. Poor (A'29–VA'39)

Associates

W. B. R. Agnew (A'44)
Morris D. Douglas (A'48)
Gaylord E. Durham (A'44)
Elmer W. Everson (A'47)
Loren B. Harrell (A'43)
Werner C. Kruger (A'48)
Richard L. Laudenslager (A'49)
Douglas S. Mackiernan (A'41)
Arthur G. Mohaupt (A'44)
Charles E. Morris (S'45–A'48)
Alexander Nadosy (A'39)
Alfred P. O'Connor (A'42)

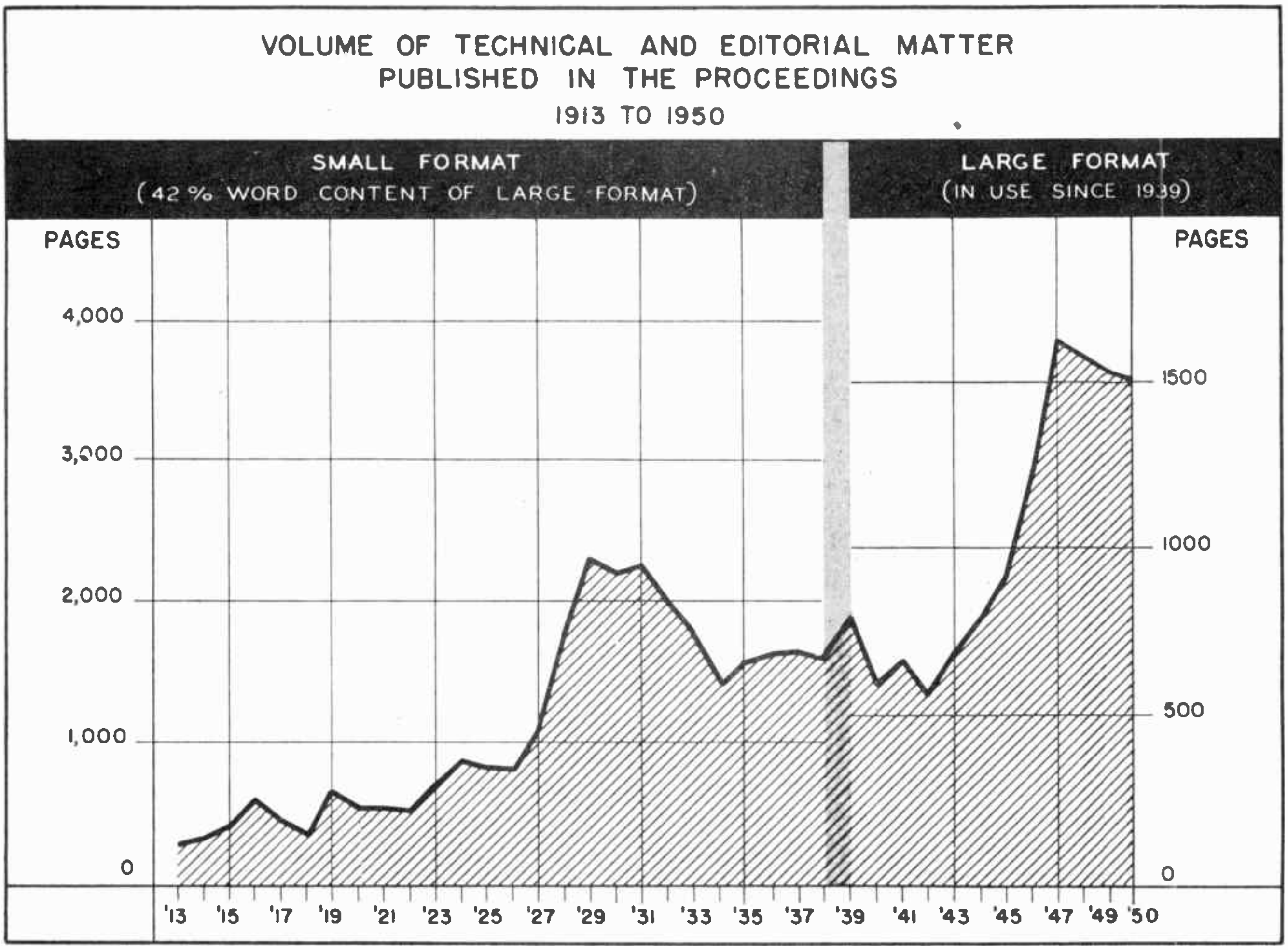

VOLUME OF TECHNICAL AND EDITORIAL MATTER
PUBLISHED IN THE PROCEEDINGS
1913 TO 1950
SMALL FORMAT
(42 % WORD CONTENT OF LARGE FORMAT)
LARGE FORMAT
(IN USE SINCE 1939)
PAGES
4,000
3,000
2,000
1,000
0
PAGES
1500
1000
500
0
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49 '50

Fig. 1

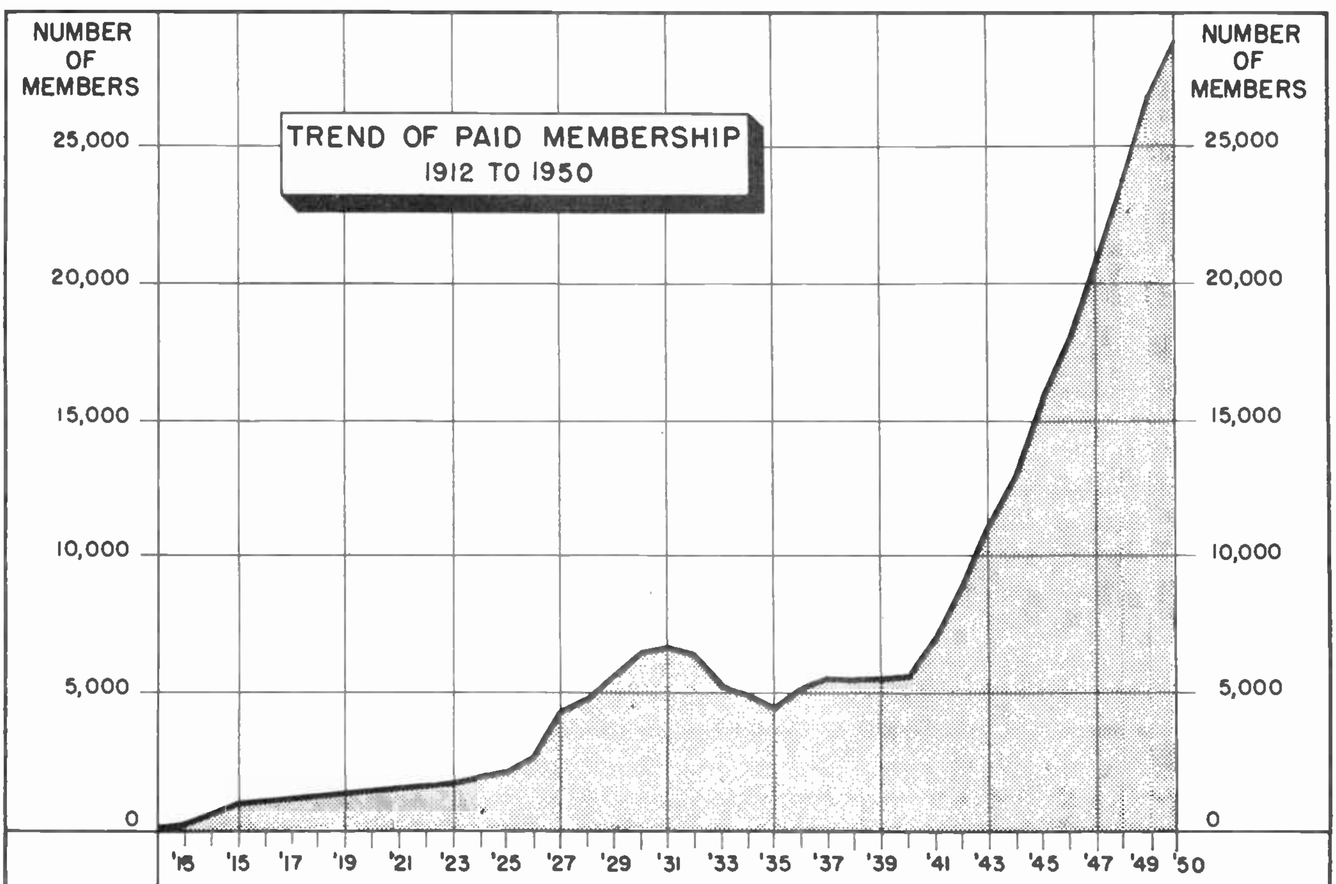

NUMBER OF MEMBERS
TREND OF PAID MEMBERSHIP
1912 TO 1950
25,000
20,000
15,000
10,000
5,000
0
NUMBER OF MEMBERS
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49 '50

Fig. 2

Henry F. Parks (A'42)
S. Sampat (A'47)
Robert H. Schott (S'49–A'50)
William F. Seeman (A'35)
C. S. Sesha-iyer (A'49)
George S. Stancu (S'46–A'47)
Max D. Weinberg (A'48)

Students

William M. Cory (S'49)
Robert W. Gardner (S'49)
Donald B. Garrett (S'47)
Henry P. Kun (S'49)
Donald W. Maas (S'49)
Nicholas Sidjak (S'49)

Fiscal

A condensed summary of income and expenses for 1950 is shown in Table III, and a balance sheet for 1950 is shown in Table IV. Income and expenses for each year since 1914 are shown graphically in Fig. 4.

TABLE III.—SUMMARY OF INCOME AND EXPENSES,

	1950	
Income		
Advertising	$211,021.40	
Members Dues and Conventions	434,596.55	
Subscriptions	50,575.45	
Sales Items, Binders, Emblems, etc.	15,811.38	
Investments Income	11,415.65	
Miscellaneous Income	282.35	
TOTAL INCOME		$ 723,702.78
Expenses		
PROCEEDINGS Editorial Pages	$170,806.71	
Advertising Pages	116,565.09	
Yearbook	49,030.14	
Section and Student Branch Rebates	28,916.62	
Sales Items	9,090.08	
Miscellaneous Printing	2,387.15	
General Operations	174,707.08	
Convention Cost	108,975.04	
TOTAL EXPENSES		$ 660,477.91
Surplus		$ 63,224.87
Reserve for Depreciation		7,106.77
NET SURPLUS		$ 56,118.10

TABLE IV.—BALANCE SHEET—DECEMBER 31, 1950

Assets		
Cash and Accounts Receivable	$234,505.04	
Inventory	13,267.59	
Total Current Assets		$ 247,772.63
Investments at Cost	488,059.14	
Building and Land at Cost	356,687.30	
Furniture and Fixtures at cost	76,014.32	
Other Assets	14,602.49	
TOTAL		935,363.25
TOTAL ASSETS		$1,183,135.88
Liabilities and Surplus		
Accounts Payable	$ 8,891.00	
Federal Taxes on Emblems, etc.	85.58	
Total Current Liabilities		$ 8,976.58
*Deferred Liabilities		260,103.96
Total Liabilities		$ 269,080.54
Reserve for Depreciation		26,878.80
Surplus—Donated	$597,281.03	
Surplus—Earned	289,895.51	
TOTAL SURPLUS		887,176.54
TOTAL LIABILITIES AND SURPLUS		$1,183,135.88

* 1951 Items, PROCEEDINGS for members and subscribers, Advertising, and Convention Service.

Editorial Department

During the year 1950 there were published in the PROCEEDINGS OF THE I.R.E. a total of 2,532 pages, including covers. Of these, 1,516 were editorial pages and 1,016 advertising pages.

The total of 2,532 pages published during the year compares with 2,404 in 1949, and 2,452 in 1948. The number of editorial pages (1,516) compares with 1,532 in 1949, and 1,592 in 1948. Advertising pages numbered 872 in 1949 and 860 in 1948. The number of editorial pages published each year since 1913 is shown in Fig. 1.

Technical papers totaling 186 were published in 1950, as against 183 in 1949. In addition, 6 IRE Standards were published in 1950 as against 4 in 1949. Authorship of these papers was by 252 individuals of whom 182, or 73 per cent, were members of the Institute. In 1949, 183 of the 252 authors, or 73 per cent, were members.

The volume of papers submitted for publication continues at a high rate. During 1949, 287 papers totaling an estimated 1,894 PROCEEDINGS pages were submitted, or an average of 24 papers and 158 pages per month. During 1950, 271 papers of 1,547.2 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 1950 consisted of 141 papers totaling 777 PROCEEDINGS pages, of which 81 papers or 436 pages had been accepted for publication, the remainder being under review. This represented a noticeable increase over the record low backlog total at the end of 1949 of 103 papers or 585 pages, but was substantially lower than the 1948 total of 141 papers or 981 pages.

In a move to consolidate the technical material in the PROCEEDINGS, the Waves and Electrons Section was discontinued with the January, 1950, issue. This did not affect the contents of the journal, but offered a more closely integrated arrangement of technical and editorial material.

An important innovation in papers procurement activities is being carried out by the Tutorial Papers Subcommittee of the IRE Committee on Education. The Subcommittee, under the chairmanship of Ernst Weber, is engaged in procuring, and recommending for publication, a series of tutorial papers on a wide variety of topics of both present and historical interest. These papers, which will appear at frequent intervals during 1951 and 1952, are to be educational in nature, of exceptional clarity, and prepared in each case by an authority in the corresponding field. It is believed that members in all grades will derive considerable benefit from this series.

During the year, the Editorial Department co-operated closely with IRE Professional Groups on the publication of technical papers and news items of interest to Group members in order to give this important Institute activity full expression in the pages of the PROCEEDINGS. Some of the steps taken by the Editorial Department are as follows: the procurement of papers of particular interest to Groups; the referral of submitted papers to Professional Groups for publication consideration; prompt publication of papers recommended by Groups; publication of 100-word abstracts of papers delivered at Group-sponsored conferences; publication of monthly columns on Professional Group News in the Institute News and Notes Section of the PROCEEDINGS; and bimonthly publication of a list of Professional Groups, together with the names and addresses of their chairmen. In addition, the Advertising Department frequently devoted the first page to publicizing Group-sponsored meetings, as well as other IRE conferences.

In recognition of the important work of the Editorial Administrative Committee in determining the acceptability of technical material for publication, the Board of Directors raised its status to that of a Standing Committee of the Institute and changed its name to Administrative Committee of the Board of Editors.

The 1950 IRE Directory (formerly known as the IRE Yearbook) was published on a considerably shortened production schedule, despite the substantial increase in size over prior years. The closing date for membership listings was changed from December 31 of the previous year to May 15 of the current year, and the mailing date was shifted from July to September 15.

The Editorial Department is directed by Editor Alfred N. Goldsmith in matters of editorial policy, content, and format, and by Executive Secretary George W. Bailey in matters of finance and administration. Both function through Technical Editor E. K. Gannett and an effective staff. It has been greatly assisted by the counsel and co-operation unstintingly given by the members of the Board of Editors, Papers Review Committee, and the Administrative Committee of the Board of Editors.

Technical Activities

Technical Committees. During 1950, 23 Technical Committees, with their subcommittees, task groups and planning committees for conferences or symposia, held 146 meetings at the Institute Headquarters, five meetings at the Hotel Commodore during the National Convention, and ten meetings at other points, a grand total of 161 meetings.

A report on the activities of the IRE Technical Committees and their Subcommittees is compiled in the office of the Technical Secretary and published each month in the PROCEEDINGS.

In addition to IRE committee meetings, the National Television System Committee held 40 meetings at Headquarters. Several of the NTSC Panels requested and received secretarial assistance from the Office of the Technical Secretary. Eleven meetings of committees jointly sponsored by the IRE and other outside organizations were also held at Headquarters, many requiring secretarial services. During the year 21 conferences, symposia, technical sessions, and national meetings were sponsored by IRE Professional Groups, by Technical Committees or co-sponsored by the IRE and other organizations. All of these functions were well attended and met with enthusiastic audience response. The major portion of the work involved in the preparation of

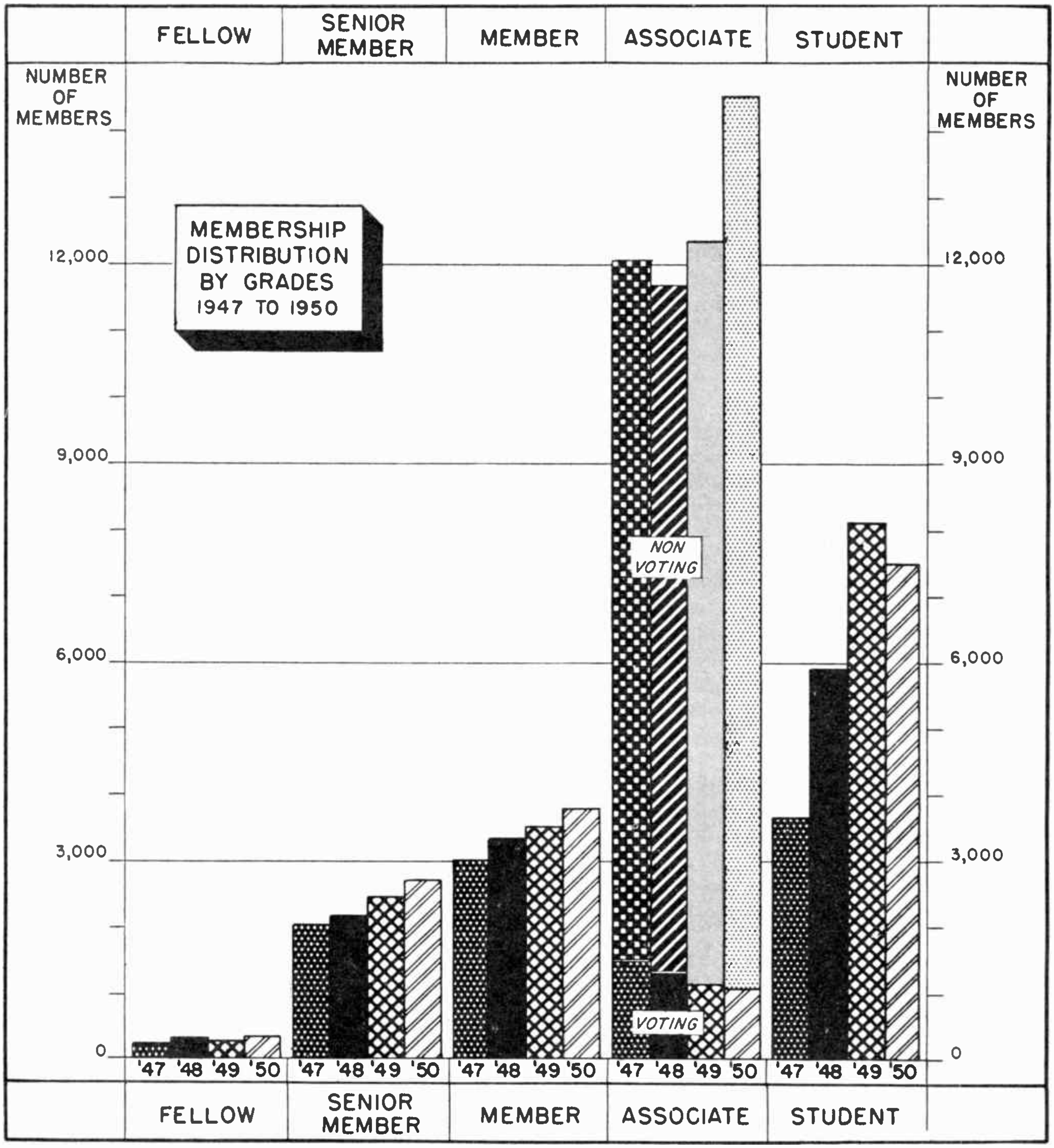


Fig. 3

these joint conferences and symposia was done by the Office of the Technical Secretary.

The policy of publishing Standards in the PROCEEDINGS was adopted late in 1949. Issues of the 1950 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. Six standards were published during 1950.

The following standards were prepared by Technical Committees and appeared in various issues of the PROCEEDINGS during 1950:

Standards on Electron Tubes: Definitions of Terms, 1950.

Standards on Electron Tubes: Methods of Testing, 1950.

Standards on Designations for Electrical Electronic and Mechanical Parts and their Symbols, 1949 (appeared Feb. '50).

Standards of Television: Methods of Measurement of Television Signal Levels, Resolution, and Timing of Video Switching Systems, 1950.

Standards on Television: Methods of Measurement of Time of Rise, Pulse Width, and Pulse Timing of Video Pulses in Television, 1950.

Standards on Wave Propagation: Definitions of Terms, 1950.

The following Standards, prepared by IRE Technical Committees, have been approved by the Standards Committee and will be published in forthcoming issues of the PROCEEDINGS.

Standards on Circuits: Definitions of Terms in Network Topology, 1950.

Standards of Electroacoustics: Definitions of Terms, 1950.

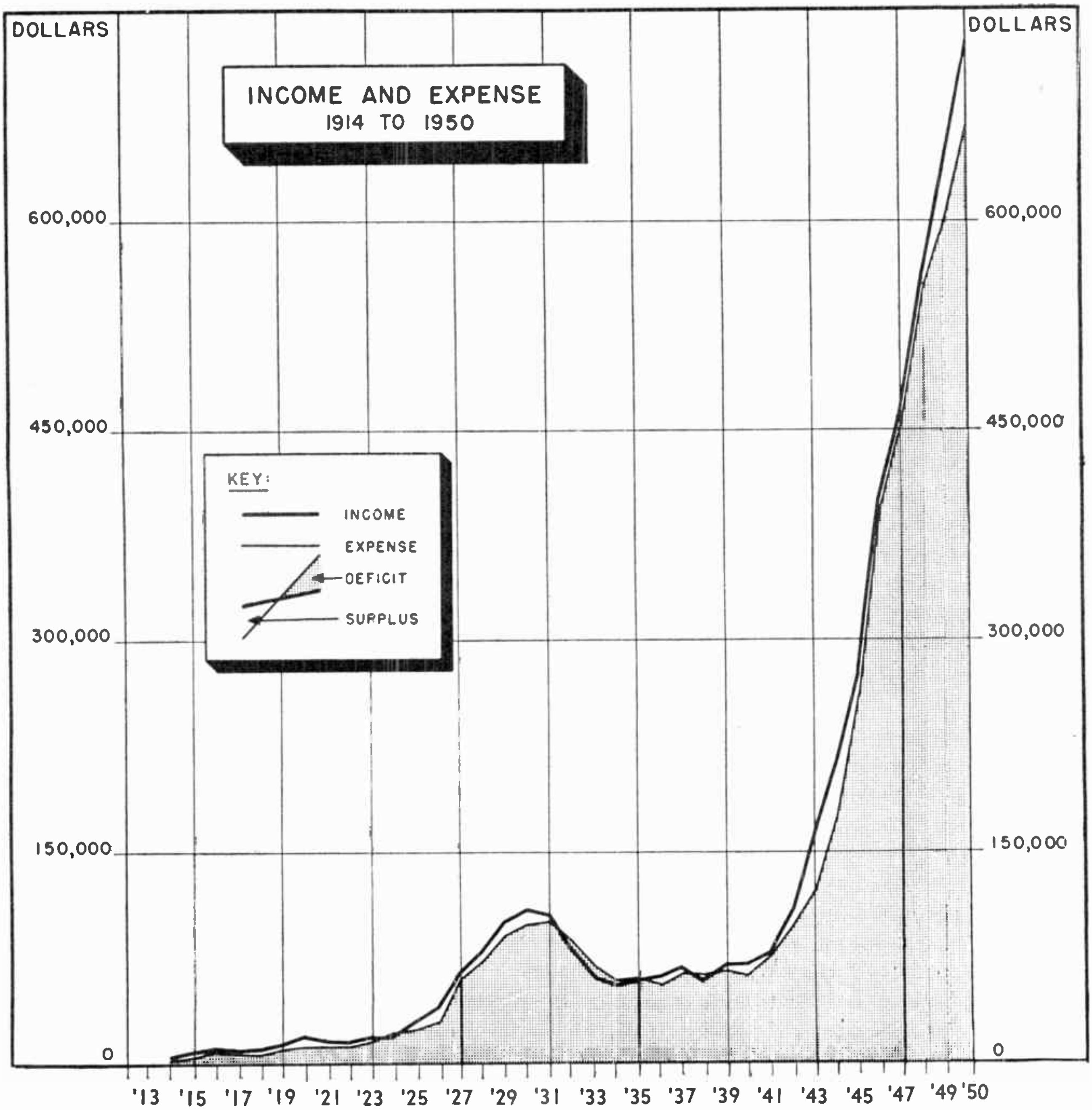


Fig. 4

Standards on Electronic Computers: Definitions of Terms, 1950.
Standards of Abbreviations of Radio-Electronic Terms, 1950.
Standards on Television: Methods of Measurement of Electronically Regulated Power Supplies, 1950.

The Annual Review Committee prepared its survey, "Radio Progress During 1950," which appeared in the April, PROCEEDINGS 1951, issue.

The Master Index of Terms, prepared in 1949 under the supervision of the Definitions Co-ordinating Subcommittee, was made available to all IRE Committees and, upon request, to other societies. It is presently being revised for the IRE Definitions Co-ordinating Subcommittee by the Office of the Technical Secretary.

The Institute approved the formation of the IRE Co-ordination Committee with International Technical Organizations. This Committee will provide liaison with CCIR and other international standardizing bodies with which the IRE is concerned. Axel G. Jensen has been appointed Chairman. This Committee will report directly to the IRE Executive Committee; it will work through the U. S. Department of State with the CCIR; and with other international bodies. When the CCIR or any other group requests IRE assistance, this new Committee will establish an ad hoc group to study the problem and provide a recommended solution. When the Standards Committee feels that a matter should be brought to the attention of an international standardizing body, it will in turn be referred to this Committee.

The Joint IRS Committee was formed by IRE, RTMA, and SMPTE to avoid duplication of effort in the preparation of standards in the fields associated with television. Axel G. Jensen, Millard W. Baldwin, Jr., and R. L. Garman are the IRE representatives on the Committee.

A Joint IRE/AIEE Committee on Noise Definitions has been organized for the specific task of critically reviewing all existing noise definitions and adding such new definitions as may be required.

The Institute has also approved the IRE participation in a newly organized committee which is known as the Joint IRE/AIEE Committee on High-Frequency Measurements. Ernst Weber, of the Polytechnic Institute of Brooklyn, is Chairman. The personnel of this Committee is composed of an equal number of IRE and AIEE members. E. I. Green is the leader of the IRE group. As a Joint Committee, it is autonomous and can engage in any activities which are of interest to Technical Committees of both sponsoring societies and to the IRE Professional Groups. It is the duty of

IRE members of this Committee to keep in contact with the IRE and AIEE activities. This Joint Committee has already sponsored two conferences.

The IRE maintains liaison with ASA Sectional Committees. During the year several IRE members were appointed to represent the Institute on various ASA Committees. The Office of the Technical Secretary is alert to the importance of maintaining adequate liaison with these committees.

Professional Group System. There are ten Professional Groups presently in existence in the following fields: Antennas and Propagation, Audio, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Instrumentation, Nuclear Science, Quality Control, Radio Telemetry and Remote Control, and Vehicular Communications. The total membership is in excess of 8,500, and is increasing steadily. It is anticipated that a petition to form the 11th IRE Professional Group on Airborne Electronics will be approved by the Executive Committee. An example of the growth of the Professional Group System is evidenced by the fact that at the 1950 National Convention, five groups sponsored Symposia, while all of the Groups sponsored symposia at the 1951 National Convention. Each group was represented 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 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. All Groups are interested in securing papers for publication in the PROCEEDINGS. The Professional Group Manual was revised in June 1950.

The Joint Technical Advisory Committee. There were 11 meetings during the year. Members of the JTAC attended demonstrations given by CBS and the Hazeltine Electronics Corporation of various matters within their fields of interest. Volume V, Adjacent-Channel Interference In Monochrome Television, was published and submitted with a letter to the Federal Communications Commission; Volume VI, (correspondence and minutes) was authorized and work commenced; the publication of a supplement to Volume IV, Comments On The Proposed Allocation of Television Broadcast Services, was prepared. A 'Supplemental Statement' prepared for the FCC in connection with Volume IV was formally presented in evidence before the Commission. The membership was appointed for the year July 1, 1950 to June 30, 1951, and J. V. L. Hogan was unanimously elected Chairman. Melville Eastham and E. K. Jett were unable to continue their duties with JTAC, and were succeeded by I. J. Kaar and T. T. Goldsmith. Several reports from technical associations and industry have been circulated for comment, and reports on current topics exchanged with other technical bodies, such as the BBC.

Section Activities

We were glad to welcome four new Sections into the Institute during the past year. They are as follows:

Evansville-Owensboro	(March) 1950
Hawaii	(February) 1950
Miami	(February) 1950
Vancouver	(September) 1950

The total number of Sections is now 57. There has been a substantial increase in membership of these Sections with a few exceptions. It should be borne in mind that most Sections with noticeable decreases in membership released substantial numbers of members to new Sections.

The Subsections of Sections now total 12, the following being formed in 1950:

Binghamton (Syracuse Section)
May, 1950
Mid-Hudson (New York Section)
November, 1950

Student Branches

The Institute's program with respect to Student Branches continued to flourish in 1950. The number of Student Branches formed during 1950 was 17, 14 of which operate as joint IRE/AIEE Branches. The total number of Student Branches is now 105, 60 of which operate as joint IRE/AIEE Branches. This increase of Student interest was accompanied by a large increase in Student members.

Following is a list of the Student Branches formed during 1950:

University of Akron, University of British Columbia, Polytechnic Institute of Brooklyn, Bucknell University, University of Connecticut, Cooper Union, University of Denver, Drexel Institute of Technology, Louisiana State University, University of Miami, Michigan College of Mining and Technology, New Mexico College of Agriculture and Mechanic Arts, Oklahoma Agricultural and Mechanical College, Rensselaer Polytechnic Institute, Stevens Institute of Technology, and University of Toronto.

Books

Radio Laboratory Handbook, Fifth Edition by M. G. Scroggie

Published (1950) by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London S.E. 1, England. 409 pages+5-page index+14-page appendix. 215 figures. 7×4¾. 15s.

This is one of a series of radio books issued by the magazine *Wireless World*, the first of which appeared in 1938. The present fifth edition of this work, published in 1950, has been thoroughly brought up to date.

It is not a textbook on radio theory, but, as its name indicates, a reference book of material useful for workers in radio, designed to be useful both to the amateur enthusiast and to the professional radio engineer.

The material treated is grouped under the main headings of sources of power and signals, indicators, standards, laboratory equipment in general, measurements of resistance, inductance and capacitance, and tests of amplifiers and receivers. A new chapter is added on very-high-frequency work, with special reference to its relation to lower-frequency procedures. A reference chapter compiling useful fundamental formulas and numerical constants is a convenient feature.

With such a wide coverage of subject matter, exhaustive treatment of any subject is, of course, out of the question. A practical working knowledge of fundamental circuit theory and of elementary vacuum-tube functions on the part of the reader is assumed. Working diagrams of special circuits and arrangements are clearly given and discussed, but for further study by those interested, reference is given to the original published articles.

Space is given to sensible practical advice on the organization of an experimental laboratory, the choice of necessary apparatus, and instructions for the building of certain devices for measurement.

The style of presentation is, throughout, easy and pleasing, the choice of type size is good, and, in spite of the fact that the volume is of pocket size, the diagrams and illustrations do not suffer.

The amateur should find this a most helpful guide, and the professional engineer a handy and comprehensive reference book.

FREDERICK W. GROVER
Union College
Schenectady, N. Y.

TV Master Antenna Systems, by Ira Kamen and Richard H. Dorf.

Published (1951) by John F. Rider Publishers, Inc., 480 Canal St., New York 13, N. Y. 352 pages+4-page index+vii pages. 234 figures. 5½×8½. $5.00.

This book was designed to be a working manual for many kinds of organizations and people concerned with master-television systems. The authors have succeeded to a remarkable degree in satisfying technical completeness and yet insuring the layman's comprehension by treating problems and their solutions. Although primarily presented for those whose interests are in manufacture, sales, installation, maintenance, and use, this is also a tutorial work of interest to radio engineers.

The reader is introduced to the subject by a discussion of basic television antenna systems. Descriptions of four nonamplified and ten amplified master-antenna systems follow. All but two of the systems discussed are replete with circuit diagrams giving component values. The two exceptions, Lynmar and Multitenna, give the dubious and provocative excuse that for "patent reasons" the information is not available. The authors may be commended for their adequate presentation in the absence of co-operation from these manufacturers.

The table of contents and the index appear to be ample for a work of this kind. The information is clearly presented in an easily readable manner. The book also covers video distribution systems, and provides an appendix with landlord agreement, survey report, and tenant letter forms.

ALOIS W. GRAF
135 S. LaSalle St.
Chicago 3, Ill.

Radiation Monitoring in Atomic Defense by Dwight E. Gray and John H. Martens

Published (1951) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. 111 pages+2-page index+19-page appendix+iv pages. 20 figures. 5½×8. $2.00.

Were the title of this little book to read simply "Radiation Monitoring," the reviewer would commend both the authors and the publishers for producing a manual useful to the uninitiated who are confronted with the problem of monitoring the radioactive hazard of a laboratory, an X-ray installation, or a similar radiation source. But the title contains the additional words "In Atomic Defense." Herein lies the fault—not in the construction of the book, but in its objective.

By aiming their book specifically at civil defense, the authors tend to dignify the status of the radiation hazard attending an atomic bomb explosion. In this respect they are not to blame, for they are merely "following the party line" established by the many medical men who have been responsible for civil defense planning. By and large, defense planners have gotten the radioactive menace of the atomic bomb completely out of focus, until it has assumed proportions overshadowing the more important blast and fire hazards of the new weapon. Many local communities have set up civil defense plans patterned to a radiological monitoring procedure which is not only disproportionate to the hazard, but may actually be dangerous to the overall success of civil defense. Apparently the thought has been that if you have a Geiger counter, you have the panacea for atomic defense.

Atomic bombs are normally exploded above the surface of the earth in order to achieve maximum effectiveness. American experience with 12 A-bombs so exploded shows conclusively that with an above-surface burst the residual radiation hazard is very small, and negligible with a high air burst. The very violence of the atomic explosion sweeps all the radioactive fragments upward and sucks them into the stratosphere, where they disperse harmlessly.

If one thinks of the book purely as a manual for radiation monitoring without reference to atomic disaster, then one is quite favorably impressed that the writers have done a creditable job. Treating the subject in two parts, they first lay out the fundamentals of atomic energy and nuclear radiation. Then they point their remarks at individual instruments (cookbook fashion as they, themselves, admit), and describe how the basic types of Geiger counter, ion chambers, and similar instruments should be used. As a matter of style, the reviewer wishes that they had been consistent in using the capitalized form of Geiger counter; they use both, but mainly the "geiger" form appears in the text.

Factually, the book appears to be quite accurate with only occasional errata cropping up. For example, the authors state: "radiation remains a serious cause of injury at distances up to 1½ to 2 miles from ground zero." This is a misinterpretation of data published by the Atomic Energy Commission, and the figures should read "up to 1 mile from ground zero."

Both the authors are on the staff of the Atomic Energy Commission. This very fact will add weight to the "official nature" of the publication. It also tends to explain the nature of the book, for there are a number of AEC officials who have been responsible for the delusion that the Geiger counter is the "open sesame" to the realms of civil defense.

R. E. LAPP
Nuclear Science Service
Committee on Atomic Energy
Washington, D. C.

Industrial Instrumentation by Donald P. Eckman

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 353 pages+9-page index+31-page appendix+vi pages. 247 figures. 5½×8½. $5.00.

This up-to-date textbook on industrial instrumentation is intended for the undergraduate student in engineering. The author assumes that the student has an acquaintance with the elements of calculus, mechanics, and thermodynamics, as well as a general course in physics.

As an introduction to the body of the text, the initial chapter covers the qualities of measurements, defines terms, and discusses the response of first- and second-order physical systems. Approximately one third of the text is devoted to temperature measurements (4 chapters). Then a chapter each is devoted to composition analysis, mechanical measurements, pressure and vacuum measurements, head and liquid level, and flowmetering. The final chapter is devoted to process instrumentation.

In general, the various types of measuring instruments are adequately discussed. Since the description is supplemented by a discussion of the basis for the measurement, some of the limitations, as well as the range of the usually available industrial instruments, are included. For many types of instruments, e.g., potentiometers and optical pyrometers, the various forms made by different instrument manufacturers are discussed.

However, because of the large number of different types of instruments discussed, there are only brief comments on many forms, such as emission spectroscopy and mass spectroscopy.

Among some of the errors noted is the implication that a second-order system composed only of resistance and condensors will oscillate when a unit step function impulse is applied. Another is the statement that the volume change of a gas due to change in temperature is relatively small. Actually, the quantity involved is the pressure developed by an enclosed fluid.

However, these errors should not detract from the usefulness of the book as a text for the undergraduate. The material is easily readable. Numerous good illustrations and the many examples which have been worked out should give the student a clear picture of the field of industrial instrumentation. The sets of problems at the end of each chapter and the many tables in the appendix should also prove helpful. In general, very little mathematics is needed, since most of the equations are stated rather than derived, and the application of these equations, which is well illustrated, is not difficult.

Thus the book fulfills the undergraduate engineering student's need for a textbook in industrial instrumentation. It may also prove useful to some engineers who are in daily contact with instruments, but who need only a precursory knowledge of the fundamental principles of their operation.

HERMAN FARBER
Microwave Research Institute
Polytechnic Institute of Brooklyn
Brooklyn, N. Y.

The Principles of Cloud-Chamber Techniques by J. G. Wilson

Published (1951) by the Cambridge University Press, 51 Madison Ave., New York 10, N. Y. 127 pages+2-page index+2-page references+viii pages. 33 figures. 5½×8½. $2.75.

One of the Cambridge Monographs on Physics, this book is addressed to nuclear physicists who use or plan to use cloud chambers, or who may find an account of the cloud chamber as an instrument of precise measurement helpful.

The purpose of the cloud chamber is to study the motion of ionizing particles from photographs of the drops condensed on ions formed along particle trajectories. The early chapters of the monograph deal with the condensation process, ionization, and the general problems of the establishment of supersaturation and its persistence, on which the technique depends. Arrangements for photographing are then discussed and methods of measurement are described in detail, with particular reference to cosmic-ray investigations. A final chapter deals with the interpretation of photographs, a topic of increasing importance. Detailed specifications for the construction of cloud chambers are not included, but the advantages and disadvantages of various constructional materials are assessed in the light of experience.

C. W. CARNAHAN
Sandia Corp.
Sandia Base
Albuquerque, N. M.

URSI—IRE Spring Meeting*

WASHINGTON, D.C.,—APRIL 16–18, 1951

SUMMARIES OF TECHNICAL PAPERS

1. RECENT MATHEMATICAL DEVELOPMENTS IN THE THEORY OF TROPOSPHERIC PROPAGATION

Bernard Friedman
(New York University, New York, N. Y.)

The theory of tropospheric propagation fits naturally into the general theory of propagation in a stratified medium. It proves to be unnecessary to assume a flat earth and a modified index of refraction; the theory can be developed using a spherical earth and the ordinary index of refraction. The usual WKB treatment of the eigenvalue equation is extended and it is shown that propagation in an atmosphere without ducts can be reduced to propagation in a constant atmosphere by a suitable change of scale. The solution is approximated for an atmosphere with ducts, by means of Hankel functions of order one-fourth instead of the customary Hankel function of order one-third.

2. MARTYN'S THEORY OF MAGNETIC STORMS AND AURORAS

H. G. Booker
(Cornell University, Ithaca, N. Y.)

The theory supposes a neutral stream of charged particles from the sun as the origin of the ring current in magnetic storms. From this ring current, particles enter the earth's atmosphere along the auroral zone acquiring some ten times the energy of the primary particles. Visual auroras are caused by these secondary particles and drift of ionization from the auroral zones produce the current systems and the diurnal storm variation of the earth's magnetic field. Certain features of ionospheric observations are explained in terms of this current system.

3. IMPEDANCE MEASUREMENTS IN THE 50 TO 1,000 MEGACYCLE RANGE

F. J. Gaffney
(Polytechnic Research & Development Co., Inc., Brooklyn, N. Y.)

Several attacks have been made on the difficult problem of designing apparatus for impedance measurements in the 50–1,000 mc frequency range. These methods use an adaptation of slotted line techniques or the properties of tee junctions. The uhf equivalent of a low-frequency bridge is also described employing a novel variable-resistance element. Accuracies of the order of 5 per cent are obtained and possibilities of development for even higher precision are pointed out.

* Co-sponsored by the IRE Professional Group on Antennas and Propagation and the USA National Committee of URSI.

4. AIR-TO-AIR TROPOSPHERIC RADIO PROPAGATION

G. B. Fanning and F. P. Miller
(Wright-Patterson Air Force Base, Dayton, Ohio)

Recent flight tests performed over the Midwest by Aircraft Radiation Laboratory, Engineering Division, AMC, Wright-Patterson Air Force Base are described. Significant features of instrumentation and extensive experimental data for 250 mc and 3,000 mc are presented. The data indicate that many rapid and deep fades in signal level occur at ranges far short of the total horizon range. It is further concluded that significant observed field strength variations do not require severe associated meteorological anomalies for their explanation.

5. THE EFFECT ON PROPAGATION OF AN ELEVATED ATMOSPHERIC LAYER OF NONSTANDARD RETRACTIVE INDEX

L. H. Doherty
(Cornell University, Ithaca, N. Y.)

Elevated layers with nonstandard gradients of refractive index may radically affect the field of a transmitter above the layer. Serious difficulty in communication between aircraft flying above the layer may result from slight deviations from the standard atmosphere which frequently occur. Using a trilinear refractive index model of the atmosphere, it is shown that a decrease of field strength above the layer results within a certain range of distances. The interference pattern in a region of interference rays following this "radio hole" is shown. The effect of some nonlinear refractive index models is indicated. Comparisons are made with experimental results.

6. EXPERIMENTAL DISCRIMINATION OF THE FACTORS IN VHF RADIO WAVE PROPAGATION

A. W. Straiton and C. W. Tolbert
(University of Texas, Austin, Tex.)

Results are reported for recent experiments in determining the relative contribution of the various possible propagation mechanisms affecting vhf radio field strengths at distances well beyond the optical horizon. Height-gain curves and their interpretation are given for distances up to 175 miles. Results of measurements of cross polarization are given providing additional information as to the relative contribution of the various propagation factors. Means of obtaining the scale of turbulence as used in the Booker-Gordon scattering theory are indicated and several measured values of this factor are given.

7. SOME CHARACTERISTICS OF 92.9-MC PROPAGATION OBSERVED AT A DISTANCE OF 127 MILES

R. J. Wagner, Jr.
(RCA Laboratories Division, Princeton, N. J.)

Vhf field strength observations of a distance of 127 miles from Boston to Riverhead are reported. For periods of weakest field strength the range of vertical angles of arrival extended up to about 26° while for stronger fields it extended only to 7° or less. During periods of rapid fading, a pronounced height diversity effect was found with field strength values ranging from 44 to 74 db below free space. When rapid fading was absent this range was 14 to 42 db below free space.

8. VALIDITY AND LIMITATIONS OF THE VAN VLECK-WEISSKOPF EQUATION FOR ATMOSPHERIC MICROWAVE ABSORPTION

T. F. Rogers
(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

A detailed study is made of the Van Vleck-Weisskopf absorption equation for a collision-broadened line. Evidence indicates the equation to be essentially correct in the intermediate pressure region of 10^2 to 10^{-2} mm of Hg. Attenuation values remote from resonance are studied and the frequency dependence of absorption discussed as a function of temperature and pressure. Below 10^{-3} mm Doppler broadening becomes the limiting line-width factor and line saturation effects occur for high incident power densities. Effects which become important at about one atmosphere, such as a shift in the resonance maximum and variations of line widths and structure with large partial pressure changes, are also studied.

9. ON THE DEFINITION OF VIRTUAL HEIGHT

Jerry Shmoys
(New York University, New York, N. Y.)

The difficulties with the geometrical optics ("classical") definition of virtual height are discussed. The more general definition of virtual height in terms of the derivative of the phase of the reflection coefficient with respect to frequency is derived. It is then shown that the former definition can be derived from the latter, if the phase integral method is used. The two definitions are compared in the specific examples of linear, rectangular, Epstein, and parabolic charge distributions.

It is demonstrated, by means of examples, that the relation between virtual height and frequency derivative of phase is not valid when the reflected wave contains more than one pulse. In this case the frequency derivative of phase cannot be interpreted as the time delay of any one of the pulses.

10. DISPERSION OF F_2-LAYER CRITICAL FREQUENCIES

M. L. Phillips and H. S. Moore
(National Bureau of Standards, Washington, D. C.)

Variations in dispersion of f^oF_2 about the monthly mean were studied with respect to solar activity, season, time of day, and geographical location. In general, the dispersion bears no simple relationship to the monthly mean f^oF_2. Comparison of the standard deviations for the five hours, 11 through 15, and 23 through 03, local standard time, individually, in summation, and in means of the five hours for each day, at three representative stations, showed a pronounced lack of independence in these data, especially for daytime hours. Evidence was found for consistent bias in the data obtained from certain stations. Nomograms were devised for applying the results of this analysis to practical transmission problems.

11. FLUCTUATIONS OF F_2 REGION BETWEEN STATIONS SEPARATED BY 100 TO 150 MILES

H. W. Wells
(Carnegie Institution of Washington, Washington, D. C.)

Analysis of ionosphere measurements made at three independent stations within a small geographical area revealed an occasional development of large differences in F_2-layer critical frequencies. Distances involved ranged from 100 to 150 miles. Mutually consistent values were observed during quiet periods by critical frequency differences in excess of 0.5 mc and on some occasions as great as 1.0 mc are observed. Cloud-like surges occurred independently at each station suggesting small physical dimensions of such disturbances. Results indicate the importance of transport by winds or redistribution by compressional-type traveling disturbances as affecting the instantaneous value of ionization density.

12. ANGLE OF ARRIVAL AND POLARIZATION AT FORT CHIMO

J. E. Hogarth
(Canadian Defence Research Board, Ottawa, Ont., Canada)

The direction of arrival of vertical incidence ionospheric echoes was measured by an interference method using four vertical-loop antennas placed at the corner of its horizontal square. The state of polarization was determined by phase measurements on a pair of mutually perpendicular loops. The measurements were designed to investigate spread echoes, echoes whose equivalent height changes rapidly, and echoes at a frequency higher than a critical frequency of the layer to which they appear to belong. Particularly important are polarization measurements on the z trace, which is uncommon in the latitude of Fort Chimo.

13. USE OF LONG-DISTANCE BACKSCATTER TO DETERMINE SKIP DISTANCE AND MAXIMUM USABLE FREQUENCY

William Abel
(Raytheon Manufacturing Company, Waltham, Mass.)

Measurements of delay time in the 5- to 30-mc band between the leading edge of the backscatter pattern and the response of a beacon transponder has shown that delay time to the leading edge of the backscatter is a measure of the skip distance. Since the frequency on which backscatter is obtained is the maximum usable frequency (muf) for the skip distance of this frequency, the muf for other distances can be obtained by conventional methods. Backscatter can be obtained using peak-power outputs as low as 200–300 w. Methods are being developed to permit instantaneous determination of skip distance using ordinary transmitters and receivers together with suitable display equipment.

14. APPLICATION OF PUNCH CARDS TO THE ANALYSIS OF MULTIPATH IONOSPHERIC PULSE PROPAGATION RECORDS

W. C. Moore
(Boston University, Boston, Mass.)

The film records obtained from multipath ionosphere pulse propagation experiments made aboard high altitude missiles must be analyzed almost frame by frame because the variable parameter *altitude* has a different value for each frame. As many as 36 individual items of information may be of interest on a single frame, or may be necessary for comparison between frames selected from different portions of the record. This mass of data can be recorded as pencil marks on *mark-sense* cards which are then automatically punched and used in machines for all subsequent operations in the evaluation of the results of the experiment. Methods and operations are given for reduction of the data and punch-card analysis.

15. LOW-FREQUENCY ANTENNAS

P. S. Carter
(RCA Laboratories Division, Rocky Point, N. Y.)

For transmitting antennas in the frequency range from 15 to 300 kc, radiation efficiency is of primary importance. For ground-based receiving antennas the problem is different. Because of the high level of atmospheric noise, signal-to-noise ratio is little affected by efficiency and a receiving antenna of low efficiency and moderate directive gain is superior to a highly efficient omnidirectional radiator. Improvement in efficiency by top loading and reduction of ground losses are discussed. Several examples are given including the characteristics of multiple-tuned antennas. Improvement of signal-to-noise ratios by the Beverage wave antenna as used at Riverhead is described.

16. ANTENNA SYSTEMS FOR RADIO DIRECTION FINDING

E. C. Jordan
(University of Illinois, Urbana, Ill.)

A brief summary is made of antenna systems useful for radio direction finding. The characteristics of systems both small and large in wavelengths are considered. Methods of antenna synthesis are described that will yield narrow beam patterns and optimum patterns for linear and circular arrays. Problems peculiar to radio-direction finding antennas in the low-frequency, medium-frequency, and high-frequency bands are reviewed and some of the solutions obtained are indicated.

17. ELECTRICALLY SMALL ANTENNAS AND THE LOW-FREQUENCY AIRBORNE ANTENNA PROBLEM

J. T. Bolljahn
(Stanford Research Institute, Stanford, Calif.)

Properties of antennas which are small relative to their operating wavelength are discussed. Such antennas have radiation patterns corresponding either to that of a short dipole or to that of a small loop. The problem of a small antenna of arbitrary shape is treated in detail and a sensitivity parameter based upon the equivalent generator circuit is defined. An experimental procedure for determining the sensitivity parameters and radiation patterns of low-frequency aircraft antennas using scale models immersed in an electrostatic field is described, and several experimental results are shown. The problem of small loop-type antennas is discussed in a less detailed fashion, and some calculations of the bearing errors indicated by a low-frequency direction-finder loop in the vicinity of an idealized aircraft structure are presented.

18. ANTENNAS IN CONDUCTING MEDIA

R. K. Moore
(Cornell University, Ithaca, N. Y.)

A chief difficulty arises from high losses in the immediate vicinity of the antennas. The input impedance to an antenna in a conducting medium is essentially due to current flowing near the antenna and bears little relation to the field at a distance. Methods for analyzing certain kinds of antennas are shown and results of analysis are compared.

19. WAVE PROPAGATION OVER ROUGH SURFACES

W. S. Ament
(Naval Research Laboratory, Washington, D. C.)

Effect of ocean waves on the propagation of microwaves is discussed. The reflection coefficient for plane waves almost normally incident is derived on the assumption that all slopes of the ocean surface are nearly horizontal and that no portion of the surface is shadowed. Reflection formulas are given for near grazing incidence where the latter assumption is invalid. These are used in explaining the observed attenuation rates of theoretically completely trapped 3-cm normal modes. Alternative mechanisms and formulations of the problem are discussed.

20. SIMULTANEOUS MOBILE MEASUREMENT OF THE FIELD STRENGTHS OF TWO VHF RADIO STATIONS OVER IRREGULAR TERRAIN

R. S. Kirby
(National Bureau of Standards, Washington, D. C.)

Mobile field strength measurements at vhf to evaluate the random variations due to buildings and rough terrain are described. Measurement techniques and analysis of the data are discussed, leading to information on the distribution of median values of field strength within given sectors and the correlation between variations in two received fields as observed simultaneously. Overwater measurements show the effect of smooth terrain near the received location in that relatively minor variations of field strength were observed.

21. SUPPRESSION OF WAVES BY ZONAL SCREENS

H. E. Bussey
(National Bureau of Standards, Washington, D. C.)

Novel diffraction screens covering half a Fresnel zone are described. These may produce a complete null of field at a point and a weak field in the surrounding shadow area, thereby providing a method for greatly reducing reception of either a direct or a reflected wave at a fixed antenna. Application of this technique to reduction of the ground reflection wave which may cause interference fading in, for example, microwave radio relays, is discussed. An experimental test over an 800-foot path using microwaves is described.

22. THE EFFECT OF LOW-LEVEL ATMOSPHERIC CONDITIONS ON OVERWATER INTERFERENCE PATTERNS AT MICROWAVE FREQUENCIES

V. R. Widerquist and J. E. Boyd
(Georgia Institute of Technology, Atlanta, Ga.)

This paper gives a simple theoretical study of the effect of low-level atmospheric conditions on overwater interference patterns for microwave frequencies. Experimental data on radar interference patterns are given. Shorter range patterns reveal evidence of slight compression of the surface lobe while longer range patterns indicate marked compression of this lobe, to widths as small as one half the standard lobe width. Comparison with meteorological soundings is given and the relationship of the measurements to anomalous propagation conditions close to the water surface is indicated.

23. VHF TROPOSPHERIC RECORDING MEASUREMENTS OF PLANE AND CIRCULAR POLARIZED WAVES IN THE GREAT LAKES AREA

J. S. Hill
(United Broadcasting Company, Cleveland, Ohio)
and
G. V. Waldo
(Federal Communications Commission, Washington, D. C.)

Results are given of measurements of characteristics of circularly polarized vhf waves propagated through the troposphere. Continuous field intensity recordings of horizontal and vertical components as well as the circularly polarized field over a 125-mile path from Columbus, Ohio, are given. Results are compared with recordings of plane-polarized wave propagation over the same path and over a different path.

24. THE EXPERIMENTAL AND THEORETICAL STUDY OF IONOSPHERIC ABSORPTION AT 150 KC

A. H. Benner
(RCA Victor Division, Camden, N. J.)

Analytical expressions for vertical incidence ionospheric absorption in E layer are obtained from a study of the complex index of refraction for 150 kc. The double parabolic approximation to the Chapman electron distribution and an exponential frequency variation with height are used. Calculations of the total absorption made by summing the contributions of the E region and the nondeviating D region indicate that the values of E-region absorption depend greatly on the characteristics of this region near the maximum of ionization. It is tentatively concluded that the D region is not a pure electron layer and that it probably does not exhibit a distinct ionization maximum. Typical vertical incidence 150-kc absorption data are presented and the results of analysis of 17 months' data are discussed.

25. THE MEASUREMENT OF THE POLARIZATION OF IONOSPHERIC REFLECTIONS AT LOW FREQUENCIES

R. A. Helliwell, A. J. Mallinckrodt, D. A. Campbell, and W. Snyder
(Stanford University, Stanford, Calif.)

A relatively simple method for measuring the polarization of pulse echoes is described not involving the use of frequency converters or phase shifters. Examples are given using typical E-layer echoes at 100 and 340 kc. Vertical incidence results indicate both directions of rotation of the polarization ellipse at 310 to 340 kc but only left-handed rotation at 100 kc. At times marked changes in polarization occurred within a few rf cycles. Cases of echo splitting in which the component polarizations are opposite in sense of rotation are considered in terms of ray theory.

26. THEORETICAL AND EXPERIMENTAL INVESTIGATION OF THE POLARIZATION OF LONG WAVES REFLECTED FROM THE IONOSPHERE

J. M. Kelso and H. J. Nearhoof
(The Pennsylvania State College, State College, Pa.)

Methods are given for determining the relations between measurable quantities of down-coming echo polarization and a complex polarization R defined by Appleton and Hartree. Calculations are made for 150 kc at State College, Pa. A cross-loop polarimeter is described. Polarization results indicate a definite diurnal trend with limiting polarization taking place for collisional frequency values greater than the critical.

27. POLARIZATION MEASUREMENTS OF LOW-FREQUENCY ECHOES

E. L. Kilpatrick
(National Bureau of Standards, Washington, D. C.)

A method for determination of polarization in a vertically incident down-coming radio wave is described using crossed dipoles and pulses. A gating arrangement permits selection of 25 microsecond samples for any desired portion of the echo, thus permitting independent observations of ordinary and extraordinary components. Experimental results indicate the polarization to be in general elliptical, with left-handed sense of rotation. The rapid time variation of polarization which sometimes occurs within a single echo suggests the presence of a composite reflection unresolved in time.

28. A METHOD FOR OBTAINING THE WAVE SOLUTIONS OF IONOSPHERICALLY REFLECTED LONG RADIO WAVES INCLUDING ALL VARIABLES AND THEIR HEIGHT VARIATION

J. J. Gibbons and R. J. Nertney
(The Pennsylvania State College, State College, Pa.)

A method is presented for obtaining approximate solutions to the wave equation in the form of a linearly independent pair involving a parameter p which is a complex function of the height. The solutions reduce to WKB solutions for regions for small partial reflection but differ considerably in region of considerable reflection. Examples are given for various configurations including derivation of reflection coefficients. Advantages include the simplicity of application and no requirement for explicit knowledge of the refractive index off the axis of reals.

29. ON THE THEORY OF ANTENNA BEAM-SHAPING

A. S. Dunbar
(Stanford Research Institute, Stanford, Calif.)

The diffraction pattern of an aperture with specified amplitude distribution and phase aberration is examined for two cases. The effects of these parameters on the beam shape and its control are discussed. A general formulation of Chu's method[1] for an amplitude distribution on a curved surface is developed. Theoretical results are applied to the design of progressive-phase antennas for

[1] L. J. Chu, "Microwave Beam-Shaping Antennas," Tech. Rpt. No. 40, Research Lab. of Electronics, MIT, Cambridge, Mass.; 1947.

both control variation of the amplitude distribution and the controlled variation of phase. Experimental results obtained with long-slot antennas, corrugated-surface antennas, slot-array antennas, and a dielectric slab on a curved surface, are presented.

30. BEAM SHAPING IN DOUBLY CURVED REFLECTOR SYSTEMS EMPLOYING QUASI-POINT SOURCES

A. E. Marston
(Naval Research Laboratory, Washington, D. C.)

Techniques for designing doubly curved reflectors for shaped beams, developed by Chu and Silver, are extended to embrace the case of the so-called quasi-point source. Such sources in general are identified with a one-parameter family of normal congruences which includes the spherical and cylindrical congruences as special cases. An explicit design procedure is given and experimental confirmation presented.

31. LACK OF UNIQUENESS IN ANTENNA PATTERN SYNTHESIS METHODS AND THE RELATED ENERGY STORAGE CONSIDERATIONS

T. T. Taylor
(Hughes Aircraft Company, Culver City, Calif.)

A modification of a nonsynthesis method for aperture-type antennas is discussed and it is shown how, within directivity limitations, high-stored energy solutions can be avoided. The synthesis problem becomes more difficult if the objective is to find an excitation function whose Fourier transform shall approximate a given function with a specified interval. This problem can be solved by methods analogous to those discussed by Whinnery and the author in regard to arrays of discrete elements. Extent of the lack of uniqueness of the solution is discussed.

32. FEED PROBLEMS IN BROADBAND ANTENNA ARRAYS

W. R. LePage and R. F. Gates
(Syracuse University, Syracuse, N. Y.)

The input admittance of an element of the array is analyzed to show the effect of coupling to adjacent elements. This is particularly important when adjacent elements are fed in different phase. The analysis is carried out by neglecting coupling to elements farther from a given element than its immediate neighbors. An equivalent circuit is evolved and analyzed for the input admittance to such an element. Also, the range to be expected for the standing-wave ratio on a line feeding a tapered section is discussed in terms of the impedance transformation of the section, and the standing-wave ratios on the various lines which it feeds. Among other things, it is shown that the standing-wave ratio on the feed to the tapered section is always less than the average of the standing-wave ratios on the lines leading from it, if the tapered section has the proper impedance ratio.

33. SECOND-ORDER BEAMS OF SLOTTED WAVEGUIDE ARRAYS

H. Gruenberg
(National Research Council, Ottawa, Ont., Canada)

An array of longitudinal shunt slot in the broad face of a rectangular waveguide is studied. It is shown that the radiation pattern of such an array will contain small second-order beams even for slot spacings less than a free-space wavelength. These beams are usually split with a minimum or even a null in the plane containing the axis of the array and the maximum of the main beam. Second-order beams may be as large as 10 per cent of the main beam (in voltage) for typical arrays about 50 wavelengths long, and even larger for shorter arrays. Methods of suppressing these beams are discussed and experimental confirmation is given.

34. VERY-HIGH-FREQUENCY PROPAGATION IN THE EQUATORIAL REGION

O. P. Ferrell
(Radio Magazines, Inc., Philadelphia, Pa.)

Radio amateur observations in the frequency range 50.0–54.0 mc are being collected throughout the Western Hemisphere. Sky-wave propagation in the geographic tropical zone may be divided into three categories, which are: (1) sporadic-E reflections; (2) F_2-layer reflections, particularly associated with ionospheric storms in the temperate latitudes; (3) post-sunset F_2-region scatter. The effect of these modes upon vhf propagation are summarized with particular emphasis placed upon the hitherto unreported post-sunset scatter. The results of the present study indicate that the scattering is observed over fairly long paths, but is restricted to those months around the equinoxes

35. SOUTHERN EXTENT OF AURORA BOREALIS IN NORTH AMERICA

C. W. Gartlein and R. K. Moore
(Cornell University, Ithaca, N. Y.)

Results are presented for the first eleven years of a study of the frequency of overhead auroras in North America as a function of latitude in a region south of the auroral zone. The data have been averaged in various ways so that monthly and annual variations are demonstrated. It appears that there is a relatively constant level of auroral activity throughout the year in the region 58°–60° geomagnetic latitude, while auroras appearing overhead south of these latitudes are more frequent during equinoctial periods. Auroras have been seen as far south as 52° during this period every month of the year. Correlation of auroral frequency with sunspot number is not high on a month-by-month or three-month running mean basis.

36. A VHF PROPAGATION PHENOMENON ASSOCIATED WITH THE AURORA

R. K. Moore
(Cornell University, Ithaca, N. Y.)

Anomalous propagation observed by radio amateurs at 28 to 148 mc during displays of Aurora Borealis is described. This phenomenon is characterized by the following features: (1) a very high fading rate, such that voice modulation is rendered unintelligible; (2) directional antennas give best results when pointed north (toward the aurora); (3) lack of skip effect; (4) little change in polarization. Results indicate this effect to be most common with auroral displays extending below 56° geomagnetic latitude.

37. PHASE VELOCITY OF VERTICALLY POLARIZED ELECTROMAGNETIC WAVES IN THE DIFFRACTION REGION AT THE SURFACE OF A SPHERE

Henry Lisman
(Signal Corps Engineering Laboratories, Fort Monmouth, N. J.)

Based on the work of Eckersley and Millington, an investigation was carried out to determine phase velocity of electromagnetic waves over the surface of a sphere. Numerical calculations were carried on for 5.305 mc on various assumptions as to values of dielectric constant and conductivity. The velocity was found insensitive to changes in ground constants. Changes in assumed radius of curvature were found to affect the velocity, thereby indicating a substantial difference over a region where the curvature varies considerably to that over a smooth earth.

38. MAGNETO-IONIC MULTIPLE SPLITTING DETERMINED WITH THE METHOD OF PHASE INTEGRATION

Wolfgang Pfister
(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

The method used is a first-order WKB approximation with an integration path in the complex height plane. Suitable paths of integration in the Riemann surface determine the reflection coefficient for the various magneto-ionic components. Numerical computations have been carried out for a Chapman distribution of electrons and an exponential decrease of collisional frequency located in a moderate magnetic latitude. The results are represented in a sweep-frequency picture of virtual height and absorption for five fundamental modes.

39. ELECTROMAGNETIC ENERGY DENSITY AND FLUX

C. O. Hines
(Canadian Defence Research Board, Ottawa, Ont., Canada)

J. C. Scott and H. G. Booker by different methods have obtained discrete formulas for energy flow in the ionosphere when absorption is considered. Investigation of the bases of the two methods revealed that the usual expressions for energy density and flux lead to difficulties which are avoided by alternative expressions based on an equally valid conservation theorem. These alternative formulas are shown to lead to much neater results, not only for the dissipative case, but for all problems in plane wave propagation, and also give the group speed for the speed of the energy.

40. ASYMPTOTIC SOLUTION OF MAXWELL'S EQUATION

Morris Kline
(New York University, New York, N. Y.)

A method is given which improves upon geometrical optics solutions of electromagnetic problems by establishing asymptotic series for E and H. The successive coefficients of the two series are obtained by solving two recursive systems of first-order ordinary differential equations. These differential equations must be solved along each system of rays. At any point in space each coefficient of the series is the sum of the solutions of the differential equations corresponding to rays passing through the point.

41. CURRENT DISTRIBUTIONS ON LARGE REFLECTING CYLINDERS

H. J. Riblet
(Microwave Development Laboratories, Inc., Waltham, Mass.)

This preliminary report describes calculations and measurements made with the intention of determining the current distribution on an infinite circular cylinder, whose radius is $12\lambda 2\pi$, when placed in a plane wave. Two cases are considered. For Case I, the incident electric vector is polarized parallel to the axis of the cylinder, while for Case II, the incident magnetic vector is polarized in that direction. Six different current distributions are compared in amplitude and phase for both cases and some general conclusions are reached.

42. APERTURE PHASE ERRORS IN MICROWAVE OPTICS

K. S. Kelleher
(Naval Research Laboratory, Washington, D. C.)

The fundamental difference between optics and microwave optics lies in the fact that the former stresses the aberrations resulting from imperfect focussing, while the latter is more concerned with aberrations in terms of the deviation of a wavefront from a plane. An indication of the relation between the phase errors which describe this deviation and the radiation pattern is made by considering the case of a paraboloidal reflector with feed displaced from the focal point and the case of a spherical cap reflector with feed on axis. In these instances, consideration is given to the amplitude of intensity across the aperture since phase errors are relatively unimportant when they occur in the region of the aperture in which the intensity is low.

43. REVIEW OF SPHERICAL REFLECTOR RESEARCH AT AFCRL

J. E. Walsh
(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

Spherical reflectors have been investigated because of their superiority to the paraboloid in wide-angle scanning potentialities, even though they are subject to spherical aberration. The optimum location of a point source feed was determined and experimental measurements made on patterns. Line source feeds with nonconstant phase gradients were also investigated, resulting in an aberrationless optical system with a favorable f/d ratio which can scan a solid angle of at least 60°. Successful experimental tests of these line sources have been made.

44. REVIEW OF RECENT METAL LENS RESEARCH AT AFCRL

John Ruze
(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

Lenses of the "constrained" type have been investigated in which the ray direction is determined by the metal-plate configuration. Lenses of this type lend themselves readily to wide-angle analysis, and can be designed to focus perfectly at two symmetrically placed off-axis points. These corrected lenses have good scanning properties and have been successfully constructed for use with both line and point sources. Work has also been done on achromatic lenses in which electromagnetic energy in the *TEM* mode of transmission is guided through the plates. Two methods of focussing by adjusting the path length through each pair of plates are described.

Institute Committees—1951

Executive

I. S. Coggeshall, *Chairman*
W. R. G. Baker, *Vice-Chairman*
Haraden Pratt, *Secretary*

S. L. Bailey, R. F. Guy
A. N. Goldsmith, J. W. McRae

Constitution and Laws

R. A. Heising, *Chairman*

J. B. Epperson, A. A. McKenzie
D. B. Harris, Haraden Pratt
H. R. Zeamans

Professional Groups

W. R. G. Baker, *Chairman*

L. L. Beranek, R. A. Heising
W. H. Doherty, Harner Selvidge
W. L. Everitt, J. R. Steen
R. F. Guy, L. C. Van Atta
Ernst Weber
(Professional Group Chairmen Ex-Officio)

Finance

W. R. G. Baker, *Chairman*

S. L. Bailey, R. F. Guy
I. S. Coggeshall, ex-officio, F. B. Llewellyn

Education

H. J. Reich, *Chairman*

R. G. Anthes, P. H. Nelson
R. E. Beam, J. M. Pettit
L. J. Black, W. H. Pickering
C. G. Brennecke, Albert Preisman
C. C. Chambers, L. R. Quarles
R. M. Fano, John Ragazzini
G. H. Fett, J. D. Ryder
R. A. Galbraith, George Sinclair
A. W. Graf, R. P. Siskind
A. E. Harrison, W. R. Smith
R. W. Hickman, Karl Spangenberg
G. B. Hoadley, F. R. Stansel
G. L. Hollander, Ernst Weber
A. H. Howell, L. E. Williams
F. S. Howes, A. H. Wing, Jr.
Glenn Koehler, C. R. Wischmeyer
A. W. Melloh, Irving Wolff

Policy Development

F. B. Llewellyn, *Chairman*

S. L. Bailey, W. R. Hewlett
W. H. Doherty, J. V. L. Hogan
A. V. Eastman, T. A. Hunter
W. L. Everitt, Haraden Pratt
Keith Henney, J. D. Reid
J. A. Stratton

Nominations

F. E. Terman, *Chairman*

S. L. Bailey, F. Hamburger, Jr.
Ralph Bown, W. R. Hewlett
R. F. Guy, T. A. Hunter
A. F. Van Dyck

Membership

T. H. Clark, *Chairman*

F. W. Albertson, W. A. Knoop
V. E. Clayton, R. M. Kreuger
J. B. Coleman, O. I. Lewis
H. C. Forbes, T. G. Morrissey
G. W. Fyler, B. F. Osbahr
R. N. Harmon, F. H. R. Pounsett
A. E. Harrison, L. C. Signon
J. H. Ivers, D. B. Smith
E. C. Jordan, W. C. White
(Chairmen of Section Membership Committees Ex-Officio)
(Regional Directors Ex-Officio)

Public Relations

R. R. Batcher, *Chairman*

W. C. Copp, R. F. Guy
W. L. Everitt, R. A. Hackbush
E. K. Gannett, G. L. Van Deusen
V. M. Graham, Lewis Winner
(Chairmen of Section Publicity Committees Ex-Officio)

Tellers

E. J. Isbister, *Chairman*

J. Z. Millar
J. H. Moore

Board of Editors

A. N. Goldsmith, *Chairman*

F. W. Albertson
William Bachman
G. M. K. Baker
W. L. Barrow
R. R. Batcher
B. B. Bauer
H. S. Black
R. M. Bowie
Ralph Bown
R. S. Burnap
O. H. Caldwell
C. W. Carnahan
C. C. Chambers
L. M. Clement
J. D. Cobine
J. W. Coltman
M. G. Crosby
R. L. Dietzold
R. B. Dome
W. G. Dow
E. W. Engstrom
W. L. Everitt
W. G. H. Finch
D. G. Fink
H. C. Forbes
I. A. Getting
A. W. Graf
F. W. Grover
A. V. Haeff
D. B. Harris
L. B. Headrick
E. W. Herold
J. K. Hilliard
J. K. Johnson
L. F. Jones
H. S. Knowles
J. D. Kraus
J. B. H. Kuper
S. C. Lawrence
D. G. Little
F. B. Llewellyn
S. S. Mackeown
Nathan Marchand
E. D. McArthur
Knox McIlwain
Donald McNicol
J. W. McRae
L. A. Meacham
G. F. Metcalf
E. L. Nelson
D. O. North
K. A. Norton
H. F. Olson
R. M. Page
H. O. Peterson
G. W. Pickard
Haraden Pratt
C. A. Priest
John Ragazzini
Simon Ramo
H. J. Reich
J. D. Reid
F. X. Rettenmeyer
P. C. Sandretto
H. H. Scott
Stuart Seeley
V. W. Sherman
L. C. Smeby
C. E. Smith
Newbern Smith
J. A. Stratton
G. C. Sziklai
W. C. Tinus
K. S. Van Dyke
E. K. Van Tassel
Ernst Weber
H. A. Wheeler
J. R. Whinnery
W. C. White
L. E. Whittemore
Jerome Wiesner
Irving Wolff
Harold Zahl
V. K. Zworykin

Papers Review

George Metcalf, *Chairman*

H. A. Affel
W. J. Albersheim
J. J. Baruch
W. R. Bennett
P. H. Betts
F. J. Bingley
D. S. Bond
D. R. Brown
Kenneth Bullington
H. A. Chinn
J. K. Clapp
S. B. Cohn
F. W. Cunningham
L. T. DeVore
A. R. D'heedene
M. J. Di Toro
H. D. Doolittle
O. S. Duffendack
R. D. Duncan, Jr.
E. H. Felix
L. M. Field
J. W. Forrester
V. H. Fraenckel
R. L. Freeman
E. G. Fubini
W. M. Goodall
O. B. Hanson
A. E. Harrison
Walter Hausz
T. J. Henry
C. N. Hoyler
P. K. Hudson
W. H. Huggins
D. L. Jaffe
Henry Jasik
E. V. Jordan
D. C. Kalbfell
A. G. Kandoian
Martin Katzin
C. R. Knight
W. E. Kock
J. G. Kreer, Jr.
Emile Labin,
J. J. Lamb
V. D. Landon
M. T. Lebenbaum
B. R. Lester
C. V. Litton
W. P. Mason
R. E. Mathes
H. F. Mayer
L. L. Merrill
H. R. Mimno
F. L. Moseley
A. F. Murray
J. R. Nelson
C. H. Page
H. W. Parker
L. J. Peters
A. P. G. Peterson
W. H. Pickering
A. F. Pomeroy
Albert Preisman
C. F. Quate
J. A. Rajchman
H. E. Roys
J. D. Ryder
H. A. Samulon
Samuel Seely
Harner Selvidge
R. E. Shelby
George Sinclair
J. E. Smith
W. M. Smith
R. L. Snyder
E. E. Spitzer
J. R. Steen
H. P. Thomas
C. E. Torsch
Bertram Trevor
W. N. Tuttle
Dayton Ulrey
L. C. Van Atta
S. N. Van Voorhis
J. R. Weiner
M. S. Wheeler
R. M. Wilmotte
H. R. Zeamans

Sections

J. F. Jordan, *Chairman*

Ben Akerman
J. B. Epperson
R. A. Heising
C. A. Norris
G. Rappaport
J. E. Shepherd
E. T. Sherwood
L. C. Sigmon
R. N. White
(Section Chairmen Ex-Officio)

Admissions

P. S. Christaldi, *Chairman*

H. S. Bennett
C. M. Burrill
A. G. Clavier
H. P. Corwith
C. A. Hachemeister
L. N. Hatfield
R. E. Mathes
H. S. Moncton
G. B. Riley
H. C. Vance

Administrative Committee of The Board of Editors

A. N. Goldsmith, *Chairman*

G. M. K. Baker
H. S. Black
R. S. Burnap
R. L. Dietzold
A. V. Haeff
E. W. Herold
F. B. Llewellyn
Knox McIlwain
Donald McNicol
K. A. Norton
F. X. Rettenmeyer
Newbern Smith
G. C. Sziklai
Ernst Weber
H. A. Wheeler
L. E. Whittemore
Harold Zahl

Awards

A. V. Loughren, *Chairman*

S. L. Bailey
K. C. Black
D. E. Chambers
M. G. Crosby
A. V. Eastman
F. S. Howes
J. F. Morrison
J. D. Reid
B. E. Shackelford
J. E. Shepherd
D. B. Sinclair

Technical Committees, May 1, 1951–April 30, 1952

Annual Review

R. R. Batcher, *Chairman*
Trevor Clark, *Vice-Chairman*

J. D. Crawford
R. T. Hamlett
G. L. Van Deusen
H. P. Westman
L. E. Whittemore

Antennas and Waveguides

A. G. Fox, *Chairman*
D. C. Ports, *Vice-Chairman*

P. S. Carter
George Deschamps
J. E. Eaton
Henry Jasik
E. C. Jordan
M. L. Kales
O. E. Kienow
George Sinclair
P. H. Smith

Audio Techniques

R. A. Miller, *Chairman*
H. H. Scott, *Vice-Chairman*

O. L. Angevine, Jr.
H. W. Augustadt
W. Lindsay Black
Harold Burris-Meyer
C. A. Cady
D. H. Castle
W. W. Dean
H. D. Harris
J. K. Hilliard
D. E. Maxwell

Circuits

C. H. Page, *Chairman*
W. R. Bennett, *Vice-Chairman*

J. G. Brainerd
A. R. D'heedene
R. L. Dietzold
R. M. Foster
W. H. Huggins
Herbert Krauss
John Linvill
P. F. Ordung
E. H. Perkins
Wolcott Smith
W. N. Tuttle
Ernst Weber
J. R. Weiner

Electroacoustics

E. S. Seeley, *Chairman*
B. B. Bauer, *Vice-Chairman*

E. C. Gregg
H. C. Hardy
W. F. Meeker
H. F. Olson
R. J. Rockwell
Vincent Salmon
P. S. Veneklasen

Electronic Computers

Nathaniel Rochester, *Chairman*
Robert Serrell, *Vice-Chairman*

D. R. Brown
R. D. Elbourn
J. W. Forrester
E. L. Harder
John Howard
E. Lakatos
B. R. Lester
G. W. Patterson
J. A. Rajchman
R. L. Snyder
W. H. Ware
J. R. Weiner
C. F. West
Way Dong Woo

Electron Tubes and Solid-State Devices

A. L. Samuel, *Chairman*
G. D. O'Neill, *Vice-Chairman*

E. M. Boone
R. S. Burnap
J. W. Clark
W. J. Dodds
W. G. Dow
George Esperson
C. E. Fay
M. S. Glass
A. M. Glover
J. E. Gorham
J. W. Greer
L. A. Hendricks
T. J. Henry
E. C. Homer
S. B. Ingram
R. B. James
J. A. Morton
I. E. Mouromtseff
L. S. Nergaard
P. A. Redhead
H. J. Reich
A. C. Rockwood
R. M. Ryder
W. G. Shepherd
R. W. Slinkman
H. L. Thorson
C. M. Wheeler

Facsimile

R. J. Wise, *Chairman*
Henry Burkhard, *Vice-Chairman*

A. G. Cooley
I. H. Franzel
John V. Hogan
Pierre Mertz
C. J. Young

Industrial Electronics

John Dalke, *Chairman*
Eugene Mittelmann, *Vice-Chairman*

G. P. Bosomworth
J. M. Cage
E. W. Chapin
J. E. Eiselein
G. W. Klingaman
H. R. Meahl
H. W. Parker
S. I. Rambo
Walther Richter
E. H. Schulz
C. F. Spitzer
W. R. Thurston
R. S. Tucker
M. P. Vore
Julius Weinberger
D. E. Watts

Measurements and Instrumentation

F. J. Gaffney, *Chairman*

Wilson Aull
C. C. Chambers
P. S. Christaldi
John Dalke
G. L. Fredendall
W. D. George
E. I. Green
W. J. Mayo-Wells
G. A. Morton
C. D. Owens
A. P. G. Peterson
J. G. Reid, Jr.
J. R. Steen
R. S. Tucker
Ernst Weber

Mobile Communications

F. T. Budelman, *Chairman*
Alexander Whitney, *Vice-Chairman*

G. M. Brown
D. B. Harris
C. M. Heiden
C. N. Kimball, Jr.
D. E. Noble
J. C. O'Brien
David Talley
George Teommey
R. W. Tuttle

Modulation Systems

W. G. Tuller, *Chairman*
J. G. Kreer, Jr., *Vice-Chairman*

L. A. DeRosa
D. D. Grieg
E. R. Kretzmer
V. D. Landon
Nathan Marchand
L. A. Meacham
Winslow Palmer
Dale Pollack
J. R. Ragazzini
H. E. Singleton
Bertram Trevor

Navigation Aids

P. C. Sandretto, *Chairman*
C. J. Hirsch, *Vice-Chairman*
H. R. Mimno, *Vice-Chairman*

W. B. Burgess
Harry Davis
R. E. Gray
Wayne Mason
H. I. Metz
K. A. Norton
Winslow Palmer
L. M. Sherer

Piezoelectric Crystals

R. A. Sykes, *Chairman*
W. P. Mason, *Vice-Chairman*

W. L. Bond
Clifford Frondel
W. G. Cady
J. K. Clapp
W. A. Edson
Edward Gerber
Hans Jaffe
P. L. Smith
K. S. Van Dyke

Radio Transmitters

M. R. Briggs, *Chairman*
A. E. Kerwien, *Vice-Chairman*

E. L. Adams
T. J. Boerner
H. R. Butler
L. K. Findley
Harold Goldberg
J. B. Heffelfinger
J. B. Knox
L. A. Looney
J. F. McDonald
John Ruston
Berthold Sheffield
I. R. Weir

Receivers

Jack Avins, *Chairman*
J. D. Reid, *Vice-Chairman*

K. A. Chittick
R. De Cola
L. M. Harris
K. W. Jarvis
J. K. Johnson
W. R. Koch
I. E. Lempert
C. R. Miner
Garrard Mountjoy
R. F. Norton
Leon Riebman
L. M. Rodgers
S. W. Seeley
S. C. Spielman
W. O. Swinyard
R. S. Yoder

Sound Recording and Reproducing

H. E. Roys, *Chairman*
A. W. Friend, *Vice-Chairman*

S. J. Begun
M. S. Corrington
R. M. Fraser
G. P. Hixenbaugh
E. W. Kellogg
A. P. G. Peterson
Harry Schecter
R. A. Schlegel
Lincoln Thompson
R. E. Zenner

Standards Committee

A. G. Jensen, *Chairman*
M. W. Baldwin, Jr., *Vice-Chairman*
L. G Cumming, *Vice-Chairman*
Ernst Weber, *Vice-Chairman*

Jack Avins
R. R. Batcher
H. G. Booker
J. G. Brainerd
M. R. Briggs
F. T. Budelman
P. S. Carter
A. G. Clavier
John Dalke
A. G. Fox
F. J. Gaffney
R. A. Hackbusch
E. A. Laport
Wayne Mason
R. A. Miller
C. H. Page
W. J. Poch
A. F. Pomeroy
N. Rochester
H. E. Roys
A. L. Samuel
P. C. Sandretto
E. S. Seeley
R. F. Shea
R. E. Shelby
R. A. Sykes
W. G. Tuller
L. E. Whittemore
R. J. Wise

Symbols

A. G. Clavier, *Chairman*
K. E. Anspach, *Vice-Chairman*

W. J. Everts
W. A. Ford
R. T. Haviland
O. T. Laube
C. D. Mitchell
C. Neitzert
A. F. Pomeroy
A. A. Powell
M. B. Reed
M. S. Smith
H. R. Terhune
H. P. Westman

Television Systems

R. E. Shelby, *Chairman*
R. M. Bowie, *Vice-Chairman*

W. F. Bailey
M. W. Baldwin, Jr.
A. H. Brolly
J. E. Brown
K. A. Chittick
C. G. Fick
D. G. Fink
C. J. Franks
P. C. Goldmark
R. N. Harmon
J. L. Hollis
A. G. Jensen
I. J. Kaar
R. D. Kell
P. J. Larsen
H. T. Lyman
Leonard Mautner
J. Minter
J. H. Mulligan, Jr.
A. F. Murray
J. A. Ouimet
D. W. Pugsley
David Smith
M. E. Strieby
A. Talamini

Video Techniques

W. J. Poch, *Chairman*
A. J. Baracket, *Vice-Chairman*

M. W. Baldwin, Jr.
P. F. Brown
R. H. Daugherty, Jr.
Stephen Doba, Jr.
V. J. Duke
G. L. Fredendall
R. L. Garman
J. L. Jones
J. E. Keister
L. L. Lewis
R. S. O'Brien
C. G. Pierce
B. F. Tyson
J. F. Wiggin

Wave Propagation

H. G. Booker, *Chairman*
H. W. Wells, *Vice-Chairman*

E. W. Allen, Jr.
S. L. Bailey
C. R. Burrows
T. J. Carroll
A. B. Crawford
A. E. Cullum, Jr.
W. S. Duttera
E. G. Fubini
I. H. Gerks
M. C. Gray
D. E. Kerr
J. E. Keto
Morris Kline
M. G. Morgan
K. A. Norton
H. O. Peterson
George Sinclair
Newbern Smith
R. L. Smith-Rose
A. W. Straiton
A. H. Waynick
J. W. Wright

Special Committees

Professional Recognition

G. B. Hoadley, *Chairman*

C. C. Chambers
H. F. Dart
W. E. Donovan
C. M. Edwards

Armed Forces Liaison

G. W. Bailey, *Chairman*

INSTITUTE REPRESENTATIVES IN COLLEGES—1951*

*Agricultural and Mechanical College of Texas: H. C. Dillingham
*Akron, University of: P. C. Smith
*Alabama Polytechnic Institute: G. H. Saunders
*Alberta, University of: J. W. Porteous
*Arizona, University of: H. E. Stewart
*Arkansas, University of: G. H. Scott
*British Columbia, University of: L. R. Kersey
*Brooklyn, Polytechnic Institute of, (Day Division): H. A. Foecke
*Brooklyn, Polytechnic Institute of, (Evening Division): A. B. Giordano

* Colleges with approved Student Branches.

*Bucknell, University of: B. H. Bueffel, Jr.
*California Institute of Technology: W. H. Pickering
*California State Polytechnic College: Clarence Radius
*California, University of: L. J. Black
California, University of at Los Angeles: E. F. King
Carleton College: G. R. Love
*Carnegie Institute of Technology: E. M. Williams
*Case Institute of Technology: J. D. Johannesen
Cincinnati, University of: A. B. Bereskin
*Clarkson College of Technology: F. A. Record
*Colorado, University of: H. W. Boehmer
*Columbia University: J. R. Ragazzini
*Connecticut, University of: C. W. Schultz
*Cooper Union: J. B. Sherman
*Cornell University: True McLean
Dartmouth College: M. G. Morgan
*Dayton, University of: Appointment later
*Delaware, University of: H. S. Bueche
*Denver, University of: Herbert Reno
*Detroit, University of: Thomas Yamauchi
*Drexel Institute of Technology: R. T. Zern
Duke University: W. J. Seeley
Evansville College: J. F. Sears

Conference at Massachusetts Institute of Technology, 1950. *See* also 2693 of 1950.

534.782 1058
An Electrical Vocal System—L. O. Schott. (*Bell Lab. Rec.*, vol. 28, pp. 549–554; December, 1950.) An outline description of apparatus arranged to form an electrical analogue of the human vocal system. The "vocal tract" comprises an artificial transmission line of 24 L sections. The "tongue hump" is an auxiliary inductor of adjustable value ("hump-height") which can be switched in series with any chosen inductive series element of the transmission line. The "lip opening" is an adjustable inductor which terminates the line and across which the loudspeaker is connected. Separate generators are provided for producing vowels and unvoiced fricatives, while further arrangements enable pitch and inflection to be varied and other vocal features to be reproduced.

534.784 1059
System-Function Analysis of Speech Sounds—W. H. Huggins. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 765–767; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. "The analysis of speech sound is facilitated if the sound is considered to be the response of a slow-time-varying linear system to appropriate excitations. The linear system is characterized by one or more system functions which may be represented most naturally as the sum of complex exponential terms having complex frequencies corresponding to the various formants. Thus, a generalized frequency analysis is made not of the speech wave itself but rather of the system function that shaped that wave. On the other hand, the excitation is best analyzed by autocorrelation methods."

534.861:534.76 1060
Stereophonic Broadcasting of Music—J. Bernhart and J. W. Garrett. (*Toute la Radio*, no. 150, pp. 353–358; November, 1950.) Discussion of methods of obtaining "relief" effect in sound reproduction, including an analysis of directional discrimination in audition and a description of the principle of "directed stereophony." Variation of intensity itself is sufficient to convey a sense of spatial movement of the sound source, particularly for low frequencies. This can be accomplished using a single microphone and two separate modulation channels the levels of which are adjusted independently: a 5-db alteration between the two levels corresponds to a radial movement of 30° angular displacement being a linear function of the intensity difference. Phase variation has a similar effect but necessitates a more complex system in practice. Displacement in depth is conveyed by use of an echo room. The effects of combination and rapidity of these adjustments are discussed with reference to an experimental broadcast transmission from Paris [*see* 1061 (Aisberg)].

534.861:534.76 1061
Improved Stereophony—E. Aisberg. (*Wireless World*. vol. 56, pp. 327–330; September, 1950.) An account of experiments in France, using simultaneous transmissions from the Parisian and Paris Inter broadcasting stations. For reception two loudspeakers spaced 5 to 7 feet apart were required, the listeners being 7 to 10 feet away. The two receivers receive signals from separate channels and by suitably controlling the modulation depth for the two channels, a realistic impression of movement of the original source of sound is given to the listeners. Modulation of the two channels was recorded on double-track magnetic tape. *See* also 1060 above.

621.314.2.029.3:621.3.018.78† 1062
The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers—Macfayden. (*See* 1087.)

621.392.52 1063
Unusual Ladder Filter. Applications in Audio and Radio Circuits—Davey. (*See* 1112.)

621.395.623.45 1064
An Ultrasonic Underwater 'Point Source' Probe—L. Fein. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 876–877; November, 1950.) An ultrasonic receiver uses a conical piezoelectric probe made of ADP. Its sensitivity over the frequency range 1.235 to 1.280 mc was determined by means of a reciprocity calibration, an average value of 0.012 μv per dyne per cm^2 being obtained. The measured directional pattern does not very with position in the field, indicating that the probe is effectively a point receiver.

621.395.625.3/.6(41) 1065
Magnetic-Sound-Film Developments in Great Britain—O. K. Kolb. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 55, pp. 496–508; November, 1950.) The introduction of this type of recording and reproducing apparatus in Great Britain is reviewed. Types of magnetic film available and their characteristics are described. The general circuit arrangements and details of apparatus are given together with a description of special apparatus that has been used for adding a visible record to the invisible magnetic sound track. Jointing processes and a method for bulk erasure of records on magnetic-film stock are discussed.

621.396.645.029.3:621.385.3/.4 1066
A Comparison of Triodee and Beam-Tetrodes as Power Output Valves in Audio Amplifiers—Bruckmann, Carey, and Fuller. (*See* 1123.)

ANTENNAS AND TRANSMISSION LINES

621.392 1067
Phase Distortion in Feeders—L. Lewin, J. J. Muller, and R. Basard. (*Elect. Commun.* (London), vol. 27, pp. 320–323; December, 1950.) Reprint. *See* 2125 of 1950.

621.392:621.3.012.1/.2 1068
Geometrical Solution of some Transmission-Line Problems—Q. C. Gupta and F. Duerden. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 525–529; December, 1950.) A description of the vector-diagram approach, which is said to provide a better physical picture than analysis based on differential equations. The method is illustrated by an analysis of stub matching and a verification of the principles of the circle-diagram calculator.

621.392:[621.3.015.71:621.314.2 1069
Pulse Transients in Exponential Transmission Lines—In 280 of March, for "E. R. Schatz" please read "E. R. Schatz and E. M. Williams."

621.392:621.3.09 1070
The Propagation of Transient and Periodic Waves along Transmission Lines—R. Pélissier. (*Rev. gén. Elect.*, vol. 59, pp. 379–399, 437–454 and 502–512; September to November, 1950.) A general theory of the mode of propagation and of the different causes of attenuation of electric waves is developed, using the symbolic calculus. The theory generalizes those of Carson and Pollaczek. Approximate formulas are derived and the limits of their applicability are discussed. Applications of the theory to cases of practical importance include, (*a*) calculation of charge and current distribution along an infinitely long line after application of a step voltage, (*b*) corresponding calculations for a line of finite length, including discussion of reflection at terminations of various types, (*c*) reflection at line discontinuities, (*d*) discussion of causes of wave attenuation, (*e*) propagation of high-frequency currents and of surges due to lightning. The theory shows that the explanation of the attenuation of such surges depending on corona effects is inexact. A new explanation is developed and confirmed by experimental results. *See* also 2972 of 1950.

621.392.011.2:621.3.012.3 1071
Transmission-Line Impedance Curves—H. A. Wheeler. (Proc. I.R.E., vol. 38, pp. 1400–1403; December, 1950.) A universal family of curves gives the wave resistance, inductance, and capacitance of transmission lines with 1, 2, or 4 wires. Each curve represents one field pattern. An expanded scale is given for the region of very low impedance. The seven field patterns are for the cases of conductors of circular cross section in or near shields such as a plane, inclined or parallel planes, a trough, or a coaxial conductor.

621.392.4 1072
A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped *LC* Circuits—van Weel. (*See* 1107).

621.396.67 1073
Theory of Collinear Antennas—R. King. (*Jour. Appl. Phys.*, vol. 21, pp. 1232–1251; December, 1950.) The analysis of a system of collinear antennas is practicable only if the transmission lines feeding the system are in the neutral plane of the electromagnetic field. Two cases are considered: a single center-driven unit, (*a*) with two symmetrical collinear parasitic elements, and (*b*) with the outer elements driven from the center unit by means of phase-reversing stubs. Approximate expressions are obtained for the currents and for the mutual and self impedances of the elements. The particular case of $\lambda/2$ dipoles is treated in detail. The advantages of the collinear array with phase-reversing stubs as compared with arrays in which each element is center driven are discussed.

621.396.67 1074
Elliptically Polarized Waves—L. Hatkin. (Proc. I.R.E., vol. 38, p. 1455; December, 1950.) Comment on 3343 of 1948 (Sichak and Milazzo).

621.396.67 1075
On the Radiation Patterns of Dielectric Rods of Circular Cross Section—the $TM_{0,1}$ Mode—C. W. Horton, F. C. Karal, Jr., and C. M. McKinney. (*Jour. Appl. Phys.*, vol. 21, pp. 1297–1283; December, 1950.) Radiation patterns were measured for five rods of diameter d equals $0.87\lambda_0$ and lengths $2\lambda_0$ to $10\lambda_0$. The theoretical radiation patterns were computed by means of equivalent-surface electric and magnetic currents. Excellent agreement is obtained when it is assumed that the diameter of the surface on which these currents are distributed is $0.65d$. Representative experimental and theoretical patterns are shown for the full 360°.

621.396.67:621.397.6 1076
Temporary Vision Aerial—F. D. Bolt. (*Wireless World*, vol. 57, pp. 13–14; January, 1951.) A brief description of a temporary folded dipole used for radiating the vision signal from the London television transmitter (frequency, 45 mc; mean power, 7.5 kw) while the permanent system was being overhauled. Two coaxial feeders, about 400 feet long, one being a half wavelength longer than the other, were used as a "binocular pair" to energize the antenna. A central element, of *Al* tubing, surrounded by three driven elenents, gave a driving-point impedance of $168\Omega \pm 10$ per cent over the frequency band 42 to 48 mc. Test transmissions produced pictures nearly as good as those from the original antenna.

621.396.67.029.64 1077
Microwave-Aerial Radiation Patterns—D. G. Kiely and A. E. Collins. (*Wireless Eng.* vol. 28, no. 328, pp. 23–29; January, 1951.) The errors involved in measuring the side-lobe amplitudes of 3-cm fan-beam antennas are discussed. Site errors and instrumental errors

are considered in detail. Experimental evidence shows that the largest errors are caused by re-radiation from trees, hedges, and similar obstacles near the measurement sites. For two typical field sites, when three slightly different frequencies were used, the standard deviation of site errors in measurements of side lobes in the radiation patterns of cheese antennas was 1.4 db. The design of antennas to conform to specifications for slide-lobe magnitude must make allowance for such errors.

621.396.677 1078

An Achromatic Microwave Antenna—N. I. Korman and J. R. Ford. (PROC. I.R.E., vol. 38, pp. 1455–1456; December, 1950.) Comment on 1345 of 1950 (Ruze).

621.396.677:538.56 1079

Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates—Berz. (*See* 1138.)

621.396.677:621.396.932.1 1080

Cheese Aerials for Marine Navigational Radar—D. G. Kiely, A. E. Collins, and G. S. Evans. (*Proc. IEE* (London), Part III, vol. 98, pp. 37–45; January, 1951.) A consideration of the problem of the suppression of side lobes in the horizontal radiation pattern of fan-beam antennas. The theory of the primary feed requirements of cheese and half-cheese antennas is outlined and the effect of the secondary radiator on the radiation of the primary feed, and hence on the distribution of energy across the aperture, is analyzed. The side lobes of a symmetrical cheese antenna cannot be reduced much more than 23 db below the main-beam level. The design and performance of a half-cheese antenna having side lobes 30 db below the main-beam level are described.

Book

621.396.67 1081

Antenna Theory and Design, Vols. 1 and 2 [Book Review]—H. P. Williams. Publishers: Pitman and Sons, London, 1950, 142 pp. and 522 pp., 21s. and 63s. (*Jour. Brit. IRE*, vol. 10, p. vii; December, 1950.) The first volume, which is independent of the second, deals with the theory of antenna. The treatment is necessarily mathematical, but numerous diagrams are used in illustration of the theory. The second volume is concerned primarily with antenna design. All types are considered and many curves, diagrams, and photographs are given. A full bibliography is included in both volumes.

CIRCUIT AND CIRCUIT ELEMENTS

537.312.6:621.315.592† 1082

Thermistors—Please note that the above UDC number is now used for thermistors in place of 621.316.86 used hitherto.

537.312.6:621.315.592† 1083

A Relationship between Resistance and Temperature of Thermistors—G. Bosson, F. Gutmann, and L. M. Simmons. (*Jour. Appl. Phys.*, vol. 21, pp. 1267–1268; December, 1950.) The equation $\log R = A + B/(T + \Theta)$ for the resistance R of a thermistor at T°K is proposed. Least-square analyses of the most precise resistance temperature data available for three different thermistor materials show that the equation is a considerable improvement over its predecessors; the standard relative errors of fit are 0.4 per cent, 0.17 per cent, and 0.91 per cent.

621.3.011.3/.4:621.3.012.3 1084

Reactance Chart—H. A. Wheeler. (PROC. I.R.E., vol. 38, pp. 1392–1397; December, 1950.) An extension of the range of the normal chart involving reactance, frequency, inductance, and capacitance, with added scales for susceptance, wavelength, and time constant. Simple geometrical patterns are given which enable the chart to be used for the direct solution of problems such as the bandwidth of a resonant circuit, wave impedance, and delay of a transmission line, and design of constant-k filter half sections.

621.3.015.7†:621.3.087 1085

A Stable Ninety-Nine Channel Pulse-Amplitude Analyser for Slow Counting—D. H. Wilkinson. (*Proc. Camb. Phil. Soc.*, vol. 46, Part 3, pp. 508–518, July, 1950.) The input pulse is converted into a series of pulses of standard height, their number being proportional to the amplitude of the input pulse. Registration is then done by passing the series of pulses into two decade counting rings in cascade; the first ring passing a pulse to the second ring each time it has itself made a complete circuit. Attached to each element of the rings is a telephone message register which operates only when the pulse counting stops at the associated element. The analyzer accepts pulses at a maximum rate of 10 per second. Its calibration is linear to within 1 per cent, and independent of pulse shape, provided the pulse rise time is between 1 and 50 μs.

621.3.016.352† 1086

Relation of Nyquist Diagram to Pole-Zero Plots in the Complex Frequency Plane—W. W. Harman. (PROC. I.R.E., vol. 38, pp. 1454–1455; December, 1950.) Two criteria for the stability of a closed-loop system consisting of an amplifier of gain A with fractional feedback β are: (a) the complex gain A to $(1-A\beta)$ must not have a pole in the right half of the complex frequency plane; (b) the Nyquist polar plot of the loop gain $A\beta$ must not enclose the point (1, 0). The equivalence of these conditions is demonstrated by reference to a 4-stage RC amplifier.

621.314.2.029.3:621.3.018.78† 1087

The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers—K. A. Macfadyen. (*Proc. IEE* (London), Part II, vol. 97, pp. 809–811; December, 1950.) "In the design of transformers, the prediction of the amount of waveform distortion from a knowledge of the properties of the core mterial becomes complicated if the preceding circuit has a high impedance. A theorem is given showing that, if suitable distortion and other measurements are made on a sample core of the material to be used and the distortion factor plotted as a function of two suitably chosen variables, then this information is sufficient to enable the distortion arising from any transformer in any resistive circuit to be calculated exactly."

621.314.22:621.395.641.1 1088

Double-Humped Resonance Curves of Repeaters—G. Schmitt and H. Schrag. (*Fernmeldetech. Z.*, vol. 3, pp. 422–427; November, 1950.) Experiment shows that the two resonance peaks in the output from the divided secondary winding of a repeater are due to the different winding capacitances of the two windings w_2 and w_3. One peak, is eliminated by shunting w_2 with a capacitor, connecting the outer end of w_2 to the inner end of w_3 and earthing the outer end of w_3. Theory of the method is outlined.

621.314.224 1089

Dielectric Admittances in Current Transformers—A. H. M. Arnold. (*Proc. IEE* (London), Part II, vol. 97, pp. 727–732. Discussion, pp. 733–734; December, 1950.) For a current transformer with appreciable capacitance between windings, five ratios may be defined without knowledge of the external circuits, but only two of these ratios can be measured directly with a simple bridge. Formulas are developed for evaluation of all the ratios from simple bridge measurements, and the formulas are tested experimentally. A simple test is described for determining quickly whether the effect of capacitance is appreciable in a transformer of unknown characteristics, and methods of construction of low-capacitance transformers are described.

621.314.224 1090

The Effect of Capacitance on the Design of Toroidal Current-Transformers—A. H. M. Arnold. (*Proc. IEE* (London), Part II, vol. 97, pp. 797–809; December, 1950.) For high-accuracy operation at currents below 5 amps or frequencies above 50 cs, transformer performance is dependent not only on the magnitude of the self-capacitances of the windings and their mutual capacitance, but also on the distribution of the capacitance between windings. The normal method of layer winding gives an unsatisfactory concentrated distribution, but a form of bank winding is described which gives a symmetrical distributed capacitance between windings, resulting in improved performance. If the thickness of the insulation between windings is increased to about one eighth of the mean depth of the secondary winding and an insulant of low permittivity is used, further improvement in performance is obtained. The performance of multi-ratio transformers is considered and good winding arrangements are suggested.

621.314.3† 1091

Feedback in Magnetic Amplifiers—A. S. Fitzgerald. (*Elec. Commun.* (London), vol. 27, pp. 298–319; December, 1950.) Reprint. *See* 2731 of 1949.

621.314.3† 1092

Magnetic-Amplifier Characteristics—L. A. Finzi and D. C. Beaumariage. (*Elec. Eng.* vol. 69, pp. 1109–1115; December, 1950.) Essential text of American Institute of Electrical Engineers 1950 Summer and Pacific General Meeting paper. Dimensionless curves are presented for the determination of the steady-state output characteristics of magnetic amplifiers in terms of design parameters. Correction factors for the feedback ratio are calculated. Results obtained by this method agree well with experimental results for amplifiers with cores of gradually varying incremental permeability.

621.314.3† 1093

A Theoretical and Experimental Study of the Series-Connected Magnetic Amplifier—H. M. Gale and P. D. Atkinson. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 302–303; November, 1950.) Discussion of paper abstracted in 2729 of 1949.

621.314.3† 1094

Magnetic Amplifiers—A. G. Milnes. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 302–303; November, 1950.) Discussion of paper abstracted in 2728 of 1949.

621.314.3† 1095

A Theory of the Series Transductor—C. S. Hudson. (*Proc. IEE* (London), Part II, vol. 97, pp. 751–755; December, 1950.) "The theory of the series transductor with natural excitation is discussed and general equations for load and control current are derived. The effect of phase shift between the supply voltage and that across the transductor is taken into account, and the transient changes which occur at the moment of transductor cut off are explained. The effect on the transductor characteristic of the shape of the B/H curve for the core material is considered, and the analysis is extended to enable an experimental B/H curve to be used in the derivation of the transductor characteristic. It is shown that, with 100 per cent self excitation, the transductor characteristic approximates in shape to the upper part of the B/H curve."

621.318.4 1096

Design of Iron-Cored Inductances carrying D. C.—N. H. Crowhurst. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 516–523; December, 1950.) A series of charts based on the properties of typical transformer iron permit the rapid design of inductors for many purposes. From the dc polarizing current and the required inductance and resistance, the optimum core size, number of turns and width of air gap may be readily determined. Data are tabulated for laminations of stock sizes.

621.318.423 1097
A 100-Kilowatt Water-Cooled Solenoid—J. M. Daniels. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1028–1034; 1st December, 1950.) Design data and construction details for an air-cored solenoid, producing a field of 14.7 kilogauss, uniform to within 0.5 per cent in a cylindrical volume of length 6 cm and diameter 4 cm.

621.318.423.011.3:621.3.012.3 1098
Inductance Chart for Solenoid Coil—H. A. Wheeler. (Proc. I.R.E., vol. 38, pp. 1398–1400; December, 1950.) A simple chart relates inductance, coil over-all dimensions, and winding density. Logarithmic scales covering several decades suffice for almost all practical applications.

621.318.423.015.33 1099
The Natural Frequencies of Single-Layer Solenoids subjected to Voltage Surges—B. Heller, J. Hlávka, and A. Veverka. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 24, pp. 951–957; 26th November, 1949. In German.) Formulas for the natural frequencies are developed, taking into account the mutual inductance between turns, for the two cases where the far end of the coil is (*a*) earthed or (*b*) free. In case (*a*) the spatial distribution of voltage and current can be represented to a close approximation by harmonic functions whose wavelengths are integral multiples of π, the hyperbolic terms being negligible. In case (*b*) the spatial fundamental frequency varies within wide limits. For values of $\lambda>1$, the fundamental frequency is to a first approximation equal to $\pi/2$, but this does not hold if $\lambda<1$, when the hyperbolic terms must also be taken into account.

621.319.45.025 1100
Study of Alternating-Current Electrolytic Capacitors—W. C. van Geel and A. J. Dekker. (*Philips Res. Rep.*, vol. 5, no. 4, pp. 250–261; August, 1950. In French.) Capacitors consisting of two Al plates coated with Al_2O_3 and immersed in an electrolyte are investigated by analogy with a system comprising two ideal capacitors in series, independently shunted by opposed rectifiers, also in series. The potential differences between the various parts of the actual system are then considered, taking into account the effect of the leakage current. Measurements confirming the theory are reported and oscillograms of voltages and currents are shown.

621.385.3:621.315.592:518.4:621.396.615
+621.396.645 1101
Graphical Analysis of Transistor Characteristics—L. P. Hunter. (Proc. I.R.E., vol. 38, pp. 1387–1391; December, 1950.) Graphical methods are described for determining the required values of circuit components of voltage amplifiers, current amplifiers, and oscillators, from the dc characteristics of the transistors used.

621.392 1102
A Note on the Synthesis of Resistor-Capacitor Networks—C. Belove. (Proc. I.R.E., vol. 38, pp. 1453–1454; December, 1950.) Comment on 1605 of 1950 (Bower and Ordung).

621.392 1103
Wide-Band Two-Phase Networks—H. J. Orchard. (*Wireless Eng.* vol. 28, No. 328, p. 30; January, 1951. Comment on 56 of February (Johannesson). It is suggested that a still better way of treating problems of wide-band networks is to reverse the Landen transformation and work from a suitable hyperbolic function. The advantage of the method is that the larger the bandwidth ratio, the easier does the computation become, only one step being necessary for bandwidths exceeding one decade.

621.392 1104
Wide-Band Two-Phase Networks—W. Saraga. (*Wireless Eng.* vol. 28, no. 328, pp. 30–31; January, 1951.) In recent papers by Darlington (1359 of 1950), Orchard (1356 of 1950), and Saraga (2,736 of 1950) on wide-band networks, elliptic functions are used for obtaining a Tchebycheff approximation to the ideal performance. The expressions given for this approximation differ very much in form, although defining the same mathematical relation. Their equivalence is here demonstrated by reference to certain standard formulas in the theory of elliptic functions and their transformation.

621.392:621.387.424 1105
Auxiliary Electronic Circuits for Geiger Counters—L. Fontes and R. Moret. (*Radio franç.*, pp. 5–15 and 18–24; November and December, 1950.) Description of quench circuits, amplifiers, stabilized hv supply units, scaling-down circuits, including decade systems, and integrating circuits.

621.392.4/.5 1106
Selective Generation, Amplification or Rejection of Alternating Voltages without the aid of Oscillatory Circuits—H. Pieplow. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 6, pp. 1031–1039; 24th December, 1949. In German.) Starting from the conditions which must be satisfied by feedback networks, rules are established for the construction of oscillator, amplifier, or rejector networks not including *LC* circuits. Examination of the way in which different couplings can be used shows that the sharpness of resonance or rejection of such networks tends toward a lower limit which is appreciably higher than in the case of oscillatory circuits and can only be lowered by special means not dependent on the circuit arrangement. It follows that complex circuits are especially not worthwhile when the frequency must be controllable.

621.392.4 1107
A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped *LC* Circuits—A. van Weel. (*Philips Res. Rep.*, vol. 5, pp. 241–249; August, 1950.) The ratio of the bandwidths obtainable in the two cases was calculated and found to depend on the product $\omega Z_0 C$ (ω_0 = resonance frequency; Z_0 = characteristic impedance of line; C = terminal capacitance.) The bandwidth of Lecher lines is always less than that of an *LC* circuit; the longer the line, the less the bandwidth. For a short-circuited $3\lambda/4$ line the bandwidth is at least four times less than for an *LC* circuit. The bandwidths of $\lambda/2$ and $3\lambda/4$ lines can be increased to that of a $\lambda/4$ line by introducing discontinuities in the characteristic impedance at certain points. A qualitative explanation of these effects is given which applies also to cavity resonators.

621.392.41 1108
Impedance Synthesis without Use of Transformers—R. Bott and R. J. Duffin. (*Jour. Appl. Phys.*, vol. 20, p. 816; August, 1949.) A method is outlined for the synthesis of an arbitrary impedance by series-parallel combinations of inductors, resistors, and capacitors.

621.392.41 1109
Impedance Synthesis distributing Available Loss in the Reactance Elements—W. Nijenhuis. (*Philips Res. Rep.*, vol. 5, pp. 288–302, August, 1950.) A study of the synthesis of 2-pole networks whose elements possess maximal losses. Difficulties arising from mutual inductances are overcome in the way suggested by Bott and Duffin (1108 above). It is shown how losses can be introduced into coils to simulate skin effects or eddy currents. A numerical example is worked out.

621.392.5 1110
Influence of Losses on Attenuation of Four-Terminal Impedance Networks—S. Brimberg. (*Kungl. tekn. Högsk. Handl., Stockholm*, no. 43, 30 pp; 1950. In English.) The influence of losses on complex image attenuation in the case of infinite idealized image attenuation and at band edges is calculated. The effect of losses on complex effective attenuation at band edges in symmetrically designed filters with symmetrical loads is also investigated and theoretical and experimental results are compared for a low-pass filter.

621.392.52 1111
Attenuation Equalization within the Pass Band of Ladder Filters—W. Wolman. (*Arch. Elektrotech.*, vol. 40. no. 1, pp. 30–36; 1950.) Analysis shows that the frequency-response curve of each section of a filter can be flattened by introducing small resistances in shunt or in series, effectively forming auxiliary meshes with complementary response curves. Measurements are reported in confirmation of the theory, and the effect of the compensation is illustrated in response curves for a band-pass filter of 6 half sections and for a low-pass filter of 4 sections. Factors affecting the choice of location of the auxiliary resistors, a matter of great practical importance, are discussed.

621.392.52 1112
Unusual Ladder Filter. Applications in Audio and Radio Circuits—F. G. G. Davey. (*Wireless World*, vol. 57, pp. 31–34; January, 1951.) The properties are discussed of a class of filters in which the impedances looking forward and backward from the central element bear a constant ratio to each other with change of frequency and maintain equal phase angles. The application is described of adjustable low-pass filters of this type between the speech coil of a loudspeaker and the corresponding transformer winding, and of band-pass versions for use as variable coupling elements at intermediate frequency.

621.392.5:621.395.44 1113
Loss-Compensated Filters in Carrier-Frequency Systems—H. Lehmann. (*Fernmeldetech. Z.*, vol. 3, pp. 415–417; November, 1950.) Discussion of the advantages of using asymmetric filters in ssb carrier systems, particularly in eliminating the need for lf equalizers, in affording a choice of design, and in improving stability. The discussion is based on experimental results for channel filters covering the range 12 to 24 kc.

621.396.611.3 1114
Rolling-Ball Analogue of Coupled Circuits—P. C. Magnusson. (*Elec. Eng.*, vol. 69, p. 1079; December, 1950.) Summary of American Institute of Electrical Engineers 1950 Fall General Meeting paper. The system consisting of two inductively coupled *LC* loops with negligible resistance has as dynamic analogue a ball rolling freely on a specially curved surface. The time integrals of the loop currents are used as displacements along horizontal co-ordinates x_1, x_2, with the loop currents themselves analogous to the velocity of the ball along those co-ordinates. The magnetic-field energy is analogous to the kinetic energy of the ball if the scales of x_1 and x_2 are proportional to the square roots of the corresponding self-inductances and x_1 is inclined to x_2 at an angle β, the cosine of which is equal to the coefficient of coupling. The height of any point of the surface, and hence the potential energy of the ball, is proportional to the dielectric field energy, contours of which are concentric ellipses. The ball path is shown which

corresponds to initial circuit conditions where both capacitors are charged and circuit currents are zero. The method of deriving an analogue should be applicable to any dynamically conservative electrical or mechanical system with two degrees of freedom, and should be useful in the interpretation of transients in such systems.

621.396.611.31.001.11 1115

Theory of Oscillations in Coupled Electromagnetic Cavity Resonators—E. Ledinegg and P. Urban. (*Acta Phys. austriaca*, vol. 2, no. 2, pp. 198–213; September, 1948. In German.) General formulas are derived by a perturbation method for the frequencies of a system of cavity resonators coupled by similar resonant members. Except for the assumption of weak coupling between the resonators, which in the undisturbed state must give rise to at least one common frequency, no essential limitations are introduced in the calculations. The theory shows that the frequencies of such coupled systems satisfy a secular determinant, whose members represent energy terms determined by the individual elements of the cavity-resonator system, and which can accordingly be calculated. The theory is applied to the practical cases of (*a*) two concentric transmission-line sections with a coaxial coupling unit, (*b*) two inductively coupled quasistationary circuits.

621.396.615 1116

The Self-Blocking Oscillator and its Applications—J. Moline. (*Radio franç.*, pp. 11–15 and 12–16; December, 1950 and January, 1951.) The development of the blocking oscillator from one having a sinusoidal wave form, by increasing the coupling between the anode and grid circuits, is described and the blocking action is explained; the effects of the different circuit elements being considered separately. Various typical practical circuits are shown and their particular advantages are mentioned. Applications described include pulse and saw-tooth generators, use in counter circuits, and so forth.

621.396.615:621.316.726:621.396.4 1117

Stabilized Master Oscillator for Multichannel Communication—E. W. Pappenfus. (*Electronics*, vol. 23, pp. 108–113; December, 1950.) Particular attention is paid to systems providing a wide range of operating frequencies but utilizing only a single reference crystal. A commercially available oscillator covering the band 2 to 4.5 mc is described. Using a 100 kc reference crystal, the oscillator provides an output of any desired frequency in the above range stabilized to within $\pm 5\times 10^{-8}$ for wide temperature and humidity variations. Details of the servo-motor afc circuit are given and discussed. Any one of ten preset output frequencies can be selected by operation of a remote switch.

621.396.615:621.316.726.078.3 1118

Frequency Stabilization of Oscillators—J. Zakheim. (*Radio franç.*, pp. 1–4; November, 1950.) If the effective output impedance Z of the amplifier section is given by $Z=R+jX$ and the transfer coefficient k of the limiter by $k=\alpha+j\beta$, the phase angle is theoretically constant and independent of the over-all slope of the maintaining amplifier when $\beta\alpha=-XR$. Practical application of the formulas derived is illustrated by numerical calculation for an oscillator of the Wien-bridge type operating at 3 kc.

621.396.615.14.029.63 1119

Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits—Ratheiser. (*See* 1280.)

621.396.615.141.2 1120

Some Aspects of Split-Anode Magnetron Operation—H. J. Reich, J. C. May, J. G. Skalnik, and R. L. Ungvary. (Proc. I.R.E., vol. 38, pp. 1428–1433; December, 1950.) Resonators and tubes at present available are unsuitable for use in wide-band magnetron oscillators using butterfly-type resonators, but they could be more successfully used if specially modified for the purpose. The negative slope of part of the anode-current voltage characteristic, explainable by back heating, is absent when the current is varied rapidly. The best method of anode-current stabilization is that in which the magnetic field is produced by an electromagnet excited by the direct anode current.

621.396.615.17:621.385.832 1121

A Simple Slow-Running Timebase—C. J. Dickinson. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 505–506; December, 1950.) A detailed description of a circuit designed to overcome inherent disadvantages of earlier timebases. The circuit requires less critical adjustment and avoids the need for a large negative supply voltage.

621.396.645:621.3.018.78† 1122

Universal Design Curves for Intermediate-Frequency Amplifiers with Particular Reference to Phase and Amplitude Distortion and their Compensation—W. Hackett. (Proc. I.R.E., vol. 38, pp. 1408–1417; December, 1950.) Graphs and abacs are provided which enable the phase and amplitude distortion to be determined for various types of coupling network, including compensation combinations. Graphical methods are indicated for assessment of performance in terms of the product of gain and bandwidth, and for the interpretation of equation parameters in terms of circuit parameters.

621.396.645.029.3:621.385.3/.4 1123

A Comparison of Triodes and Beam-Tetrodes as Power Output Valves in Audio Amplifiers—D. Bruckmann, W. S. Carey, and D. J. Fuller. (*Trans. S. Afr. Inst. Elect. Eng.*, vol. 41, pp. 258–266. Discussion, pp. 266–267; September, 1950.) Tests made on balanced push-pull amplifiers show that the curve representing the change of harmonic distortion with increase in load resistance is parabolic for beam tetrodes, while for triodes it resembles a rectangular hyperbola. With reactive loading the increase in distortion is greater for the tetrode. As the load resistance rises, the intermodulation distortion for triodes drops to a negligible value, while for tetrodes it increases rapidly. The effects with reactive loading are similar to those observed in the case of harmonic distortion. The relative merits of triodes and tetrodes for use in af amplifiers are summarized.

621.396.645.371 1124

Amplifiers with Selective Negative Feedback—H. Boucke and O. Schmidt. (*Fernmeldetech. Z.*, vol. 3, pp. 417–420; November, 1950.) Basic circuits are described for suppressing either a harmonic or the fundamental in the output of an amplifier by incorporating a resonant circuit in the feedback line. Turning this circuit to the first harmonic reduces the distortion appreciably. A typical attenuation curve for a frequency-multiplier circuit shows a drop of 28 db in output level for the 800 cps fundamental below that obtained with a non selective cathode circuit.

621.396.662.085.4 1125

Bandspreading and Scale Equalization for RC Tuning Networks—C. F. Van L. Weiland. (*TV Eng.*, vol. 1, pp. 12–15, 28–30 and 21, 28; July, August, and October, 1950.) A continuous-selector type of circuit is described. By use of suitable tuning elements in series and parallel, a threefold scale elongation is achieved, with increased spread at the hf end of the scale. Circuits providing partially equalized and completely equalized band-spreading are shown and design formulas given.

621.396.69+621.38]:061.5(058.7) 1126

Alphabetical Listings of All Components, Complete Units, Allied Products, used in Electronic Equipment for All Purposes—(*Electronics, Annual Buyers' Guide Issue*, vol. 23, pp. D1-D164; Mid-month June, 1950.) A list giving the names of more than 2,500 manufacturers of electronic components and equipment. Products are arranged alphabetically under generic names, with subdivisions. A trade-name index and an index of manufacturers are also included.

GENERAL PHYSICS

519.2:621.3.015.2 1127

On a First-Passage-Time Problem—F. L. H. M. Stumpers. (*Philips Res. Rep.*, vol. 5, pp. 270–281; August, 1950.) An investigation of the probability that a function starting at a time $t=0$ with a value E_0 will not exceed a value E_1 in the time interval 0-t_1. This probability is calculated (*a*) from the Fokker-Planck equation with suitable boundary conditions. (*b*) from an integral equation derived by Schrödinger. The theory has application to the pulse charging of a capacitor and also to fluctuation problems.

534.1+538.56]:517.93 1128

On the Forced Vibrations of Quasi-Linear Systems—Friedlander. (*See* 1177.)

534.26+[535.42:538.56 1129

Theory of Diffraction at a Screen—W. Franz. (*Z. Phys.*, vol. 128, pp. 432–441; 3rd October, 1950.) The author's successive-approximation method for diffraction calculations (3118 of 1949 and 323 of March) yields, for sharp-edged screens, functions which are singular at the edge and in the higher approximations give rise to divergent integrals. The singularities can be eliminated and the method made useful for higher approximations by assuming the integrands of the Kirchhoff integrals to fall to zero gradually, instead of discontinuously, beyond the edge. *See* also 2183 of 1950 (Braunbek).

535.42 1130

A Note on the Diffraction of a Cylindrical Wave by a Perfectly Conducting Half-Plane—P. C. Clemmow. (*Quart. Jour. Mech. Appl. Math.*, vol. 3, pp. 377–384; September, 1950.) The most easily derived solutions in diffraction theory are for the case of incident plane waves. The solution for a line source can then be obtained by a further integration. This is illustrated by the example of a line source in the presence of an infinitely thin, perfectly conducting half plane. The introduction of Hankel functions is avoided, and the solution appears in a useful form, equivalent to that given by Macdonald and analogous to Sommerfeld's solution for an incident plane wave.

537.52 1131

Retroaction and Similitude Considerations in Relation to the Starting of the Electrical Gas Discharge—W. Fucks. (*Arch. Elektrotech.*, vol. 40, no. 1, pp. 16–30; 1950.)

537.523.5:621.317.33.029.64 1132

A Microwave Study of the High-Pressure Arc—J. D. Cobine, E. P. Cleary, and W. C. Gray. (*Jour. Appl. Phys.*, vol. 21, pp. 1264–1267; December, 1950.) The Impedance of an arc at atmospheric pressure was measured at 1 kmc by means of a coaxial line and standing-wave detector. The arc reactance and resistance, essentially a lumped load at the end of the line, increase with arc length. The resistance decreases with increasing current and is approximately the same as the dc resistance. The reactance is capacitive and nearly independent of the current. Arcs in air, argon, and helium, with direct currents of 1 to 4 amps, were studied.

537.525:621.396.822 1133
Noise Temperature of a D.C. Gas-Discharge Plasma—S. Kojima and K. Takayama. (*Phys. Rev.*, vol. 80, no. 5, p. 907; December 1, 1950.) At 14 mc, the temperature of the plasma noise of discharge tubes containing argon at pressures of 1 mm and 3 mm Hg is found to be proportional to the electron temperature. The ratio of plasma-noise temperature to electron temperature is approximately 4.5:1 at 1 mm and 2:1 at 3 mm Hg. The noise temperature is nearly the same at 600 kc as at 14 mc, indicating that the noise observed is mainly thermal.

537.525.029.64:538.69 1134
The Effect of Magnetic Field on the Breakdown of Gases at Microwave Frequencies—B. Lax, W. P. Allis, and S. C. Brown. (*Jour. Appl. Phys.*, vol. 21, pp. 1297–1304; December, 1950.) Results are given of measurements of the electric field required to produce hf breakdown in helium at low pressure in the presence of a magnetic field of 0 to 3,000 gauss. Two relevant theories, the average electron theory and the Boltzmann theory, are presented, and the correspondence between them is discussed.

537.525.5 1135
Techniques for Measuring the Dynamic Characteristics of a Low-Pressure Discharge—B. T. Barnes and S. Eros. (*Jour. Appl. Phys.*, vol. 21, pp. 1275–1278; December, 1950.)

537.527 1136
The Minimum Voltage and the Discharge Process on Flashover with Alternating Voltage in Humid Compressed Air—H. Böcker. (*Arch. Elektrotech.*, vol. 40, no. 1, pp. 37–44; 1950.) Measurements were made of the Flashover voltage in air at pressures up to 24 atm with water-vapor content close to saturation, cylindrical glass insulators 0.6 to 3 cm long being inserted between the dished electrodes of the plane parallel gap. The lowest value of the reduced flashover voltage observed in the region of saturation is the "minimum voltage." Results are shown graphically and discussed.

538.3 1137
Wheeler and Feynman's Theory of Electromagnetic Interaction—O. Costa de Beauregard (*Rev. Sci.* (Paris) vol. 88, no. 3305, pp. 34–40; January to March, 1950.) Analysis in the light of other theories leads to criticism on the grounds that conditions for the irreversibility of the radiation process are contained in the postulates of the theory.

538.56:621.396.677 1138
Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates—F. Berz. (*Proc. IEE* (London), Part III, vol. 98, no. 51, pp. 47–55; January, 1951.) A mathematical analysis of the problem based on Maxwell's equations and on periodicity and continuity considerations. "Formulas are obtained giving the nature, direction, phase, and amplitude of the reflected and transmitted waves. The reflected waves follow the usual laws found in the grating theory. In the case when only one non-evanescent plane wave is reflected, the transmission and reflection power coefficients are those obtained at the junction of two semi-infinite transmission lines corresponding to the free space and the plate medium respectively. The analogy with transmission lines is more complex when two non-evanescent waves are reflected. For the usual angles of incidence and plate spacings, the transmission power coefficient is high (maximum 99.5 per cent), and the phase shift on transmission is small."

538.561 1139
Radio-Frequency Emission due to the Gyromagnetic Effect in a Discharge—M. Laffineur and C. Pecker. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231. no. 25, pp. 1446–1447; December 18, 1950.) The source of electrons used was a Penning vacuum gauge (1525 of 1937) with two cold cathodes, one on each side of a common ring-shaped anode. This was evacuated, placed in the field of an electromagnet producing fields up to 300 gauss, and a voltage of 2 kv was applied. The em waves emitted were detected by means of a receiver tuned to a wavelength of 55 cm and connected to a pickup loop at the side of the apparatus. The best results were obtained at a pressure of 10^4 mm Hg. A curve shows the voltage at the receiver terminals, less the background noise (equivalent to about 50 μv at the input), for fields from 184.5 to 194.5 gauss. This curve has a very sharp peak at 193.5 gauss, the corresponding gyromagnetic frequency being precisely that for which the receiver was tuned. A much smaller peak at about 190 gauss is attributed to an image frequency of the receiving apparatus.

549.321.13:537.311.3:537.533.9 1140
Equilibrium Currents induced in Zinc Blende by Electron Bombardment of Negative Electrode—M. F. Distad. (*Phys. Rev.*, vol. 80, no. 5, pp. 879–886; December 1, 1950.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7:538.122 1141
Visual Investigation of the General Magnetic Field of the Sun—G. Thiessen. (*Z. Astrophys.*, vol. 26, no. 1, pp. 16–30; July 31, 1949.) The method of measurement and the results obtained during 1947 to 1948 with the 60-cm refractor of the Hamburg observatory are described. Measurements on various *Cr* and *Fe* spectral lines show that at the time of the observations the pole field strength in the direction given by Hale, could not have a value >5 gauss; the results rather indicate a very weak field in the reverse direction.

523.7+523.854]:621.396.822 1142
Electromagnetic Waves radiated by Fast Protons in Strong Magnetic Fields, and Correlation between Cosmic Radiation and Radio Noise from the Sun and the Galaxy—B. Kwal. *Compt. Rend. Acad. Sci.* (Paris), vol. 231, no. 20, pp. 1057–1059; November 13, 1950.) The classical theory of radiation from a charged particle in a magnetic field shows that, in fields of the order of 10^3 to 10^4 gauss, protons of energy 10^9 to 10^{10} ev radiate meter and centimeter waves. Hence it seems possible to associate radio noise with the acceleration of cosmic rays in sunspots and in similar spots occurring in stars.

523.752:[551.510.535+537.591 1143
The Chromospheric Flare of 19th November 1949 and its Geophysical Consequences—R. Bureau and A. Dauvillier. (*Ann. Géophys.*, vol. 6, no. 2, pp. 77–103; April to June, 1950.) Close correlation was observed between ionospheric and cosmic-radiation effects associated with this flare. The rarity of such a correlation is emphasized. The two authors describe respectively their detailed investigations of the two effects. Observations from various parts of the world are reported and discussed, especially in relation to possible explanations of the delay of the corpuscular effects with respect to the electromagnetic effects. See also 1911 (Ellison and Conway), 2206 (Müller), and 3039 (Clay and Jongen) of 1950.

550.362:621.317 1144
Theoretical Basis of a Method for Determining the Specific A. C. Resistivity of the Ground—Weber. (*See* 1193.)

550.386/.387]:621.331:625.1 1145
Disturbances of Terrestrial Magnetic Field and Currents by Electrified Railways—G. Dupouy. (*Ann. Géophys.*, vol. 6, pp. 18–50; January to March, 1950.) Disturbances observed in the magnetograms recorded at Chambon la Forêt have been found to be caused by the Paris-Orléans line 28 km away. The effect is particularly noticeable when the continuity of the electric cables is broken, either accidentally or for purposes of maintenance. For horizontally stratified ground the disturbances can be calculated as a function of the leakage currents along the track. Agreement with measurements is satisfactory.

551.510.534/.535 1146
Stratospheric-Ionospheric Relationships—N. C. Gerson. (Proc. I.R.E., vol. 38, p. 1456; December, 1950.) Investigations reveal no correlation between either the temperature or pressure at the tropopause and the critical frequency of any of the ionospheric regions, and also no correlation between temperatures at a height of 13 km and critical frequencies. Some slight correlation was found in several instances between the pressure at a height of 13 km and the critical frequencies of the E and F_2 regions, but no correlation with sporadic E conditions.

551.510.535 1147
Magneto-Ionic Triple Splitting of Ionospheric Waves—O. E. H. Rydbeck. (*Jour. Appl. Phys.*, vol. 21, pp. 1205–1214; December, 1950.) A full theoretical discussion of the third magneto-ionic component, or z trace, which has been observed frequently at the new Kiruna observatory. Strong coupling often exists between the ordinary and z components at a critical level. The z wave becomes strongly excited when the collisonal frequency ν approaches the critical value ν_c of Appleton and Builder. When ν is larger than ν_c, the ordinary echo rapidly disappears and only the z and extraordinary components remain. A useful expression for the oz wave transmission coefficient is derived. Transmission of appreciable strength is possible only at high latitudes, where ν_c is fairly small. Experimental results obtained at Kiruna (magnetic latitude 65°) are reported.

551.510.535 1148
Preliminary Results of Ionospheric Observations made at Dakar—F. Delobeau, E. Harnischmacher, and F. Oboril. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 17–20; January to March, 1950.) Discussion of records obtained during the period May to September, 1949. Monthly mean values of f_0F_2 conform to the general shape of curves plotted for comparable latitudes, but extreme values are more pronounced: the highest critical frequencies in May were estimated at about 15 mc; median midday values for June and July were 13.8–13.5 mc and the night value for June was 4.2 mc. This lends support for the revision of zone demarcation proposed by Oboril and Rawer (2794 of 1949). Strong diffusion effects observed cannot be correlated with magnetic disturbances. The presence of an intermediate layer between F_1 and F_2 is indicated; the true height of the F_2 layer deduced from virtual-height measurements may reach 500 km. Values of k and n in the formula $f_0E=k(\cos\chi)^n$ are tabulated. Records of E_s layer characteristics are also shown and discussed.

551.510.535:523.745 1149
Long-Term Relation between F_2-Layer Ionization and Solar Activity, over the Whole World—R. Gallet and K. Rawer. (*Ann. Géophys.*, vol. 6, nos. 2 and 3, pp. 104–114 and 212–213; April to June and July to September, 1950.) Close correlation is found between the monthly mean value of the F_2 layer maximum electron density and the relative sunspot number, on comparing the running means over 13 months. The coefficients of the linear correlation equations for 12 stations in different parts of the world are plotted against, (*a*) geographic latitude, (*b*) geomagnetic latitude, and (*c*) magnetic inclination; the last method gives the most satisfactory picture of their systematic variation. Abnormal relations are observed at the magnetic equator; observations from stations in that region are reserved for separate study.

551.510.535:551.594.5 1150
Ionospheric Disturbance caused by the Aurora Borealis of 20th–21st February 1950—R. Eyfrig. (*Ann. Géophys.*, vol 6, pp. 70–73; January to March, 1950.) A direct observation of an aurora borealis, relatively rare in latitudes as low as 48°, was made at Freiburg, starting at 1845 universal time on February 20. Unusually high electron concentration in the F_2 layer preceded and accompanied the observation. Further abnormalities of the *E* and *F* layers are shown in a series of h'f records for the period from 1830 to 2200. The disturbance is interpreted as a horizontally travelling dip in the ionosphere. f_0F_2 was low throughout the night and until the next afternoon. A marked ionospheric disturbance was also observed on the antarctic on board the *Commandant Charcot* on February 20.

551.510.535:551.594.51 1151
New Results of Importance in the Study of Auroral Spectra and the Physics of the Ionosphere—L. Vegard. (*Ann. Géophys.*, vol. 6, pp. 157–163; July to September, 1950.) Results more complete and more precise than ever before have been obtained at Oslo, using a new spectrograph for analysis of the aurora of the 23rd to the 24th of February, 1950. Fifty-four new bands or lines were observed. The presence of molecular oxygen was proved for the first time. Many lines are due to nitrogen and oxygen, neutral and ionized, as previously observed. The temperature of the luminous region was estimated to be −54° centigrade.

551.510.535:621.3.087.4 1152
An Automatic Ionospheric Recorder for the Frequency Range 0.55 to 17 Mc/s—Naismith and Bailey. (*See* 1191.)

551.594.11 (988) 1153
Measurements of the Electric Field of the Atmosphere in Greenland between Sea Level and the Centre of the Ice-Cap—P. Pluvinage and G. Taylor. (*Ann. Géophys.*, vol. 6, p. 69; January to March 1950.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:526.9 1154
The Application of Radar to Geodetic Surveying—J. Warner. (*Aust. Jour. Appl. Sci.*, vol. 1, pp. 133–146; June, 1950.) A modified shoran radar equipment is used for measuring distances between 160 and 310 miles to an accuracy within about 7 parts in 10^5. The method of measurement, equipment, used reduction of observations, and errors involved are described. The greatest errors were instrumental; it is considered that modifications of the equipment would reduce them to about 2 parts in 10^5.

621.396.9:551.578.1/.4 1155
Radar Echoes from Meteorological Precipitation—J. E. N. Hooper and A. A. Kippax. (*Proc. IEE* (London), Part 1, vol. 97, no. 108, pp. 303–394; November, 1950.) Discussion of paper abstracted in 2222 of 1950.

621.396.9:621.396.11 1156
Some Adverse Influences of Meteorological Factors on Marine Navigational Radar—Saxton and Hopkins. (*See* 1228.)

621.396.932.1:621.396.677 1157
Cheese Aerials for Marine Navigational Radar—Kiely, Collins, and Evans. (*See* 1080.)

621.396.933.2 1158
Rho-Theta System of Air Navigation—P. C. Sandretto. (*Elec. Commun.* (London), vol. 27, pp. 268–276; December, 1950.) Description of equipment designed to fulfil I.C.A.O. recommendations for short-range navigational aids. Azimuth (theta) is obtained by reception of signals from a ground vhf transmitter radiating (*a*) a carrier with 30 cps tone modulation from an antenna having a circular horizontal radiation pattern and (*b*) an unmodulated carrier of the same frequency radiated from a dipole rotating at 1,800 rpm. The phase relations of the two carriers remain rigidly fixed. Azimuth is obtained from the measurement of the phase difference between the tone from the rotating system and that from the omnidirectional system. Distance (rho) is determined from the time taken by a pulse radiated from the aircraft at uhf to travel to a ground station and be returned, at a different uhf, to the aircraft.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.12.088 1159
Reading Errors with the McLeod Gauge—P. Romann. (*Le Vide*, vol. 3, pp. 522 530; November, 1948.)

533.56 1160
Study and Realization of a New Rotary Molecular Vacuum Pump—H. Gondet. (*Le Vide*, vol. 3, pp. 513–521; November, 1948.) Description of a pump of the Holweck type in which a disk is used instead of the usual drum, so that an increased output is obtained. The optimum dimensions for the various channels are calculated and pumping rates are given for an experimental pump with a duralumin disk, of diameter 40 cm and thickness 1.2 cm, rotating between two duralumin castings with spiral channels of cross section rapidly decreasing toward the center. Clearance between the disk faces and the partitions separating the channels is of the order of 0.03 mm. The driving motor is fitted within the part of the machine connected to the backing pump.

537.311.33:546.289 1161
Pressure Dependence of Resistance of Germanium—J. H. Taylor. (*Phys. Rev.*, vol. 80, pp. 919–920; December, 1950.) The application of hydrostatic pressure increases the width of the forbidden band and hence increases the resistivity. Measurements, at pressures up to 4,500 pounds per inch², of the change of resistivity with pressure for several samples leads to the value $10.2 \pm 0.4 \times 10^5$ atm for the pressure coeffi ient of resistance at 300° K. The temperature rate of change of width of the forbidden band then follows as 0.87×10^4 ev° Centigrade.

537.311.33:546.289:539.164.9 1162
Changes in Conductivity of Germanium Induced by Alpha-Particle Bombardment—W. H. Brattain and G. L. Pearson. (*Phys. Rev.*, vol. 80, pp. 846–850; December 1, 1950.) The bombardment of *n*-type Ge by α particles first removes the conduction electrons at the rate of 78 per α particle. After these electrons are gone, conducting holes are introduced at the initial rate of 8.6 per α particle. There are 6.6 of these conducting holes that decay with a half life of 13 hours at room temperature and the remaining two holes are permanent at this temperature. Conductivity is changed only to the depth of penetration of the particles, namely 1.9×10^3 cm. The distribution of holes with depth is not uniform.

537.312.5:546.46.86 1163
Photoconductivity in Magnesium Antimonide Layers—T. S. Moss. (*Proc. Phys. Soc.*, (London), vol. 63, no. 372B, pp. 982–989; December 1, 1950.)(Layers of Mg_3Sb_2 evaporated in vacuo act as semiconductors and are photoconductive. Current to voltage and radiation intensity to voltage relations, response time, spectral dependence of photosensitivity, and the resistance to temperature relation were investigated. Threshold wavelengths of 1.5μ; 2.6μ and 3.5μ were found, the first corresponding to intrinsic conductivity (ϵ=0.81ev) and the latter two (ϵ=0.48ev and 0.35ev) to impurity levels. Recent Russian work is in agreement with these results [*Zh. tekh. Fiz.*, 1949 vol. 18, pp. 1459–1477 (Boltaks and Zhuze)].

537.312.8:669.15 1164
A Study of the Magneto-Resistance of Silicon-Iron—R. Parker. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 996–1004; December 1, 1950.) Curves for $\Delta\rho,\rho$ against field strength were obtained for single crystals and polycrystalline specimens, for both longitudinal and transverse magnetization. Strong fields give a linear decrease less than 0.02 times that for Ni and Co. Weak fields give unusual features. It is concluded that the small magneto resistance of polycrystalline silicon iron is due to the small contribution of individual crystallites and not to an averaging-out process.

537.312.8.029.63:546.87 1165
Magnetoresistance of Bismuth at 3,000 Mc/s—C. W. Heaps. (*Phys. Rev.*, vol. 80, no. 5, pp. 892–893; December 1, 1950.) Using a resonant re-entrant cavity machined from cast Bi, fed through a slotted coaxial waveguide, swr measurements indicated that at 3 kmc the magnetoresistance of Bi is not more than half the dc value. This is possibly due to the skin depth being of the same order as the mean free path of electrons in the metal.

538.221 1166
Oxide Ferromagnetic Materials—E. Flegler. (*Arch. Elektrotech.*, vol. 40, pp. 4–16; 1950.) An account is given of the experimental investigation of various ν iron oxides and ferrites subjected to alternating fields. The variations with frequency of the conductivity, permittivity, permeability, and loss factors are shown and conclusions are drawn regarding the causes of these variations. The different uses of the term "complex permeability" are briefly discussed.

538.221 1167
High-Frequency Permeability of Ferromagnetic Materials—F. V. Webster and K. S. Driver. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, p. 1040; December 1, 1950.) Preliminary results obtained with more accurate apparatus than that previously used [913 of April (Millership and Webster)] are reported. The general conclusions remain unaffected.

538.221:538.214 1168
Investigations on the Reversible Susceptibility of Ferromagnetics—R. S. Tebble and W. D. Corner. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1005–1016; December 1, 1950.) The reversible susceptibility κ_r of long wire specimens was measured by applying small alternating fields and using a mutual-inducto-meter bridge. Results are discussed. The minimum contribution of reversible processes to the total change in magnetization varies from 10 per cent to 20 per cent for iron and nickel specimens. Gans' statement, that κ_r is a unique function of intensity of magnetization and independent of magnetic history has definite limitations.

549.514.51:621.396.611.21 1169
Quartz Crystals free from Harmonics—J. J. Vormer. (*Tijdschr. ned. Radiogenool.*, vol. 15, pp. 339–346; November, 1950.) Overtones of Y' oscillations in X-cut crystals can be suppressed by shaping the width of the electrodes to a sine curve; this has the effect that the areas subjected to opposed fields are exactly equal for any harmonic. Experiments made to test the theory indicated that the intensity of all harmonics could certainly be reduced to less than 1 per cent of the fundamental. Alternatively, a field distributed sinusoidally in the Y direction may be used with ordinary electrodes to achieve the same effect.

621.314.634:537.311.33 1170
Forming Processes in Thallium-Containing Selenium Rectifiers—A. Hoffmann. (*Z. Phys.*) vol. 128, no. 3. pp. 414–431; October 3, 1950. A new method of measuring the capacitance of a dry rectifier, due to Lehovec, makes it pos-

sible to estimate the concentration of impurity centers at different production stages. For a certain Tl content of the counterelectrode there is (a) a reduction of concentration of impurity centers in the boundary zone, (b) an increase of path resistance, and (c) an increase of blocking power; (a) and (b) are quantitatively related. The influence of the Tl is markedly dependent on whether the counterelectrode is absent or present during the final heat treatment at 210° centigrade. The increase of blocking power cannot be explained simply by the reduction of the boundary field strength, but must be mainly due to other causes.

621.318.2 1171

Calculation and Construction of Permanent-Magnet Systems—E. Steinort. (*Radio Tech.* (Vienna), vol. 26, pp. 545–548; November, 1950.) The design affording maximum flux in the air gap with minimum use of material is discussed. Formulas are given for optimum length and cross section of the magnet, the correction factor for length and the leakage flux as a function of air-gap inductance are shown graphically for ticonal and alnico materials. When the leakage characteristic of the material is not known, optimum length and cross section may be determined by an experimental method using an electromagnet.

621.791.3 1172

Solders and Soldering: Some Recent Advances—H. C. Watkins. (*Metallurgia*, (Manchester) vol. 42, no. 254, pp. 372–376; December, 1950.) Survey of materials and techniques, with special reference to wiping solders, the soldering of aluminum, special solders for automatic processing, and flux-cored solders. The physical effects of surface roughening and of absorption of atoms of the liquid metal are briefly considered.

666.1.037.5 1173

Glass-to-Metal Seals—J. A. Pask. (*Prod. Eng.*, vol. 21, pp. 129–134; January, 1950.) A discussion of the properties of various metals and special alloys used for sealing to glass. Only direct seals, with no intermediate material other than the metal oxide, are considered. Thermal expansion curves are given and the stresses arising from imperfect matching between metal and glass are analyzed. Other important factors are the electrical and thermal conductivities of the metal, and the thickness of the oxide film allowed to form on its surface.

621.317.2:537.72 1174

Techniques Générales du Laboratoire de Physique: Vol. II [Book Review]—J. Surugue (Ed.). Publishers: Centre National de la Recherche Scientifique, Paris. and H. K. Lewis and Co., London, 336 pp., paper cover 42s., cloth 44s. (*Jour. Sci. Instr.*, vol. 27, p. 341; December, 1950.) Nine sections, contributed by different authors, deal respectively with: measurement of infrared radiation; technique of thin quartz fibers and Wollaston wires; casting thin cellulose films; electrometers; electrometer tubes and circuits; ionization counters; investigation of fields by means of electrolyte tanks; "micro-forge"; automatic regulation of temperature.

MATHEMATICS

512.831:518.6:681.177 1175

Calculation of the Eigenvalues of Matrices by means of Punched-Card Machines—P. Henrici. (*Z. angew. Math. Phys.*, vol. 1, pp. 185–189; May 15, 1950.)

517.9 1176

Natural Eigenvalue Problems—E. Stiefel and H. Ziegler. (*Z. ang w. Math. Phys.*, Vol. 1, pp. 111–138; March 15, 1950. Eigenvalue problems, rising from questions of stability or in connection with oscillations may be approached (a) as problems of the calculus of variations, (b) by establishing their differential equations, (c) by means of their integral or integrodifferential equations. Up to the present, method (b) has been developed further than the others, but it has the disadvantage that it is not applicable to problems whose eigenvalue is explicitly involved in the boundary conditions. To overcome this difficulty, it is advisable to use method (a) which, from a physical point of view, has the advantage of facilitating the mathematical interpretation of mechanical restrictions. The authors intend to generalize the theory of eigenvalues and to establish relations between the three methods for the general case of so-called "natural" problems. In the present paper the relation between methods (a) and (b) is established for the case of one independent and one dependent variable.

517.93:[534.1+538.56 1177

On the Forced Vibrations of Quasi-Linear Systems—F. G. Friedlander. (*Quart. Appl. Math.*, vol. 3, part 3, pp. 364–376; September, 1950.) Discussion of the differential equation: $\ddot{x}+\omega^2x=kf(x, \dot{x}, t)$ where ω, k are parameters, k being small, and $f(x, \dot{x}, t)$ is periodic with period 2π in t. Using an approximate solution due to Kryloff and Bogoliuboff, it is shown that, if $f(x, \dot{x}, t)$ is a polynomial in x, $\dot{x}$, $\cos t$, and $\sin t$, the amplitude $(x^2+\dot{x}^2/\omega^2)^{\frac{1}{2}}$ settles down to an asymptotic range of variation of order k as $t\rightarrow\infty$ which depends on the initial conditions, unless ω differs only by order k from any one of a set of critical rational values. For these values, subharmonic resonance occurs. In this case the upper limit for the asymptotic range of variation of the amplitude is of order $k/|\omega\text{-}\omega_0|$, where ω_0 is the critical frequency nearest to ω. The extension of the argument to an equation in which $f(x, \dot{x}, t)$ is not a polynomial in x, $\dot{x}$, $\cos t$, and $\sin t$, but still periodic in t, is indicated. The asymptotic behavior of the phase $\tan^1 (x\omega/\dot{x})$ is also considered briefly. The reduction of the case of subharmonic resonance to an auxiliary differential equationof the first order is outlined. if either $\omega_0=m/n$ where m and n are large relatively prime integers, or $k/|\omega\text{-}\omega_0|$ is small, the results obtained in this way are in agreement with those obtained when subharmonic resonance does not occur.

517.942.82:517.942.6/.9 1178

Determination of Eigenvalues using a Generalized Laplace Transform—M. D. Friedman. (*Jour. Appl. Phys.*, vol. 21, pp. 1333–1337; December, 1950.) The classical eigenvalue problems of mathematical physics are solved by means of a Laplace transform extended, not over the interval (o ∞) but over the interval of interest for the differential equation. The method is applied to the Hermite, Laguerre, and Bessel equations and to the equation for the hypergeometric polynomials which include the Legendre, Tchebycheff, and Jacobi polynomials as special cases.

519.272.119:534.78 1179

Correlation Function Analysis—L. G. Kraft —(*Jour. Acous. Soc. Amer.* vol. 22, pp. 762–764; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. Discussion of theory on which the MIT digital electronic correlator is based, and presentation of speech correlation curves obtained with it.

519.272.119:621.39.001.11 1180

The Auto-Correlation Function—D. A. Bell. (*Wireless Eng.*, vol. 28, no. 328, pp. 31–32; January, 1951.) An explanation is given of what the auto-correlation function is and what it can be used for, and a list of papers and books bearing directly on the subject is included.

681.142 1181

Electronic Computers: General Design and Construction—L. Kosten. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 313–338; November, 1950.) An outline is given of the main parts of digital computers, that is, the input and output units, the memory, the arithmetic unit, and the external control. Magnetic, supersonic, and electrostatic memory systems are discussed Two adding units are described, working respectively on a parallel and on a series basis. A simplified block diagram of the external control of the EDSAC is shown, together with a schedule of orders.

681.142 1182

Electronic Computers: Mathematical Bases—I. Boxma. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 299–310; Discussion, p. 311. November, 1950.) An examination is made of the conflicting considerations imposed in the choice of a calculating system by considerations of accuracy, speed and scope on the one hand and cost on the other. The advantages and disadvantages of the binary digital system are discussed, and the elementary mathematical operations required in that system are described.

681.142 1183

Special Devices for Differential Analyzers—A. C. Cook, L. K. Kirchmayer, and C. N. Weygandt. (*Elec. Eng.*, vol. 69, p. 1080; December, 1950.) Summary of American Institute of Electrical Engineers 1950 Fall General Meeting paper. The range of the General Electric Company's differential analyser is being extended by the development of special devices for the generation of auxiliary functions, thus effectively increasing the number of integrators available for the solution of a given problem. Recent additions include, (a) a gear-ratio selector, (b) a remote indicator, (c) a transportation-lag device, (d) a recording counter, (e) dead-band units, (f) disconnect units; their functions are described.

681.142 1184

The Physical Realization of an Electronic Digital Computor—A. D. Booth. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 492–498; December, 1950.) Details are given of circuits designed to convert an existing relay computer with magnetic storage (3179 of 1949) to fully electronic operation. The units discussed are the memory control, the digit amplifier, the shifting register, the magnetic gate circuits, and the adding equipment.

681.142 1185

Marginal Checking as an Aid to Computer Reliability—N. H. Taylor. (Proc. I.R.E., vol. 38, pp. 1418–1421; December, 1950.) The presence of deteriorating components, which are liable to fail and cause errors in a digital computer, can be detected by a variation of circuit voltages, thus inducing actual failure, while a test program is performed. The failing components can then be systematically isolated. Trials have shown the value of this "preventive maintenance" in improving reliability.

681.142:519.272.119:621.39.001.11 1186

A Digital Electronic Correlator—Singleton. (*See* 1237).

681.142:621.383 1187

Photoelectric Analog Computer—E. C. Koenig. (*Electronics*, vol. 23, pp. 124, 126; December, 1950.) Description of equipment, making use of units comprising a single lamp for excitation of ten photocells arranged radially and having separate apertures and outputs. The photocells used are of the vacuum type with $AgC_{s_2}OC_s$ cathode. Advantages of the computer are enumerated.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.7):621.396.615.17/.18+621.317.761 1188

Frequency Generation and Measurements—W. S. Mortley: H. J. Finden. (*Electronic Eng.* (London), vol. 22, No. 274, p. 531

December, 1950. Comment on 1961 of 1950 and author's reply.

621.3.018.41(083.74)+621.316.726.078.3 1189
The Evolution of Frequency Control—Booth. (*See* 1271).

621.3.018.41 (083.74):621.396.81 1190
Coverage of Standard-Frequency Station WWVH—E. L. Hall. (*TV Eng.*, vol. 1, pp. 16–18 and 20–27; August and October, 1950.) A report of reception tests at several dozen stations in The United States, Alaska, Australia, China, Japan, and islands in the Pacific Ocean. The results are plotted and the resulting curves show reception efficiency as dependent on signal frequency, time of day, season of year, and sunspot conditions. The times of sunrise and sunset at the receiving station and at the transmitter have a definite effect on the percentage usability of the signals; the best reception is obtained for an all-darkness path. This is particularly the case for the 5-mc signals. The 10-mc signals are usable for more time than the 5-mc signals, while the 15-mc signals are usable still longer and may give uninterrupted service throughout the 24 hours at distances up to nearly 4,000 miles. The results in general indicate that for Alaska and the Pacific areas the transmissions from WWVH have resulted in an improvement of 87 per cent in the standard-frequency and time-signal service, though in The United States some confusion or interference with WWV transmissions, amounting to 2.5 per cent, has been caused.

621.3.087.4:551.510.535 1191
An Automatic Ionospheric Recorder for the Frequency Range 0.55 to 17 Mc/s—R. Naismith and R. Bailey. (*Proc. IEE* (London), Part III, vol. 98, pp. 11–18; January, 1951.) A full description of automatic equipment giving a photographic record of ionospheric characteristics in the form of a graph of height of reflection at vertical incidence against transmitted frequency. A pulsed self-oscillator coupled directly to the transmitting antenna is tuned from 0.55 to 17 mc in five bands. The receiver tuning capacitor is driven by a magslip motor whose speed is controlled by a discriminator circuit so as to keep the receiver in tune with the transmitter. The height and frequency calibrations and the date and time are recorded automatically on each record. The equipment has been in service for six years and has recorded 98 per cent of the hourly values of the critical frequency of the F_2 layer.

621.3.087.4:551.510.535 1192
A Single-Band 0-20-Mc/s Ionosphere Recorder Embodying Some New Techniques—T. L. Wadley. (*Proc. IEE* (London), Part III, vol. 98, pp. 45–46; January, 1951.) Discussion on 392 of 1950.

621.317:550.362 1193
Theoretical Basis of a Method for Determining the Specific A.C. Resistivity of the Ground—H. Weber. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 257–260; July 1, 1950. In French and German.) Description of a method, similar to that of Collard (1136 of 1936), making use of a family of curves for the magnetic field produced by current in an overhead line with earth return.

621.317.3:621.392.5 1194
Automatic Transmission Measuring Set—J. M. Hudack. (*Bell Lab. Rec.*, vol. 28, pp. 538–541; December, 1950.) An outline description of the equipment and its principles of operation. It provides, in about 30 seconds, a record of the amplitude transmission characteristic of the apparatus under test as a function of frequency within the alternative bands 20 cps to 20 kc or 100 cps to 100 kc. Portions of these bands may be selected if required. Other uses for the equipment include the measurement of harmonic generation within the equipment under test and examination of its noise spectrum.

621.317.311+536.58:621.38+621.397.62 1195
Electronics in Britain—J. H. Jupe. (*N.Z. Elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The direct current or voltage to be measured is applied to a circuit including a short thin Pt wire whose resistance is varied periodically by application of hf current modulated at 50 cps. From the resulting fluctuations of the dc an alternating voltage is derived, amplified, and applied to an indicator calibrated in terms of direct current or voltage.

621.317.329 1196
Rheographic Study of Laplace Fields with Helicoidal Structure—H. Dormont. (*Philips Res. Rep.*, vol. 5, pp. 262–269; August, 1950. In French.) Two methods are described for determining the equipotentials of a spirally wound system of conductors from measurements in an electrolyte tank. *See* also 1206 and 1266.

621.317.334/.335.087.3:546.3 1197
A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator)—A. Hersping. (*Arch. Elektrotech.*, vol. 40, pp. 57–75; 1950.) The method is based on the phase variation within the pull-in range of a pair of coupled tube oscillators. A description is given of a metal locator embodying the principle. The theory of the pull-in action is given briefly; formulas derived are confirmed by experiments.

621.317.725.029.45/.5 1198
Study of a H. F. Millivoltmeter—(*Radio franç.*, pp. 1–6; December, 1950.) Description of a Philips instrument. *See* 2285 of 1950 (Lindenhovius and all).

621.317.755:621.3.011.6 1199
Measurement of Time-Constants—R. Aschen. (*Radio franç.*, pp. 16–19; November, 1950.) Description of the principle of operation of a cro circuit. Two EL41 tubes are connected in a multivibrator circuit which provides the horizontal sweep voltage and also controls the current through a third EL41, in the anode circuit of which is the component or circuit on test. When the tube is conducting, the scanning spot is held to the right of the screen; during flyback the tube anode current is cut off and the discharge characteristic of the component is displayed on the screen. Typical traces and complete circuit details are shown.

621.396.622.63.001.4 1200
Mixer-Crystal Checker—P. D. Strum. (*Electronics*, vol. 23, pp. 94–97; December, 1950.) A method for quick determination of conversion loss from dc measurements. The theory of the method is outlined and a practical instrument described; its accuracy is estimated by comparison with a series of direct measurements of conversion loss made at wavelengths of 3 and 10 cm. The basis of the method is that minimum conversion loss for a crystal can be predicted from its static voltage-current curve.

621.396.645.001.4 1201
Test and Alignment Procedures for Video Amplifiers—F. E. Cone and N. P. Kellaway. (*Broadcast News*, no. 61, pp. 28–33; September to October, 1950.) General procedures are outlined and specific alignment methods are described for each component of the RCA television-terminal equipment.

621.397.6.001.4 1202
Signal Sources for Television Testing—D. W. Thomasson. (*Jour. Brit. IRE*, vol. 10, pp. 369–390. Discussion, pp. 391–392; December, 1950.) A detailed discussion of the performance requirements of signal sources. The rf source must be within 0.1 per cent of the correct frequency. The video-frequency generator must give an output approximating to the actual television signal. Tolerances for the timing of the various parts of the synchronizing and picture signals are given. Suitable circuits are described for the complete generator.

621.317.3.029.6 1203
Micro-Wave Measurements [Book Review] H. M. Barlow and A. L. Cullen. Publishers: Constable and Co., London, 339 pp., 30s. (*Jour. Sci. Instr.*, vol. 27, p. 342; December, 1950.) A book of permanent value not only to communication engineers, for whom it is primarily intended, but also to research physicists. Clear explanations are provided of the principles involved in the measurements. The longest chapters are those dealing with standing-wave measurements and with matching and transmission systems.

621.396.6.001.4 1204
Testing Radio Sets [Book Review]—J. H. Reyner. Publishers: Chapman and Hall, London, 5th edn., 215 pp., 22s 6d. (*Jour. Sci. Instr.*, vol. 27, p. 341; December, 1950.) A practical book which can be recommended.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.316.79.076.7:621.369.3 1205
Electronic Control of Home Heating—J. M. Wilson. (*Electronics*, vol. 23, pp. 84–87; December, 1950.) Description, with complete circuit details, of an ac resistance bridge and associated amplifier controlling the heating system of a building, greater heat being switched on when the resistance variation of an outdoor element indicates a drop in temperature.

621.317.311+536.58:621.38+621.397.62 1206
Electronics in Britain—J. H. Jupe. (*N. Z. elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The temperature-control unit operates by means of a Wheatstone-bridge circuit with a special diagonal branch including a resistor forming part of an amplifier circuit, a galvanometer, and two thermocouples connected in opposition whose action controls a servomechanism. *See* also 1195 and 1266.

621.317.334/.335.087.3:546.3 1207
A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator)—(*See* 1197).

621.365.44† 1208
Load-Matching Dielectric Heaters—R. H. Hagopian. (*Electronics*, vol. 23, pp. 98-101; December, 1950.) Increased efficiency is obtained by use of transmission lines for coupling rf generators to the loads, as matching problems are simplified.

621.38.001.8 1209
Electronics Symposium and Exhibition—(*Nature*, (London), vol. 166, no. 4234, pp. 1062–1063; December 23, 1950.) Summaries of the lectures at the opening ceremony of the conference arranged by the Scientific Instrument Manufacturers Association, London, September, 1950, and of the papers presented at the various sessions, with a note of the exhibits. For another account *see* 420 of March.

621.38.001.8:786.6 1210
The Baldwin Electronic Organ—A. Douglas. (*Electronic Eng.*, (London), vol. 22, no. 274, pp. 507–511; December, 1950.) An instrument in which the tones are generated entirely by tube oscillators. The basic waveform is a sawtooth, and square waves are generated by

the addition of two sawtooth waves. Instrument tones are simulated by the use of resonant filters. Details are given of the basic circuits and of various refinements designed to provide tone quality closely resembling that of a pipe organ.

621.383+621.385.38]:796.357 1211
Thyratron-Controlled Photoelectric Umpire—R. F. Shea. (*Electronics*, vol. 23, pp. 74–77; December, 1950.) Description of an automatic device that indicates passage of a baseball through the strike zone and also the speed of the ball as it crosses the plate. Three sets of photocells look at the sky through narrow slits, thus defining two vertical planes and an inclined plane intersecting the other two at knee and arm-pit height respectively. Passage of the ball through these planes in the correct sequence triggers an interconnected trio of thyratrons and operates an indicator.

621.384.611.2† 1212
The Design of Electron Synchrotrons—F. K. Goward, J. D. Lawson, J. J. Wilkins, and R. Carruthers. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 320–333. Discussion, pp. 334–339; November, 1950.)

621.384.611.2† 1213
The Design and Operation of a 30-MeV Synchrotron—D. W. Fry, J. Dain, H. H. H. Watson, and H. E. Payne. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 305–319. Discussion, pp. 334–339; November, 1950.)

621.384.612.1† 1214
Beam Oscillations in an F. M. Cyclotron—T. Teichmann. (*Jour. Appl. Phys.*, vol. 21, pp. 1251–1257; December, 1950.)

621.385.1.001.8:531.768.087 1215
The Improved Ramberg Vacuum-Tube Accelerometer—(*Tech. Bull. Nat. Bur. Stand.*, vol. 34, pp. 180–182; December, 1950.) Improved performance is obtained by, (*a*) an aging treatment resulting in more constant emission, (*b*) double gettering, (*c*) a 25-fold increase of sensitivity. See also 812 of 1949 and 2528 of 1947.

621.385.832:621.3.012:518.4 1216
The Graph-Scope, an Electronic Graph Plotter and Graphical Computer—A. L. Thomas, Jr. (*Elec. Eng.* vol. 69, pp. 1097–1100; December, 1950.) Description of an instrument for displaying on the screen of a cathode-ray tube a graph corresponding to any set of data, together with co-ordinate scales which can be either linear, logarithmic, or hyperbolic. A keyboard like that of a computer is used for setting the point-plotting circuits.

621.385.833 1217
A New Ratiometer for the Electron Microscope—R. Strauss. (*Arch. Elektrotech.*, vol. 40, pp. 49–56; 1950.) The object of this ratiometer is to make possible the full exploitation of the high resolving power of electron microscopes with magnetic lens systems, by accurately measuring and maintaining the ratio of the accelerating voltage to the square of the lens current, which ratio determines the image definition. The ratiometer comprises a long magnetic coil, within whose field the path of an electron beam is a helix; variations of the path are displayed on a fluorescent screen, and adjustment can be performed automatically or by hand. The accuracy attained for voltage variations is to within 2 parts in 10^4.

621.385.833 1218
Focusing in Electron Microscopy—R. S. M. Revell and A. W. Agar. (*Jour. Sci. Instr.*, vol. 27, p. 337; December, 1950.) In high-resolution work, personal errors in judging optimum focus can be avoided by taking a through-focus series of micrographs.

621.385.833 1219
The Correction of the Spherical Aberration of Electron Lenses Using a Correcting Foil Element—U. F. Giancla. (*Proc. Phys. Soc.* (London), vol. 63, no. 472B, pp. 1037–1039; December 1, 1950.) A physical explanation of the action of foil elements, introduced into the active region of an electron lens, in correcting the positive aberration of the lens, is confirmed by rigorous analysis.

621.387.4† 1220
An End-Window Alpha Scintillation Counter for Low Counting Rates—C. W. Reed. (*Nucleonics*, vol. 7, pp. 56–62; December, 1950.)

621.387.424†:621.392 1221
Auxiliary Electronic Circuits for Geiger Counters—Fontes and Moret. (*See* 1105.)

778.37:537.523.4 1222
A High-Intensity Short-Duration Spark Light Source—J. A. Fitzpatrick, J. C. Hubbard, and W. J. Thaler. (*Jour. Appl. Phys.*, vol. 21, pp. 1269–1271; December, 1950.) Summary noted in 2604 of 1950 (Fitzpatrick and Thaler).

778.37:621.397.331.2 1223
An Iconoscope Electro-Optical Shutter for High-Speed Photography—H. A. Prime and R. C. Turnock. (*Proc. IEE* (London), Part II, vol. 97, No. 60, pp. 793–796; December, 1950.) The system described makes use of an iconoscope in which the photo-emission is controlled by the application of suitable voltages to the collector electrode. The charge image, formed during a selected interval, is stored on the mosaic and subsequently scanned. Examples of exposures of duration ranging down to 20 μs are shown. The mechanisms of spurious-image formation during the "shutter-closed" period are discussed. Performance is compared with that of the Kerr-cell type of shutter.

621.387† 1224
Ionization Chambers and Counters [Book Review]—D. H. Wilkinson. Publishers: Cambridge University Press, New York, N. Y., 1950, 255 pp., $4.50. (Proc. I.R.E., vol. 38, p. 1465; December, 1950.) The treatment is focused "upon the principles of instrumentation, particularly upon the Geiger counter and ion chamber, to the exclusion of any description of the crystal counter and some of the more recent nuclear detection devices. None the less, the book is very complete and well organized with respect to the more conventional counting devices."

PROPAGATION OF WAVES

621.396.11+535.222 1225
The Velocity of Propagation of Electromagnetic Waves derived from the Resonant Frequencies of a Cylindrical Cavity Resonator—L. Essen. (*Proc. Roy. Soc. A*, vol. 204, no. 1077, pp. 260–277; December 7, 1950.) The length of the resonator was varied by mean of a piston whose change of position was measured by calibrated slip gauges. This change was measured for the H_{01n} resonance modes in vacuo near 9 and 6 kmc up to $n=8$ and $n=2$, respectively. The frequency was measured by comparison with a quartz-crystal standard, using an interpolation oscillator. Measurements made at different frequencies allowed the diameter of the cavity to be determined without knowing c; the value obtained was greater by 2×10^4 cm than that determined by metrological methods; the difference is attributed to a film of AgS, which would also account for the 25 per cent difference between the measured and calculated values of Q.

From a survey of all possible sources of error the maximum estimated error is ±3 km per second and the final value of $c=299792.5$ kms. *See* also 1751 of 1950, and 324 (Bergstrand) and 430 (Bol) of March.

621.396.11:551.510.535 1226
The Path of a Ray in a Curved Ionosphere Layer—K. Bibl. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 27–29; January to March, 1950.) Calculation of the effective ground-path length (D_1) for a reflected ray can be simplified by considering, not the angle of incidence on the lower surface of the layer, but the angle which the nonrefracted ray makes with the median plane of the layer, and by starting integration from this plane. The expression derived for D_1 indicates that the plane-layer formula is applicable in the case of a curved layer, the maximum error not exceeding 25 km. For nearly horizontal propagation the second-order (frequency correction) terms cannot be neglected.

621.396.11:621.396.812.4 1227
Comparison of Tropospheric Reception at 44.1 Mcs with 92.1 Mc/s over the 167-Mile Path of Alpine, New Jersey, to Needham, Massachusetts—G. W. Pickard and H. T. Stetson. (Proc. I.R.E., vol. 38, p. 1450; December, 1950.) Summary only of paper abstracted in 977 of May.

621.396.11:621.396.9 1228
Some Adverse Influences of Meteorological Factors on Marine Navigational Radar—J. A. Saxton and H. G. Hopkins. (*Proc. IEE* (London), Part III, vol. 98, pp. 26–36; January, 1951.) The effect of absorption and scattering of centimeter radio waves by atmospheric gases, uncondensed water vapor, and various forms of precipitation on the range obtained by marine radar equipment is discussed. For the wavelength of 3.2 cm now used for marine radar, the attenuation by atmospheric gases and uncondensed water vapour is hardly significant; absorption and scattering by rain are likely to cause most of the appreciable reductions in range of detection. It appears that, at a wavelength of 3.2 cm, for targets having echoing areas greater than 2,000 m² (that is ships of more than about 10,000 tons), attenuation in rainstorms will cause reduction in range, whereas for smaller targets masking by echoes from the rain will often be the more serious factor. Although very intense snowstorms can produce echoes sufficiently strong to be troublesome, the rate of precipitation required is such that its frequency of occurrence is unlikely to be great. In dense fogs reduction in detection range may be appreciable. Tables and graphs are given showing the effect on detection range of various types of precipitation in polar, temperate, and tropical regions for ships of 10,000 tons and of 1,000 tons, and for small boats or buoys.

621.396.11.029.62 1229
The Propagation of Metre Waves—F. C. Saic. (*Elektrotech. u. Maschinenb.*, vol. 67, pp. 325–332; November, 1950.) Critical discussion of the formulas of van der Pol and Bremmer and of Eckersley for propagation over land and over sea, both within and beyond the "visibility" range, taking account of transmitter height. Experimental results for wavelengths in the range 3 to 8 m are shown graphically.

621.396.81:621.3.018.41(083.74) 1230
Coverage of Standard-Frequency Station WWVH—Hall (*See* 1190.)

621.396.812 1231
Radio Propagation between Noumea and Adélie Land—M. Barre, K. Kawer, and E. Argence. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 21–26; January to March, 1950.) A series of graphs shows the relative strengths at various times of day of communication signals at different frequencies received at Noumea and in the steam ship *Commandant Charcot* about 5,500 km away in the antarctic. The optimum time for communication was about 1100 GMT (2100 local time), using frequencies in the range 8 to 17 mc. Conditions generally

remained good during the night but deteriorated abruptly at dawn. Factors used in predicting usable frequencies for the voyage are considered; due account was taken of auroral effects and of records formerly obtained at stations in the corresponding northern latitudes. The predictions show remarkable agreement with observations.

RECEPTION

621.396.812 1232

Diversity Effects in Spaced-Aerial Reception of Ionospheric Waves—E. N. Bramley. (*Proc. IEE* (London), Part III, vol. 98, pp. 19–25; January, 1951.) A mathematical analysis is given of the phase and amplitude of the signals induced in two spaced antennas by a continuous angular spectrum of downcoming rays, all lying in the vertical plane through the two antennas. It is assumed that the amplitudes of the individual rays are constant and that their phases vary in a random manner. If the energy is concentrated in a small range of angles the extent of the angular distribution can be estimated from observations of the variation of either the amplitudes or the phase difference of the signals in the spaced antennas. A similar analysis is made for a specularly reflected component superimposed on an angular distribution of waves having random phase. If observations are made of both the amplitudes and the phase difference of the signals in the two antennas, the width of the angular distribution and the ratio of the amplitude of the specularly reflected component to that of the angular distribution (the signal noise ratio) may be found.

Some daytime measurements of first-order reflections from the ionosphere at nearly vertical incidence at frequencies between 4 and 7 mc are most satisfactorily explained in terms of a steady specularly reflected component, superimposed on a noise background. Values of signal noise ratio of about 2 or 3 have been obtained with a noise background having an angular spread of about 1 degree.

621.396.823 1233

Car Ignition Radiation—C. C. Eaglesfield. (*Wireless Eng.*, vol. 28, no. 328, pp. 17–22; January, 1951.) Theory previously given (3741 of 1946) for the radiation from car ignition systems is developed, assuming that the impulse voltage is applied to the sparking plug through a radiating inductor. Allowance is made for resonances in the ignition system and for the addition of suppression resistors. Satisfactory agreement is obtained with the experimental results of Pressey and Ashwell (1779 of 1949) for the latter case.

621.396.822 1234

Threshold Signals [Book Review]—J. L. Lawson and G. E. Uhlenbeck. Publishers: McGraw-Hill, New York, N. Y., 1950, 388 pp., 42s. 6d. (*Electronic Eng.*, vol. 22, No. 274, p. 532; December, 1950.) Volume 24 of the Radiation Laboratory Series. "... presents a connected account of the work at the Radiation Laboratory on the discrimination of signals in the presence of noise, with particular emphasis on pulsed signals, visually displayed in the presence of random noise.... The volume can be highly recommended both to the newcomer and to the specialist."

STATIONS AND COMMUNICATION SYSTEMS

621.39:001.8 1235

Storage Devices for Communications—A. J. Lephakis. (*Electronics*, vol. 23, pp. 69–73; December, 1950.) The advantages to be gained by systematically recording messages prior to transmission over a communications link are considered. For such purposes storage systems are required in which the rate of extraction of data is not at all times equal to the input rate; possible systems are indicated. Other uses of storage devices considered include (*a*) radar display gear designed to respond only to moving targets and (*b*) equipment in which a desired relation between input and output signals (transfer characteristic) is synthesized.

621.39.001.11:519.272.119 1236

The Auto-Correlation Function—Bell. (*See* 1180.)

621.39.001.11:681.142:519.272.119 1237

A Digital Electronic Correlator—H. E. Singleton. (*Proc. IEE* (London), vol. 38, pp. 1422–1428; December, 1950.) 1950 IRE National Convention paper. "The relation between correlation functions and the general theory of communication is presented, and this relation leads to a technique for electronic computation of correlation functions and to the design of a machine for carrying out the computation. Because of the requirements of great accuracy and long storage, the machine makes use of binary digital techniques for storage, multiplication, and integration. Descriptions of the more unusual circuits in the machine are given, and circuit diagrams are included. A number of experimental results obtained by the machine are presented."

621.392.001.11 1238

The Information-Theory Point of View in Speech Communication—R. M. Fano. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 691–696; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. A nonmathematical discussion of the Wiener and Shannon theory of information, which is used to estimate the rate of transmission of information in speech communication.

621.395.44:621.392.52 1239

Loss-Compensated Filters in Carrier-Frequency Systems—Lehmann. (*See* 1113.)

621.396.4:621.396.615:621.316.726 1240

Stabilized Master Oscillator for Multichannel Communication—Pappenfus. (*See* 1117.)

621.396.41:621.396.619.13 1241

The Application of Frequency Modulation to V.H.F. Multichannel Radiotelephony—J. H. H. Merriman and R. W. White. (*Proc. IEE* (London), Part III, vol. 98, p. 56; January, 1951.) Discussion on 3513 of 1948.

621.396.619.13:621.3.018.42† 1242

Bandwidth of a Sinusoidal Carrier Wave, Frequency-Modulated by a Rectangular Wave with Half-Sine-Wave Build-Up—W. A. Cawthra and W. E. Thomson. (*Proc. IEE* (London), Part III, vol. 98, pp. 69–74; January, 1951.) "The problem of determining the frequency spectrum of a sinusoidal carrier, frequency modulated by a modified rectangular wave, is considered and a general solution obtained. The modification of the rectangular wave consists in replacing the vertical sides by a noninstantaneous transition in the form of a half-cycle sine wave. It is shown that for the determination of bandwidth it is reasonable to use, instead of the spectrum proper, a function which forms an envelope to it; this function is simpler and, in particular, it is independent of the on or off ratio of the rectangular wave. Curves are given for various values of the parameters involved."

621.396.65:[621.317.083.7+621.396.5+621.398 1243

Field Testing a Microwave Channel—D. R. Pattison, M. E. Reagan, S. C. Leyland, and F. B. Gunter. (*Elec. Eng.*, vol. 69, pp. 1092–1097; December, 1950.) Essentially full text of AIEE 1950 Summer and Pacific General Meeting paper. A description is given of microwave channel and af multiplex terminal equipment installed by the Pennsylvania Electric Company for simultaneous relaying, telemetering, supervisory control, and voice communication. A reflector is used to provide a continuous microwave path between terminal stations where direct line-of-sight communication is not possible.

621.396.65:621.396.828 1244

Reflected-Ray Suppression—H. E. Bussey. (PROC. I.R.E., vol. 38, p. 1453; December, 1950.) Destructive interference between the direct and ground-reflected rays may impair line-of-sight microwave communication. The reflected signal may be removed by means of an opaque screen on the ground at the geometrical reflection point, which blocks out one half of the first Fresnel zone. Confirmatory tests for an 800 foot transmission path operating on a frequency of 4.5 kmc are described.

621.396.65.029.63 1245

A Simplified Method of Planning Decimetre-Wave Line-of-Sight Links, taking into Account the Earth's Curvature and the Vision Ellipse—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 3, pp. 408–414; November, 1950.) Description, with relevant curves and abacs, of a geometrical method based on constructing the ellipse representing the first Fresnel region. The method is applied to a 50-km link using a wavelength of 12.5 cm.

621.396.712 1246

WBSM, WBSM-FM, New Bedford, Massachusetts—O. F. A. Arnold. (*Broadcast News*, no. 61, pp. 9–15; September to October, 1950.) Illustrated description of studios, control room, and transmission system.

621.396.712 1247

WKPT and WKPT-FM—T. Phillips. (*Broadcast News*, no. 61, pp. 22–27; September to October, 1950.) Illustrated description of studios, control rooms, and transmission system at Kingsport, Tennessee.

621.396.712(94) 1248

Engineering Aspects of the National Broadcasting Service—R. J. Boyle. (*Telecommun. Jour. Aust.*, vol. 7, no. 1, pp. 16–30; June, 1948.) An outline of the establishment of the National Broadcasting Service in Australia, and discussion of the problems involved, and the methods used to solve them. The planning of the transmitter locations, site selection, and field-strength surveys with test transmitters, are first considered. Typical transmitter equipment is then described, with construction details of the top-loaded vertical radiators used at Cumnock, NSW, Dooen, Victoria, and Wagin, Western Australia. As standard equipment, all transmitting stations are provided with a beat-frequency oscillator, noise and distortion meter, and modulation monitor; all have stand-by power plant.

Practically all the program material broadcast from National stations originates in, or is routed through, the studios in the capital cities. A general account of the studio and program-switching arrangements is illustrated by reference to the Sydney studios. Relay systems for interstate and National broadcasting, and the arrangements at Liverpool, NSW, for the reception and rebroadcasting of overseas programs, are also described.

SUBSIDIARY APPARATUS

621-526 1249

Relay Servomechanisms: The Shunt-Motor Servo with Inertia Load—T. A. Rogers and W. C. Hurty. (*Trans. Amer. Soc. Mech. Engr.*, vol. 72, pp. 1163–1172; November, 1950.) The theory is developed in terms of dimensionless motor parameters. Operating curves are drawn in the phase plane illustrating the effect of parameter changes on the stability of the servo for three forms of input signal.

621.314.634.011.4 1250

The Capacitance of Selenium Rectifiers—R. E. Burgess. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1036–1037; December, 1950.) Comment on paper abstracted in 744 of April (Cooper).

621.316.722 1251
Stability of Voltage Regulators—F. E. Bothwell. (*Elec. Eng.*, vol. 69, p. 1090; December, 1950.) Summary of AIEE 1950 Fall General Meeting paper. The presence of nonlinear control elements in many voltage-regulator circuits has prevented a complete mathematical analysis. Theory of Liapounoff, showing that stability of the continuous periodic solutions of a set of nonlinear differential equations is determined by the stability of its linear perturbation equations, is used to solve the problem of stability of electric circuits subjected to small displacements from equilibrium. As an example, characteristic equations are developed for a voltage-regulated *n* stage dc generator, and the stability boundary is obtained for the case of a step-resistor control element.

621.316.722.1 1252
Stabilized Power Pack—R. E. Harvison. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 512–513; December, 1950.) An analysis of the operation of two stabilizing tubes in parallel to provide high current output.

621.316.722.1.076.7 1253
Precision A.C. Voltage Stabilizers—G. N. Patchett. (*Electronic Eng.* (London), vol. 22, nos. 271–274, pp. 371–377, 424–428, 470–473, 499–503; September to December, 1950.) Critical study of different types and circuit arrangements providing stabilization to within about 0.01 per cent. *See* also 227 of February.

621.319.45.025:621.314.64 1254
Rectifier Properties of the System Al-Al_2O_3-Electrolyte Subjected to an Alternating Voltage—A. J. Dekker and W. C. van Geel. (*Philips Res. Rep.*, vol. 5, pp. 303–314; August, 1950. In French.) An account is given of an experiment showing that the static and dynamic current or voltage characteristics of this system are different. A bridge method is described for measuring the dynamic characteristic; the curve exhibits a loop in the region corresponding to forward current. The effect on the loop of frequency and temperature variations is investigated experimentally, and it is concluded that the structure of the oxide coating is not stable, but depends on the magnitude and direction of the applied voltage.

621.316.722 1255
Voltage Stabilizers [Book Review]—F. A. Benson. Publishers: Electronic Engineering, London, 1950, 125 pp., 12s. 6d. (*Electronics*, vol. 23, pp. 154, 158; December, 1950.) "A concise monograph . . . this book will be an extremely handy reference for an engineer or technician."

TELEVISION AND PHOTOTELEGRAPHY

621.397 1256
G.E.C. Picture-Telegraph Equipment—(*Electronic Eng.* (London), vol. 22, no. 274, pp. 503–504; December, 1950.) A general description, with diagrams, of the principles of operation of transmitting and receiving equipment, providing the highest possible definition within CCIT limits. At normal speed a 10×8 inch picture is transmitted in 14 minutes; this time can be halved at the expense of picture quality.

621.397.24 '.26:621.396.65 1257
Portable Microwave Television Link—(*Elect. Commun.* (London), vol. 27, pp. 295–297; December, 1950.) For another account *see* 459 of March.

621.397.331:535.623 1258
Photoelectric Method of Scanning and Chromatic Selection for Designs in Colour—M. Lange. (*Toute la Radio*, no. 150, pp. 343–348; November, 1950.) The pattern from which the various colors are to be selected is fixed to a revolving cylinder moving along its own axis, and is scanned by a brightly illuminated spot. An optical system making use of the toothed-wheel principle derives two rays of light, each of which is directed on to a prism spectroscope. Parts of the spectra so obtained are applied to three (panchromatic) photocells, arrangements being made for varying the parts selected. The outputs from the three cells are amplified and mixed in such a way that the resultant voltage is zero for the color to be selected, increases as this color fades, and reaches a threshold value for white; black and any color other than that required produce a voltage above the threshold value. This voltage actuates a galvanometer mirror reflecting a pencil of light of diamond-shaped cross section through an aperture of the same shape. A beam of light of width, depending on the position of the mirror, is directed on to a light-sensitive film carried by the receiving drum. As the drum rotates, a band is recorded on the film. An interposed revolving slotted disk transforms this band into a series of dots. Various refinements affecting the reproduced pattern are described.

621.397.5:535.624 1259
Subtractive Color TV—I. Kamen. (*TV Eng.*, vol. 1, pp. 12–13, 28; October, 1950.) The image is produced by passing light from a projection lamp through three dark-trace tubes in succession, the screens being of different materials corresponding to absorption of the three primary colors. A sequential-frame transmission system is used, and the receiver tubes are fed with the appropriate color signals. A greater luminous efficiency is claimed for the system.

621.397.5(494) 1260
Present State of Television in Switzerland—(*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 25, pp. 1001–1003; December 10, 1949. In German.) General review, with discussion of the technical and financial problems involved.

621.397.6:621.396.67 1261
Empire State Television—(*Broadcast News*, no. 61, pp. 41–45; September to October, 1950.) A description, with many photographs, of the replacement of the old 67-foot antenna surmounting the Empire State Building by a 217-foot multiple antenna system for stations WCBS TV, WABD, WJZ TV, WPIX, and WNBT.

621.397.6:621.396.67 1262
Temporary Vision Aerial—Bolt. (*See* 1076.)

621.397.6.001.4 1263
Signal Sources for Television Testing—Thomasson. (*See* 1202.)

621.397.611.2 1264
Television Camera Tubes—L. H. Bedford. (*Wireless Eng.*, vol. 28, no. 328, pp. 4–16; January, 1951.) Paper read at the International Television Congress, Milan, 1949. The effect of a high sensitivity tube in a television camera is to make it possible to use a smaller lens aperture for the same object illumination and thus obtain a greater depth of field. Curves showing depth of field against illumination are drawn for present day camera tubes, the minimum possible illumination being governed by the largest practical aperture which can be used with the tube. The minimum illumination is a criterion of sensitivity. A dimensionless form of the criterion, independent of arbitrarily chosen factors, can be found by taking the ratio of this illumination to that required by an ideal tube, the "Quanticon." Other properties of camera tubes discussed and tabulated are color response, resolution, spurious signal generation, signal and noise ratio, contrast, image persistence, geometrical accuracy, stability, and life.

621.397.611.2:778.2 1265
Flying-Spot Camera, Type TK-3A—C. R. Monro. (*Broadcast News*, no. 61, pp. 16–21; September to October, 1950.) Equipment is described for the production of video signals from 35 mm slides by the flying-spot method. The control and operating features are explained and typical applications are illustrated by photographs.

621.397.62+621.317.311+536.58:621.38 1266
Electronics in Britain—J. H. Jupe. (*N. Z. elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature, control, and extension units for television.

The Cossor extension unit for television comprises a coupling unit which reduces the modulation from the receiver to 0.5 v, together with the main unit which includes a loudspeaker and cathode ray tube. The unit is designed to give perfect linearity, so that the extension signal is not affected by any nonlinearity existing in the main receiver. The picture may be much larger than that of the main set and the unit can be made to cover more than one type of television system, with negligible increase in cost. For example, units made for use with the British 405 line 50 frames per second system can be adjusted in a few minutes to work with an American 525 line 60 frames per second receiver. *See* also 1216 of 1949 (Zeluff).

621.397.62 1267
The Simplification of Television Receivers—W. B. Whalley. (Proc. I.R.E., vol. 38, pp. 1404–1408; December, 1950.) Reprint. *See* 1271 of 1950.

621.397.62 1268
The Video Output Stage—E. T. Emms and E. Jones. (*Electronic Eng.* (London), vol. 22, nos. 272 and 273, pp. 408–413 and 454–460; October and November, 1950.) Values of t, the signal build-up time for the various elements of the video stage, are derived from the required over-all build-up time T estimated from the build-up time τ for the transmitted wave; picture definition being classified broadly in terms of T/τ. Different anode compensation circuits are shown and the indicial response (response to unit-step input) is plotted for each circuit. The design of a critically damped cathode compensation circuit is indicated. For equal output and equal t values, the input of such a stage must be 1.4 times that of a first-order shunt-compensated stage; but since the build-up time of the input signal is finite, this is not necessarily a disadvantage. Required characteristics of the video tube are tabulated. Class-A and class-B modes of operation of this tube as an anode bend demodulator are discussed and theoretical efficiency curves are derived. Secondary factors in design are considered and a design procedure is described for each of the first two types of circuit, component values being calculated for circuits using Type EF80 tubes.

621.397.62:621.396.662 1269
A Variable-Inductance T.V. Tuner—D. R. DeTar and H. T. Lyman, Jr. (*Electronics*, vol. 23, pp. 102–106; December, 1950.) Description of a tuner covering all vhf television channels without the use of switches or sliding contacts. It comprises a rf amplifier, mixer oscillator, and first detector. The wide frequency range is obtained by the use of only four variable inductors, each of which consists of a specially wound solenoidal coil with fixed core, the tuning being effected by sliding an Al sleeve between coil and core. Performance details are given and comparison is made with conventional designs.

621.397.7 1270
WOR's TV Studios—N. F. Smith. (*Broadcast News*, no. 61, pp. 46–73; September to

October, 1950.) A very comprehensive and detailed description of the New York studios and control rooms, including switching arrangements, film-projection equipment, master control system, and monitoring facilities.

TRANSMISSION

621.316.726.078.3+621.3.018.41(083.74) 1271

The Evolution of Frequency Control—C. F. Booth. (*Proc. IEE* (London), Part III, vol. 98, pp. 1–10; January, 1951.) A historical survey is made of the frequency stability of radio transmitters and the frequency tolerances specified by successive international conferences. The theoretical bandwidth needed by a single channel telegraph transmitter at the higher end of the 4- to 30-mc band is a small fraction of the tolerance (±0.003 per cent) assigned to it at the Atlantic City conference. Any improvement in stability will thus increase the number of stations possible. Simple crystal-controlled drive units capable of achieving the present tolerances are described and drive units for synchronized mf transmitters, frequency standards, and quartz clocks capable of much higher stability are discussed.

TUBES AND THERMIONICS

535.215.1/.2:621.383.2 1272

Some Properties of Complex Photoelectric Layers—A. Lallemand and M. Duchesne. (*Z. angew. Math. Phys.*, vol. 1, pp. 195–201; May 15, 1950. In French.) From experimental results on photoelectric layers of the type Ag to Cs_2O to Cs, it is concluded, (a) that such complex layers, although not obeying Richardson's law, have a well defined work function which can be determined from Fowler's theory; (b) that the layers are sensitive to wavelengths in the neighborhood of 15,000 Å, but that the photoelectric emission, though not completely explained by Fowler's theory, is not appreciably greater than that to be expected from the theory; (c) that the observed thermionic emission can be explained on the assumption that the layers consist of a large number of elements whose work function can be determined from Fowler's theory, or of a very small number of elements with a much smaller work function of the order of 0.4 v.

535.215.4:621.383.4 1273

A Relationship between the Refractive Index and the Infrared Threshold of Sensitivity for Photoconductors—T. S. Moss. (*Proc. Phys. Soc.* (London), vol. 63, no. 363B, pp. 167–176; March 1, 1950.) "It is shown that for photoconductive compounds the long wavelength "threshold" of sensitivity λ should be related to the refractive index n of the photoconductor, and analysis of the available data shows that for the more highly refractive compounds $n^4\lambda$ always has a value close to 77. As a consequence of this relation, an explanation can be given for the change of the threshold wavelength with temperature which is observed for lead sulphide and similar materials. Suggestions are made as to what materials ought to be suitable for photoconductive detectors at longer wavelengths in the infra red, and some confirmation of the ideas is shown by results obtained with cadmium arsenide, the properties of which are briefly described."

537.525.92:537.533.7:621.396.822 1274

Experimental Verification of Space-Charge and Transit-Time Reduction of Noise in Electron Beams—C. C. Cutler and C. F. Quate. (*Phys. Rev.*, vol. 80, pp. 875–878; December 1, 1950.) A cavity, resonant at 4.2 kmc, was arranged to be moved axially along a parallel electron beam from a Pierce gun. The observed variation of noise power in the cavity with distance along the beam is in fair agreement with the sinusoidal variation predicted theoretically. The greatest reduction of noise power obtained was 20 db below shot noise. That the minimum noise power is not zero is largely due to residual partition noise.

537.525.92:621.385 1275

Congruent Space-Charge Flow—G. B. Walker. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1017–1027; December 1, 1950.) For irrotational flow, propositions are established regarding, (*a*) rectilinear motion, (*b*) motion with constant current density along lines of flow, and (*c*) representation of lines of flow by the level lines of a harmonic function. Two cases of curvilinear flow from a unipotential cathode are discussed and shown to possess important features regarding magnification and transit time.

621.385.029.63/.64 1276

Interaction between a Travelling Electromagnetic Wave and a Beam of Electrons Moving in a Cylindrical System Perpendicular to Steady Crossed Electric and Magnetic Fields—R. Warnecke and O. Doehler. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231, pp. 1132–1134; November 20, 1950.) A traveling-wave tube is described in which a steady magnetic field is produced by the passage of a current I_m through the axial element of a coaxial delay line, and a steady electric field by applying a constant voltage v_0 between the inner and outer elements of the line. Relatively high gain and high efficiency are claimed to result. Two or more waves are propagated, depending on the operating conditions. For a particular system operating at about 1.5 kmc, with $I_m=500$ A and $v_0=1.2$ kv, calculation gives a useful power of 100 w, efficiency of 40 per cent, and gain of the order of 2.5 db per cm.

621.385.3:537.212 1277

On the Electric Field in a Single-Grid Radio Valve—G. B. Walker. (*Proc. IEE* (London), Part III, vol. 98, pp. 57–63; January, 1951.) The problem of calculating the electrostatic field in a planar triode is solved by assuming a system of line charges at the center of each grid wire and hence deriving a potential function which satisfies all the boundary conditions. This function is expressed in series form, but can be readily used to compute field characteristics to any required accuracy. Tables of the more important characteristics are provided.

621.385.3:621.315.592:518.4 1278

Graphical Analysis of Transistor Characteristics—Hunter. (*See* 1101.)

621.385.5:537.212 1279

On the Electric Field in a Multigrid Radio Valve—G. B. Walker. (*Proc. IEE* (London), Part III, vol. 98, pp. 64–67; January, 1951.) The question of the extent to which field disturbances caused by the individual wires of a grid penetrate an adjacent grid is examined by considering interelectrode capacitance relations. The effect of such disturbances can be measured, using an electrolyte tank, and can also be computed. Only space charge free fields are considered. and the electrode system is assumed either planar or concentric. It is concluded that Dow's simplified treatment (1904 of 1941) may always be used, at least as a first step, in the analysis of a tube containing two or more grids, the field in the region of one particular grid being afterwards investigated by a more rigorous triode analysis.

621.396.615.14.029.63 1280

Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits—L. Ratheiser. (*Radio Tech., Vienna*, vol. 26, pp. 519–524; November, 1950.) The special features and the use of the disk-seal triode with glass envelope and of the Telefunken metal ceramic type are described. Advantages of the grounded-grid circuit are stressed, and methods of overcoming the low interelectrode capacitance in this case are considered. These are, (a) "disk-edge" feedback, in which the output and input circuits are coupled capacitively to the grid through an insulating layer; (b) feedback windows, which are recesses in the wall of the grid cylinder; these give stability but limit bandwidth; (c) additional external capacitive feedback as in the Philips Type EC55 tube. Drawings of different assemblies are shown and their construction is discussed, particularly the methods used for obtaining good mechanical and electrical contacts.

621.396.615.141.2 1281

Some Aspects of Split-Anode Magnetron Operation—Reich, May, Skalnik, and Ungvary. (*See* 1120.)

621.396.615.141.2:537.525.92 1282

Effects of Space Charge on Frequency Characteristics of Magnetrons—H. W. Welch, Jr. (PROC. I.R.E., vol. 38, pp. 1434–1449; December, 1950.) "Properties of the magnetron space-charge swarm which affect the propagation of electromagnetic waves are defined in terms of an effective dielectric constant. The space charge is found to be doubly refractive in nature, the velocity of propagation depending on the direction of propagation of the wave, polarization of the wave, and frequency. Effects of synchronism of the rotating space-charge swarm are discussed qualitatively. Experimental results which check parts of the theory are presented. The relationship of the space charge to the circuit is discussed in terms of the nonoscillating and the oscillating magnetron."

621.396.822 1283

Threshold Signals. [Book Review]—Lawson and Uhlenbeck. (*See* 1234.)

MISCELLANEOUS

621.3.012.3 1284

Reference Sheets—(*Electronics*, *Annual Buyers' Guide Issue*, vol. 23, pp. R1-R40; Mid-month June, 1950.) A collection of 21 of the graphical and tabular sheets that have appeared in recent years in *Electronics*, with a complete index of all published since April, 1930. Many of those now included have been revised.

621.396.69+621.38]:061.5(058.7) 1285

Alphabetical Listings of All Components, Complete Units, Allied Products, Used in Electronic Equipment for All Purposes—(*See* 1126.)

621.38/.39 1286

Electronic Engineering Master Index, 1949. [Book Review]—Publishers: Electronics Research Publishing Co., New York, N. Y., 1950, 296 pp., $17.50. (*Jour. Frank. Inst.*, vol. 250, p. 587; December, 1950.) The third supplement, with coverage increased to nearly 400 periodicals, and so forth. Asterisks indicate Russian articles of which an English translation is available at Brookhaven National Laboratory.

621.396 1287

Einführung in die Funktechnik (Introduction to Radio Technology). [Book Review]—F. Benz. Publishers: Springer Verlag, Vienna, 4th edn, 736 pp., 78s. (*Wireless Eng.*, vol. 27, nos. 325/326, p. 273; October and November, 1950.) "First published in 1937; it has been much enlarged and is a very comprehensive and up to date text book of radio engineering."

621.396 1288

Fortschritte der Funktechnik und ihrer Grenzgebiete, Band 7/8 (Progress in Radio Technology and Related Fields, Vol. 7/8) [Book Review]—H. Richter. Publishers: Franckh'sche Verlagsbuchhandlung, Stuttgart, 1950, 387 pp., DM 60. (*Arch. elekt. Übertragung*, vol. 4, p. 340; August, 1950.) Includes surveys of the following subjects, contributed by different authors: broadcasting receiver design, radio wave propagation, fm, television, hf and industrial measurements, radio prospecting, cathode ray oscillography, and magnetic-tape recording.

Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

November, 1951
Volume 39 Number 11

Chicago Telephone Supply Corporation

MINIATURIZED "ALL-WEATHER" COMPONENTS

Pictured above is a miniaturized variable resistor designed for stable operation over extreme ranges of temperature and humidity. With the advent of jet planes, guided missiles and the like, the restrictions imposed on the size, weight, and performance requirements of electronic equipment become more exacting.

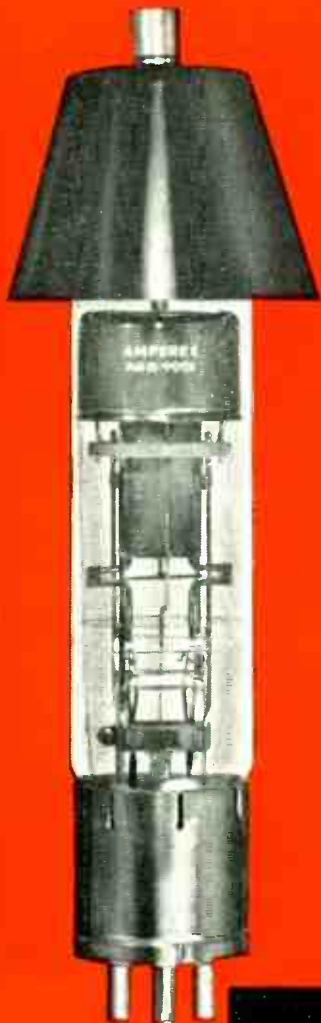
PROCEEDINGS OF THE I.R.E.

Table of Contents, Indicated by Black-and-White Margin, Follows Page 32A

The Institute of Radio Engineers

NEW AMPEREX tubes

Specifically designed for grid-control operation at peak anode voltages as high as **27,000 v.** for heavy duty INDUSTRIAL uses and high power TRANSMITTERS with outputs to **150 KW.** (3 phase full wave)

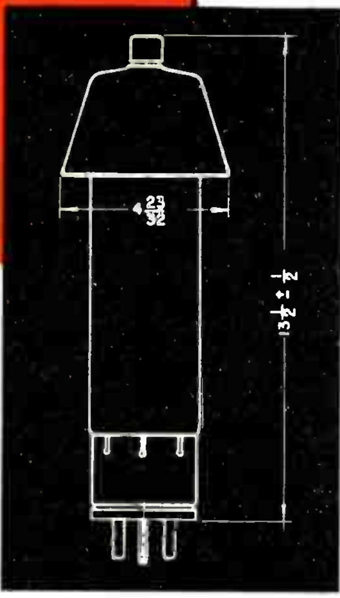


PROVEN LIFE

AGR-9951/5870

$110.00

	AGR-9951/5870		AGR-9950/5869	
CATHODE Directly Heated, Oxide Coated				
MAXIMUM PEAK ANODE VOLTAGE				
Inverse	27,000	10,000	13,000	10,000
Forward	27,000	10,000	13,000	10,000
CONDENSED MERCURY TEMPERATURE LIMITS (centigrade)	+30° to +40°	+25° to +60°	+25° to +55°	+25° to +60°
MAXIMUM PLATE CURRENT (Amperes)				
Peak	10		4	
Average	2.5		1	
FREQUENCY RANGE (cps)	25 to 150		25 to 150	
FILAMENT VOLTAGE	5.0		5.0	
FILAMENT CURRENT (amperes)	15		6.5	
TUBE VOLTAGE DROP (volts, approx.)	14 (Ib = 10 amperes)		15 (Ib = 4 amperes)	



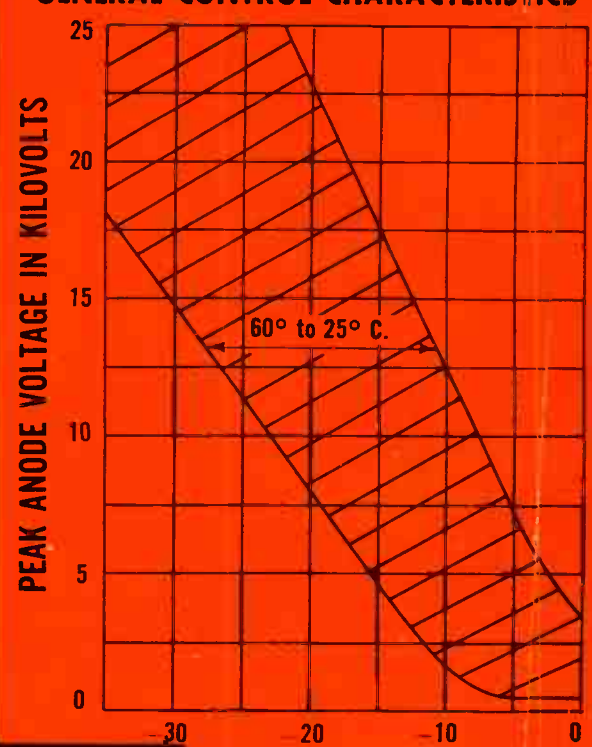
Re-tube with AMPEREX

THREE-ELECTRODE, MERCURY VAPOR

RECTIFYING TUBES

with NEGATIVE CONTROL characteristics

GENERAL CONTROL CHARACTERISTICS



AGR-9951/5870

D-C CONTROL-GRID VOLTAGE IN VOLTS

Data sheets and charts available on request

AMPEREX

AMPEREX ELECTRONIC CORP.

25 WASHINGTON STREET, BROOKLYN 1, N. Y.

In Canada and Newfoundland: Rogers Majestic Limited

11-19 Brentcliffe Road, Leaside, Toronto, Ontario, Canada

Stiff assignment you've been given, Mr. Designer! Fortunately General Electric's brand-new 17RP4 picture tube enables you to carry out your instructions word-for-word.

CHASSIS COSTS ARE TRIMMED because the 17RP4, electrostatic in design, requires no fixed magnet or focus coil with potentiometer. Convert either of these, plus labor, into retail pricing, and you have the desired mark-down in your new receiver.

CRITICAL MATERIALS SAVED! Needing neither fixed magnet nor focus coil, G.E.'s new 17RP4 eases your requirements for cobalt, nickel, and copper . . . thus helps assure steady TV production in your plant.

NO FOCUS CONTROL REQUIRED! The 17RP4 has zero focus voltage, which eliminates a receiver focusing knob or internal adjustment . . . meaning simpler TV operation and extra customer appeal for your set.

Picture performance is equal to other tubes *or better!* Exhaustive tests have proved this. Phone, wire, or write for descriptive Bulletin ETD-102, just off the press! *General Electric Company, Electronics Division, Section 7, Schenectady 5, New York.*

17RP4

Electrostatic picture tube with zero focus voltage.

RECOMMENDED OPERATING CONDITIONS

Anode No. 2, voltage	14,000 v
Anode No. 1, voltage for focus	0 v
Grid No. 2, voltage	300 v
Grid No. 1, voltage for spot cut-off	−33 to −77 v
Ion-trap field intensity (single-field), approximate	35 gausses

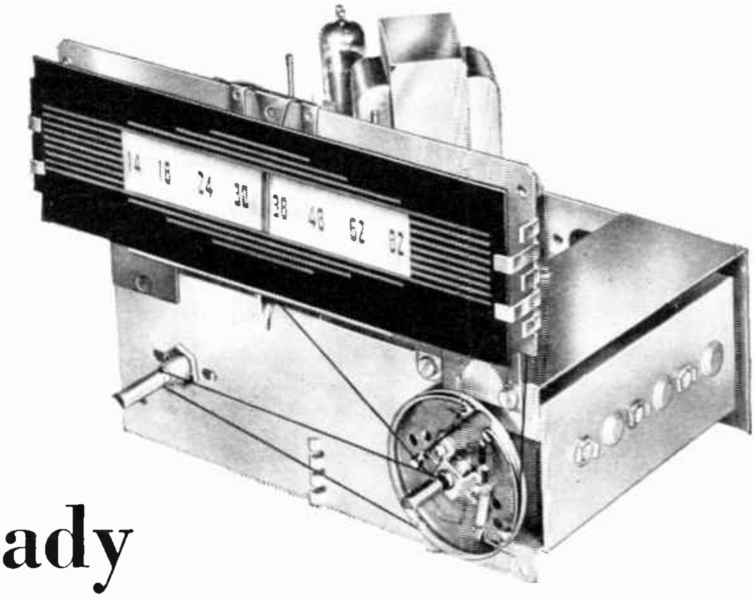
You can put your confidence in—

GENERAL ELECTRIC

181-K5

Proceedings of the I.R.E. November, 1951, Vol. 39, No. 11. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October, 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

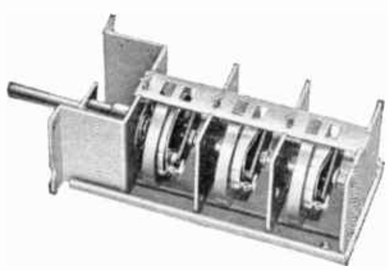
Table of Contents will be found following page 32A



UHF

Mallory Is Ready

to equip any receiver for UHF channels



Mallory UHF Tuner

A new version of the continuously variable Mallory Inductuner®, consisting of three sections of variable inductance. Covers the range between 470 and 890 megacycles with approximately 2 micromicrofarads of shunt capacity and in 270° of shaft rotation. Selectivity is excellent over the entire band.

Available now for assembly in your converter or as an auxiliary UHF tuner in your receiver.

Now In Development
A combination VHF-UHF tuner.

The Mallory UHF converter has been designed to permit the tuning of *all* UHF channels by *any* TV receiver, with no sacrifice of VHF reception. Connection to the receiver involves only the power line and antenna leads—no internal adjustments are required. Check the characteristics listed below and in the panel at the left describing the basic tuner.

Physical dimensions 8⅛″ x 6¼″ x 5$\frac{3}{16}$″

Built-in IF amplifier operating at the conversion frequency (channels 5 and 6) makes up for conversion and tuning losses

Temperature compensation and stabilization prevents frequency drift after initial warm-up

Low noise figure

High image and IF rejection ratios

The converter chassis is now available to set manufacturers for assembly with cabinets, dial plates and knobs of their design. Complete technical literature will be sent promptly on request.

Television Tuners, Special Switches, Controls and Resistors

SERVING INDUSTRY WITH

Electromechanical Products
Resistors *Switches*
TV Tuners *Vibrators*

Electrochemical Products
Capacitors *Rectifiers*
Mercury Dry Batteries

Metallurgical Products
Contacts *Special Metals*
Welding Materials

P.R. MALLORY & CO., Inc.
MALLORY

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

NEW Sperry Signal Source

operates both low and high voltage klystrons

A new *Microline* instrument Model 555 Klystron Signal Source, is an extremely well-regulated power supply. It features a continuously adjustable beam supply from 250 to 3600 volts. In addition, a reflector power supply is continuously variable from 0 to 1000 volts, and a control electrode supply is continuously variable from 0 to 300 volts. The versatility of this signal source permits operation of low voltage as well as high voltage klystrons.

Several types of modulation are provided with this instrument: sine wave at 60 cps, 0-300 volts peak to peak; saw tooth wave continuously variable from 600 to 1050 cps, 0-300 volts peak to peak with 15 microseconds decay time; and square wave continuously variable from 600 to 1050 cps, 0-300 volts peak to peak with 5 microseconds maximum rise and fall time. A modulation selector switch on the front panel permits external choice of type of modulation.

Write our Special Electronics Department for further information on Model 555 as well as other *Microline* instruments.

USABLE KLYSTRONS WITH MODEL 555 SIGNAL SOURCE

2K22	3K23	QK-277
2K25	3K27	QK-289
2K26	707B	QK-290
2K28	723A/B	QK-291
2K29	726A,B,C	QK-292
2K33	QK-140	QK-293
2K39	QK-141	QK-294
2K41	QK-142	QK-295
2K42	QK-143	QK-306
2K43	QK-159	6BL6
2K44	QK-226	6BM6
2K48	QK-227	SRX-16
2K56	QK-246	X-12
2K57	QK-269	X-13
		X-21

SPERRY GYROSCOPE COMPANY

DIVISION OF THE SPERRY CORPORATION

GREAT NECK, NEW YORK · CLEVELAND · NEW ORLEANS · BROOKLYN · LOS ANGELES · SAN FRANCISCO · SEATTLE
IN CANADA — SPERRY GYROSCOPE COMPANY OF CANADA, LTD., INTERNATIONAL AVIATION BUILDING, MONTREAL

PAA adds 3 stations to global radio net

The Collins high frequency radio equipment in this 18,000-pound load which went aboard at La Guardia field has now extended PAA radiophone operations throughout Africa. The three new stations are set up at Leopoldville in the Belgian Congo, Salisbury in Southern Rhodesia, and Johannesburg, South Africa.

Included for each station are a Collins 231D-20 Autotune* transmitter and four 51N-2 fixed frequency receivers with remote control units, now standard throughout the Pan American World Airways system.

*Reg. U.S. Pat. Off.

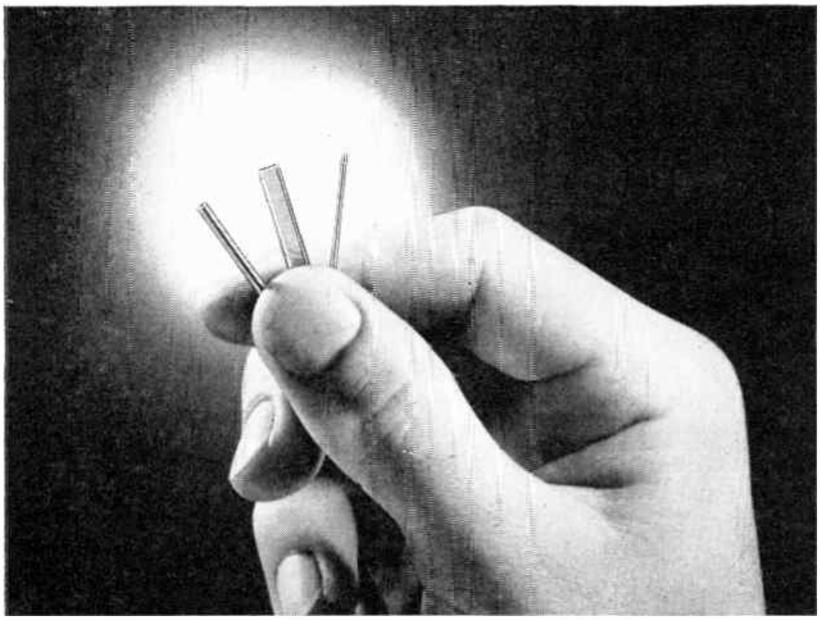
For land based radio communications, it's . .

COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 W. 42nd St., NEW YORK 18 **1937 Irving Blvd., DALLAS 2** **2700 W. Olive Ave., BURBANK**

What Superior Electroneering Does for You

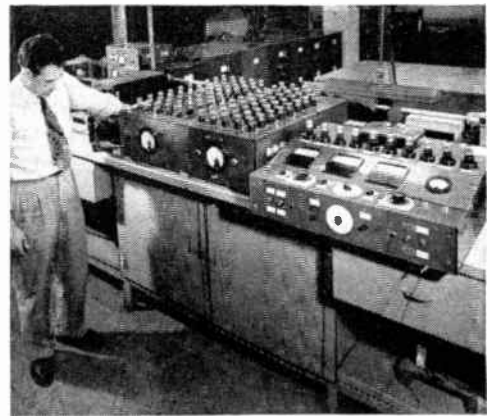


● Cathodes, three types of which are shown above, are one result of Superior engineering for the Electronics Industry—Electroneering. This is one of our big jobs. It is also one in which we take considerable pride and pleasure.

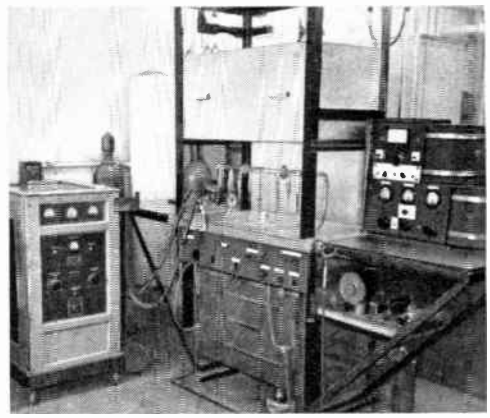
Among the usual run of our operations in this field: melt approval tests, raw material inspection, chemical analysis, testing of emission characteristics, physical characteristic tests, customer specification investigation and many others, we find time to dig well beneath the surface of the field.

For example, we continuously examine satisfactory products in an effort to improve their quality, shorten the required fabrication time, cut the costs to you, make it easier for you to assemble into finished parts or give you better service in any way.

Another example is our customer service which goes well beyond the limits of supplying good parts on schedule. We frequently work hand-in-hand with customers' engineers to solve their problems involving tubular parts. We are glad to consult with them at any time about the design or materials required for a new part or application. And, although we do a good bit of this, we'd like to do more. If you have a problem, why not let us help you find a solution with our combination of Electroneering and production know-how about cathodes and other parts for television, radio and other vacuum tubes. Write Superior Tube Company, Electronics Division, 2506 Germantown Ave., Norristown, Pennsylvania . . . no obligation of course.



Life testing of standard diodes which compares various cathode alloys.



Trolley exhaust for sealing-off standard diodes.

Microscopic examination studying surface conditions of cathodes.

This Belongs in Your Reference File
. . . Send for It Today.

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

Superior
THE BIG NAME IN SMALL TUBING

All analyses .010" to ⅝" O.D.
Certain analyses (.035" max. wall) Up to 1⅛" O.D.

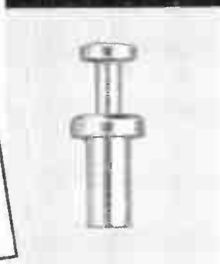
SUPERIOR TUBE COMPANY • Electronic Products for export through Driver-Harris Company, Harrison, New Jersey • Harrison 6-4800

NOW! Fill Your Needs

FOR ALL TYPES OF ELECTRIC AND ELECTRONIC COMPONENTS

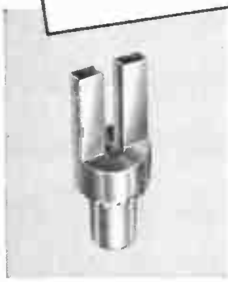
custom or standard the guaranteed components

SILVER-PLATED BRASS TERMINAL LUGS



Short Lugs. For low "headroom" applications. Mounted heights from $\frac{3}{32}$". In shank lengths for 6 board thicknesses, starting with $\frac{1}{64}$".

Turret Lugs. With 2 soldering spaces for 2 or more connections. Sizes range from $\frac{1}{32}$" to ¼" terminal board thicknesses. Mounted heights from $\frac{7}{32}$".

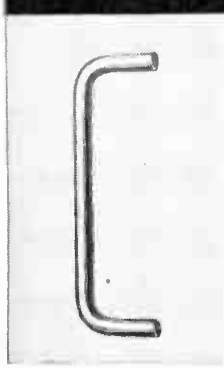


Split Lugs. For potted units where later soldering is advisable. Also standard applications. Hole through shaft allows top or bottom wiring. Fit standard board thicknesses from $\frac{1}{64}$" through ¼". Mounted heights from $\frac{5}{32}$".

Double End Lugs. Provide terminal posts on both sides of board. Through connection for easy wiring. For board thicknesses from $\frac{1}{32}$" to ¼". Mounted heights from $\frac{5}{32}$".

Combination Lug. Removable screw permits mounting components directly to screw end. Also provides removable link connections at screw end. 3 sizes, $\frac{5}{16}$", $\frac{11}{32}$", ⅜" diameter. Bright alloy plated for easy soldering.

HARDWARE



Handles in nickel-plated brass are available in 3 sizes ranging from $3\frac{5}{16}$" length to 6¾" length. Black alumilite aluminum handle available in 4⅜" length. Ferrules available on brass and aluminum handles.

Other Hardware includes tube clamps, panel and thumb screws, combination screw and solder terminals, shaft locks, terminal board brackets, standoff mounts, etc.

INSULATED TERMINALS

Phenolic. ¼" diameter, in rivet or screw stud type. Voltage breakdown from 4800 — 11,000 V at 60 cycles RMS.

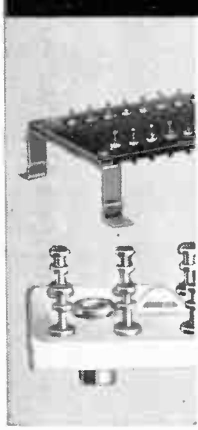
Ceramic. Silicone impregnated. 5 lengths of dielectric. Voltage breakdown ratings up to 5800 V. Over-all heights range from ⅜", including lug. For high electrical stresses over a broad humidity range. Cadmium plated studs. Brass terminals plated for soldering.

INSULATED FEED THROUGHS

Phenolic. Approved XXX material. Brass bushings, nickel plated. Brass through-terminals, silver plated for easy soldering. Rugged, withstand shock and vibration. Two sizes: for ¼" and ⅜" mounting holes.

Ceramic. Silicone impregnated. Threaded for ¼" hole mounting. O.A. length ⅞". Voltage breakdown 4800 RMS at 60 cycles.

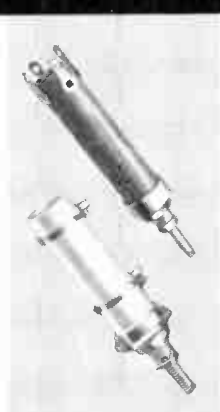
TERMINAL BOARDS



Phenolic. Available in various widths and terminal arrangements from ½" wide to 3" wide. Thicknesses: $\frac{3}{32}$"; ⅛"; $\frac{3}{16}$". All boards in 5 sections scribed for easy separation. Special boards made to your specifications.

Ceramic. Silicone impregnated. Type X1986 with 8 lugs staked in two rows. Standoffs riveted and soldered to ground strap for good grounding at R. F. frequencies. 1¼" long, ⅞" wide. All metal parts plated. O.A. mounted height: $\frac{35}{64}$".

SLUG TUNED COILS



Phenolic. 3 sizes: $\frac{27}{32}$"; 1⅛", and 2" high. 5 standard windings — also special windings or as high-quality phenolic coil forms.

Ceramic. Silicone impregnated. 5 sizes, mounted heights from $\frac{19}{32}$" to $1\frac{11}{16}$", diameters from $\frac{3}{16}$" to ½". Spring lock for slug. Cadmium plated mounting studs. Complete with mounting hardware and high, medium or low frequency slug.

R. F. CHOKES

LHC. High Q iron core with 6-32 mounting stud. 8 values from 2.5 mh to 125.0 mh. Wax impregnated.

LAB. Pie wound on phenolic core with cotter pin terminals. 8 windings .75 mh to 15.0 mh. Current rating 125 ma.

CUSTOM ENGINEERING — to your specifications or standard government specifications. C.T.C. Engineers will design all types of Boards, Coils, and Terminal Lugs for production in quantity to fill your needs. No extra charge for this service

WRITE FOR NEW CATALOG — For complete data and engineering drawings on these and other C.T.C. Electrical and Electronic Components and Hardware. No obligation.

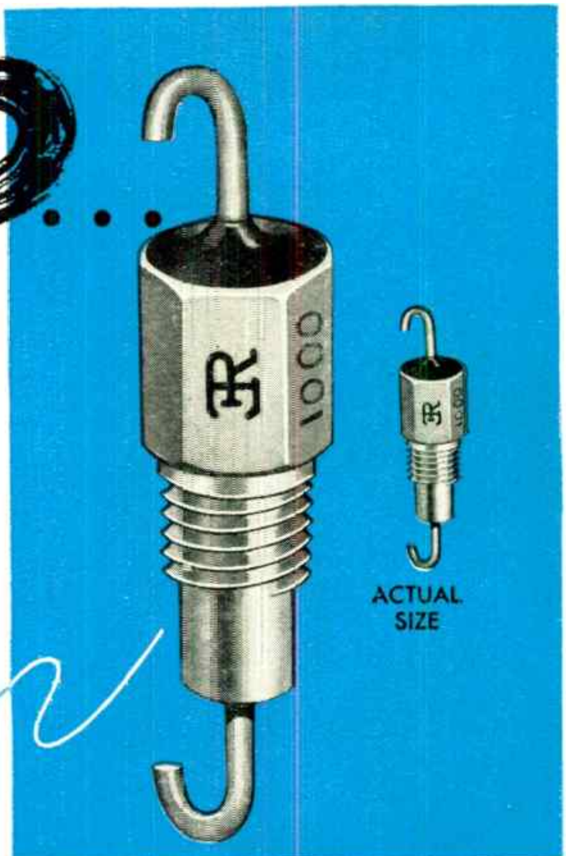
CAMBRIDGE THERMIONIC CORPORATION, 456 CONCORD AVE., CAMBRIDGE 38, MASS.

West Coast Stock Maintained by: E. V. Roberts, 5068 West Washington Blvd., Los Angeles, California

RUGGED...

is the word for the new Erie Style 327

FEED-THRU CERAMICON®



ERIE adds another outstanding capacitor to the most complete line of ceramic by-pass units available. Style 327 Feed-Thru design is a further result of continued Erie development in accomplishing ruggedness in components to meet severe military requirements and to give trouble-free service in other electronic applications. It embodies the following outstanding features:

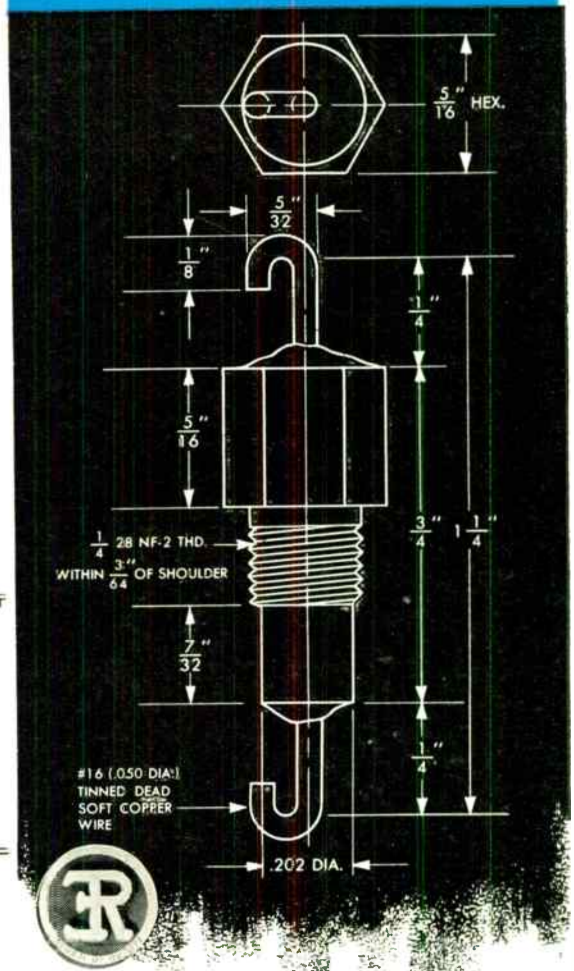
1. Mechanically rugged. Tubular ceramic capacitor is sealed at both ends in thermosetting low loss insulation.
2. Very low and uniform inductance path to ground.
3. Electrical shielding is provided by means of the grounded metal case.
4. All internal connections are soldered; no pressure contacts.
5. Hook type terminals provide sturdy connection tie points; also facilitate precision spacing of leads from other components where required in VHF and UHF circuits.

Specifications:

Standard capacitance values, mmf: 10, 33, 47, 68, 82, 100, 470, 680, 1000

Capacitance tolerance: ± 20% or + 80%, —20%

Rated voltage: 500 WVDC (values through 100 mmf also available in 1000 WVDC rating)



OTHER ERIE FEED-THRU CERAMICONS:

Style 357, rigid hooked wire lead, maximum capacitance 1000 mmf.
Style 362, #20 straight pig-tail wire lead, maximum capacitance 1500 mmf.
Style 2416, rigid wire lead, cadmium plated shell for solder mounting, maximum capacitance 1500 mmf.
Style 2418, no center lead, cadmium plated shell for solder mounting, maximum capacitance 1500 mmf.

Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND . . . TORONTO, CANADA

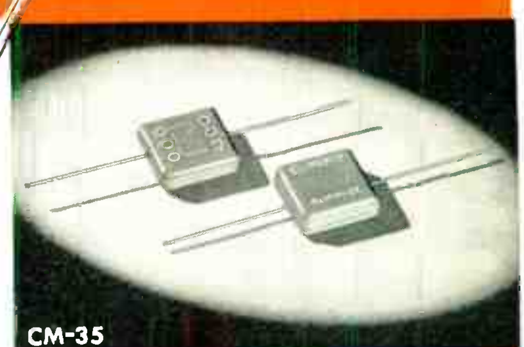
PERFORMANCE...

COMES IN...

From the midget champ which spins around a ¼ mile oval in 19 seconds . . . to the Indianapolis winner which clocks 157 m.p.h. on the straightaway . . . performance is the keynote in auto racing. In Electronics El-Menco Silvered-Mica Capacitors set the space. From the tiny CM-15 (2-525 mmf. cap.) to the mighty CM-35 (3300-10000 mmf. cap.) . . . unexcelled performance is paramount.

Precision workmanship, fine materials, careful design . . . these are the qualities which produce peak performance in racing cars and in El-Menco Capacitors. There is an El-Menco Capacitor for every specified military capacity and voltage. Each unit is factory-tested at *double* its working voltage. You are assured of dependability for every application.

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



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

Write on your business letterhead for catalog and samples.

MOLDED MICA **El-Menco** MICA TRIMMER

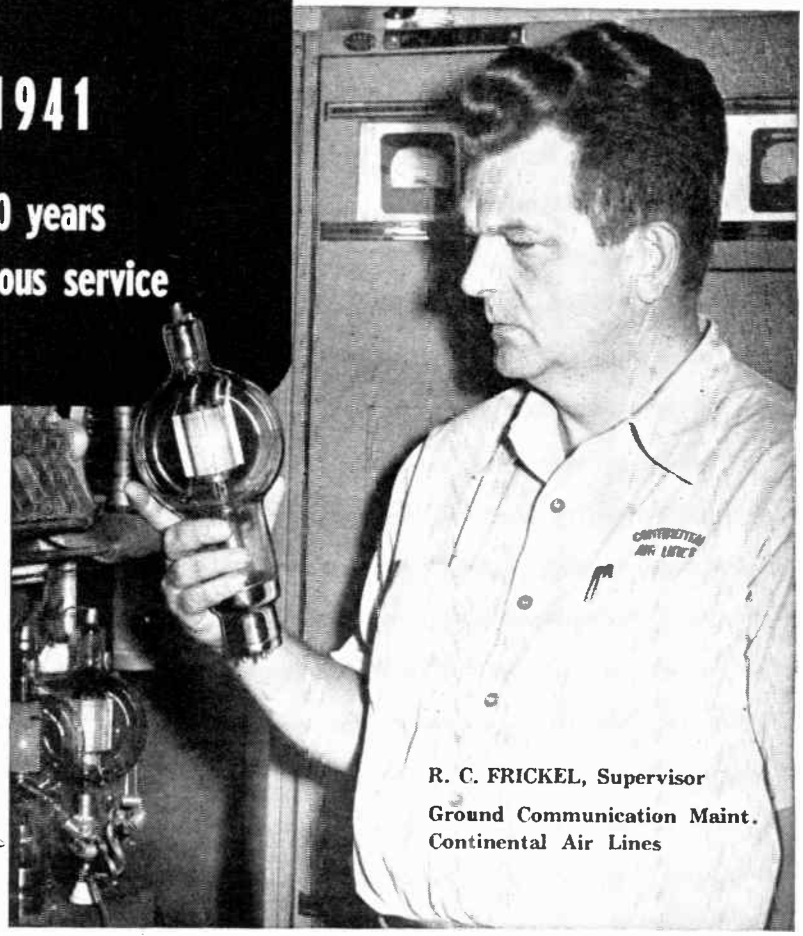
CAPACITORS

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

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

INSTALLED . . . 1941

Old P-167 gave 10 years of almost continuous service



R. C. FRICKEL, Supervisor
Ground Communication Maint.
Continental Air Lines

CONTINENTAL AIR LINES, Inc.

STAPLETON AIRFIELD
DENVER 7, COLORADO

August 9, 1951

Eitel-McCullough, Inc.
San Bruno, California

Gentlemen:

We finally had to replace old "P-167." This tube had been in continuous use at Continental Air Lines so long that it almost seemed like the passing of an old friend.

We here at Continental are very proud of our 17 year safety record and we know that dependable plane-to-ground radio communications have played an important part in the maintaining of this perfect record of safety.

Old P-167 was installed on June 17, 1941, and was removed July 28, 1951. During these 10 years the tube has seen almost continuous use at Continental's Denver transmitter, which is the communications control center of the airline's plane-to-ground radio contact.

The dependable performance of your tubes, as demonstrated by old P-167, is all the evidence we need as to where to buy our tubes. We will continue to use Eimac Tubes, as in the past.

Cordially,

R. C. Frickel, Supervisor
Ground Communications Maintenance

RCF:wrs
Attch.

P.S. Tubes P-117 and P-118, which were installed at the same time as P-167 in the same transmitter are still going strong.

The feelings expressed in Continental's letter are not unlike the feelings of thousands of other users of Eimac tubes. Top performance and a low cost to life ratio always make for customer satisfaction.

The new Eimac 450T that replaced "Old P-167" in Continental's transmitter should, because of improved vacuum tube materials and techniques, give even more satisfactory service.

Eimac tubes invariably do a job better . . . and at lower cost. Take advantage of their almost two decades of proved performance.

A new "Quick Reference" catalog on Eimac's Wide Variety of Tube Types is yours for the asking.

Follow the Leaders to Eimac TUBES The Power for R-F

EITEL-McCULLOUGH, INC.
San Bruno, California
Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

Now!

Transient-free test voltages down to 0.01 cps!

Versatile, general purpose generator for subsonic and audio work! Continuously variable, 0.01 to 1,000 cps, 5 bands. High stability, distortion less than 1%. Radical new circuitry offers sine, square and triangular waves.

-hp- 202A Low Frequency Function Generator

SPECIFICATIONS
-hp- 202A Low Frequency Function Generator

FREQUENCY RANGE: 0.01 to 1,000 cps in five decade ranges.

DIAL ACCURACY: Within 2%.

FREQUENCY STABILITY: Within 1% including warm-up drift.

OUTPUT WAVEFORMS: Sinusoidal, square, and triangular.

MAXIMUM OUTPUT VOLTAGE: At least 30 volts peak-to-peak across rated load for all three waveforms.

DISTORTION: Less than 1% RMS distortion in sine wave output.

OUTPUT SYSTEM: Can be operated either balanced or single-ended. Output system is direct-coupled; dc level of output voltage remains stable over long periods of time. Adjustment available from front panel balances out of any dc.

FREQUENCY RESPONSE: Constant within 1 db.

HUM LEVEL: Less than 0.1% of maximum output.

SYNC PULSE: 5 volts peak, less than 10 μsec duration. Sync pulse occurs at crest of sine and triangular wave output.

POWER: 115-volt, 50/60 cycles, 175 watts.

DIMENSIONS: 10½" high, 19" wide, 13" deep.

PRICE: $450.00 f.o.b. Palo Alto, California. End frames, for table use, $5.00 per pair f.o.b. factory. (Specify No. 17.)

Data subject to change without notice.

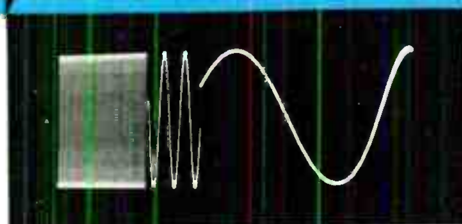


Figure 1. Oscillogram shows freedom from transients as output frequency is changed.

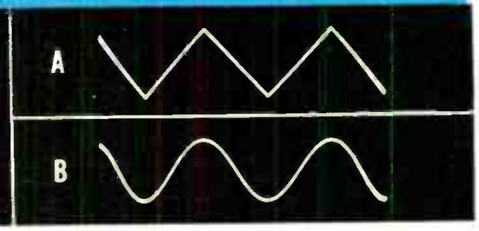


Figure 2. Oscillogram of (a) triangular wave applied to shaping circuit and (b) resulting sine wave.

-hp- 202A Low Frequency Function Generator offers you a compact, convenient and versatile source of transient-free test voltages between 1,000 and 0.01 cps. It provides virtually distortion-free signals for vibration studies, servo applications, medical and geophysical work, and other subsonic and audio problems. For such applications, the equipment generates 3 wave forms: sine, square and triangular. (Desired wave form is selected on front panel switch.) Output is 30 volts peak-to-peak for all 3 wave forms.

NEW CIRCUIT CONCEPT

-hp- 202A differs from conventional low-frequency oscillators in that the sine wave is electronically synthesized. A controlled bi-stable circuit generates a rectangular wave. This wave is passed through a special integrator providing a true triangular wave (Figure 2a). The triangular wave then enters a shaping circuit developed by *-hp-*. Here 6 duo-diodes modify or "shape" the peaks and provide a true sine wave with distortion of less than 1% (Figure 2b). This synthesizing circuit provides virtually transient-free output even when range switch is operated or frequency is rapidly varied. This circuit also maintains the amplitude constant under all conditions. It is not necessary to wait long periods for the circuit to stabilize at a new level as with conventional oscillators.

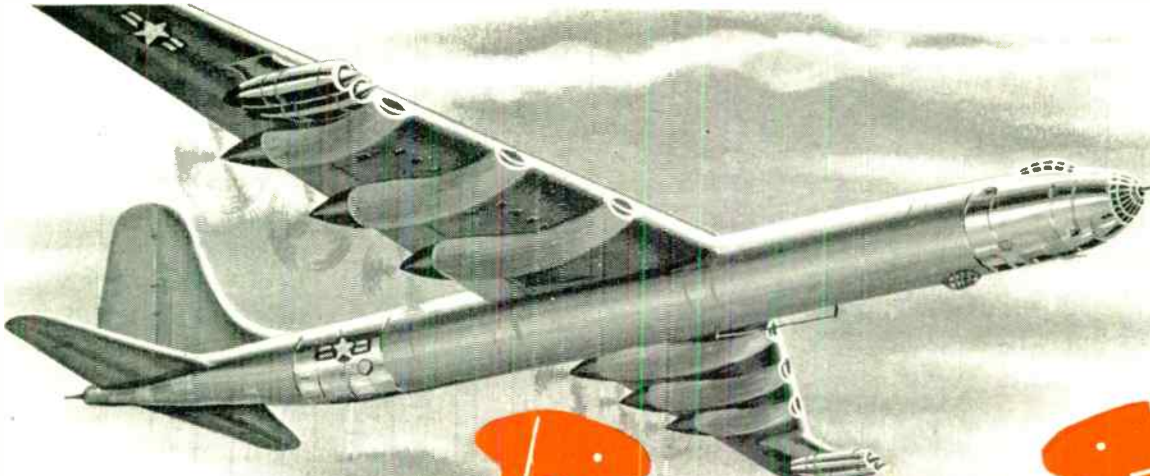
OTHER FEATURES

The output system of *-hp-* 202A is fully floating with respect to ground. May be used to supply a balanced voltage or either terminal may be grounded. It will deliver 10 v RMS to a load of 5,000 ohms or greater; internal impedance, however, is only 100 ohms. There are no coupling capacitors in the output system, and a high degree of dc balance is achieved by means of a special circuit.

-hp- field engineers, in most major cities, have complete details. Or, write direct.

HEWLETT-PACKARD CO.
2325D Page Mill Road • Palo Alto, Calif.
Export: FRAZAR & HANSEN, LTD.
San Francisco • Los Angeles • New York

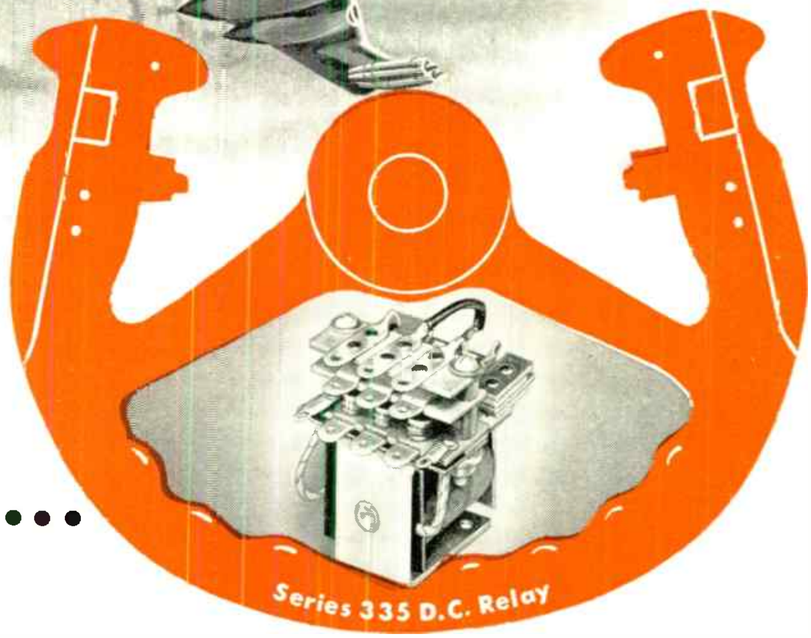
HEWLETT-PACKARD hp INSTRUMENTS



U. S. AIR FORCE B-36 BOMBER

One

OF A LINE....



APPROVED BY THE AIR FORCE FOR *Dependable Control*

"GUARD-A-SEAL"

Guardian engineering developments of hermetically sealed aluminum containers, designed specifically for aircraft, include: The AN Connector — Screw Terminal—Lug Header — Octal Plug standard types, plus a variety built to government specifications.

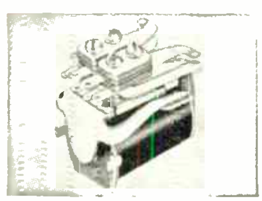
A. N. CONNECTOR

GET GUARDIAN'S NEW CATALOG OF HERMETICALLY SEALED RELAYS!

GUARDIAN RELAYS— Standard equipment in virtually all U.S. warplanes—exemplify the highest degree of *electronic precision* required by the U.S. Air Force, by government agencies and aircraft manufacturers. Guardian Relays and Solenoids—specified for timing, fusing and releasing bombs . . . firing guns . . . controlling radios . . . floodlights . . . landing gears . . . navigation aids . . . turrets . . . further establish Guardian's reputation for *electronic precision.* The Guardian Series 335 D. C. Relay, illustrated above, is but *one of a complete line* of Guardian Relays designed to permit more control in less space . . . more room for armament, power and personnel. Sensational Guardian developments include the famous "Guard-A-Seal" units specifically designed for aircraft and portable equipment, *sealed in aluminum.* They incorporate heavier frames, larger contacts, higher capacities, yet qualify under all AN weight requirements because the weight is in the relay—not the can!

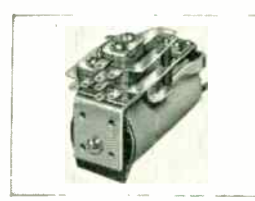
AN-3320-1 D. C.

AN-3324-1 D. C.



Series 595 D.C.

Series 610 A.C.—615 D.C.



Series 695 D.C.

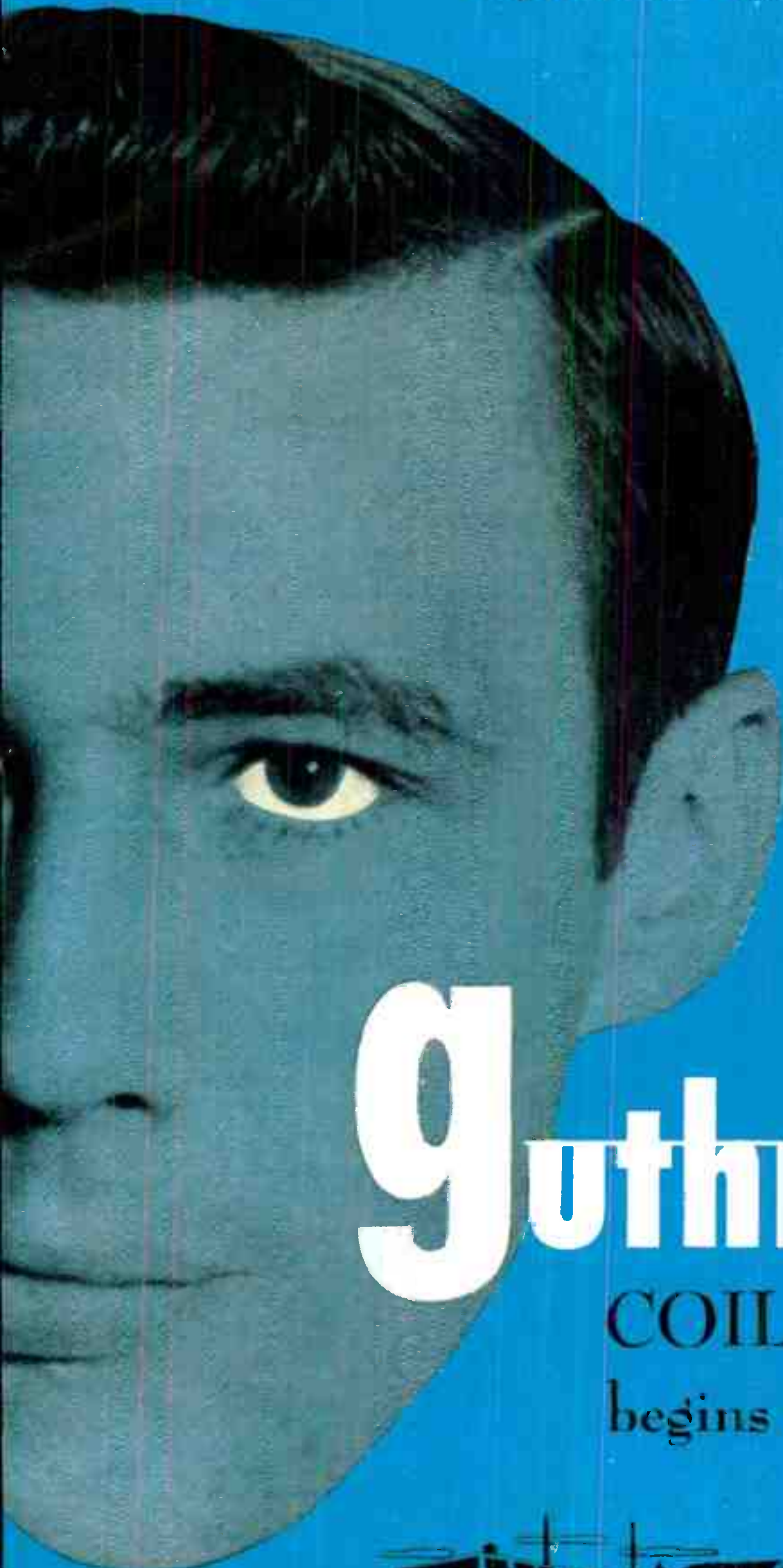
ALSO MINIATURE AND SUB-MINIATURE CONTROLS—WRITE

GUARDIAN ELECTRIC

1628-M W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

yokes • toroids • shield cans • loop antennas • oscillator coils
i. f. transformers • metal frame padder condensers • magnet and litzendraht-
textile served wire • compression type mica trimmers • dual compression trimmers
output transformers • antenna coils • assemblies • r f tuners



Besides specialized quality control, custom built plant equipment and exclusive production techniques, there is yet another important reason why Guthman is the World's largest independent maker of coils and basic electronic components —*research personnel!* Eager for the challenge of new coil development are engineers like Richard Huber whose academic background and practical training under the guidance of long experienced Guthman research experts eminently qualifies him to solve today's coil problems. So we confidently urge you to incorporate Guthman components into your next design . . . products engineered and fabricated within the Guthman plant like the BC348Q oscillator, detector, R. F. and antenna stages—and Guthman engineers will gladly design components especially for you. For a comprehensive descriptive brochure of Guthman facilities, write Dept. F.

guthman

COIL QUALITY
begins with laboratory design

edwin i. guthman & co., inc.

15 s. throop st. chicago 7, CH 3-1600   also attica, indiana

introducing
richard huber

Harry M. Neben, *Chief, Electrical*
AMPHENOL *Testing Laboratory*

Simpson RECOMMENDS Model 303

VACUUM TUBE VOLT-OHMMETER

Says Harry M. Neben: "I understand the 303 was developed to be of particular use to television service men for aligning sets in the field—so it's designed to perform a lot of test functions and is compact and easy to carry around. These same features make it quite a valuable laboratory and production tool here at Amphenol."

In the photo, Mr. Neben is using the Simpson 303 in conjunction with an Amphenol test fixture to measure insulation resistance between one wire and all other wires of a cable assembly.

SPECIFICATIONS

DC VOLTAGE: Ranges 1.2, 12, 60, 300, 1200 (30,000 with Accessory High Voltage Probe). Input Resistance 10 megohms for all ranges. DC Probe with one megohm isolating resistor. Polarity reversing switch.

OHMS: Ranges 1000 (10 ohms center). 100,000 (1000 ohms center). 1 megohm (10,000 ohms center). 10 megohms (100,000 ohms center). 1000 megohms (10 megohms center).

AC VOLTAGE: Ranges 1.2 12, 60, 300, 1200. Impedance (with cable) approx. 200 mmf. shunted by 275,000 ohms.

AF VOLTAGE: Ranges 1.2, 12, 60. Frequency Response Flat 25 to 100,000 cycles.

DECIBELS: Ranges −20 to +3, −10 to +23, +4 to + 37, +18 to +51, +30 to +63. Zero Power Level 1 M. W., 600 ohms.

GALVANOMETER: Zero center for FM discriminator alignment and other galvanometer applications.

R. F. VOLTAGE: (Signal tracing with Accessory High Frequency Crystal Probe). Range 20 volts maximum. Frequency Flat 20 KC to 100 M.C.

LINE VOLTAGE: 105-125 V. 50-60 Cycles.

SIZE: $5\frac{1}{4}$"x7"x$3\frac{1}{8}$" (bakelite case). Weight: 4 lbs. Shipping Wt.: $6\frac{1}{2}$ lbs.

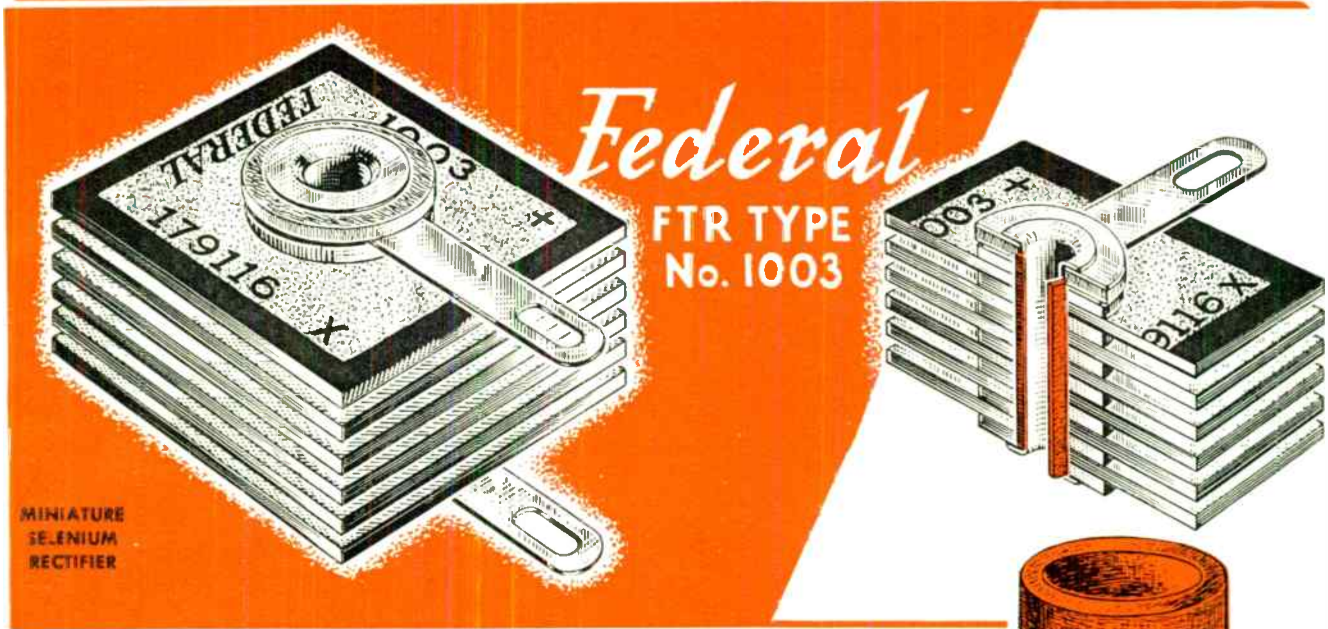
STILL AT THE SAME NET PRICE: Model 303, including DCV Probe, ACV—Ohms probe and Ground Lead with Operator's Manual—**$58.75**
Accessory High Frequency Probe, **$7.50**
Accessory High Voltage Probe, **$9.95**
Also available with roll top case, Model 303RT—**$66.70**

Available through your Parts Jobbers

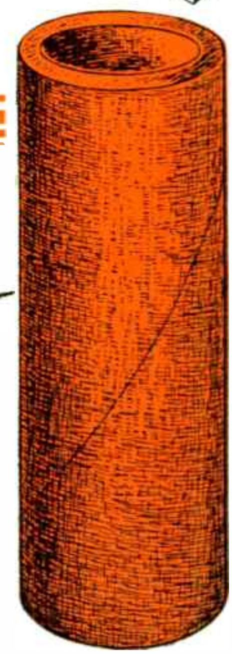
Simpson ELECTRIC COMPANY
5200 W. Kinzie St., Chicago 44, Illinois Phone: COlumbus 1-1221
In Canada: Bach-Simpson, Ltd., London, Ont.
World's Largest Makers of Electronic Test Equipment

BURTON BROWNE ADVERTISING

For the Best .. Call Cleveland



CLEVELITE*



is used in Federal Selenium Rectifiers to insulate the live electrical parts of the rectifier from the central eyelet upon which the rectifier is mounted.

Federal Telephone and Radio Corporation is known as America's oldest and largest manufacturer of Selenium Rectifiers.

CLEVELITE and COSMALITE* high quality . . . low cost Phenolic Tubing is the first choice of the Radio and Television Industries.

Low Moisture Absorption . . . Dimensional Stability . . . High Dielectric Strength . . . Low Loss . . . Great Physical Strength . . . Good Machineability . . .

Why Pay More?

Our tubing is available in . . .
Diameters 1/8 inch to 8 inches.
Lengths 1/8 inch to 8½ feet.
Wall thicknesses, .008" to .250".
Colors: Natural and Black.
FREE SAMPLE made to your exact specifications gladly furnished.

*Reg. U.S. Pat. Off.

Your inquiry will receive immediate attention.

The CLEVELAND CONTAINER Co.
6201 BARBERTON AVE. CLEVELAND 2, OHIO

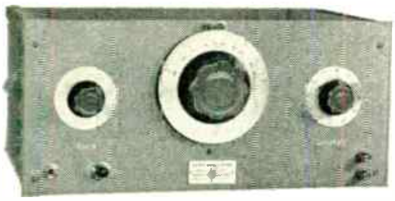
PLANTS AND SALES OFFICES at Plymouth, Wisc., Chicago, Detroit, Ogdensburg, N.Y., Jamesburg, N.J.
ABRASIVE DIVISION at Cleveland, Ohio
CANADIAN PLANT: The Cleveland Container, Canada, Ltd., Prescott, Ontario

REPRESENTATIVES

NEW YORK AREA R. T. MURRAY, 604 CENTRAL AVE., EAST ORANGE, N. J.
NEW ENGLAND R. S. PETTIGREW & CO., 10 N. MAIN ST., WEST HARTFORD, CONN.
CHICAGO AREA PLASTIC TUBING SALES, 5215 N. RAVENSWOOD AVE., CHICAGO

Accurate ac test voltages ½ to 10,000,000 cps

Complete Coverage

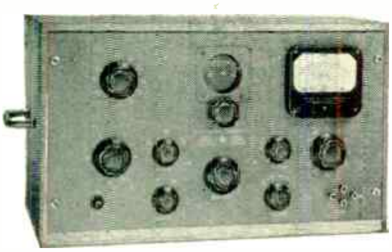


-hp- 200 Series Audio Oscillators

Six standard models, –*hp*– 200A and 200B have transformer-coupled output delivering 1 watt into matched load. –*hp*– 200C and 200D have resistance-coupled output and supply constant voltage over wide frequency range. –*hp*– 202D is similar to 200D, with lower frequency range. –*hp*– 200I is a spread-scale oscillator for interpolation or where frequency must be known accurately.

-hp- 650A Resistance-Tuned Oscillator

Highly stable, wide band (10 cps to 10 mc), operates independently of line or tube changes, requires no zero setting. Output flat within 1 db. Voltage range 0.00003 to 3 volts. Output impedance 600 ohms or 6 ohms with voltage divider.



-hp- 206A Audio Signal Generator

Provides a source of continuously variable audio frequency voltage with less than 0.1% distortion. Very high stability, accuracy 0.2 db at any level. Specially designed for testing high quality audio circuits, checking FM transmitter response and distortion, broadcast studio performance or as a low distortion source for bridge measurements, etc.

INSTRUMENT	PRIMARY USES	FREQUENCY RANGE	OUTPUT	PRICE
-hp- 200A	Audio tests	35 cps to 35 kc	1 watt/22.5v	$ 20.00
-hp- 200B	Audio tests	20 cps to 20 kc	1 watt/22.5v	$ 20.00
-hp- 200C	Audio and supersonic tests	20 cps to 200 kc	100 mw/10v	$150.00
-hp- 200D	Audio and supersonic tests	7 cps to 70 kc	100 mw/10v	$175.00
-hp- 200H	Carrier current, telephone tests	60 cps to 600 kc	10 mw/1v	$350.00
-hp- 200I	Interpolation and frequency measurement	6 cps to 6 kc	100 mw/10v	$225.00
-hp- 201B	High quality audio tests	20 cps to 20 kc	3 w/42.5v	$250.00
-hp- 202B	Low frequency measurements	½ cps to 50 kc	100 mw/10v	$350.00
-hp- 202D	Low frequency measurements	2 cps to 70 kc	100 mw/10v	$275.00
-hp- 204A	Portable, battery operated	2 cps to 20 kc	2.5 mw/5v	$175.00
-hp- 205A	High power audio tests	20 cps to 20 kc	5 watts	$390.00
-hp- 205AC	High power tests, gain measurements	20 cps to 20 kc	5 watts	$425.00
-hp- 205AH	High power supersonic tests	1 kc to 100 kc	5 watts	$550.00
-hp- 206A	High quality high accuracy audio tests	20 cps to 20 kc	+ 15 dbm	$550.00
-hp- 650A	Wide range video tests	10 cps to 10 mc	15 mw/3v	$475.00

Data subject to change without notice. Prices f. o. b. factory

Whatever ac test voltage you need—whatever frequency or magnitude you require—there is an *-hp-* oscillator or generator to provide the exact signal desired.

-hp- oscillators offer complete coverage, ½ cps to 10,000,000 cps. They are dependable, fast in operation, easy to use. They bring you the traditional *-hp-* characteristics of high stability, constant output, wide frequency range, low distortion, no zero set during operation.

-hp- oscillators and audio signal generators are used by manufacturers, broadcasters, sound recorders, research laboratories and scientific facilities throughout the world. For complete details on any *-hp-* instrument, see your *-hp-* sales representative or write direct.

HEWLETT-PACKARD COMPANY

2250D Page Mill Road **Palo Alto, California, U.S.A.**

Sales representatives in principal areas.

Export: Frazar & Hansen, Ltd., San Francisco, New York, Los Angeles

2250

HEWLETT-PACKARD *hp* INSTRUMENTS

Quality control *for resistors too!*

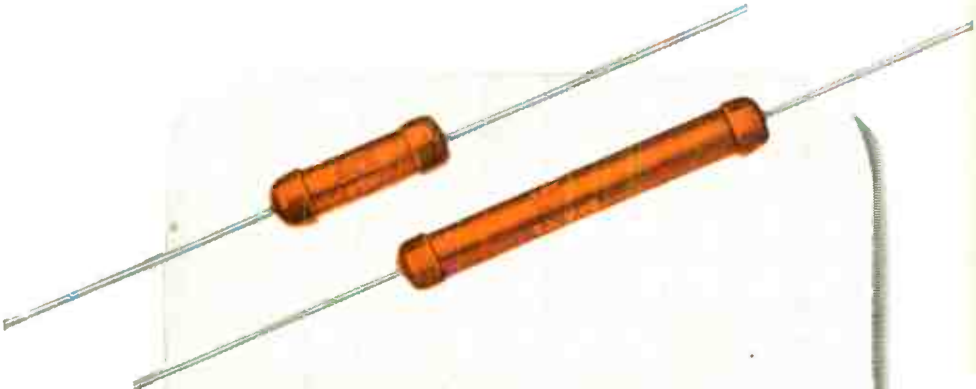
You can test a Quality Control Program in several ways . . . Does it begin *outside* your plant—with rigid specification of materials? Does it include detailed, continuous production-line inspections? Are these supplemented by laboratory and field-testing of your parts and products? And finally—do you reject on the basis of minor flaws? *IRC answers "YES" to all these yardsticks!* But the real proof of our Quality Control Program is the multitude of customers who specify IRC resistors—year after year.

Typical products of quality control, IRC Advanced BT Resistors meet and surpass JAN-R-11 Specifications. In standard RTMA ranges, Advanced BT's are designed to operate with moderate temperature rise and provide efficient power dissipation. Reason is the combination of IRC's filament-type resistance elements with exclusive construction features. Resistance material is permanently cured and bonded to special glass. Leads extend into filament for rapid heat dissipation. Molded bakelite seals element against moisture and prevents grounding. Advanced BT's are available in ⅓, ½, 1 and 2 watt ratings. Send for full details in 12-page technical data Bulletin B-1.

Meet and beat JAN-R-11 Specifications.
⅓, ½, 1 and 2 watts—available in ±5%, ±10% and ±20% tolerance
Low noise level and temperature coefficient

is essential

Quality control assures maximum uniformity in IRC's small 15/16'' Type Q Controls. Construction features one-piece dual contactor of special alloy—simplified single-unit collector ring—molded voltage baffles—special brass element terminals that will not loosen when bent or soldered. Type Q Controls have unusual durability and efficiency—adapt to a great variety of small-space applications. Send coupon for full details in Catalog A-4.



Where accuracy and economy are desired in high frequency applications and circuits requiring high stability and close tolerance, use IRC PRECISTORS. IRC makes 2 sizes of PRECISTORS *to customers' specifications*, rather than to standard RTMA values (subject, of course, to maximum and minimum values for each type). You'll find PRECISTORS excellent where carbon compositions are unsuitable or wire-wound precisions too expensive. Coupon brings full particulars in Catalog B-4.

Exacting tests and inspections control every step in the processing of IRC Power Wire Wound Resistors—assure balanced performance in every characteristic. In essential electrical and mechanical characteristics, these rugged wire wounds are ideal for heavy duty applications. Steatite forms are uniformly wound with high-grade alloy wire, and coated with special heat-dissipating cement. PWW's have been specified for over 14 years by leading industrial, commercial, broadcast, maritime and aircraft users. Catalog C-2 contains full information.

Your supplies of standard resistors for pilot runs, experimentation, or maintenance need never get out of control. When you need resistors in a rush, simply phone your local IRC Distributor. We keep his shelves filled with the most wanted types of standard resistors; he can give you prompt, 'round-the-corner delivery. If you don't know his name and address just ask us.

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

IRC *Wherever the Circuit Says*

INTERNATIONAL RESISTANCE COMPANY

Philadelphia 8, Pennsylvania

In Canada: International Resistance Company, Limited, Toronto, Licensee

INTERNATIONAL RESISTANCE COMPANY

405 N. Broad St., Philadelphia 8, Pa.

Send me additional data on the items checked below:

- ☐ Advanced BT Resistors
- ☐ Q Controls
- ☐ Deposited Carbon PRECISTORS
- ☐ Power Wire Wounds
- ☐ Name and Address of local IRC Distributor

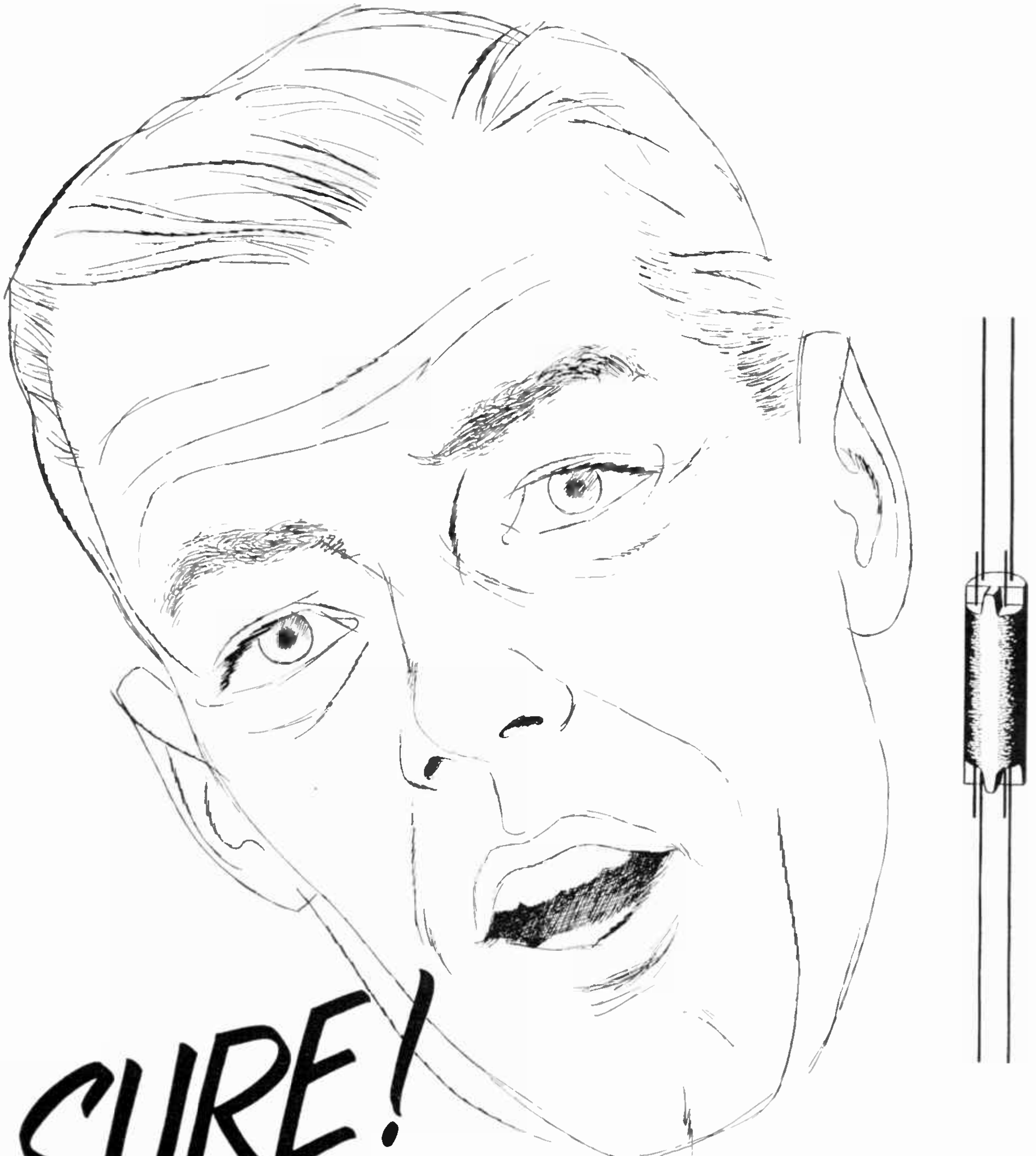
NAME ....

TITLE ....

COMPANY ....

ADDRESS ....

J. F. ARNDT & CO., ADV. AGENCY



SURE!

Buy SPEER when you buy coil forms

...for the exclusive notch for really anchored leads

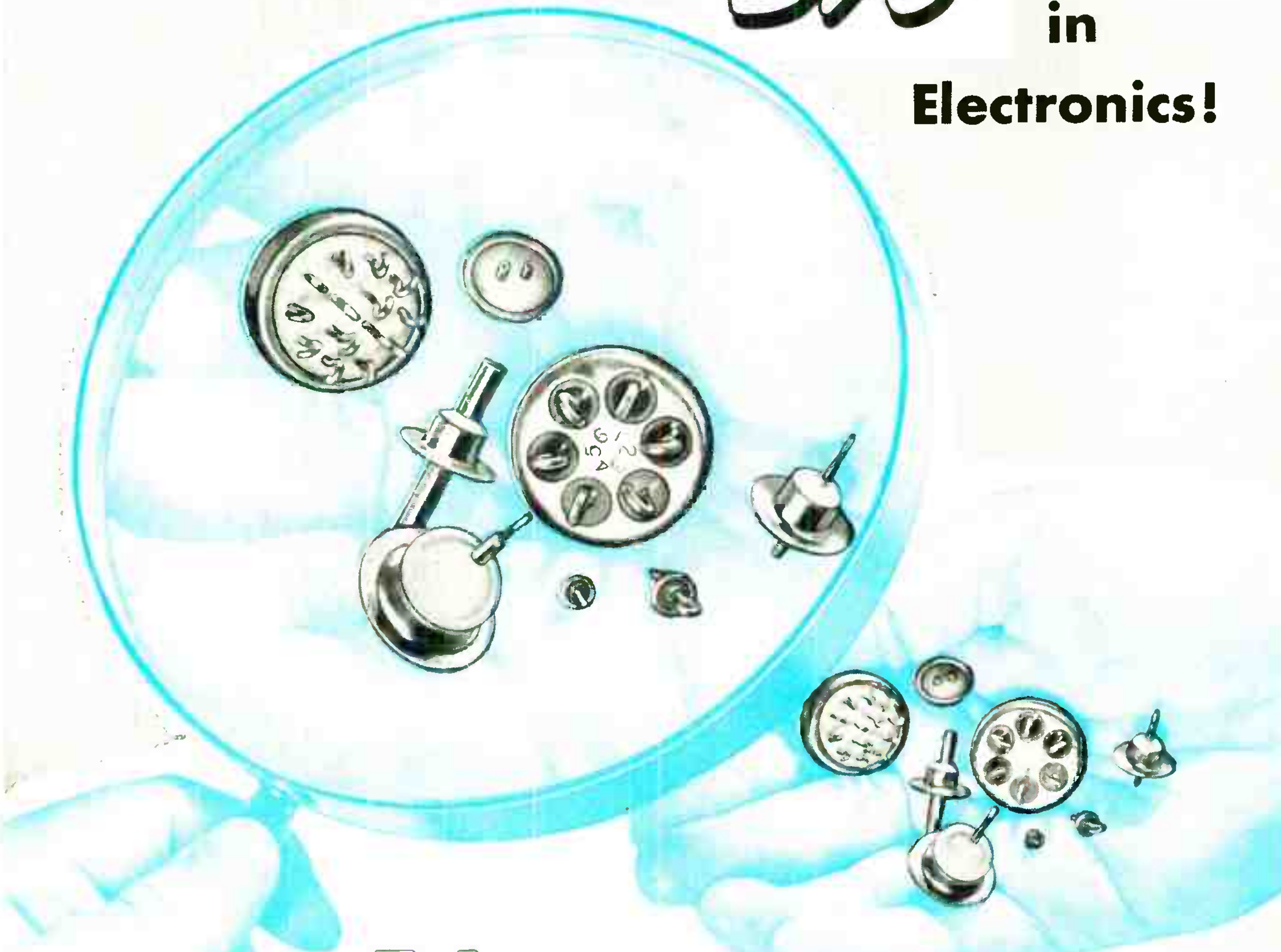
SPEER Resistor CORP.
A SPEER CARBON COMPANY SUBSIDIARY
ST. MARYS, PENNSYLVANIA

Complete your circuits with R. F. coils...chokes...tubular and disc capacitors...high voltage condensers...capristors by
JEFFERS ELECTRONICS, INC.
DU BOIS, PA.
Another Speer Carbon Co. Subsidiary

7568

SMALL PARTS

—figuring *BIG* in Electronics!



E-I *Hermetically Sealed*
MULTIPLE HEADERS and SEALED LEADS

COMPLETELY ILLUSTRATED

LITERATURE

AVAILABLE NOW!

Detailed data on standard designs is contained in new literature just off the press. Engineering assistance on special problems is also available without obligation. Please address requests on company letterhead.

Hermetically sealed multiple headers and leads are vital parts of countless electronic and electrical assemblies. E-I offers these important components in over 100 different standard types with a variety of optional features. Thus, E-I offers a quick, economical solution to most terminal problems. **For specialized applications,** E-I engineers can design and produce multiple headers and sealed leads to meet your requirements at a practical cost. If your problem involves the hermetic sealing of terminals and leads, consult us today!

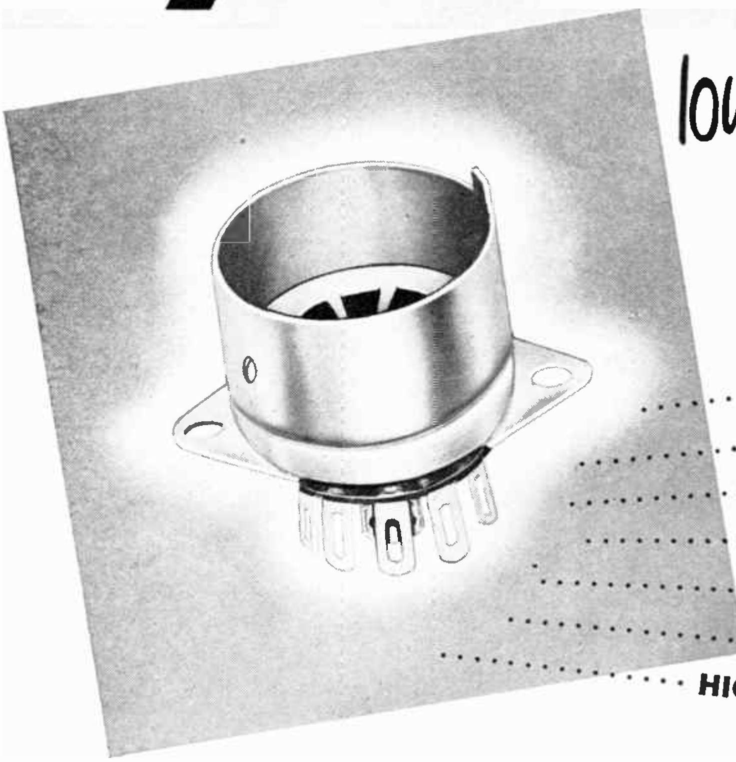
EI INC

ELECTRICAL INDUSTRIES • INC.

44 SUMMER AVENUE, NEWARK 4, NEW JERSEY

MANUFACTURERS OF SPECIALIZED ELECTRONIC PRODUCTS AND EQUIPMENT

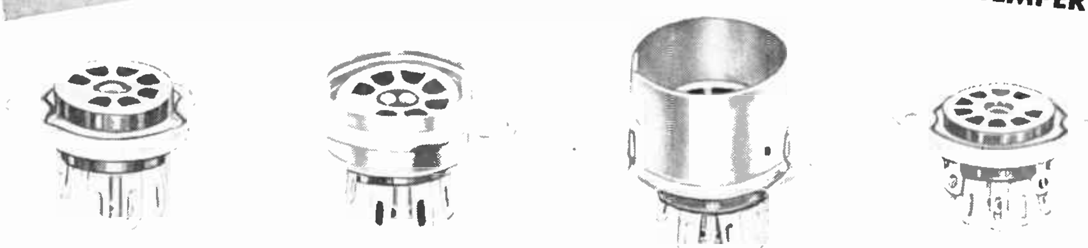
MYCALEX



low-loss miniature
TUBE SOCKETS

OFFER ALL THESE ADVANTAGES:

- CLOSER TOLERANCES
- LOWER DIELECTRIC LOSS
- HIGH ARC RESISTANCE
- HIGH DIELECTRIC STRENGTH
- GREAT DIMENSIONAL STABILITY
- IMMUNITY TO HUMIDITY
- HIGH SAFE OPERATING TEMPERATURE



-cost no more than
PHENOLIC TYPES

These glass-bonded mica sockets are produced by an exclusive MYCALEX process that reduces their cost to the level of phenolic sockets. Electrical characteristics are far superior to phenolics while dimensional accuracy and uniformity exceed that of ceramic types.

MYCALEX miniature tube sockets, available in 7-pin and 9-pin types, are injection molded with great precision and fully meet RTMA standards. They are produced in two grades, described as follows, to meet diversified requirements.

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

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

MYCALEX TUBE SOCKET CORPORATION
Under Exclusive License of
MYCALEX CORPORATION OF AMERICA
30 ROCKEFELLER PLAZA, NEW YORK 20, N.Y.

SINCE 1919
MYCALEX
THE INSULATOR
TRADE MARK REG US PAT OFF

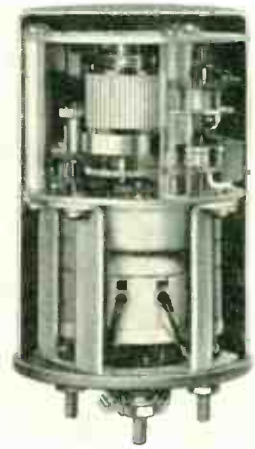
MYCALEX CORPORATION OF AMERICA
Owners of 'MYCALEX' Patents and Trade-Marks
Executive Offices: 30 ROCKEFELLER PLAZA, NEW YORK 20 — Plant & General Offices: CLIFTON, N.J.

News—New Products

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

Two New Relays

The R. W. Cramer Co., Inc., Centerbrook, Conn., announces the development of two new devices as additions to their line of timing equipment.



The Type HRR circuit-reclosing relay has been designed as a protective device which automatically tests the continuity of a fault for a given number of times at specific spaced intervals. Should the fault be cleared during any one of the trial tests, the control is restored to its normal operating condition. However, should the fault persist, the protective device automatically locks out the control circuit.

The Type HTI miniature time-delay Relay is intended primarily for use in electronic circuits, and can also be used in any application where it is desired to delay the operation of one circuit in relation to a second.

Type HTI Relays are available with maximum time ranges of 30 seconds, 60 seconds, 2 minutes, 5 minutes, 15 minutes and 30 minutes.

For complete information, write manufacturer for bulletin.

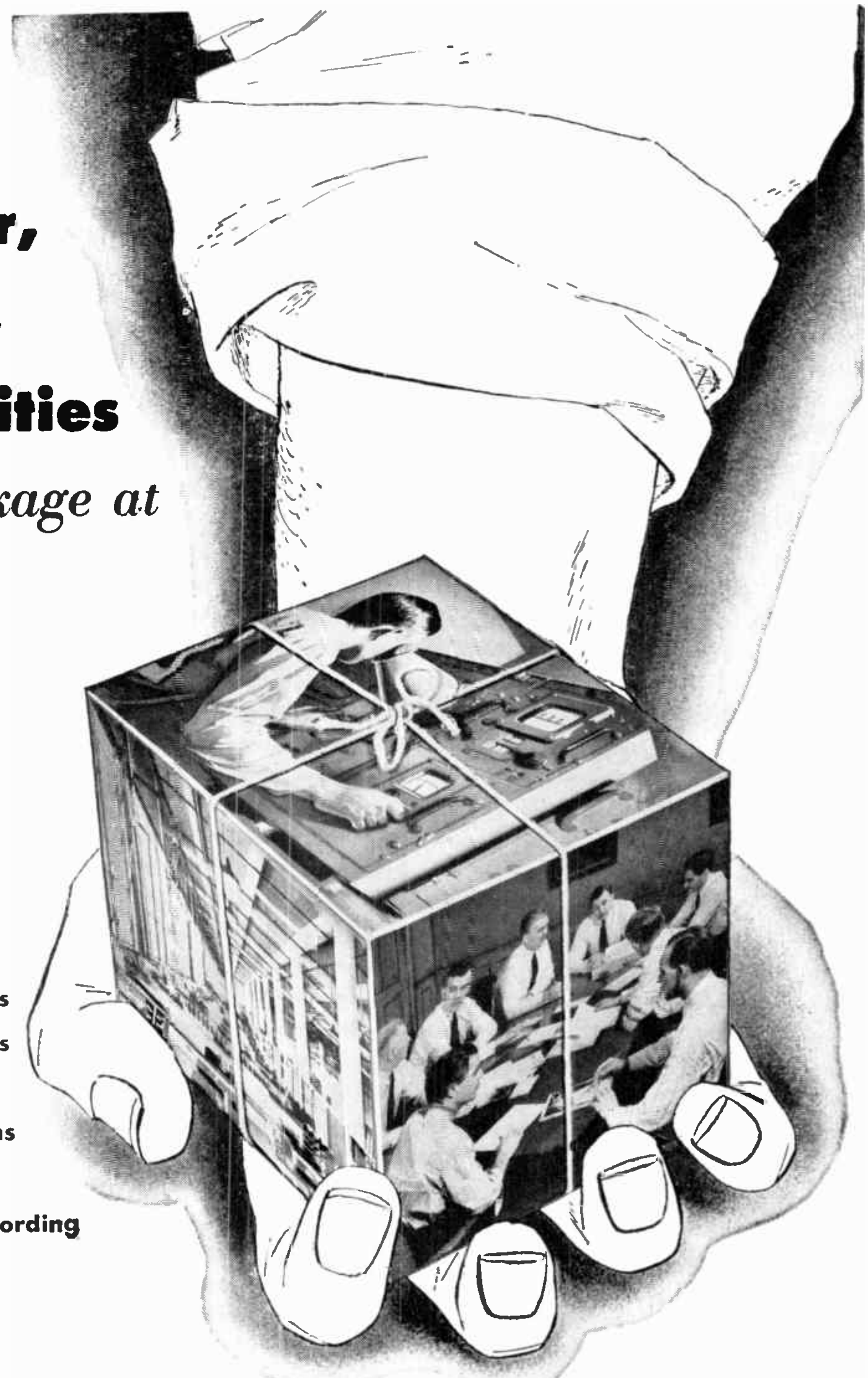
Recent Catalogs

• • • **Remler Co. Ltd.**, 2101 Bryant St., San Francisco 10, Calif., has three data sheets available. One on tube clips and silastic shock mounts, the second on tube sockets, and the third on the Servislide, a cabinet or rack mounted sheet metal chassis which permits servicing of equipment either top or bottom while in operation. *(Continued on page 38A)*

Brain power, experience, superb facilities

all in one package at

GRAY RESEARCH



- Television
- Video and display systems
- Audio and communications
- Teleprinter techniques
- Precise electro-mechanisms
- Aeronautic control devices
- Data transmission and recording
- Facsimile

In each of the defense-important fields listed here, the Gray organization has recently solved important problems. These facilities are available to prime contractors and to the military services as our contribution to the national effort in furtherance of communications, engineering or electro-mechanical designing. A booklet telling more of the Gray story will be sent for the asking.

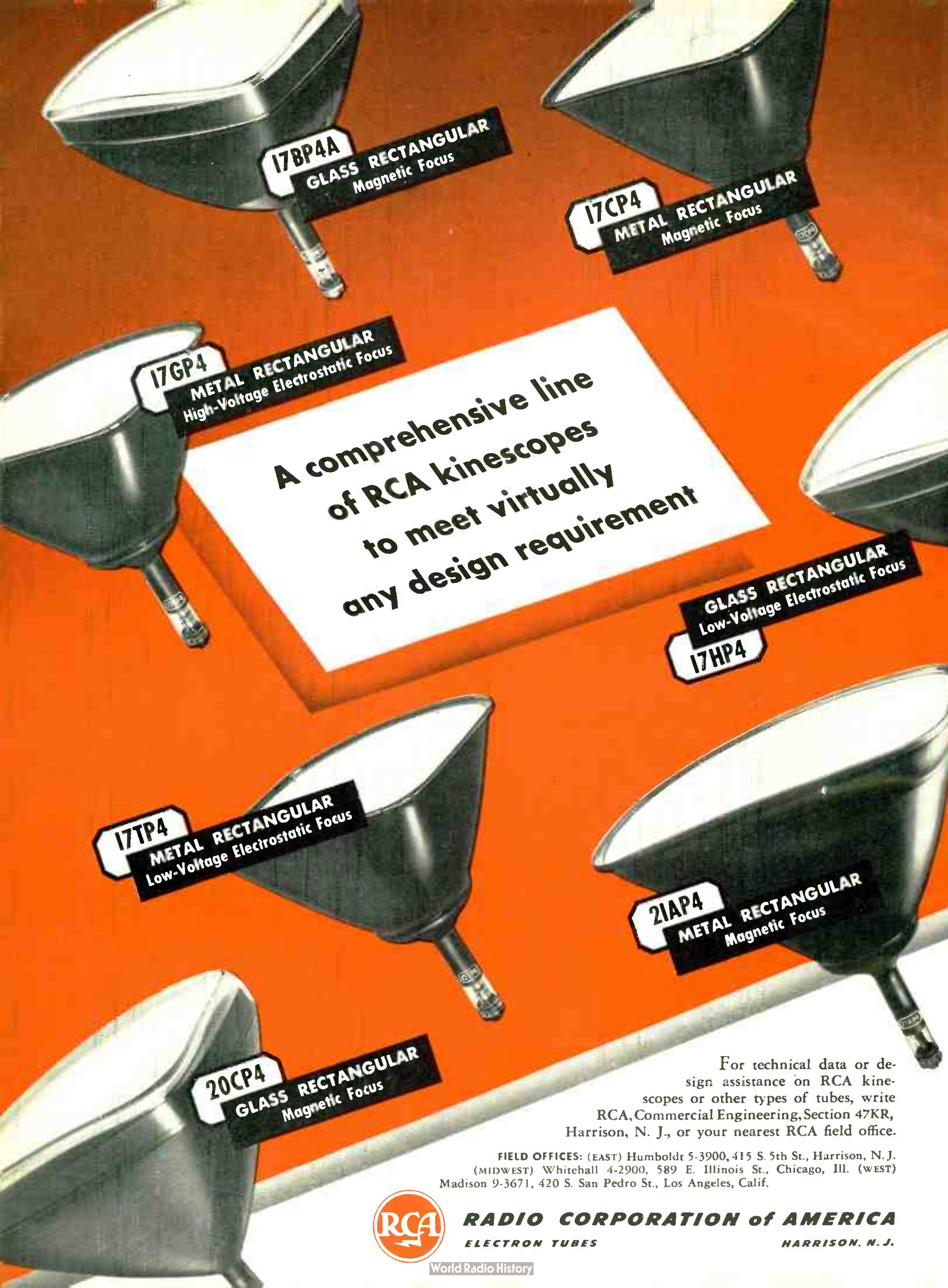
● Please write for Bulletin RE-11 describing the above equipment

GRAY RESEARCH

AND DEVELOPMENT CO., INC., 16 ARBOR ST., HARTFORD 1, CONN.
Division of The GRAY MANUFACTURING COMPANY—
Originators of the Gray Telephone Pay Station and the Gray Audograph

AUDOGRAPH

President



For technical data or design assistance on RCA kinescopes or other types of tubes, write RCA, Commercial Engineering, Section 47KR, Harrison, N. J., or your nearest RCA field office.

FIELD OFFICES: (EAST) Humboldt 5-3900, 415 S. 5th St., Harrison, N. J. (MIDWEST) Whitehall 4-2900, 589 E. Illinois St., Chicago, Ill. (WEST) Madison 9-3671, 420 S. San Pedro St., Los Angeles, Calif.

RADIO CORPORATION of AMERICA

ELECTRON TUBES — HARRISON, N. J.

PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 39 — *November, 1951* — NUMBER 11

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

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

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES

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

W. M. Rust, Jr.

BOARD OF DIRECTORS, 1951-1952

William Rust, Jr. attended the Rice Institute in Houston, Texas, where he received the degree of Doctor of Philosophy in 1931. His fields of study were in mathematics of physics and electrical engineering.

After a year of post-graduate study at the Charlottenburg Polytechnic Institute in Berlin, Germany, Dr. Rust spent a year as an instructor in mathematics at Harvard University. He was then employed as a research geophysicist by the Humble Oil and Refining Company in 1934, and was made Chief of their Geophysics Research Section in 1937. His work with this group was concerned with the problem of improving or devising new equipment and techniques for obtaining information of the deeper formations of the earth's crust, as a guide to the drilling of wells to discover new deposits of oil and gas.

During World War II, Dr. Rust was a consultant to Division II, NDRC, dealing with problems related to the effect of subterranean explosions. As Humble's representative under contracts with the Radiation Laboratory, MIT, Committee on Propagation of NDRC, Texas University, and other organizations, he was responsible for work ranging from radar components to plane-to-plane fire control.

Dr. Rust was President of the Society of Exploration Geophysicists in 1944, and as such represented the geophysical industry in the subsequent allocation proceedings of the Federal Communications Commission. Since the establishment, by the FCC, of the Petroleum Radio Service, he has been active in the industry committees concerned with the problems of this service. He has participated in numerous allocation and rule-making procedures of the FCC. He is a member of the American Petroleum Institute-Control Committee on Radio Facilities and assisted in the organization of the National Petroleum Radio Frequency Co-ordinating Association.

Dr. Rust holds a number of U. S. and foreign patents in the fields of seismic prospecting and electrical well logging. He is a member of a number of technical societies including the American Mathematical Society, American Physical Society, American Institute of Mining and Metallurgical Engineers, American Association of Petroleum Geologists, and the American Geophysical Union. He became a Senior Member of the IRE in 1947. Dr. Rust is a charter member of the Houston Section of the IRE and is at present the Regional Director for Region Six.

The IRE Professional Groups and the Institute

BENJAMIN B. BAUER

In order further to increase the value to IRE members of The Institute of Radio Engineers, the establishment of the IRE Professional Groups was authorized by the Board of Directors. The results of this action have, up to the present, been encouraging. The number of the Professional Groups steadily grows, and their activities are being thoughtfully and constructively expanded.

Since the success of the Groups is of vital interest to the IRE members, the following significant guest editorial from the Chairman of the IRE Professional Group on Audio, who is as well vice-president and chief engineer of Shure Brothers, Inc., in Chicago, is of particular and timely interest.—*The Editor.*

The rapid expansion of the sciences of communication and electronics into new fields has confronted the Institute with numerous problems. Inevitably, the professional interests of one or more groups of members have become temporarily overlooked. Consider, for example, the field of Audio Technology: To convey intelligence audibly to the sense of hearing is one of the most important end objects of electronic instrumentation. Nevertheless, of the 177 papers published in the PROCEEDINGS during the calendar year of 1950, only three titles relate to Audio. Similarly, on the Institute Section level, the tendency has been to pay court to the technology predominant in the particular locality, resulting in neglect of the interests of a smaller, though important segment of local membership.

Thus it is seen that occasional gaps have taken place in the technical services rendered by the Institute to various groups of members. The IRE Professional Groups are intended to bridge these gaps. The formation and evolution of Professional Groups is a tribute to the power of adaptability which the Institute shares with the industry which bears its name. In recognition of the growing complexity and extension of the fields encompassed by the Institute, the Professional Groups have been created within the framework of the IRE to bring together and serve members with common interests in specialized aspects of the Profession.

The Professional Group on Audio, for example, has been formed for the purpose of serving the professional interests of the IRE members concerned with Audio Technology. In the beginning, the Group pursued this objective by sponsoring sessions on Audio at various IRE meetings and conventions. Soon after its inception, the Group started issuing a NEWSLETTER to keep the members up to date regarding the activities of the Group, and including news and program notes. Beginning with the Fall of 1950, the Group began mailing to its members reprints of technical papers presented at the Audio Sessions which could not have been published promptly in the PROCEEDINGS.

Recently the NEWSLETTER has been improved and expanded. With the July, 1951 issue, a new series of technical editorials covering Sound Reproduction has been started. Each editorial is written especially for the NEWSLETTER by some outstanding authority. The first of the series, by L. L. Beranek of the Massachusetts Institute of Technology, is entitled, "Design of Loudspeaker Grilles." The September issue carries an Editorial by H. F. Olson of the RCA Laboratories dealing with, "Selection of a Loudspeaker." It is contemplated that this series will be followed by another, dealing with studio acoustics and sound transmission. A third series will deal with the problems in sound reinforcement and public address. The NEWSLETTER has also opened its pages to the Sectional Professional Groups on Audio, helping them to voice and solve their mutual problems. The first article dealing with this subject, in the July, 1951 issue, is by S. L. Almas, of the K.L.A. Laboratories Incorporated, Chairman of the Detroit Section PGA. Plans are afoot, stemming from a suggestion by A. B. Jacobson, Chairman of the Seattle Section PGA, to record on tape and distribute "tapescripts" of important talks on Audio to Sectional Audio Groups.

Thus, the Professional Group on Audio, along with other Professional Groups is rapidly becoming an integrated part of the whole which constitutes the IRE. In future years, the IRE historians may well agree that the creation of Professional Groups has been one of the most significant events in the history of the Institute.

Using Tests to Select Engineers*

WARREN G. FINDLEY†

Summary—**Experience in selecting students for admission to undergraduate engineering colleges provides a clear outline for a program of tests and related procedures that should prove helpful in identifying potential engineering talent early in high school. Qualified students may then be guided toward adequate preparation for engineering training. Such a program would include the following as a basic minimum: the students' average grades, tests of mathematical aptitude, reading comprehension, spatial visualization, and interest inventories.**

A recently revised program of examinations are now available for selecting students for graduate study in engineering. Research is being undertaken which gives promise of the development, in the measurable future, of a means of detecting creative talent for scientific research.

When Are Tests Helpful?

The value of standardized tests for selecting engineers must be judged by the extent to which they improve selection over what can be done without tests by using other information routinely available or readily obtainable. That standardized tests will distinguish between superior and inferior applicants for engineering training or employment is not sufficient. Such tests must do the job better or add to what can be done by other methods (i.e., ready-made tests or evaluation of previous school records).

We may illustrate this point graphically by reference to Fig. 1. In the cross-hatched bottom line of each set of three show how well the five engineering colleges, by using the averages of high-school grades alone, can predict whether their students will make average grades or better. Thus, enrolled engineering freshmen whose high-school averages were in the top 4 per cent of the averages of all their classmates had 84 chances in 100 of doing average or better work in the first term of engineering college; those who stood in the next 12 per cent with respect to high-school averages had 75 chances in 100 of doing average work or better; and those who stood in the lowest 4 per cent with respect to high-school averages had only 16 chances in 100 of doing average work or better in the first semester of engineering college. The added effectiveness of prediction attained by using two mathematical tests is shown by the black lines in the same charts. Those who were in the top 4 per cent on a weighted composite based on high-school averages and scores on the two tests had 97 chances in 100 of doing average work or better in first-semester engineering; those who were in the next 12 per cent on the weighted composite had 88 chances in 100 of doing above average work in the first semester; while the bottom 4 per cent with respect to the composite had only 3 chances in 100 of doing average or better work during the first semester.

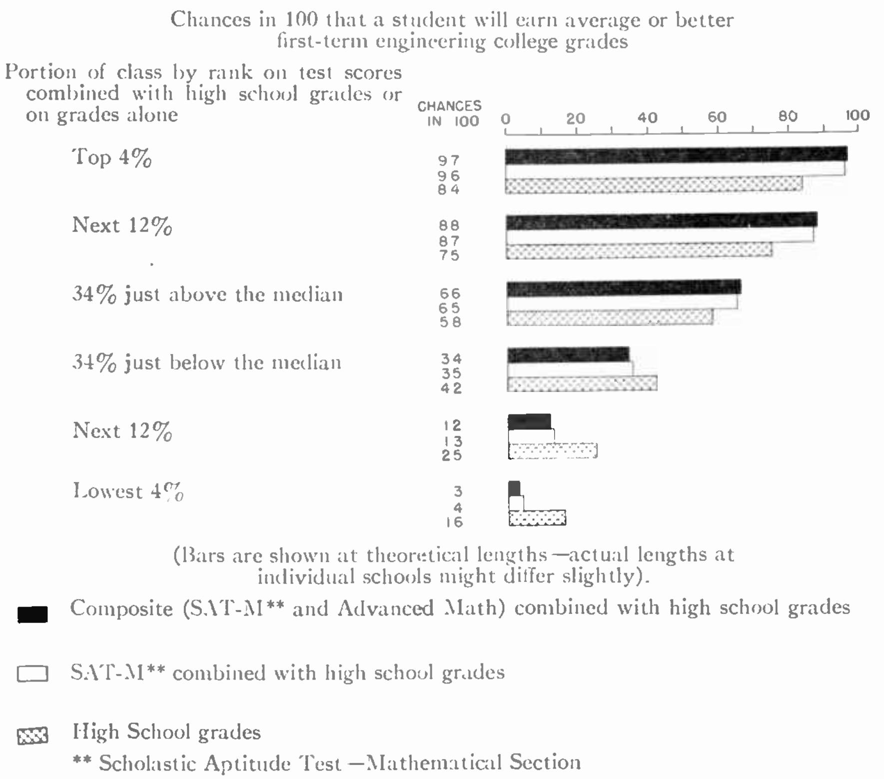


Fig. 1—Prediction of scholastic success in a group of five engineering colleges by college board test scores combined with high school grades.[1]

A corollary to the preceding statements is that standardized tests are particularly helpful where other sources of information about applicants for engineering training or employment, such as school records, do not afford comparable data on all applicants. This would apply especially to applications for graduate study in engineering at a university or applications for junior professional employment in a large engineering organization which are received from undergraduate colleges of all sorts with varied grading systems.

A second general proposition, applicable especially in these days of shortages of engineers and engineering students, is that tests and other information about applicants are most useful when the number of applicants is large relative to the number of openings for training or employment. In other times than these the author might be expected to devote a relatively large part of his remarks to selection for graduate training and employment because of their more natural place in the thinking of practicing engineers. These problems will be mentioned and discussed in their place. But the problem of the moment in selecting engineers is in identifying, at an early stage, potential engineering ability in high-school boys so that they may be guided into pre-engineering study. And so, confident that relations prevailing at the college entrance level of selection are most pertinent to the problem of selection (identification) in early high school, we turn to recent recorded experience in selecting students for undergraduate engineering colleges.

What Tests Help Most?

Over and above what can be ascertained by scrutiny of an applicant's previous school record, the most significant single factor to measure by tests is mathematical aptitude. This has been found consistently in studies of the Pre-Engineering Inventory (produced for the Measurement and Guidance Project in Engineering Education, sponsored jointly, since 1943, by the Engineers' Council for Professional Development and the American Society for Engineering Education), the American Council on Education Psychological Examination, the Engineering and Physical Science Aptitude Test, and the examinations of the College Entrance Examination Board.

A set of difficult arithmetic reasoning problems will serve very well. Additional problems in the fundamentals of algebra and geometry, though helpful, are not strictly necessary. The greatest contribution of the latter is that they require less reading time than corresponding arithmetic reasoning exercises. More advanced mathematics should be tested only if all applicants may be presumed to have studied approximately

* Decimal classification: R070. Original manuscript received by the Institute, March 29, 1951.

† Director of Test Development for the Educational Testing Service, Princeton, N. J.

[1] Data, courtesy of Dr. W. B Schrader, for 721 enrolled engineering freshmen tested during their first week in the Fall of 1948 at Carnegie Institute of Technology, Cornell University, Lehigh University, Rutgers University, and the University of Pennsylvania.

the same subjects. But in all cases the problems should be kept short, be based on the major topics of instruction, and require ingenuity in solution rather than memory of formulas or standard processes of solution. One writer has put it this way: "(A mathematical aptitude test) provides a measure of the slope of the student's learning curve on the subject, and success in freshman mathematics appears to be predictable from a knowledge of this slope regardless of the actual position on the curve." Of course, insofar as school records are lacking and certain formal mathematics training is presumed as a base for more advanced instruction, the above statement would have to be modified to include assessing the position as well as the slope. But when exposure to certain mathematical concepts is clearly indicated by academic records, the slope is the determining factor. In the study summarized in Fig. 1, the mathematical section of the Scholastic Aptitude Test, consisting of short arithmetic, algebra, and geometry problems, showed as good or slightly better correlation with mathematics grades and grade averages in engineering than did the test of corresponding length in advanced mathematics.

The second most generally significant factor to test in potential engineers is the ability to comprehend and interpret scientific reading material and data. Tests of this nature appear in most programs used in selecting or guiding potential engineers. They vary from tests of general reading comprehension in the humanities and social sciences as well as in the natural sciences to tests designed especially to include scientific passages and material presented in tables, graphs, and diagrams. Tests of ability to interpret quantitative data accurately and logically and to make reasonable inferences, deductions, and generalizations from data are in increasing demand.

Beyond these two major aptitudes a variety of other measures are useful for selective purposes whenever certain subjects may be presumed to have been studied by all applicants. For example, if a year of physics has been required of all applicants for entrance to an undergraduate engineering course, basic problems in that subject may well be set as part of the selective process.

Success in descriptive geometry, engineering drawing, surveying, and the like is especially dependent on abilities reflected in spatial-relations tests. Tests of comprehension of mechanical principles or movements in pictorially presented situations also add to our understanding of the applicants' background abilities, and in some measure improve the selection for engineering training.

If the task of selection is conceived more broadly as that of selecting intellectually and socially well-rounded persons for eventual admission to the engineering profession, tests like the one in the Pre-Engineering Inventory on Understanding Modern Society should be considered. At the graduate entrance level the so-called Profile Tests of the Graduate Record Examination and the Cooperative General Culture Test serve a similar purpose in measuring general understanding outside of one's immediate professional field.

It is perhaps worthy of special mention here that the Medical College Admission Test, required of candidates for admission to institutions in the Association of American Medical Colleges, includes a special section on Understanding Modern Society. Although this section will not approve prediction of success in medical training, the medical colleges' support it strongly because it will indicate a breadth of viewpoint important in professional practice.

Further extension of the concept of desirable general characteristics should lead to the inclusion of measures of skill in human relations. At present no established group tests of emotional stability or personal adjustment can be recommended for general use although many promising instruments are effectively used by clinical psychologists in individual diagnosis. Studies of how well individual students are accepted by their contemporaries and chosen for leadership may well lead to the development of procedures that would permit objective statements about personal adjustment and leadership qualities. Ability to direct or supervise others would be especially worth exploring.

Advanced Tests

Recently, under the stimulus of the National Research Council's need for an advanced test in engineering for candidates for Atomic Energy Commission Fellowships, the so-called Advanced Test in Engineering of the Graduate Record Examination has been revised and expanded by a committee of five professors of engineering, nominated by the American Society for Engineering Education. Now a 3-hour examination, it consists of 100 general engineering problems followed by four groups of 25 difficult problems in the four chief traditional branches of engineering training; these latter sections are clearly labeled so that each examinee can start in the section in which he feels most competent. The more limited previous test proved useful in predicting success in graduate study of engineering; however, the revised test should provide a more searching, better-balanced instrument of selection and guidance.

The Advanced Test in Engineering is only one segment of the program of examinations developed for the National Research Council's program under the auspices of the Educational Testing Service. Corresponding tests in mathematics, physics, chemistry, biology, and geology have been developed by committees of outstanding graduate professors. A high-level aptitude test yielding separate verbal and mathematical aptitude scores is included. All these tests and the Profile Tests are available in the Graduate Record Examination programs for selection, appraisal, or guidance of graduate students in engineering.

Interests

For many years interest inventories have been widely accepted for use in guiding students and graduates into the broad areas of curriculum or employment. Such inventories are helpful when responses are given in a spirit of unbiased self-exploration. If used for selection in a single field, they are readily faked. Interest inventories, in conjunction with aptitude tests, should prove especially useful in identifying potential engineering talent early in secondary school.

The most widely used of these tests are the Strong Vocational Interest Blank and the Kuder Preference Record. The Strong Vocational Interest Blank, based on a careful study of the interests of men well established in their professions, has been in use since 1927 and has stood up well. The base group in engineering involved over 500 members of the engineering societies. The Kuder Preference Record is a relative newcomer, first published about 1940. It has been widely used in secondary schools, and an increasing number of studies show that its more general categories of interest (mechanical, computational, scientific, persuasive, literary, artistic, musical, social service, clerical, and outdoor activity) lend themselves to interpretations similar to those derived from the Strong Vocational Interest Blank. In addition, it has a handy self-scoring format. Both inventories suffer somewhat from the limitation of requiring an adult vocabulary. It is also true that interests change during adolescence, and become relatively more fixed at 18 or so. Experimental study of such variations might well be made an integral part of any major guidance program in secondary schools.

The effectiveness of these interest measures in counseling will depend considerably on the breadth of the counselor's background and the amount of additional information he has available on each student. As one example, a student who shows strong interest in science and has adequate mathematical ability might show secondary interest and ability in art. An alert counselor could point out the prospect of combining these talents in industrial design or architectural engineering.

New Approaches

Much is being said nowadays, and rightly, about the importance of identifying creative talent, or competence for scientific research. Instruments and procedures for this purpose have just begun to be developed. Work during World War II on selecting pilots for the United States Air Force led to confidence that certain elements of biographical information can predict success in training and leadership. This approach is being applied to certain kinds of scientific personnel in special studies. It depends on finding enough discriminating items of information from the testing of large numbers thought to be possibly related. In the case of Air Force pilots certain items about participation in sports and hobbies, items of specialized information that could only be gained from active participation in such activities as flying, driving an automobile, hunting, and the like, proved useful, when enough were brought together, in predicting pilot success.

Another approach is being made in a study by the American Institute for Research under the auspices of the Office of Naval Research. Descriptions have been secured from research personnel of critical incidents which led them to judge other research workers either outstandingly effective or downright inept. Aptitude tests have now been built based on 36 categories of types of critical incidents, and these may be found under the following major headings: formulating problems and hypotheses, planning and designing the investigation, conducting the investigation, interpreting research results, preparing reports, administering research projects, accepting organizational responsibility, and accepting personal responsibility. In the course of a few years' time these tests will be validated against measures of productive effectiveness.

Other tests being experimented with include one in which the examinees are asked to describe "what would happen if" some fundamental change in the physical or biological laws were to transpire. Another requires the examinee to show a fluency of ideas by classifying 25 items of a list into as many sets of four as can be identified with any basic principle in science. This latter test is based on the theory that creative research demands, among other talents, an ability to conceive and consider great numbers of possible relationships very rapidly. Still another type of test poses the question, "Do you have enough data to solve?," rather than merely "Solve the following," as evidence of ability to grasp the essential features of problems and of willingness to pronounce tough-minded judgments; this is in contrast to the usual requirement of ingenuity in operations.

Our greatest hopes are that these and other research studies will reveal basic intellectual and motivational factors which are not measured in current tests but are useful in identifying outstanding scientific and engineering talent.

Acknowledgments

The author wishes to express his indebtedness to A. P. Johnson of the Educational Testing Service for data and references used in preparing this paper and to E. M. Rickard for aid in digesting the references and preparing the bibliography.

Annotated Bibliography (Arranged Chronologically)

1. H. S. Rogers and G. W. Holcomb, "Inventory of engineering motives," *Jour. Eng. Ed.*, vol. 22, pp. 851–860; June, 1932.

   This early biographical inventory was designed to reveal the "intensity of motives" in respect to:
   1. "Constructive Instinct"
   2. Scientific Curiosity in Mechanisms
   3. Abstract Ideas of a Quantitative Nature
   4. Precise Observation and Description of Physical Things
   5. Practical Application
   6. Planning and Organization
   7. Manual Activity and Craftsmanship.
2. W. V. Bingham, "Aptitudes and Aptitude Testing," Harper and Bros., New York, N. Y.; 1937.

   Part I includes a broad discussion of aptitude, intelligence, and interest. In Part II, in discussing the field of engineering, attention is focused on the following:
   1. General Scholastic Aptitude
   2. Aptitude for Mathematics
   3. Aptitude for Thinking about Space Relations
   4. Aptitude for Understanding Mechanisms
   5. Aptitude for Mastering Physical Sciences.

   The statement is made that superiority in tests known to be indicative of these aptitudes, as well as a liking for engineering work, the necessary health, and constancy of purpose, indicate a high probability of success. Low scores should not be construed as barring an engineering career but as warning signals.

   The book contains a very complete appendix of representative tests and interest schedules.
3. I. L. Kandel, "Professional aptitude tests in medicine, law, and engineering," Bureau of Publications, Teachers College, Columbia University, New York, N. Y.; 1940.

   This book reviews the early development of tests for engineering ability, and the part taken by the Society for the Promotion of Engineering Education and the Engineers Council for Professional Development.
4. E. N. Brush, "Mechanical ability as a factor in engineering aptitude," *Jour. Appl. Psych.*, vol. 25, pp. 300–312; 1941.

   The conclusion is reached that mechanical ability, as measured by several tests studied, may be regarded as a component of engineering aptitude, but the actual predictive power of most single tests of mechanical ability is not great.
5. C. H. Griffin and H. Borow, "An engineering and physical-science aptitude test," *Jour. Appl. Psych.*, vol. 28, pp. 376–387; 1944.

   This describes a battery of 6 tests developed to assess suitability for training in technical work on the college leve. It states that multiple correlations between the weighted battery of test parts and course achievement were as high as 0.79.
6. J. P. Guilford and J. I. Lacey, Ed., "Printed Classification Tests," Report No. 5 of AAF Aviation Psychology Program Research Reports, U. S. Government Printing Office, Washington, D. C.; 1947.

   Chapter 27 on Biographical Data summarizes the most significant research prior to United States Air Force experimentation, including aeronautical and industrial sources.
7. H. A. Edgerton, S. H. Britt, and R. D. Norman, "Later achievements of male contestants in the first annual science talent search," *Amer. Scien.*, vol. 36, pp. 403–413; 1948.

   The article describes briefly the first annual science talent search sponsored by the Westinghouse Electric Corporation, a program that has been carried on each year since, and it discusses the follow-up study of the later achievements of the contestants. Other pertinent references to the program are given at the end of the article.
8. E. C. Koerper, "Milwaukee plan of aptitude testing," *Mech. Eng.*, vol. 70, p. 750; September, 1948.

   This outlines a plan to develop the following:
   1. Test Batteries
   2. Interview Forms
   3. Job Specification System
   4. Employee Progress Appraisal Methods.
9. D. B. Stuit, G. S. Dickson, T. F. Jordan, and L. Schloerb, "Predicting Success in Professional Schools," Amer. Council on Ed., Washington, D. C., 1949.

   Chapter II summarizes detailed research findings about predicting success in engineering schools.
10. D. E. Super, "Appraising Vocational Fitness," Harper and Bros., New York, N. Y.; 1949.

    This is a general manual of tests and testing. The discussion of interest inventories, personality measures, and biographical data blanks is particularly good because of the author's experimental studies in these areas.
11. W. T. Donahue, C. H. Coombs, and R. M. W. Travers, Eds., "The Measurement of Student Adjustment and Achievement" (A collection of papers presented at a conference of guidance workers), University of Michigan Press, Ann Arbor, Michigan; 1949.

    A paper, entitled "Significant research on the prediction of academic success," by R. M. W. Travers points out that tests to predict success should be given as early in the student's career as possible Studies summarized indicate that predicted grades in the first 2 years of engineering school may be expected to correlate between 0.6 and 0.7 with actual grades if predictions are made from tests in mathematical and scientific fields. A bibliography of 272 studies is given.

    This volume also contains pertinent material in a paper entitled "Problems in guidance in vocational and technical training," by G. K. Bennett.
12. L. W. Guth, "Discovering and developing creative engineers," *Mach. Design*, vol. 21, pp. 89–94; March, 1949.

    This article describes tests used in a survey conducted at the General Electric Company. The tests, designed for selection of creative engineering talent, attempt to appraise adaptive inventiveness, design sense and perception, and reasoning power Solution of the problems is dependent on application and ingenuity rather than on memory of formulas, which are supplied where needed.
13. J. C. Flanagan, "Critical Requirements for Research Personnel," Amer. Instit. for Research, Pittsburgh, Pa.; March, 1949.

    This describes the first phase in a broad program aiming to develop tests and evaluation procedures for research personnel for the Office of Naval Research. (The same program is referred to under M. H. Weislogel in reference 16.)
14. E. J. Riegel, "Evolution of math tests," *College Board Rev.*, vol. 1, p. 95, New York, N. Y.; May, 1949.

    This includes a description of a validity study made in cooperation with five large engineering schools.
15. O. W. Eschbach, "Report of the committee on selection and guidance," *Jour. Eng. Ed.*, vol. 40, pp. 112–119; October, 1949.

    This is a fairly recent report on the testing program of the Measurement and Guidance Project in Engineering Education, sponsored jointly, since 1943, by the Engineers' Council for Professional Development and the American Society for Engineering Education. During the year 1946–1947, 32,098 students were tested in this program.
16. M. H. Weislogel, "The Development of a Test for Selecting Research Personnel," Amer. Instit. for Research, Pittsburgh, Pa.; January, 1950.

    This report describes test items that have been pretested and put into final form available for use in the later phases of a long-range research program for the Office of Naval Research. (The same program is referred to under J. C. Flanagan in reference 13.)
17. F. Lord, J. T. Cowles, and M. Cyna-

mon, "The pre-engineering inventory as a predictor of success in engineering colleges," *Jour. Appl. Psych.*, vol. 34, pp. 30–39; February, 1950.

The analyses indicate that the Pre-Engineering Inventory Composite Score performs very satisfactorily, on the whole, as a predictor of engineering school success.

18. R. F. Berdie and N. A. Sutter, "Predicting success of engineering students," *Jour. Ed. Psych.*, vol. 41, pp. 184–190; March, 1950.

Report is made on a study of 1,019 freshmen in the Institute of Technology of the University of Minnesota.

19. A. P Johnson, "Tests and testing programs of interest to engineering educators," *Jour. Eng. Ed.*, vol. 41, pp. 277–283; January, 1951.

The author reviews recent activities under the Measurement and Guidance Project in Engineering Education and the current problems in finding prospective engineers and selecting engineering students, including selection at the graduate level.

Fundamentals of Secondary Electron Emission*

MARTIN A. POMERANTZ† AND JOHN F. MARSHALL†

This paper has been secured by the Tutorial Papers Subcommittee of the IRE Committee on Education as a part of a planned program of publication of valuable tutorial material. It is here presented with the approval of that Subcommittee.—*The Editor.*

Summary—**Secondary electron emission is of great importance to the physicist because of its bearing upon the problem of the interactions between fundamental particles and to the radio engineer because of its applications, as well as its effect, upon the operation of electronic tubes. A complete theoretical picture capable of accounting quantitatively for all observed phenomena does not exist. Secondary electron emission differs from other modes of emission in many respects. The essential characteristics can best be evaluated by considering a typical experimental arrangement for investigating the phenomenon. Three categories of emitted electrons are recognized. The yield may depend upon various factors, such as the primary energy, collector voltage, target temperature, time, angle of incidence, atomic properties of target, and the composition of the target.**

The difficulties of propounding a satisfactory theory are evident from an individual consideration of each of the various processes involved. The primary interaction, primary energy loss, escape of secondaries, and integration over the range of the primary must each be treated to arrive at a final solution. In several previous attempts at formulating a theory, only the most loosely bound electrons in the solid have been regarded as constituting the source of secondary electrons. Normalizations are required for comparison of the results with existing experimental data. There are cogent reasons for regarding bound electrons as a very important source of secondaries. The probability of ionization in gases exhibits the same general dependence upon primary energy as secondary electron emission, and this resemblance suggests a possible model for secondary emission based upon detailed considerations of primary ionization probabilities.

I. Introduction

WHEN A SOLID BODY is subjected to bombardment by electrically charged particles, some electrons which may be detectable under suitable circumstances are always emitted. Although this process, commonly designated "secondary electron emission," has been observed to occur in various forms, by far the most widely investigated type is that in which an electron beam falling upon the surface of a target in a vacuum causes the emission of a stream of electrons from the surface upon which it impinges. It should be emphasized, however, that this variety of secondary emission is not endowed with any intrinsically greater significance than any other. Rather, the distinction arises solely from the geometrical and practical circumstances that in this case the phenomenon is readily observable and is, in fact, involved in the operation of common electronic devices.

In the field of radio engineering, secondary electron emission originally manifested itself only as a source of annoyance which seriously interfered with the satisfactory functioning of vacuum tubes. The problem was solved by the addition of the suppressor grid or its equivalent to the tetrode; after the advent of the pentode, the effect received relatively little attention. In recent years, however, successful attempts, both unintentional and conscious, have been made to utilize the phenomenon to some advantage. It is now generally realized that secondary electron emission is inherent in the operation of a cathode-ray tube. Devices such as the magnetron and certain types of reflex klystron depend upon secondary electron emission for their high output capabilities. In dynatron and photomultiplier tubes, secondary electron emission constitutes the fundamental principle of operation.

Electron emission may be divided into four principal categories: (1) thermionic emission, (2) photoelectric emission, (3) field emission, and (4) secondary emission. Despite the radical differences among these modes of emission, certain similarities exist among the first three. In general, satisfactory theories have been formulated (although complicated systems, such as activated

* Decimal classification: R138. Original manuscript received by the Institute, July 23, 1951. Supported by the Office of Naval Research.

† Bartol Research Foundation of the Franklin Institute, Swarthmore, Pa.

barium-strontium oxide-coated cathodes, have been the subject of considerable controversy) despite the fact that they may be unable to account for all observed phenomena quantitatively (as in the case of the photoelectric effect). Nothing resembling a complete theoretical picture of secondary electron emission exists up to the present time, nor is it likely that an all-inclusive model can be developed in the light of the limitations of our present grasp of physics of the solid state.

The most striking characteristic of secondary emission as contrasted with the other types is the similarity of behavior over a wide variety of materials. The range of values of yields encountered is quite limited as compared with thermionic or photoelectric yields. The nature of the dependence of secondary emission upon the work function of the surface differs appreciably from that in the other cases, and the yield is actually quite insensitive to the nature of the barrier. At first sight it might be expected that secondary emission is more closely related to photoelectric emission than to thermionic emission. This resemblance, which hapens to be valid in a certain sense for rather subtle reasons, cannot be accorded much significance upon closer scrutiny. Whereas photoelectric emission from a solid is primarily a surface effect, this is certainly not the case for secondary emission. A single photoelectron absorbs all of the energy of the incident quantum $h\nu$, and hence the situation is dominated by the work function ϕ. On the other hand, each secondary electron absorbs only a small fraction of the energy of the primary which may penetrate a considerable distance into the target material. Thus, we are here concerned with a combination of volume and surface effects.

From the standpoint of tube engineering, the operational problems associated with secondary electron emitters are to some extent somewhat less complicated than those encountered in applications utilizing thermionic emission. It is appropriate to emphasize at this time that the generalizations to which the discussion in this paper will be confined are occasionally subject to exceptions. In the event that these are not specifically mentioned, it should not be tacitly assumed that the broad statements are necessarily all-inclusive. For example, there are certain specific applications for which the requirements of reproducibility and long-time stability of specially prepared targets present very difficult technical problems. Except for a few composite surfaces with certain desired characteristics, activation procedures are not involved. As in the case of thermionic emitters, there are two general classes of materials which are useful as a source of electrons, namely, metals and semiconductors. A third class, broadly termed "insulators" for lack of a less ambiguous designation (note that semiconductors under some conditions are included in this group), is also of importance primarily because of practical applications, although the first two present greater theoretical interest.

II. Basic Experimental Concepts

The most direct approach to an evaluation of the significant characteristics of secondary electron emission is to consider a typical experimental arrangement for investigating this phenomenon. Fig. 1 is a schematic

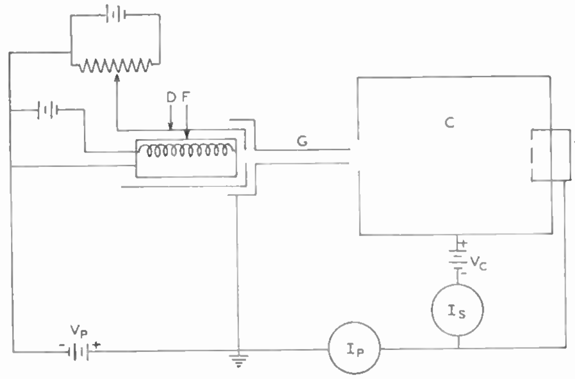


Fig. 1—Schematic diagram of a typical experimental arrangement for measuring secondary electron emission.

diagram of such an apparatus. It consists essentially of an electron gun G serving as the source of a beam of primary electrons I_p which, after acceleration through a difference of potential V_p, bombards the target T. The secondary electrons I_s leaving the target are then attracted to the collector C, owing to the presence of the positive voltage V_c applied to the collector with respect to the target.

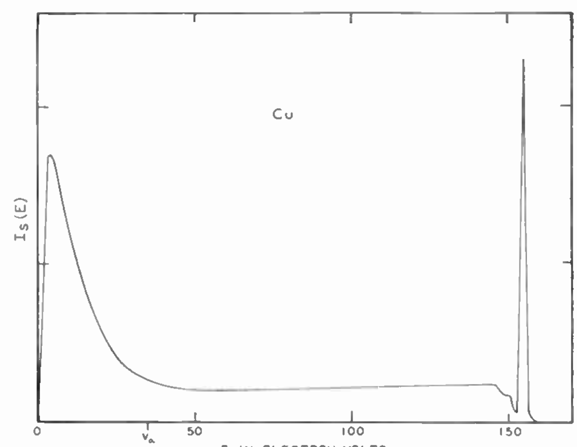


Fig. 2—Typical energy distribution of secondary electrons. This particular curve, obtained by a method utilizing a transverse magnetic analyzer, refers to a Cu target. (See E. Rudberg, *Phys. Rev.*, vol. 50, p. 138; 1936.)

The yield δ is defined as the number of secondaries *collected* per incident primary, given by the ratio I_s/I_p. It should be noted that this is a purely empirical definition, but nevertheless one of practical interest in any event. Some of the primary electrons are directly reflected, whereas others are scattered with some loss of energy. These do not constitute true secondary electrons which have actually been knocked out of the target by an impinging primary. It is practically impossible to distinguish between true secondaries and primaries reflected after suffering energy losses, and this is indeed unfortunate from the point of view of obtaining theoretically useful experimental data. However, the distinction is unnecessary as regards obtaining "emission" from a target regardless of the origin of the "emitted" electrons.

It has been stated that the secondary electrons emitted by a material bombarded by a primary beam can be ascribed to three different mechanisms. This classification is made on the basis of the energy distribution of collected electrons, which is obtained by applying retarding potentials $-V_c$ to the collector with respect to the target, taking into account the contact difference in potential between the surfaces of these two electrodes. As is evident in Fig. 2, there is a sharp distinguishable peak at the energy of the incident beam corresponding to eV_p, and these electrons are, of course, to be identified as elastically reflected primaries. At the other end of the spectrum a group of slow electrons may be observed. The average energy of these electrons is only a few ev when the primary energy is of the order of hundreds of ev. Between these electrons and the elastically reflected primaries there is an intermediate group which results principally from inelastic scattering in the lattice. The slow electrons arise from collisions between the primaries and the atomic electrons of the target in which sufficient energy is transferred to the latter so that they can penetrate to the surface and emerge from the solid material. These electrons are therefore true secondaries; although most possess low energy, it is certainly not warranted, however tempting, to assign an arbitrary upper limit, such as V_s, above which no true secondaries appear and below which scattered primaries are prohibited.

Various factors in addition to the nature of the target material may affect the magnitude of the yield of secondary electrons. These will be discussed briefly in a general manner, and certain special situations will be mentioned.

A. Primary Voltage V_p

All known secondary emitters manifest the same qualitative dependence of yield upon primary energy. Starting at low voltage, the yield rises smoothly until a maximum value, often in the neighborhood of 400 to 600 volts, is attained. Thereafter, the yield decreases slowly and may approach a more or less constant value at very high energy. A typical curve is shown in Fig. 3. The maximum value of the yield, δ_{max}, is often cited, probably because it is convenient, at least from the practical point of view and for some purposes, to know the highest multiplication which could be expected under optimum conditions. The corresponding voltage is designated $V_{p_{max}}$. Although δ_{max} and $V_{p_{max}}$ alone are not necessarily of any fundamental theoretical significance, it is interesting that at least in the case of metals a universal curve which fits the available data within the experimental errors is obtained by applying a

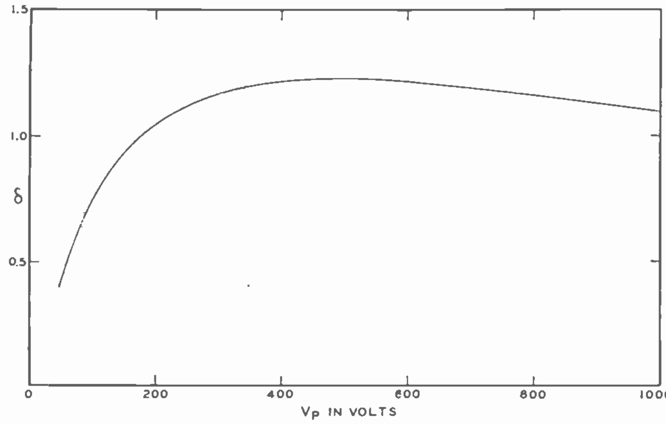


Fig. 3—Dependence of secondary electron emission upon primary bombarding energy for a typical metal, in this case Ni.

normalization in which the ratio δ/δ_{max} is plotted as a function of $V_p/V_{p_{max}}$. This is shown in Fig. 4. The shape of the *yield versus energy* relationship can be accounted for, at least qualitatively, as will be described in the following section.

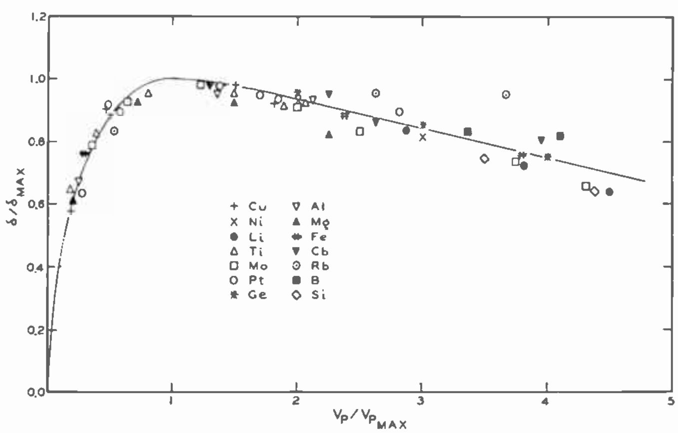


Fig. 4—Normalized yield, δ/δ_{max}, plotted as a function of normalized primary energy, $V_p/V_{p_{max}}$, for various elements. This is the so-called universal curve[5] of secondary electron emission.

Other values of primary voltage sometimes cited because of practical considerations, particularly in the case of insulators, are the two crossover points, in the δ versus V_p plot, at which the yield attains the value unity. It is evident that if an insulator is subjected to electron bombardment the surface will charge negatively, as long as the yield is less than one secondary per incident primary, until it approaches the cathode potential, thereby effectively reducing the primary bombarding energy. When the yield exceeds unity, between the so-called "lower and upper sticking potentials" a positive charge is acquired by the surface, thereby reducing the effective collector voltage V_c until the measured yield approaches one. Above the upper crossover voltage, the surface becomes negatively charged until the yield again approaches unity.

B. Collector Voltage V_c

In the case of metals, the secondary current is independent of the collector voltage as long as this is positive. For substances with lower conductivity, such as certain oxides, on the contrary, an increase of yield with increasing collector voltage is sometimes observed,

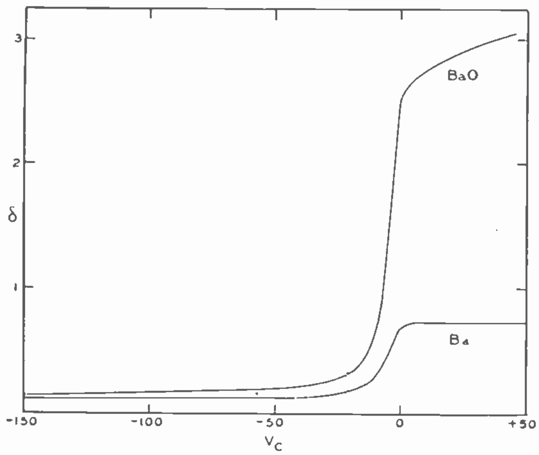


Fig. 5—Dependence of yield upon collector voltage, illustrating lack of saturation[3] in the case of BaO. (See H. Bruining and J. H. De Boer, *Physica* (The Hague), vol. 6, p. 823; 1939.)

as is illustrated in Fig. 5. This effect may be qualitatively understood when it is recognized that by virtue of a yield larger than unity more electrons leave the emitter than enter. As a consequence of the high resistance of the oxide layer, a positive charge will appear on the outer surface and a negative charge on the inner surface, thereby forming a double layer. The resulting field promotes a phenomenon analogous to cold emission. The dependence upon collector voltage arises from the fact that the potential on the outer surface of the oxide layer is limited to values smaller than the collector voltage. An increase in the latter is accompanied by an increase in the internal field.

C. Temperature of Target

The secondary electron emission from metallic materials is not ostensibly dependent upon the temperature of the target, at least in any fundamental manner. Actually, changes in the nature of the surface layer or in the crystal structure may be introduced by heat treatment, thereby producing only small changes in secondary emission. In the case of at least one semiconducting medium, which in particular has been investigated rather extensively because of its great practical importance—the so-called oxide-coated cathode—, appreciable variations with temperature do occur.[1,2]

D. Time

At least in the case of metals, there are no essential changes of secondary electron emission with time, except for obvious consequences of structural changes which the target surface may undergo during the lifetime of the tube. Specialized emitters consisting, for example, of thin films of aluminum oxide on an aluminum base, with an outer layer of Cs, operating by virtue of the so-called "Malter Effect,"[3] are capable of emitting thousands of electrons per bombarding primary. This phenomenon, also termed "thin-film field emission," is produced by an extreme manifestation of the mechanism already described in Section II B. It is evident that it might be anticipated from the nature of the process that the maximum emission is not attained until the primary beam has been on for some time and, furthermore, that emission may persist after the beam has been turned off; this is quite contrary to the situation with ordinary secondary electron emission which displays no detectable time delays.

E. Angle of Incidence

The data plotted in Fig. 3 were obtained with the beam striking the target surface perpendicularly. As might be expected from the general nature of the processes involved, the yields for oblique incidence are somewhat larger inasmuch as the secondary electrons are formed closer to the surface and are consequently absorbed to a lesser extent before reaching the surface barrier.

For nearly grazing incidence, the yield may be increased by a factor of as much as three, depending upon the primary voltage and the composition of the target. In general, the maximum value of the yield and the energy at which it occurs are both higher, which is consistent at least with qualitative expectations.

F. Atomic Properties of Target

No simple correlation between the secondary yield and the known atomic properties of the target exists, as in the case of other types of electron emission. In some instances, trends which are certainly suggestive are revealed, although the relationships are evidently indirect. For example, the curve in Fig. 6 shows the correlation between δ_{max} and ϕ. The positive slope is opposite that which would be expected at first sight, and the suggestion of a correlation is a consequence of the fact that the work function changes along with some other atomic property which really predominates the second-

[1] M. A. Pomerantz, "Secondary electron emission from oxide-coated cathodes," *Jour. Frank. Inst.*, vol. 241, p. 415; vol. 242, p. 41; 1946.
[2] J. B. Johnson, "Secondary electron emission from targets of barium-strontium oxide," *Phys. Rev.*, vol. 73, p. 1058; 1948.
[3] L. Malter, "Thin-film field emission," *Phys. Rev.*, vol. 50, p. 48; 1936.

ary emission process. Actually, a reduction in ϕ, introduced without alteration of the bulk material, would result in an increase in the yield.

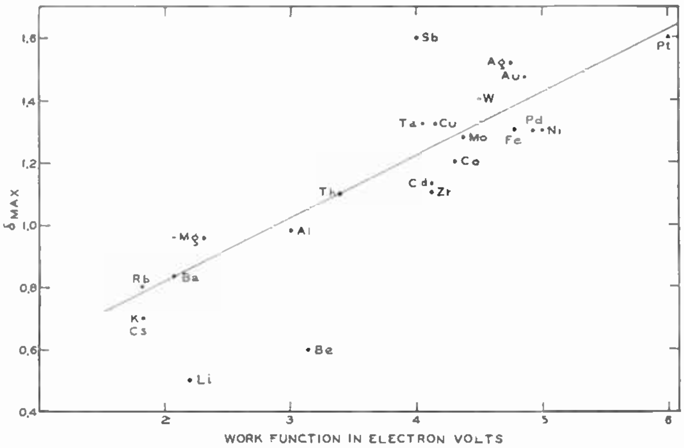


Fig. 6—Correlation between maximum secondary electron emission and work function for various metals.[3a]

Another illustration of the sort of trend which may be exhibited is shown in Fig. 7, where the yield is plotted as a function of the first atomic ionization potential of the target element. Now obviously, this characteristic itself retains no significance when the atoms are brought together to form a solid. However, it is not unreasonable to assume that it may reflect the effective number of conduction electrons in the metal (high-ionization potentials corresponding to small effective numbers) and hence the number of electrons which will impede the progress of an outgoing secondary. Thus, the addition of an electron to the outer shell should decrease the yield. It has not yet been possible to test this hypothesis experimentally because of the inavailability of the requisite data.

G. Composition of Target

It was noted in the introduction that secondary electron emission is unique as contrasted with the other types, in view of the relative insensitivity to the nature of the emitting material. It is indeed remarkable that values of δ_{max} (includingspecial unstable cases of high yields) vary only by a factor somewhat greater than one order of magnitude. It is appropriate to include herewith at least a general statement regarding the values of yield which are encountered in practice.

Secondary electron emitters may be classified into four categories as follows:

(1) Elements: This group includes all chemical elements normally in the solid state for which measurements have been reported. The maximum values of the yield range from about 0.5 to 1.6 for clean surfaces, regardless of the specific nature of the conductivity of the solid.

(2) Compounds: This group includes all chemical compounds which are designated either as semiconductors or as insulators at room temperature. The maximum values of the yield range from aproximately 1.0 to 7.5.

(3) Composite surfaces: These are various complicated systems, sometimes designated "photocathodes" because they are characterized by very high photoelectric sensitivity, usually prepared by evaporating layer upon layer of different materials in a vacuum and by performing other special operations. For example, the notation [Ag]—Cs_2O, Ag—Cs refers to an electrode consisting of a silver base covered with a Cs_2O layer (also containing Ag a toms), on the surface of which Cs atoms are absorbed. The yields range between 3 and 10, in general.

(4) Activated alloys: Certain alloys, for example several per cent Mg with Ag, when "activated" by what appears to be an oxidation procedure, produce yields as high as 18, without any stability. Yields as high as 4 to 5 can be maintained.

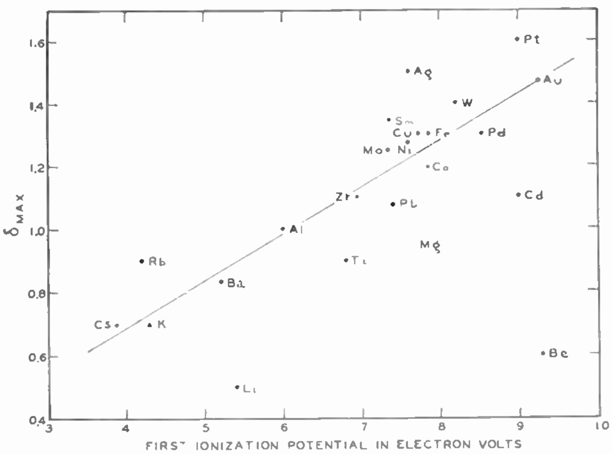


Fig 7—Correlation between maximum secondary electron emission and first atomic ionization potential for various elements.

III. Theory of Secondary Emission

In contrast to the theory of thermionic emission, the theory of secondary emission is in a very unsatisfactory state. This is a consequence of the fact that whereas the problem of thermionic emission can be treated in a fairly straightforward manner by standard techniques of statistical mechanics[4] that of secondary emission is highly complex, involving several processes none of of which is well understood. The difficulties involved become obvious from an examination of the following outline which lists the steps in the computation of the secondary yield from a solid.

A. The Primary Interaction

The problem here is to compute the probability that a primary electron of a given energy will interact with one of the electrons in the solid to produce a sec-

[3a] K. G. McKay, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. I; 1948.

[4] W. E. Danforth, "Elements of thermionics," Proc. I.R.E., vol. 39, p. 485; May, 1951.

ondary electron with sufficient energy to emerge. Our present state of knowledge of solids is such that it is certainly impossible to solve even this very fundamental problem in any exact way. It is not even clear, for example, whether the sources of secondary electrons in a metal are the conduction electrons or the more tightly bound inner-shell electrons, a complication that does not exist in thermionic emission since it is known in that case that only the conduction electrons can be involved. Any attempt to discuss the question of primary interaction thus involves the introduction of certain simplifying assumptions about the solid, and even then the problem is generally extremely complicated.

B. Primary Energy Loss

The probability of producing a secondary electron of a given energy certainly depends upon the energy of the primary that produces it. Consequently, one must have some knowledge of the energy of the primary as a function of the depth to which it has penetrated. Since there is no experimental evidence available on the rate of energy loss of low-energy primaries, it must be computed theoretically. Inasmuch as the primary can lose energy in several ways (production of secondaries, excitation of bound electrons, and the like), such a computation is very difficult.

C. Escape of Secondaries

In order for a secondary to be observed, it must escape from the surface of the solid. To accomplish this, it must move through the body of the solid from the point at which it was created to the surface, retaining sufficient energy to penetrate the surface potential barrier. Although the problem of the penetration of the barrier can be treated quite adequately, the motion of a slow secondary through the solid is not at all well understood. Several possible assumptions can be made, such as exponential absorption, random diffusion, uniform energy loss, and the like, but whether any of these assumptions are valid is not at all clear.

D. Integration over the Range of the Primary

The solution of the above problems will furnish information regarding the number of emergent secondaries which are produced at a given depth in the solid. In order to obtain the total yield, it is necessary to integrate this result over all possible depths, i.e., from zero to the total range of the primary. If steps (A), (B), and (C) have been solved, this process can always be accomplished numerically if necessary, and consequently this does not represent an essential difficulty.

As may be concluded from the above discussion, the general problem of secondary emission is very complicated, and at present the only feasible approach involves adopting various simplifying assumptions and investigating the agreement between theories based on these approximations and experiment. Several theories of secondary emission from metals have been developed in this manner. Of these we shall discuss the theories of Baroody and Wooldridge which, though not the only attempts, are quite typical of the usual approach, since both assume that the loosely bound valence electrons constitute the principal source of secondaries.

Baroody's theory,[5] which is somewhat simpler, employs the Sommerfeld model of the metal, as is done in the theory of thermionic emission. The interaction between the primary and secondary electrons is treated in a purely classical manner, and the secondaries are assumed to be absorbed exponentially in their passage to the surface. This theory, in common with all existing theories of secondary emission, involves several parameters whose magnitudes are unknown, and consequently the values of the secondary yields to be expected from metals cannot be computed in absolute terms. A relation between yield and energy is obtained, however. Baroody points out the existence of the universal curve shown in Fig. 4, relating experimental values of the secondary yield and energy, and compares his theoretical results with it. Unfortunately, although the theoretical curves have the same general form, the quantitative agreement is very poor. For example, for $V_p/V_{p\max}=4.5$, the experimental value of $\delta/\delta_{\max}$ is three times the computed value. This lack of agreement should not cause great concern, however, inasmuch as this application of the Fermi gas model is exceedingly questionable, and the aim of the investigation was primarily to demonstrate certain qualitative features of secondary emission.

Wooldridge's theory[6] is similar to Baroody's in that it considers only the valence electrons as potential secondary electrons, and assumes an exponential law for their absorption. It differs from Baroody's theory in treating the primary interaction quantum mechanically and taking into account the interaction of the valence electrons with the lattice. Again, absolute values of yields cannot be determined, and for comparison with experiment, the theoretical value of $\delta_{\max}$ is set equal to the empirical value. Within the expected range of validity of the formulas, the yield curves thus obtained agree with experiment quite well for the dense metals, such as silver and copper; but for less dense substances, such as lithium and aluminum, the agreement is rather poor. Wooldridge assumes that the primary loses energy only by the production of secondaries, and attributes the disparity with experiment in the case of light elements to the neglect of other types of energy loss. It is quite possible, however, that this disagreement arises from errors inherent in the basic assumptions.

Both of the aforementioned theories are based upon the hypothesis that valence electrons are the principal source of secondaries, and at sufficiently low primary

[5] E. M. Baroody, "A theory of secondary electron emission from metals," *Phys. Rev.*, vol. 78, p. 780; 1950.

[6] D. E. Wooldridge, "Theory of secondary emission," *Phys. Rev.*, vol. 56, p. 562; 1949.

energies, this is undoubtedly the case. At very high primary energy, on the other hand, bound inner-shell electrons are the principal source since their binding is then very small compared to the energy of the incident primary and since they are much more numerous than the valence electrons. At intermediate energies (approximately 100 to 2,000 volts) it is not at all obvious that one group of electrons or the other necessarily plays the dominant role. The work of Baroody, Wooldridge, and others shows that by proper choice of parameters, the yield versus energy curves for metals can be explained qualitatively on the valence electron assumption, and for some metals, as mentioned above, good quantitative agreement is even obtained.

The hypothesis that the bound electrons may be the principal source of secondaries has not as yet been investigated quantitatively; however, there are cogent qualitative arguments indicating that this may be the case. (1) Bound electrons which could be available as secondaries are much more numerous than the valence electrons in most metals. (2) The ionization of gases has been investigated experimentally, and it has many features in common with secondary emission from metals. In particular, if the probability of ionization σ of a gas molecule is plotted as a function of the energy W of the incident electron, a curve having the same general form as the secondary yield curve shown in Fig. 3 is obtained. Furthermore, if this curve is normalized by plotting σ/σ_{max} against W/W_{max}, a universal curve which is very similar to the universal emission curve results, as is seen in Fig. 8.

This resemblance suggests a possible model for secondary emission. If the bound electrons in a solid are the principal source of secondaries, the production of an internal secondary will be essentially an ionization process. Hence the probability of production should have much the same dependence on energy as the probability of ionization of gases. Although many factors affect the shape of the secondary yield versus energy relationship, it is quite possible that the primary interaction is the dominant factor, in which case the universal secondary emission curve and the universal ionization curve should be very similar. An examination of Fig. 8 reveals that this is indeed the case. Furthermore, rough calculations indicate that if one modifies the gas curve to take into account the fact that only those electrons with sufficient energy to penetrate the surface barrier can be observed as secondaries, the resultant curve will be in even better agreement with the secondary emission curve.

Consequently, it seems important to investigate in more detail the possibility that the bound electrons may be an important source of secondaries. It should be pointed out that whether or not the conduction electrons in a metal are a copious source of secondaries they are certainly of great importance in the emission process since it is very likely that interactions between internal secondaries and the conduction electrons are responsible for most of the secondary absorption which so drastically limits the secondary yield of metals. It might be expected that, in the case of metals, filling an inner-shell in progressing through the periodic system should increase the secondary electron emission, whereas the addition of an electron to the outer-shell should decrease the yield. The absence of conduction electrons in insulators thus accounts for their high secondary yields.

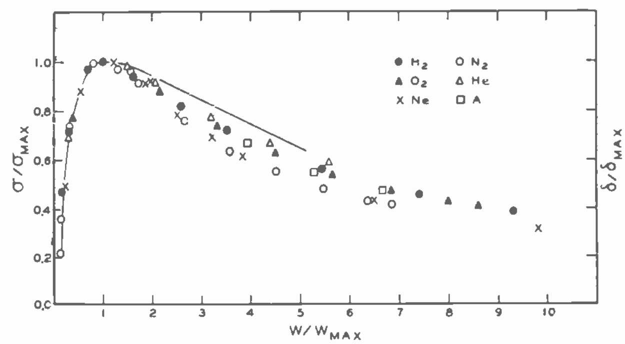


Fig. 8—Normalized primary ionization probabilities, σ/σ_{max}; plotted as a function of normalized primary energy, W/W_{max}, for various gases. The solid curve is the universal curve of secondary electron shown in Fig. 4.

IV. Conclusion

Many specialized features of secondary electron emission have not been mentioned in this paper, and it has been possible to include only a brief summary of the salient features of the phenomenon, in conformity with the primary purpose of this review. Frequent references and inclusion of detailed descriptions of experiments have necessarily been avoided in the interests of stressing general principles rather than presenting an encyclopedic survey of the literature. For this reason, descriptions of applications of the phenomenon have likewise been omitted. For additional facts, the reader is referred to the lengthier articles containing rather comprehensive bibliographies.[7,8]

In conclusion it seems appropriate to hazard a prediction regarding the course of future progress in this field. Certainly the known engineering goals are well defined; efforts to obtain high yields with good stability will undoubtedly be continued, together with the search for new applications of the effect. From the scientific point of view, there is need for a self-consistent series of reliable measurements on all chemical elements in the periodic system which can be made into suitable targets. The disagreement among the results of different experiments is such as to preclude many crucial comparisons which could cast light upon the nature of the mechanisms involved in the process of secondary emission.

[7] K. G. McKay, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. 1; 1948.

[8] H. Bruining, "Die Sekundär-Electronen-Emission fester Körper," Julius Springer, Berlin; 1948.

Propagation at 412 Megacycles from a High-Power Transmitter*

IRVIN H. GERKS†, SENIOR MEMBER, IRE

Summary—**Extended measurements are reported which indicate the existence of pronounced nocturnal superrefraction during an appreciable percentage of the summer and of very persistent scattering by atmospheric turbulence near the surface in all seasons. The measurements were taken over rolling midwestern terrain at a distance of about 100 miles. Mobile road tests were made to supplement the fixed-point measurements and to provide an approximate indication of the relation between field strength and distance. Aerial tests were made to show the effects of antenna height at large distance. Graphs are provided which show the effects of distance, terrain, antenna height, and time upon the field strength. The practical significance of the results in the broadcast and communication fields is indicated.**

Introduction

THE CALCULATION of tropospheric field strength was first satisfactorily treated by making the simplifying assumptions of a smooth, spherical earth and a linear variation of refractive index of the medium with height above the surface.[1,2] As a result of numerous short-wave propagation measurements made prior to about 1940, it was recognized that atmospheric irregularities may produce major anomalies in the field strength determined by the above methods.[3–6] During World War II both experimental and theoretical studies were made which revealed the extensive occurrence of nonstandard refraction, particularly over ocean areas, and provided improved mathematical methods for treatment of the simpler cases of nonlinear gradient of refractive index near the surface.[7–11] More recent studies of propagation, mainly on overland paths, have illustrated the complicated nature of atmospheric refraction in the presence of turbulence and the effects of terrain irregularities.[12–16] Frequent instances of abnormally strong fields in the diffraction region have been observed, and the mechanism of scattering from atmospheric irregularities, or "blobs," has been postulated to explain such fields.[17]

It is the purpose of this paper to report the results of a series of tests made at 412 mc with a transmitter having sufficient power output to permit measurements of field strength somewhat farther into the nonoptical region than has been possible customarily.

Experimental Conditions

The tests to be described were conducted during the summer of 1948 and during the fall, winter, and spring of 1949–1950. A small amount of work was done at short range to determine the effect of terrain. The great bulk of the work was done at ranges in excess of 80 miles. Since low antennas were used for surface work at both ends of the path and since the terrain was relatively smooth, the receiving antenna was located several thousand feet below line-of-sight at each of the remote receiver sites. Reception of the ground wave was therefore impossible except during the rather rare occurrence of strong superrefraction. Thus, most of the results represent reception of tropospheric waves, presumably resulting from scattering by air masses in the lower troposphere having a refractive index differing slightly from the average.

The transmitter was located at the Cedar Rapids Airport, and the transmitting antenna was mounted at a height of about 40 feet on the roof of the Collins Radio Company hangar. Two types of transmitting antennas were used, one a biconical horn with omnidirectional radiation in the horizontal plane and a power gain of about 5 relative to an isotropic source, the other a pyramidal horn with an approximately square aperture and a power gain of 28. The former was arranged so as

* Decimal classification: R113.23. Original manuscript received by the Institute, January 9, 1951; revised manuscript received June 1, 1951.

† Collins Radio Company, Research Div., Cedar Rapids, Iowa.

[1] C. R. Burrows and M. C. Gray, "The effect of earth's curvature on ground-wave propagation," PROC. I.R.E., vol. 29, pp. 16–24; January, 1941.

[2] K. A. Norton, "The calculation of ground-wave field intensities over a finitely conducting spherical earth," PROC. I.R.E., vol. 29, pp. 623–639; December, 1941.

[3] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," PROC. I.R.E., vol. 21, pp. 427–463; March, 1933.

[4] C. R. Englund, A. B. Crawford, and W. W. Mumford, "Further results of a study of ultra-short-wave transmission phenomena," *Bell Sys. Tech. Jour.*, vol. 14, pp. 369–387; July, 1935.

[5] C. R. Burrows, A. Decino, and L. E. Hunt, "Stability of two-meter waves," PROC. I.R.E., vol. 26, pp. 516–528; May, 1938.

[6] C. R. Englund, A. B. Crawford, and W. W. Mumford, "Ultra-short-wave transmission and atmospheric irregularities," *Bell Sys. Tech. Jour.*, vol. 17, pp. 489–519; October, 1938.

[7] M. Katzin, R. W. Bauchman, and W. Binnian, "3- and 9-centimeter propagation in low ocean ducts," PROC. I.R.E., vol. 35, pp. 891–905; September, 1947.

[8] J. B. Smyth and L. G. Trolese, "Propagation of radio waves in the lower troposphere," PROC. I.R.E., vol. 35, pp. 1198–1202; November, 1947.

[9] J. S. McPetrie, B. Starnecki, H. Jarkowski, and L. Sicinski, "Oversea propagation on wavelengths of 3 and 9 centimeters," PROC. I.R.E., vol. 36, pp. 243–257; March, 1949.

[10] C. L. Pekeris, "Wave theoretical interpretation of propagation of 10-centimeter and 3-centimeter waves in low-level ocean ducts," PROC. I.R.E., vol. 35, pp. 453–462; May, 1947.

[11] H. G. Booker, "The mode theory of tropospheric refraction and its relation to wave-guides and diffraction," "Meteorological Factors in Radio-Wave Propagation," The Phys. Soc. (London), pp. 80–127; 1947.

[12] C. W. Carnahan, Nathan W. Aram, and Edward F. Classen, Jr., "Field intensities beyond line of sight at 45.5 and 91 megacycles," PROC. I.R.E., vol. 35, pp. 152–159; February, 1947.

[13] G. S. Wickizer and A. M. Braaten, "Propagation studies on 45.1, 474 and 2800 megacycles within and beyond the horizon," PROC. I.R.E., vol. 35, pp. 670–679; July, 1947.

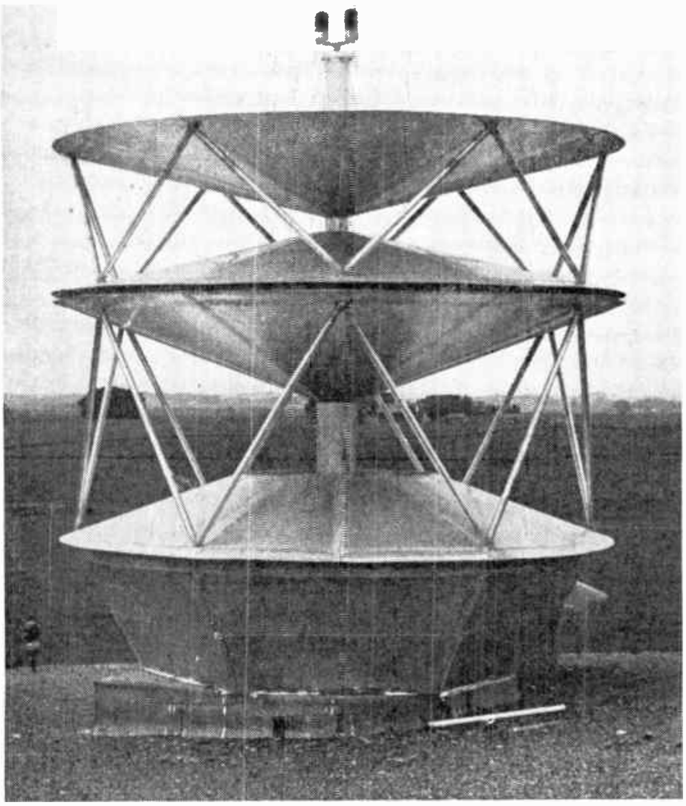
[14] W. L. Carlson, "Simultaneous field-strength recordings on 47.1, 106.5, and 700 mc," *RCA Rev.*, vol. 9, pp. 76–84; March, 1948.

[15] G. H. Brown, J. Epstein, and D. W. Peterson, "Comparative propagation measurements; television transmitters at 67.25, 288, 510, and 910 mc," *RCA Rev.*, vol. 9, pp. 177–201; June, 1948.

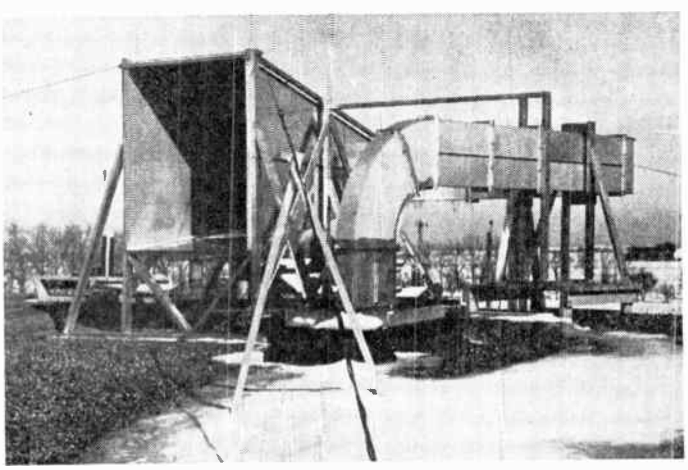
[16] J. P. Day and L. G. Trolese, "Propagation of short radio waves over desert terrain," PROC. I.R.E., vol. 38, pp. 165–175; February, 1950.

[17] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," PROC. I.R.E., vol. 38, pp. 401–412; April, 1950.

to produce either vertically or horizontally polarized radiation. The latter produced only horizontally polarized waves (see Fig. 1).



(a)



(b)

Fig. 1 (a) and (b)—View of two types of transmitting antennas. Radiation from upper biconical horn is vertically polarized, that from lower is horizontally polarized. Elevation is approximately 40 feet.

The transmitter consisted of a 50-kw resnatron oscillator coupled by means of special waveguide gear to the antenna. The transmitter output power was unmodulated and was held at a value of approximately 30 kw. A frequency in the range of 406 to 420 mc was used, with the bulk of the measurements being made at 412 mc.

The receiving antenna used for all of the fixed-point measurements was a 10-foot paraboloid, with a power gain of 140. A corner-reflector antenna with a power gain of about 10 was used for the surface mobile and exploratory tests (see Fig. 2). The receiving antenna was located about 10 feet above the surface. For aerial measurements, horizontal and vertical dipoles backed by a ground screen were mounted on the nose of a DC-3 airplane. The receiver was originally connected to an Esterline-Angus 1-ma recorder, operated at a speed of 12 inches per hour. In later tests, this record was supplemented by hourly photographs of a special counter panel indicating the time in minutes during which each of several levels was exceeded. The latter record greatly reduced the effort and time required for subsequent analysis.

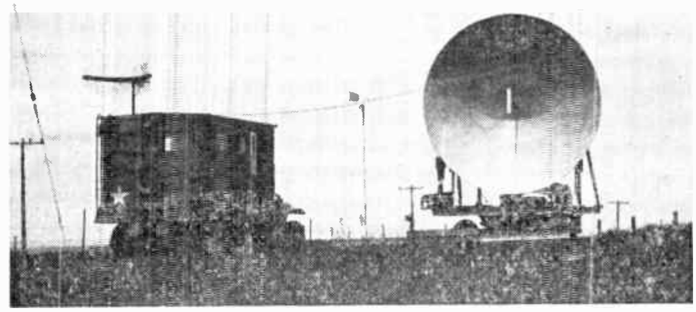


Fig. 2—View of receiving station showing 10-foot parabolic antenna and corner reflector antenna mounted on truck. Elevation of antennas is approximately 10 feet.

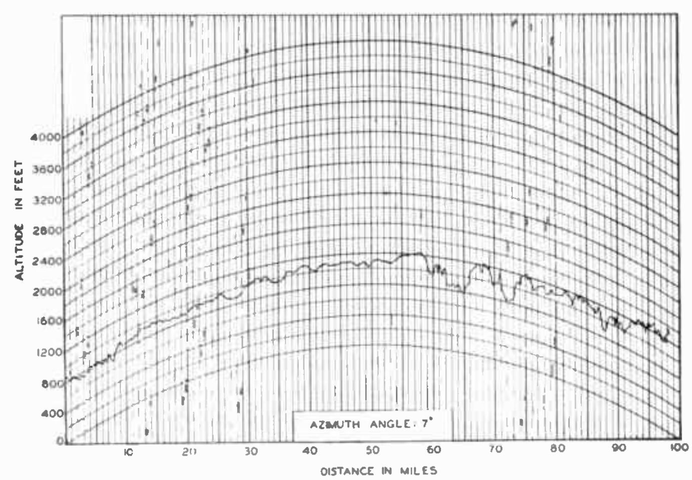


Fig. 3—Terrain profile of propagation path from Cedar Rapids to Waukon.

Extended measurements were made at three receiving sites, as follows: Waukon, Iowa, distance 98 miles, summer 1948 and fall 1949; Mitchellville, Iowa, distance 86 miles, winter 1949–1950; Quincy, Illinois, distance 134 miles, spring 1950. The terrain is gently rolling, with few wooded areas. A typical profile, that for the Waukon path, is shown in Fig. 3. The co-ordinate system is so chosen that a straight line represents a ray path in a standard atmosphere. The terrain on the other two paths was generally similar to that shown from 0 to 50 miles in Fig. 3. In addition to these fixed measurements, several mobile measurements were made with a truck and an airplane to provide more complete information regarding the effects of terrain and distance.

Mobile Tests

To determine the effects of terrain and distance upon the field strength, a truck was driven along an approximately radial road and a recording was made relating field strength with road distance. Portions of this record were analyzed to determine the statistical relation between field strength and location. The results are shown

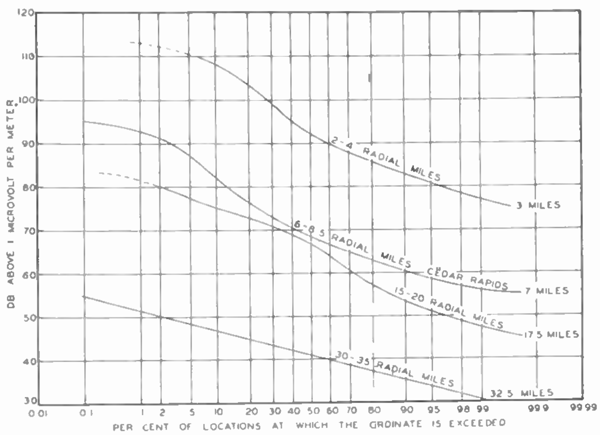


Fig. 4—Statistical relation between field strength and percentage of locations at which this field strength is exceeded. Measurements were made along an approximately radial road. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Height 40 feet. Vertical polarization. Receiving antenna height 10 feet.

in Fig. 4. In each interval, an average distance was selected, as shown at the right of the curve, and field strength values for other distances in this interval were subjected to an appropriate correction for distance. Thus, these curves show only the effect of terrain, buildings, trees, and the like.

Though the data are too limited to warrant firm conclusions, several results are striking: Each of the upper three curves shows a range between the 1 and the 99 per cent values of about 36 db. It appears that the variation of field strength due to change of location in any one limited area is on the order of ± 20 db within a distance of about 20 miles and with a receiving antenna about 10 feet above road level. A second noteworthy feature of Fig. 4 is the moderate depression of the curve obtained for the Cedar Rapids area below the position to be expected for a corresponding range in open country. This is only 5 to 10 db, indicating that the field strength in urban areas, even with low receiving antennas, is only moderately lower than in more open, rural areas. A third feature of Fig. 4 is that the terrain effect is only a little over 20 db for the curve corresponding to 32.5-miles distance. This could be only a coincidence if it were not corroborated by other tests at greater distances. It appears that the terrain effect is reduced with increasing distance, attaining a value of about ± 10 db at a range where the tropospheric wave is strongly predominant.

The road trip was continued to a distance somewhat in excess of 200 miles. A plot of field strength versus distance is shown in Fig. 5 for that portion of the distance where reasonably reliable measurements could be made. In this, each vertical line represents the range of field strength measured in 3/16 mile of road distance, and the dot indicates the estimated median value. The dashed line represents an inverse-square variation of field strength with distance. This curve appears to fit the data reasonably well, considering terrain effects, out to 25 or 30 miles, except at 25 miles, where a deep river valley caused a depression of field strength. Between 30 and 50 miles the measured medians drop well below the inverse-square curve as we should expect when diffraction loss is considered. However, beyond 50 miles the measured field strength again appears to drop no faster than the inverse-square curve if we make allowances for various river valleys. Since diffraction of the

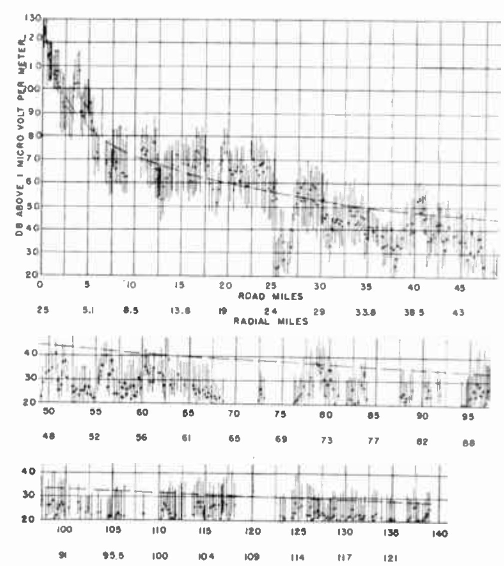


Fig. 5—Relation between field strength and distance. Measurements were made along an approximately radial road. Each line represents range of field strength observed in 3/16 mile of road distance and each dot represents estimated median. Dashed line is an inverse-square curve applying theoretically at short distances. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Height 40 feet. Vertical polarization. Receiving antenna height 10 feet.

ground wave is such as to produce quite a rapid decrease in field strength with increasing distance beyond the radio horizon, we must suppose that a tropospheric wave becomes effective beyond 40 or 50 miles so as to extend the effective range. Beyond about 125 miles the field strength was too low to permit effective measurement with the corner-reflector antenna used for the mobile test.

In Fig. 6 we see the results of measurements made with an airplane flown on a radial course at three different elevations. Very roughly, the elevation above terrain may be taken as 1,000 feet less than the elevation above sea level although terrain clearance varied as much as 500 feet during the flight. Each vertical line represents the range of field strength in a radial distance of 4 miles and the dot again indicates the median value. The dashed curves represent simple analytical approximations, in general, an inverse-distance relation at short distance and an exponential relation at longer distance. Within the range where the direct ray and ground-reflected ray, or rays, interfere to produce a lobe structure, the results are somewhat erratic, but show generally that the measured field strength approximates the free-space field strength. Beyond this range there is a diffraction region where the field strength drops rapidly with increasing distance. This extends 30 to 40 miles beyond the radio horizon. Near the outer limit of this region, the fluctuation range of the signal gradually increases, indicating the onset of contribution from the tropospheric wave. Beyond this region, the curves show a reduction in slope, which is characteristic of the region where the tropospheric wave is predominant. This third

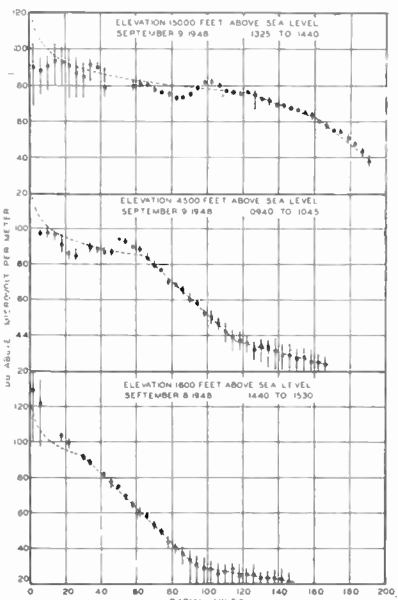


Fig. 6—Relation between field strength and distance as measured at three altitudes in an airplane. Each line represents range of field strength observed in 4 miles of radial distance and each dot represents estimated median. Dashed lines represent simple functions of distance chosen for best fit with measurements. For terrain elevation, see Fig. 3. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Horizontal polarization. Azimuth angle 7 degrees.

region was never reached at 15,000 feet within the limits of the flight. The constancy of the exponential coefficient in both the diffraction region and the tropospheric-wave region is a rather striking effect which facilitates moderate extrapolation. It will be observed, however, that the variation becomes somewhat more complicated at high altitudes.

The results just described are assembled for comparison with each other and with the theoretical results in Fig. 7. Here a logarithmic distance scale is used to permit extension to large distances. Also, the field strength has been converted to an equivalent value obtained with 1 kw radiated from a half-wave dipole by subtracting 10 log $(5/1.64\times 30)=19.6$ db from all measured values. Curves A and C were computed from the theory applying in the case of a smooth spherical earth, with the radius increased by a factor of 4/3 to take into account average atmospheric refraction. Curves B, D, E, and F were drawn as smooth curves representing the

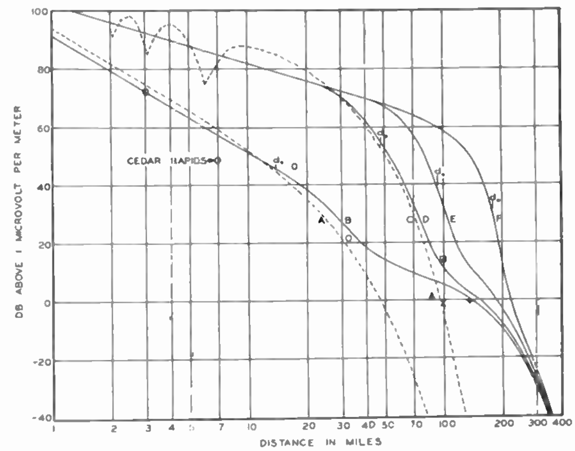


Fig. 7—Variation of median field strength with distance for various receiving antenna heights. Frequency 410 mc. Effective radiated power 1 kw. Transmitting antenna height 40 feet.
- A Theoretical curve for receiving antenna height 10 feet.
- B Experimental curve for receiving antenna height 10 feet.
- C Theoretical curve for receiving antenna height 1,000 feet.
- D Experimental curve for receiving antenna height 1,800 feet msl.
- E Experimental curve for receiving antenna height 4,500 feet msl.
- F Experimental curve for receiving antenna height 15,000 feet msl.
- d_0 Radio optical distance.
- o Median values taken from Fig. 4.
- x Median value from Curve A, Fig. 12 (Waukon, Fall, 98 miles).
- △ Median value from Curve B, Fig. 12 (Mitchellville, Winter, 86 miles).
- + Median value from Curve C, Fig. 12 (Quincy, Spring, 134 miles).
- ⊡ Median value from Curve D, Fig. 12 (Waukon, Summer, 98 miles).

estimated best fit with experimental data. Extension of the curves beyond a distance of about 150 miles was accomplished partly by extrapolating the curves in Figs. 5 and 6 and partly by utilizing the results of limited surface measurements made at a distance of 225 miles. The optical range for a spherical earth and standard refraction as calculated by the formula

$$d_0 = \sqrt{2h_t} + \sqrt{2h_r} \tag{1}$$

is shown on each of the curves. The median experimental values taken from Fig. 4 are shown by circles. The scatter of these points about curve B illustrates the prominence of the terrain factor even after purely local fluctuations have been removed. No effort was made to show lobe structure in Curves D, E, and F because no regular lobe structure could be observed. We should note finally that all data for Fig. 7 were obtained during those daylight hours which were found to yield minimum anomalies due to superrefraction.

It is observed that the experimental results, except for terrain effects, agree rather well with calculations made for a standard refractive index gradient in the atmosphere out to a distance beyond optical distance varying from about 20 to about 40 miles. The most striking feature of the experimental curves is the point of inflection 20 to 30 miles beyond optical distance. Here the tropospheric wave becomes rapidly more prominent with increasing distance, the attenuation becomes less, and fading begins. The decreased attenuation of the tropospheric wave leads to an amazing increase in range, particularly for low antennas. This increase in range may be as much as several hundred miles for low antennas and a high-power transmitter. The increase in range becomes relatively small when one antenna is sufficiently elevated, probably because waves scattered at low levels, where turbulence is most intense, do not greatly increase the optical distance, and because scattering at high levels is relatively poor. With both antennas at an elevation of several thousand feet, the effect of scattering is probably negligible. A secondary effect, shown by Fig. 7, is the convergence of the curves for various elevations at large distances. This indicates a pronounced decrease in the height gain of the antenna over certain ranges of height. Since height gain is intimately associated with diffraction and since the tropospheric wave can reach the receiving antenna with less diffraction around the curve of the earth than the ground wave, this effect is to be expected. The larger grazing angle of the tropospheric wave at the receiver also tends to account for the reduced terrain effect at large distances mentioned earlier.

The median values obtained at various fixed sites and at various seasons are also shown in Fig. 7 by the symbols x, $\triangle$, $+$, and $\boxdot$. The points representing summer results at Waukon lie above Curve B, probably because the site, which was on a carefully selected high point, was better than a median site. The greater prevalence of nocturnal superrefraction in the summer must also be considered. The point representing fall results at Waukon is based on measurements during one week only when rather persistent stormy weather caused abnormal depression of the field strength. The point representing winter results at Mitchellville lies below Curve B, probably because the receiving site in this case was poorer than the median. The receiving antenna was partly shadowed by a gentle rise in the ground facing the transmitter. The point representing results at Quincy is seen to lie very close to Curve B. Since these points are based on about 2,300 hours of data, their rather close grouping about Curve B serves to lend considerable validity to this curve in the region near 100 miles. In fact, when the results secured at the three sites are suitably corrected for distance and are combined in a manner which will be explained later, a median value of 4.5 db above 1 μv per meter is obtained at 100 miles. This is only 0.5 db less than the value shown by Curve B.

Measurements at a Fixed Point

An extensive series of field-strength recordings was made at three sites as follows:

Location	*Distance*	*Period*	*No. of Hours*
Waukon, Iowa	98 miles	June, July, August, 1948	500 approx.
Waukon, Iowa	98 miles	October, November 1949	156
Mitchellville, Iowa (WHO)	86 miles	November, 1949 to March, 1950	1,162
Quincy, Illinois (WTAD-FM)	134 miles	April, May, 1950	517

In each case, measurements were made with a receiving antenna only about 10 feet above the surface. In general, the records cover operation for 24 hours a day for intervals of about 6-days duration spaced fairly uniformly throughout the periods mentioned. Thus, reasonably representative samples are available for all seasons of the year.

During the tests made at Waukon and during the mobile tests, a pair of biconical transmitting antennas was used which were capable of radiating either vertically polarized or horizontally polarized waves. It was found that the strength of received signals at large distances varied so little when the direction of polarization was changed that this variation could not be detected in the presence of the strong fluctuations characterizing the signal at nearly all times. During the tests made at Mitchellville and Quincy, a pyramidal horn was used at the transmitter to improve the signal level at the receiver, to provide greater purity of polarization when horizontal polarization was used, and to assure a more reliable figure for the power gain. In all tests made with the pyramidal horn, horizontal polarization was used.

The field strength measured at the fixed sites was usually characterized by a rapid scintillation of large amplitude. The rapidity of the scintillation was frequently so great that the record produced by an Esterline-Angus recording milliammeter, operated with a chart speed of 12 inches per hour, showed a solid band of ink with irregular upper and lower borders. During summer nights, superrefraction was quite common. This had the effect of increasing the field strength quite markedly and reducing the rapidity of fading. Occasionally, the superrefraction became so pronounced that the fading almost disappeared and the field strength resembled that within a few miles of the transmitter. On a few occasions, peak levels of about 80 db above 1 μv per meter were recorded, roughly 40 db above the daytime median value. Such anomalies were associated, as one might expect, with relatively clear skies and strong radiation cooling of the surface. However, no simple correlation between field strength and meteorological conditions could be determined, except in the case of the sunrise maximum. This effect was a characteristic duct of unusual strength appearing

1 to 3 hours after sunrise. The explanation for this appears to be the creation of high-surface humidity by the evaporation of dew. In all cases, the duct disappeared with amazing rapidity when the evaporation was essentially complete and turbulent mixing of the air became well-established. Duct conditions during the winter were rare or nonexistent. Storms along or near the propagation path generally caused pronounced depression of the field strength, probably because of better mixing of the air at all heights.

No conclusive results are available to indicate the angle of arrival of the wave at the receiver. The rapid fluctuation of field strength and the relatively broad antenna beam (approximately 15 degrees) made it difficult to determine the angle of arrival with any accuracy from a test in which the antenna was slowly rotated in a vertical plane. The wave appeared to arrive so nearly in a horizontal direction that all tests were made with the antenna directed horizontally. However, it was quite evident that the antenna beam was appreciably broader horizontally in the presence of a scattered field, indicating scattering from a considerable volume of space. Also, the reduced shadowing effect of terrain obstacles at large distances, as well as the reduced height gain at the receiver, indicated an effective source at least several degrees above the horizon. It may be argued that the effective power gain of the receiving antenna may be much less with a scattered signal than with a single plane wave. Hence, the true field strength at the receiver may be appreciably greater than that measured with an antenna having a beam width as narrow as that used in these tests. This argument is justified to a certain extent. However, successive measurements made with a corner reflector (power gain = 10) and a 10-foot paraboloid (power gain = 140) indicated no significant differences either in the measured field strength or in the character of the fading. We must conclude that the scattered waves in this case had a sufficiently small deviation from the general direction of arrival so that the 10-foot paraboloid gave results differing by only a very few db from the true field strength. The best estimate is that scattered waves arrived principally from directions varying from 0 to about 10 degrees vertically and from -10 to $+10$ degrees horizontally. This narrow beam of the scattered field appears to indicate atmospheric irregularities of dimensions generally large compared with the wavelength.

The statistical distribution of field strength for each hour of recording was determined in a manner described elsewhere.[18] The observed values taken from the recorder chart or the time totalizer were plotted on Rayleigh graph paper, and a smooth curve was drawn through these points. In the majority of cases, the hourly distribution curve drawn on Rayleigh paper was nearly linear, especially at the low field-strength end. This fact was utilized to justify extrapolation of the curve to values of field strength lower than those which could be reliably measured. It proved of great value for periods of unusually weak fields when the received signal dropped below the receiver noise level during an appreciable fraction of the time. Two sets of readings

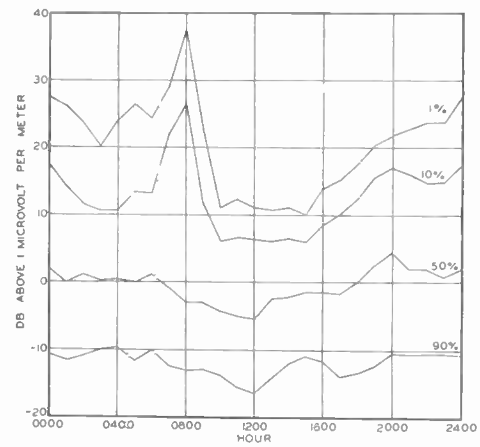


Fig. 8—Diurnal variation of field strength at Waukon, Iowa, averaged over 156 hours during October and November, 1949. Number opposite curve represents percentage of time ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

were taken from these smoothed curves: The first set consisted of the values of field strength exceeded during an arbitrary percentage of the time, such as 1 per cent, 10 per cent, 50 per cent, and so on. These values could then be plotted against time to show the variation of the hourly field strength for the period of measurement.

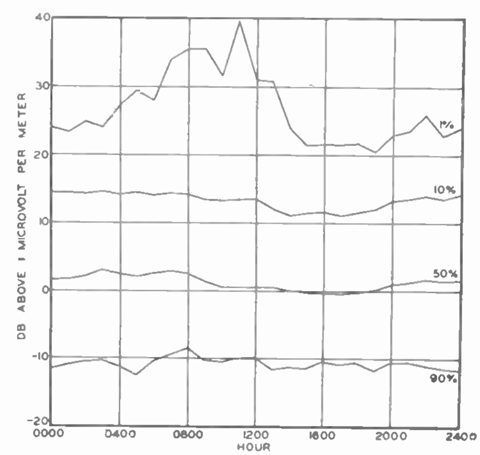


Fig. 9—Diurnal variation of field strength at Mitchellville, Iowa, averaged over 1,162 hours during November–March, 1949–1950. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

Space does not permit inclusion of these curves here. The second set consisted of the percentage of time during each hour that each of a number of arbitrary levels was exceeded. This tabulation was subsequently utilized in determining the percentage of time during which each of these levels was exceeded for a particular hour of the day for an entire period of measurement.

These results, showing the diurnal variation of field

[18] R. P. Decker, "Notes on the analysis of radio propagation data," Proc. I.R.E., pp. 1382–1388; this issue.

strength at four locations at various times of the year, are plotted in Figs. 8 to 11. Fig. 8 is based upon a very limited amount of data obtained in the fall, during a period when frequent storms caused abnormal depression of the field strength. However, sufficient fair, mild weather occurred to produce characteristic nocturnal superrefraction during a significant percentage of the time, as indicated by the two upper curves. The sunrise maximum occurring between 0,700 and 0,900 is quite evident. It should be noted that the hour indicated on the abscissa scale represents the beginning of the hour for which the ordinate value applies. A somewhat less pronounced rise of field strength occurred during the first half of the night while radiation cooling was most prominent. Fig. 9 indicates an almost complete absence of diurnal variation during the winter months. The forenoon rise in the 1-per cent curve may indicate an effect similar to that observed shortly after sunrise during the

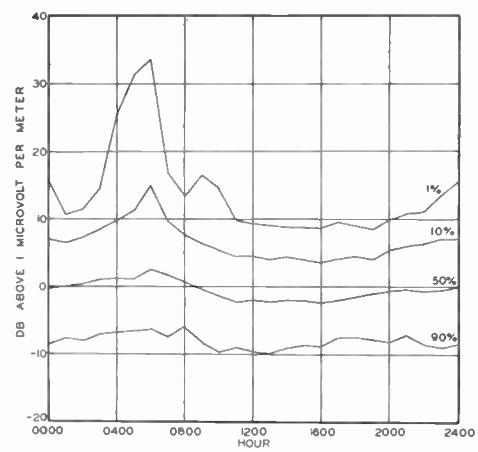


Fig. 10—Diurnal variation of field strength at Quincy, Illinois, averaged over 517 hours during April and May, 1950. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

summer. However, the absence of this rise in the 10-per cent curve indicates that superrefraction is quite rare in the winter. Fig. 10 shows the reappearance of the sunrise maximum in the spring, with only very moderate improvement in field strength during the other nighttime hours. Fig. 11 shows the strong diurnal variations occurring during summer months. The sunrise maximum is very pronounced. Another somewhat less prominent maximum occurs during the evening hours while radiation cooling is most rapid. Because of the extreme anomalies occurring during summer nights, and the rather limited period of measurement, the results shown in Fig. 11 are necessarily subject to greater uncertainty than those shown for the winter months in Fig. 9.

When the results for an entire period of measurement are summarized, we obtain the results shown in Fig. 12. We see that Curves A and B are nearly coincident in spite of different locations and seasons. The favorable location and the moderate superrefraction at site A were largely offset by an unusual incidence of stormy weather so that the over-all results resemble closely those secured during the winter months at site B. Curve C, representing spring results at an increased distance, has a shape which is not readily explained. Except for the

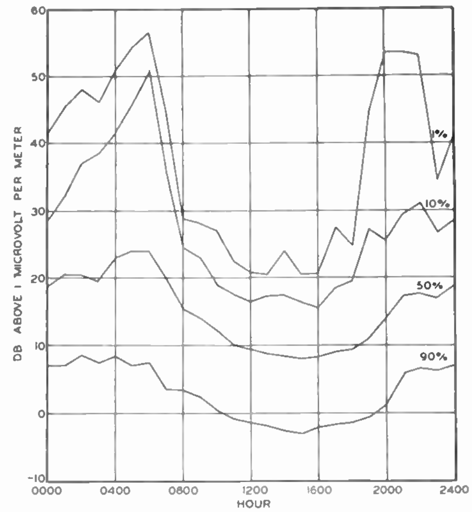


Fig. 11—Diurnal variation of field strength at Waukon, Iowa, averaged over approximately 500 hours during June, July, and August, 1948. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

portion of the curve below an abscissa of 2 per cent, which can be explained on the basis of nocturnal superrefraction, this curve shows an abnormally small fading range. Curve D indicates a generally high field strength, partly because of a favorable receiving location and partly because of rather persistent favorable propaga-

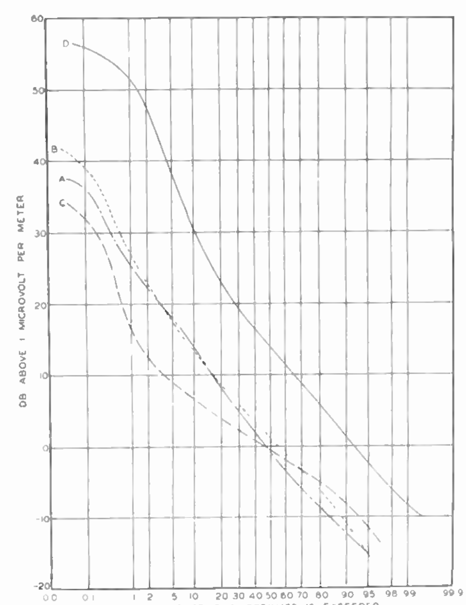


Fig. 12—Statistical variation of field strength at large distances. Frequency 410–412 mc. Effective radiated power 1 kw. Transmitting antenna height 40 feet. Vertical polarization, Curve D, horizontal polarization, Curves A, B, C. Receiving antenna height 10 feet.

A Waukon, Iowa, 98 miles, 156 hours, October–November, 1949.
B Mitchellville, Iowa, 86 miles, 1162 hours, November–March, 1949–1950.
C Quincy, Illinois, 134 miles, 517 hours, April–May, 1950.
D Waukon, Iowa, 98 miles, approximately 500 hours, June–August, 1948.

tion conditions occurring during the summer. It is interesting to note that the field strength range included between the 1-and 90-per cent levels varies from 38 db in the winter to 50 db in the summer. We should note also the very considerable interference of a 400-mc signal during a summer evening. Fig. 11 shows a level of about 54 db above 1 μv per meter exceeded 1 per cent of the time during the hours 2,000–2,300, whereas Curve D (Fig. 12) shows a level of 51.5 db exceeded 1 per cent of the time for the whole day.

In Fig. 13, an effort has been made to combine the four curves of Fig. 12 to show the field strength to be expected at a distance of 100 miles. The data for the

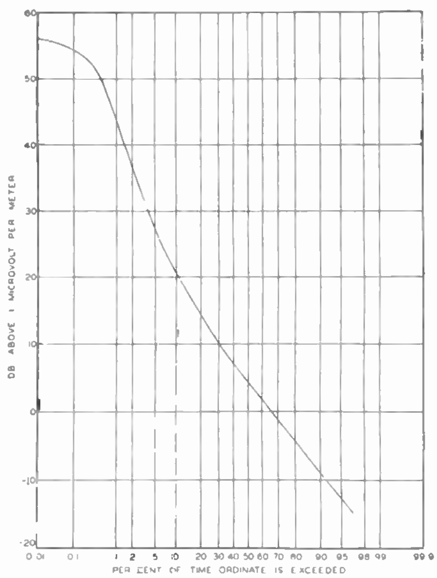


Fig. 13—Statistical variation of field strength at 100 miles. The results of Fig. 12 have been combined as explained in the text. See Fig. 12 for conditions. Effective radiated power 1 kw. Transmitting antenna height 40 feet. Receiving antenna height 10 feet.

various locations were weighted generally in accordance with the number of hours of observation, the number of months covered, and the estimated reliability. Arbitrary weighting factors were assigned as follows:

Data for Site	*Season*	*Weighting Factor*	*Distance*	*Add db*
Waukon	Fall	1	98 miles	0
Mitchellville	Winter	5	86 miles	−2
Quincy	Spring	2	134 miles	4.5
Waukon	Summer	4	98 miles	0

Distance corrections were taken from Curve B of Fig. 7. The weighting factor may be considered also as the number of months to which the data are assumed to apply. In this curve, representing all-year results, we observe a range between the 1-and 90-per cent levels of 52 db, a median value of 4.5 db (agreeing well with Fig. 7, Curve B), and a nuisance signal of 43 db exceeded 1 per cent of the time. This figure also embodies the effects of favorable and less favorable terrain in that the highest values of field strength are based entirely on the favorable site at Waukon, whereas the lowest values of field strength are considerably influenced by the results at the less favorable sites at Mitchellville and Quincy, with combined weighting factors of 7.

The diurnal variation taken from these composite results is shown in Fig. 14. We observe again the morning and evening maxima, the former being more persistent as shown in the 10-per cent curve. The median curve shows a diurnal variation of only about 6 db, whereas the 90-per cent curve shows a negligible diurnal variation.

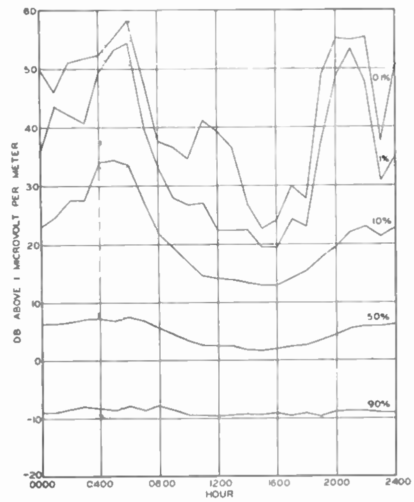


Fig. 14—Diurnal variation of field strength at 100 miles averaged over approximately 2,340 hours during all seasons. See Fig. 12 for conditions. Effective radiated power 1 kw. Low ground-based antennas.

On numerous occasions, when airplanes were observed to be flying over the propagation path, the observed field strength fluctuated rhythmically in amplitude, sometimes over a range of more than 10 db, with a period varying from a short value of less than a second to a long value of several seconds and then again to a short value. The duration of such a disturbance varied from a few seconds to more than a minute. At times, this effect caused a buildup as high as 20 db in the field strength. Though the net effect on observed field strength was slight on the propagation paths chosen in these tests, it is believed that a significant increase in the interference value of the signal could occur on a path where air traffic was heavy and the routes were generally parallel to the path.

Conclusions

We find that at 400 mc the field strength at a median location may be computed with reasonable accuracy in the conventional manner, that is, by assuming a spherical earth and standard atmospheric refraction, provided that the distance does not exceed the radio-optical distance by more than about 20 miles. Even within this range, nocturnal radiation cooling and morning surface evaporation may increase the field strength quite markedly for a small percentage of the time during the warm season. At greater distances, the diurnal variation due to nocturnal superrefraction is even more prominent. However, the field strength also develops strong, rapid fluctuations and decreases more slowly with increasing distance as a result of wave scattering in the lower troposphere. This scattering is largely independent of the time of day or the season. It is greatly reduced only by certain frontal conditions which induce more thorough mixing of the air. Pronounced superrefraction occurs only during summer nights, but this effect is sufficiently persistent to cause a significant field strength increase during as much as 5 per cent of the time for the entire year.

The degree of the anomaly produced by atmospheric turbulence and superrefraction for very low antennas at 100-miles distance can be appreciated by comparing a field strength of about -60 db above 1 μv per meter computed for a spherical earth and standard refraction with a value of about 5 db representing the median measured value. This discrepancy is reduced when either antenna is raised. With one antenna at 1,000 feet, the corresponding values are about -5 db and 11 db.

Since a high transmitting antenna may produce a greatly extended service range but little more interfering field strength at great distances, it is evident that a transmitting antenna height of about 1,000 feet is desirable in any proposed high-quality uhf broadcasting network. For communication grade of service, it appears entirely feasible to operate a 100-mile link with low, directional antennas and about 10 kw of transmitter power with a probability of satisfactory field strength more than 90 per cent of the time. In fact, quite effective use of the scattered wave can be made to a distance of 200 miles, as indicated by Fig. 7. It must be recognized, however, that the results reported here are strictly applicable only to Iowa and adjacent regions, and that differing terrain and meteorological conditions in other regions may modify the results materially.

Because of multipath propagation occurring on such long paths, a certain degree of distortion must be expected in the reception of modulated waves. Since the relative delay on tropospheric paths is probably small compared with a modulation-frequency period, such distortion should be much less severe than on ionospheric paths.[19] Very brief tests with speech modulation tend to confirm this conclusion. The signal was almost always more readable than that received over a low-power link operating in the 5,000-kc range.

Acknowledgment

The work reported here was supported in part by the Central Radio Propagation Laboratory under Contract CST-10783.

The assistance of members of the Research Division of the Collins Radio Company, who conducted the tests and analyzed the data, and the generous co-operation of the personnel of Radio Station WHO and WTAD-FM are gratefully acknowledged.

[19] Irvin H. Gerks, "An analysis of distortion resulting from two-path propagation," Proc. I.R.E., vol. 37, pp. 1272–1277; November, 1949.

Notes on the Analysis of Radio-Propagation Data*

R. P. DECKER†

Summary—**This paper deals with the reduction of radio-propagation data. The primary aim is to present a clear picture of signal-strength variation with a minimum amount of computational work. A newly developed recording device is described and illustrated, together with an effective method of data analysis.**

I. Introduction

AN INCREASING number of ultra-high-frequency and very-high-frequency radio-propagation studies have been carried out lately by investigators in different countries. This work is certain to continue, and a proportionate amount of data will have to be recorded and analyzed. Many hours of tedious work have, so far, been required to present a statistical picture of the variation of field strength over a given period of time. Thus, it may be of some value to describe the newer recording techniques together with an effective method of data reduction and presentation. This paper is based chiefly on the author's experience with high-power uhf propagation at 410 mc.[1] Modifications would have to be made in a similar treatment for lower frequencies.

The problem is then to analyze the data with a minimum amount of work and to summarize the results by means of curves and charts. The amplitude variation of the signal is translated into data which will have the most significance for statistical analysis. The representation will be most effective when the percentage of time is graphed during which various signal strengths are exceeded. The basic interval is taken to be the hour, since it is short enough to preclude, in general, a change in propagation conditions, and long enough to smooth out the random variations which are of little or no interest.

A recording instrument (usually of the Esterline-Angus type), is used to record continuously the output

* Decimal classification: R531.8. Original manuscript received by the Institute, January 9, 1951; revised manuscript received, June 6, 1951.

† Collins Radio Co., Cedar Rapids, Iowa.

[1] Measurements were made under the auspices of the Central Radio Propagation Laboratory during the winter and spring of 1949–1950. The transmitter, a 400-mc resnatron with an average power output of 24 kw, was located at the airport in Cedar Rapids, Iowa. The receiving sites were at station WHO near Mitchellville, Iowa (86.1 mi.) and at station WTAD-FM, Quincy, Illinois (133.9 mi.).

or avc voltage of a field-intensity meter. Additional equipment consists of a signal level versus time indicator or time totalizer, a counter panel with camera, and a signal generator with a suitable attenuator for calibration purposes. The signal level versus time indicator and the counter panel, which have been developed recently, will be described in some detail later.

The calibration procedure, which is of primary importance, will be considered first.

II. Calibration Procedure

The calibration of the receiver relates output meter reading to signal strength. The relation between field strength and receiver input voltage is given by the following equation:

$$E^2 = \frac{1.64(480\pi^2)e^{2\alpha l}}{P_t G_t \lambda^2 G_r Z_0} E_r^2 \left(\frac{\text{volts}}{\text{meter}}\right)^2, \tag{1}$$

where

E = field strength in volts/meter,

E_r = received input voltage,

α = attenuation in nepers per axial foot of the cable connecting the receiving exciter to the receiver input,

l = length of the cable in feet,

G_r = power gain of the receiving aperture referred to an isotropic source,

λ = wavelength in meters,

G_t = power gain of the transmitting aperture referred to an isotropic source,

P_t = power flow in kw through the transmitting aperture, and

Z_0 = surge impedance in ohms of the cable.

In decibel notation (1) may be written as follows:

$$\begin{aligned} 20 \log E = & -10 \log P_t - 10 \log G_t + 38.91 - 20 \log \lambda \\ & -10 \log G_r - 10 \log Z_0 \\ & + 8.69\alpha l + 20 \log E_r \text{ (decibels)}. \end{aligned} \tag{2}$$

Once the constants have been evaluated, (2) becomes

$$E' = 20 \log E_r - 20 \log \lambda - 10 \log P_t + \text{constant}. \tag{3}$$

E' is then expressed in db above 1μv per meter referred to a kw radiated from a half-wave dipole. The voltage required of the signal generator to produce a given output meter deflection can immediately be translated into field strength by (3).

As a rule, the frequency is stable and can be included as a constant in (3). The transmitted power, however, will possibly vary, especially when the resnatron is used in high-power uhf work.

With the aid of (3) a calibration curve of output meter reading versus db above 1 μv per meter is plotted. When a change in power level occurs, it is only necessary to add the appropriate amount of db for each meter reading.

The question which frequently arises is how often receiver calibrations should be made. It is evident that the accuracy of the analysis increases with the number of receiver calibrations. However, too much time spent in calibration could possibly make the data for the particular hour unreliable if abrupt changes in propagation occur, such as presented by an early morning duct, for example. It is therefore advisable to calibrate as infrequently as possible when such conditions prevail. since the shape of the distribution curve (see Section V) could be seriously affected, especially in the low- or high-percentage region. The minimum number of calibrations permissible must be determined by the stability of the receiver. The exact length of time during which the receiving and recording equipment were inoperative must be known in order that corrections for the percentage of time can be made.

III. Esterline-Angus Analysis

The percentage of time during which various preselected levels were exceeded may be obtained by an analysis of the Esterline-Angus chart. The chart speed should be fast enough to observe normal fading; 16 divisions or 12 inches per hour is usually adequate. The analysis has been accomplished by estimating the percentage for each division. The average of the 16 divisions yields the average for the hour.

Accurate plotting of the distribution curves becomes easier when the levels are chosen so that there is a constant db difference between them. The range of the receiver in db divided by 9 results in the db increment for 10 levels. Recourse to the calibration curve will give the 10 corresponding output meter readings to be used in the analysis.

Under normal propagation conditions, an hour of the Esterline-Angus chart can be completely analyzed in 50 minutes by a person with some experience. However, some practice is necessary before reasonable accuracy (within 2 per cent) can be obtained. Levels which were exceeded from 0.01 to 2 per cent and from 98 to 99 per cent of the time under normal conditions are not easily or accurately detected by this method. At these levels, the width of a pen line for a very short peak or dip in signal is not an accurate indication of the percentage.

IV. The Signal Level Versus Time Indicator (SLVTI)

In the preceding section it was noted that the errors involved in a visual analysis of the Esterline-Angus charts were about 2 per cent. When the results of this analysis are plotted on log log or on probability paper, an extreme scattering of points could occur, especially in the low- or high-percentage regions. It also requires an excessive amount of time for analysis, and results in eye fatigue. To overcome these disadvantages, a device called the "Signal Level Versus Time Indicator"[2] (Fig. 1) has been developed which indicates the time or per-

[2] For a description of a similar unit, see R. W. George, "Signal-strength analyser," *Electronics;* January, 1951.

centages of time during which a number of preselected levels were exceeded. The indicator shown in Fig. 1 works in conjunction with a counter panel (Fig. 3), upon which synchronous timing motors and revolution counters are mounted.

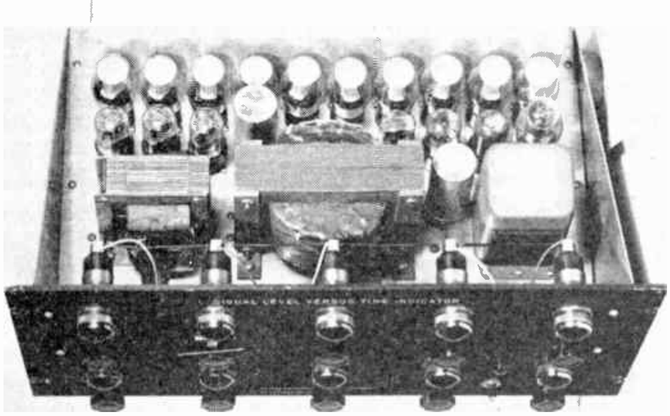


Fig. 1—Signal level versus time indicator.

The circuit diagram for such a 10-channel device is shown in Fig. 2. The 6SN7 stages are so biased by the selenium rectifier and voltage-divider circuit that an increasingly negative avc voltage applied to the grid causes the voltage across the 100,000 ohm plate resistor to go positive with respect to the thyratron cathode. As soon as the critical voltage is reached, which is determined by the 400-ohm voltage dividers, the thyratron will fire and close its plate-circuit relay. This, in turn, energizes a timing motor which drives a revolution counter. In order to make the thyratrons self-extinguishing, ac voltage is used on the plate. A panel light is shunted across each motor for level adjustments.

Under conditions of rapid fading, a relay is energized as often as several times a second. Thus the synchronous timing motor which drives the revolution counter must start and stop instantly to prevent cumulative errors. If the speed of the counter is 10 rpm, the counter will indicate the time in tenths of minutes for which its level was exceeded. However, if the counter speed is $166\frac{2}{3}$ rpm, the counter will indicate hundredths of a per cent. The latter speed eliminates the process of division in the calculation of percentages. This represents a saving of about 2 man hours per hundred hours of data. It also means increased accuracy in plotting the distribution curves, since a fraction of a per cent error in the low- or high-percentage region of Rayleigh paper

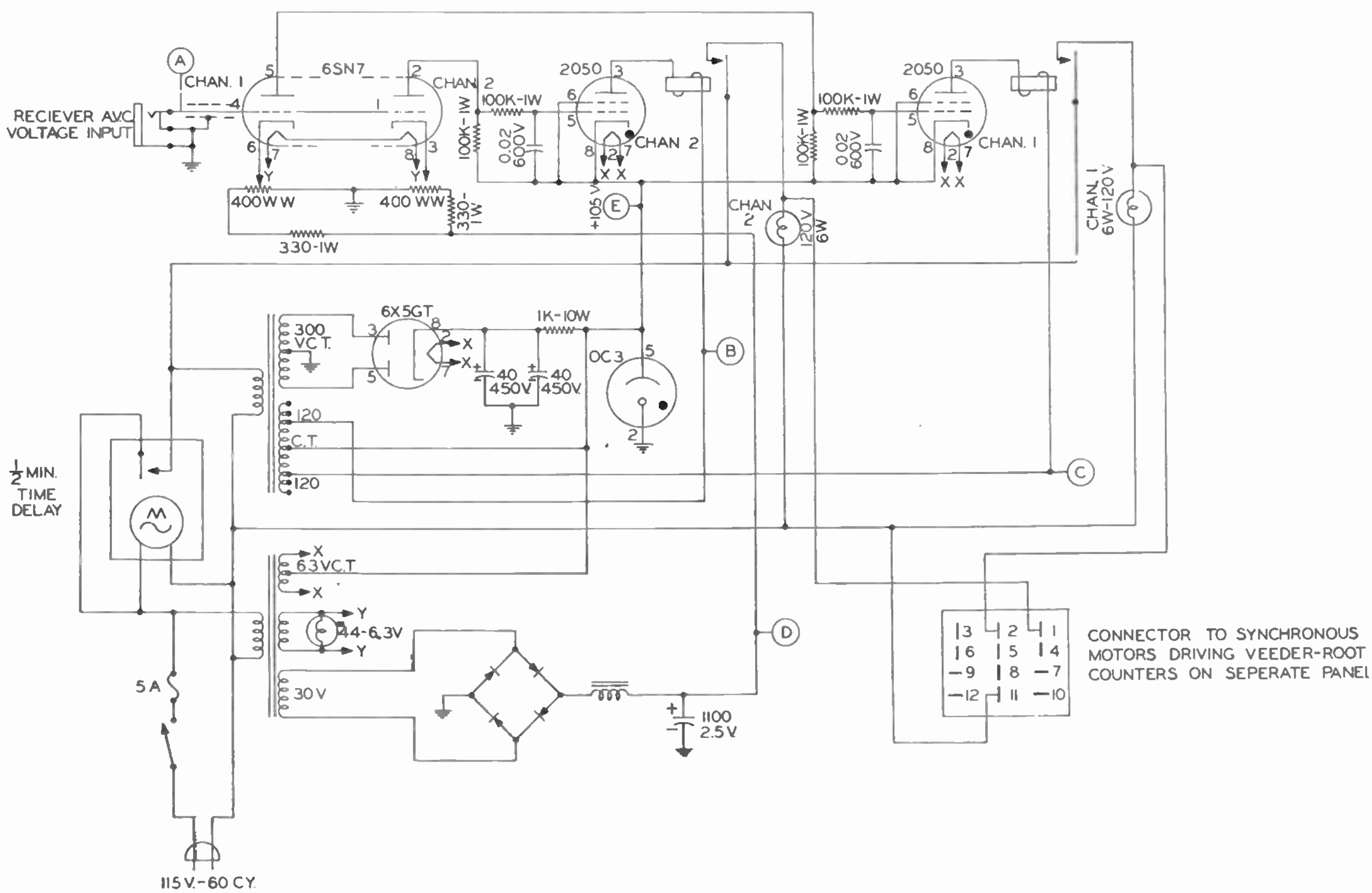


Fig. 2—Circuit diagram of the signal level versus time indicator. All resistances in ohms; all capacitances in μf. Note: Only 2 of 10 channels are shown. A=Input to all channels. B=Only to channels 2, 4, 6, 8, and 10. C=Only to channels 1, 3, 5, 7, and 9. E and B=All channels.

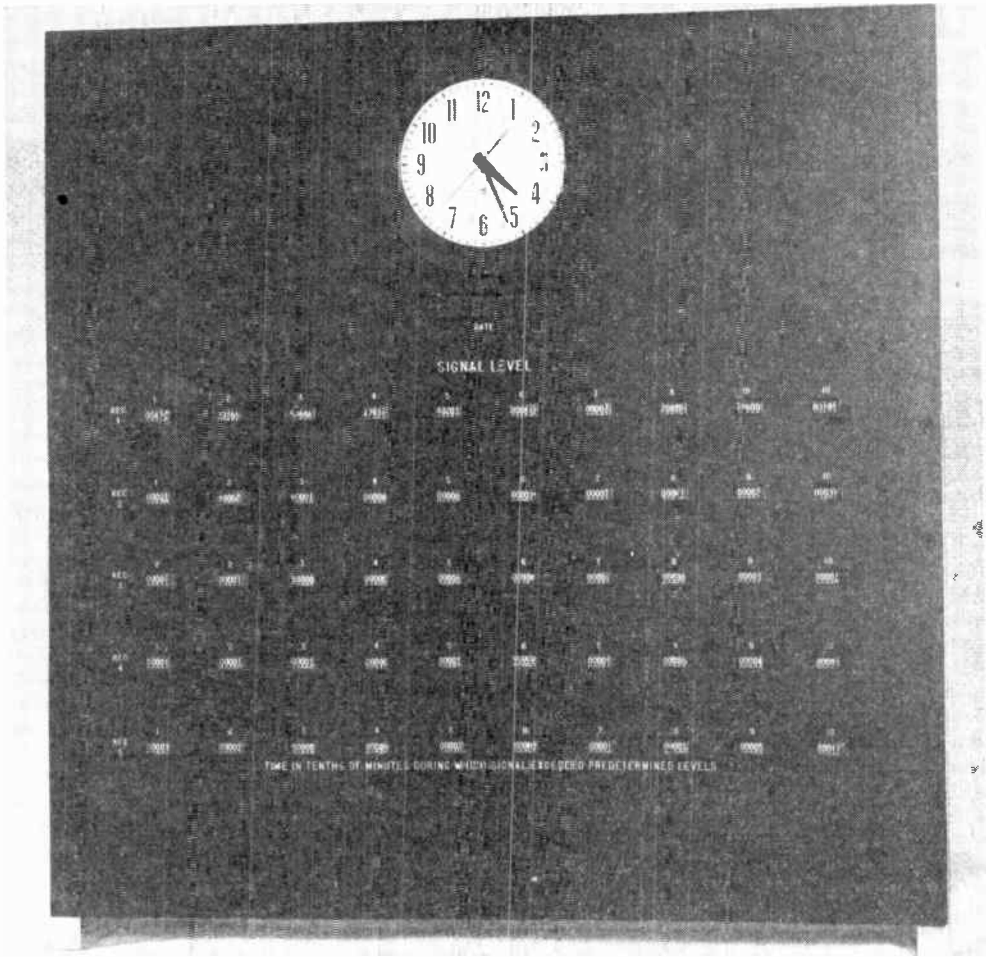


Fig. 3—Counter panel.

could distort the shape of the curve (as is possible with a slowly driven counter). As the data will eventually be plotted on Rayleigh paper, in which the abscissa distance from the origin is proportional to $\log 4 - \log [2 - \log P]$ (where P is percentage), it is easily seen that percentage increments near 0 or 100 represent relatively large distances.

The model shown in Fig. 2 must be externally regulated on account of the bias and thyratron plate voltages. Good regulation is required so that any voltage changes occurring within the SLVTI are negligible with respect to the 0–6v input for which this unit is designed.

The threshold of potential of the 2050 thyratrons has been found to be about 0.02 volt wide. This, as well as imperfect regulation, will cause the thyratron to fire and extinguish within ± 0.05 volt of the intended level. The circuit would, therefore, be improved by a well-regulated bias voltage and an additional stage of amplification ahead of the 6SN7's for greater firing accuracy with respect to the input voltage.

The circuit has been field tested and has proved to be entirely satisfactory. One indication of its performance is the exactness with which the resulting points lie on a smooth curve. A recent analysis of about 1,500 hours of data has shown that the points fit exceptionally well in each instance.

The ten counters with white on black numerals, which are mounted on a separate black panel (Fig. 3), are photographed automatically every hour on the hour. When the field run is over, the film (35 mm.) is inserted in a suitable viewer and the counter reading is recorded for each level and hour. Subtracting the counter readings of two successive hours yields the time or percentage of time (depending on the counter speed) during which the particular level has been exceeded.

As in the case of an Esterline-Angus analysis, the levels are best chosen in such a manner that there is an equal db increment between them. In calibrating, it is only necessary to adjust the signal generator until the desired level just clicks in, and to record the number of microvolts required.

The advantages of this method lie in the saving of tedious work in analysis, and in the increased accuracy. With the crude method of the Esterline-Angus chart analysis a full hour was required to analyze and draw the curve for an hour's data; but with the SLVTI the analysis may be done much more accurately in 6 minutes—a speed-up of ten to one.

V. Curve Plotting

The percentage of time during which a number of preselected levels was exceeded according to the SLVTI and the value of these levels in db above 1 μv/m/kw, as determined by the calibration, have now been established. For purposes of condensation and interpolation, the points are plotted on suitable paper. It is well known that log log or Rayleigh paper is customary for this purpose. Norton,[3] starting with Lord Rayleigh's original work,[4] has shown that the percentage of time during which a field strength will theoretically be exceeded is given by

$$P = 100e^{-E^2/(E_t^2+E_g^2)} \qquad (E_g \ll E_t), \tag{4}$$

where

P = percentage of time,
E = field intensity of resultant of sky wave plus ground wave, which is exceeded P per cent of the time,
E_t = rms of tropospheric wave, and
E_g = rms of ground wave.

The Rayleigh distribution holds normally at large distances in which the ground wave E_g is small compared to the tropospheric wave E_t. In addition, duct or layer-type propagation conditions would not be "Rayleigh-distributed" with time.

If k log log $(100/P)$ is represented as the abscissa distance and k' log $E/\sqrt{E_t^2+E_g^2}$ as the ordinate distance, it is easily seen that (4) will plot as a straight line. Since the field strength is in db above 1 μv/m/kw, it follows that on a paper with a k log log $(100/P)$ abscissa for percentage of time during which the field strength is exceeded, and with a linear ordinate for the db values, the theoretical distribution curve plots as a straight line. This graph paper is generally known as "Rayleigh distribution paper."

The points for each hour are successively plotted on Rayleigh paper. If a SLVTI is used, the points should lie on a perfectly smooth curve.

The point farthest into the 99-per cent region is determined by the setting of the lowest level on the SLVTI and by the sensitivity of the receiver. It is, of course, desirable to have a receiver sensitive enough under all conditions to run the lowest level counter almost continuously so that a 100-per cent point is always available to shape the right side of the curve correctly. This level also serves as a time check for interruptions in transmitter operation. When no points are available in this region, the extrapolation is best accomplished by drawing a line tangent at the last point of the curve. Evidence supporting this method of extrapolation was obtained in the analysis of about 1,500 hours of SLVTI data recorded at stations WHO and WTAD-FM during the winter and spring of 1949–1950.

Once the curve has been drawn, the percentages are tabulated for equally spaced db values by grouping according to the hour of the day (Table I). Equal db increments result in an easier and more accurate drawing of the average distribution curves and in greater uniformity of the tabulated data. For a picture of the inverse variation, the field strength values, which were exceeded by 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent of the time, are tabulated (Table II, see page 1387).

At WHO, for example, the lowest field intensity used in the analysis was chosen as −20 db. The successive points were spaced 2.5 db apart. It is interesting to consider the possibility of setting the SLVTI levels exactly at these values. This procedure would reduce the time spent in data reduction by about 30 per cent,

[3] K. A. Norton, "Advances in Electronics," Academic Press Inc. New York, N. Y., vol. 1, pp. 381–423; 1948.
[4] J. W. S. Rayleigh, *Phil. Mag.*, vol. 10, pp. 73–78; 1880.

TABLE I

Tabulation of Data by Hours.

Percentage of Time that the Signal Exceeded Each of a Number of Levels Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole.

Date	LEVEL IN DB −20	−17.5	−15	−12.5	−10	−7.5	−5	−2.5	0	2.5	5	7.5	10	12.5	15	17.5	20
								0000									
May																	
12	98.5	97.4	95.3	92	86	77	62	42	22	7.0	0.5						
13	100	99.83	99.6	99	97.4	94	85	67	39	18	5.3	0.7	0.02				
14	99.83	99.68	99.31	98.6	97	94	87.7	77	55	30	10	3.1	1	0.35	00.14	0.03	
15	99	98.3	97	95	91.5	86	76	65	46	28	10	0.3					
16	100	100	100	100	100	99.75	99.1	97.2	93	87.5	78	66	53	40	25	11	2
17	100	100	99.9	99.73	99.22	97.8	93.5	83	56	31	11	3.2	1	0.4	0.16	0.04	
								0100									
12	99.62	99.3	98.6	97.2	94.8	90	81	67	41	15	3.2	0.45	0.02				
13	100	100	100	99.85	99.3	97	86	65	43	22	7.5	2.0	0.5	0.25			
14	99.83	99.67	99.33	98.7	97.3	95	90	81	65	46	25	10	1.3				
15	98.6	97.7	96.2	94	90.2	85	76	65	49.5	31	15	5.5	0.8	0.01			
16	100	100	100	99.9	99.67	99	96.8	90.5	75	57	36	18	5	0.4			
17	100	100	100	100	99.7	98.8	95.1	84	64	44	22	7.5	1.1	0.03			
18																	

as it would eliminate the interpolation of percentages for the equally-spaced db values. However, this can be done only if the sensitivity of the receiver is exceptionally stable, and if the transmitter power is constant to within ±0.25 db. (This value of 0.25 db results from the estimation of the probable error in curve drawing.) However, the interpolation of the db values corresponding to 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent, would not be eliminated.

VI. Condensation of Data

The tabular data of db above 1 μv/m for 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent, are used to draw a graph of the continuous hourly variation of these percentages versus field intensity, during the length of time for which the data were recorded (Fig. 4). Thus, one may see at a glance the exact behavior of the field strength. The difference in db between the 1 and 99 per cent values is indicative of the fading range and is included in Table II.

TABLE II

Tabulation of db Values for Given Percentages.

Signal Level Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole as a Function of the Percentage of Time that This Signal Level Was Exceeded.

Date	Hour	E_1	E_{10}	E_{50}	E_{90}	E_{99}	E_1-E_{99}
May 11	1500	5.0	0.5	−4.9	− 6.8	−10.4	15.4
	1600	4.0	0	−3.0	− 5.7	− 8.9	12.9
	1700	8.0	1.9	−2.4	− 3.9	− 4.5	12.5
	1800	3.2	−0.9	−4.0	− 6.6	− 9.8	13.0
	1900	3.8	0.3	−1.9	− 5.1	− 9.0	12.8
	2000	5.0	−0.2	−1.6	− 3.8	− 6.2	11.2
	2100	11.0	0	−4.2	− 8.8	−14.3	25.3
	2200	2.0	−1.6	−4.2	− 6.7	− 9.5	11.5
	2300	1.0	−2.0	−5.8	−10.7	−16.9	17.9
May 12	0000	4.5	2.0	−3.4	−11.5	−21.6	26.1

The percentages for the equally spaced db values, which were previously grouped, are now averaged for each hour of the day, and are retabulated. This procedure results in an average distribution curve for each hour of the day; Fig. 5 (see page 1388) is a typical example. When the values for these 24 curves are weighted and averaged, the grand average distribution curve of the entire propagation run (Table III, below) is obtained.

To obtain the average diurnal variation of field strength, the 24 hourly distribution curves are examined and a graph is made of db versus hour of the day for 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent (Fig. 6, see page 1388).

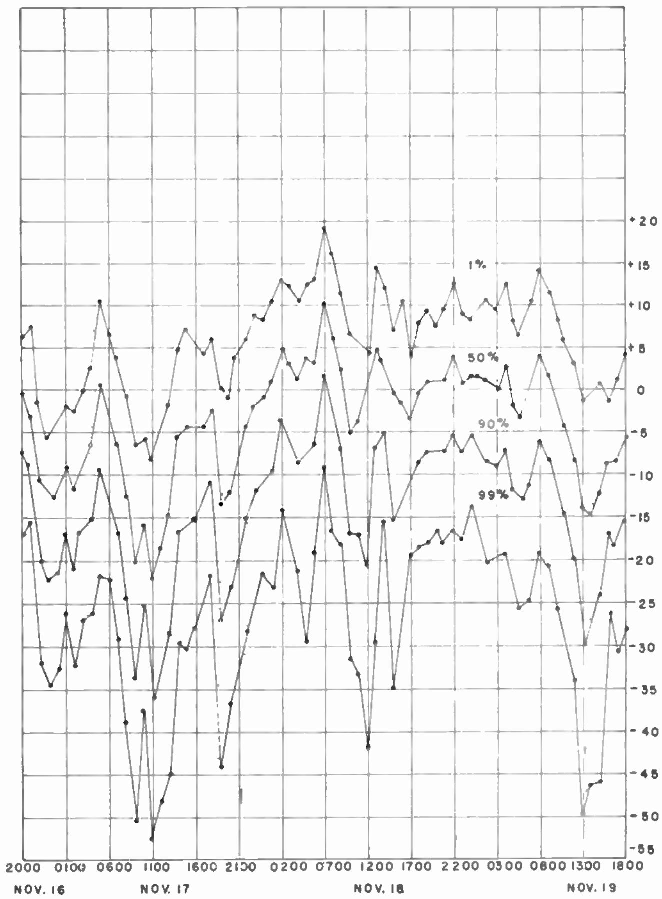


Fig. 4—An example of the continuous hour-by-hour variation of field strength. Field strength measured at WHO tower near Mitchellville, Iowa, November 16–23, 1949. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=86.1$ mi, and $f=410$ mc. Figure on curve represents percentage of time ordinate value is exceeded.

TABLE III

Typical Summary Table.

Hourly Summary of Field Strength at WTAD-FM Tower Near Quincy, Illinois, May 22–28, 1950. Percentage of Time that the Signal Exceeded Each of a Number of Levels Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole.

	LEVEL IN DB								
Hour	−20	−17.5	−15	−12.5	−10	−7.5	−5	−2.5	0
0000	98.77	98.05	96.85	94.93	91.53	86.75	78.50	66.92	50.83
0100	98.98	98.37	97.45	96.00	93.43	89.30	81.17	69.02	52.75
0200	99.02	98.42	97.50	95.88	92.93	88.30	80.35	68.92	53.33
0300	99.41	99.03	98.12	97.29	95.28	91.50	84.58	72.58	55.33
0400	99.26	98.76	98.02	96.67	94.32	90.17	82.30	70.17	51.67
1900	99.80	99.59	99.16	98.28	96.35	92.77	84.80	71.83	49.33
2000	99.71	99.46	98.97	97.90	95.85	91.80	83.33	68.58	47.67
2100	99.91	99.80	99.53	98.92	97.44	93.87	85.52	70.63	49.50
2200	98.62	97.78	96.58	94.51	91.47	87.12	79.77	69.88	55.33
2300	98.83	98.05	96.87	94.86	91.64	86.63	78.63	68.37	52.83
A	99.49	99.13	98.52	97.38	95.25	91.35	83.01	68.05	47.58

It is seen that the hour-by-hour variation, the average hourly distribution curves, the summary curve, and the diurnal variation, present a most effective picture of field-strength variation. The probable over-all accuracy of the analysis may be calculated by finding the standard deviation of the various sources of errors. A representative list of such sources is the following:

1. Transmitter power
2. Signal generator attenuator
3. SLVTI
4. Curve drawing and interpolation

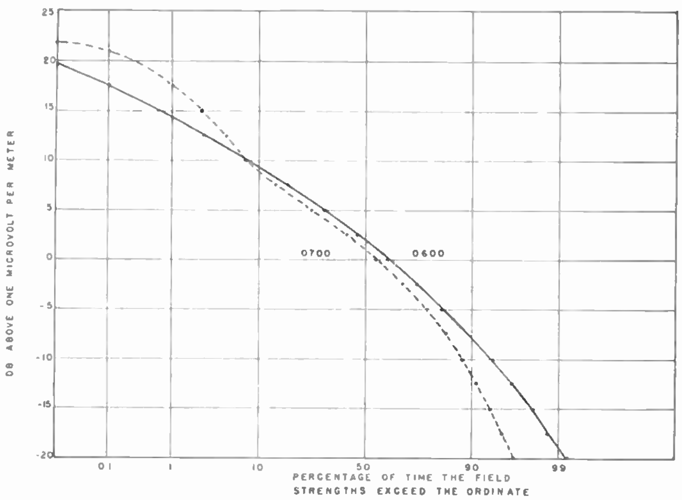


Fig. 5—Typical average hourly distribution curve. Field strength measured at WTAD-FM tower near Quincy, Illinois, April, 1950. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=133.9$ mi, and $f=412$ mc. Dotted and solid lines show average for ten days.

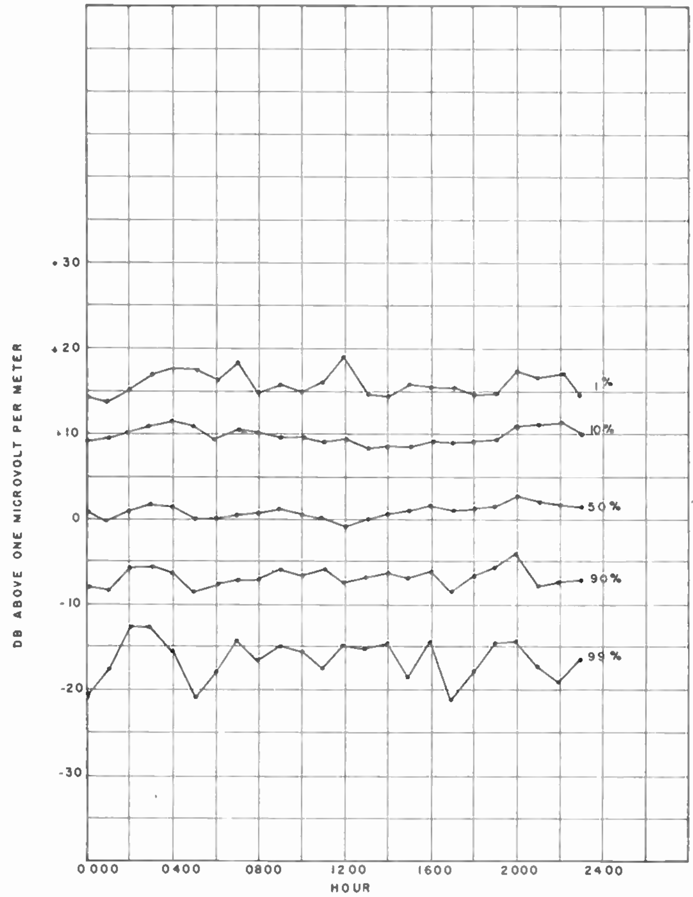


Fig. 6—An example of diurnal variation. Field strength measured at WHO tower near Mitchellville, Iowa, January, 1950. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=86.1$ mi, and $f=412$ mc. Figure on curve represents percentage of time ordinate value is exceeded.

VII. Conclusion

The work in question relates to the fundamental processes involved in the analysis of propagation data. The primary aim in data reduction is to present an effective picture of field-intensity variation with a minimum amount of interpolation and tabulation. The methods and equipment described above are believed to be accurate and reliable in ascertaining the percentage of time for which various field strengths were exceeded.

If the power output of the transmitter and the sensitivity of the receiver are constant, a considerable amount of time may be saved in analysis. The Esterline-Angus analysis should, of course, be avoided, unless the SLVTI or counter panel should fail. It is seen that they afford a 10-to-1 speed-up in analysis time.

Although the SLVTI and counter-panel system, with the suggested improvements in the circuit, would seem ideal, two additional improvements are possible. The first is the use of counters which can be reset electrically every hour on the hour. This would eliminate the process of subtracting the counter readings for successive hours. Unfortunately, a small, accurate revolution counter, which can be reset electrically, is still not available on the market. A unit could conceivably be constructed using standard resetable counters, although it would be extremely cumbersome and unreliable. The second improvement is an interpolation computer in which the db and percentage values are inserted and the intermediate values electronically determined. Using, for instance, the Gregory-Newton interpolation formula, weighted for log log $(100/P)$, such a device could be designed with a few hundred vacuum tubes. The initial cost, however, would probably outweigh its advantages.

Acknowledgment

Many thanks are due I. H. Gerks for his encouragement and advice. Mr. Gerks, who originally suggested that this paper be written, has prepared a discussion of the results of the 400-mc field-intensity measurements recorded at stations WHO and WTAD-FM.

Artificial Dielectrics for Microwaves*

W. M. SHARPLESS†, SENIOR MEMBER, IRE

Summary—**This paper presents a procedure for measuring the dielectric properties of metal-loaded artificial dielectrics in the microwave region by the use of the short-circuited line method. Formulas, based on transmission-line theory, are included and serve as guides in predicting the approximate dielectric properties of certain loading configurations.**

INTRODUCTION

FOLLOWING THE PUBLICATION of a recent paper on metallic delay lenses,[1] interest has been stimulated in the use of metal-loaded artificial dielectrics in the microwave field. Several papers[2–4] dealing with the theoretical analysis of metal-strip delay structures may be found in the recent technical literature. One paper[5] suggests a method of measuring the dielectric properties of such materials by the use of a microwave interferometer.

It is the purpose of this paper to present a procedure for measuring the dielectric properties of metal-loaded artificial dielectrics in the microwave region by the short-circuited coaxial-line method.[6] The use of an interferometer, a lens, or a prism structure to determine the properties of artificial dielectrics was discarded because these methods require the construction of a large sample for measurement.[7] By the short-circuited line method it is possible to measure small samples of artificial dielectric material over an extended frequency range in a single coaxial cavity, provided that the sample is prepared in such a way as to eliminate edge effects.

Simple formulas, based on transmission-line theory, are also presented and may be used to estimate proximity and frequency effects in strip-loaded artificial dielectric material.

Calculations and measurements of the dielectric constant and transmission loss are given for a few typical metal-loaded dielectric configurations. Fig. 1 shows a sketch of a strip-loaded type of dielectric material that is useful for microwaves. The measurements and calculations are based principally on this type. The measurements were all made in the wavelength range between 3 and 8 cm.

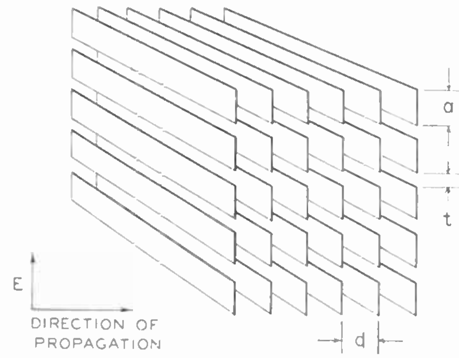


Fig. 1—Thin metallic strip artificial dielectric material.

PREPARATION AND MEASUREMENT OF TEST SAMPLES

Referring to Fig. 1, it will be noticed that if the dielectric were infinite in extent, thin metal plates might be passed through it at right angles to the voltage vector and midway between the metallic loading elements without affecting the field distribution within the dielectric. This means that a single metal-bounded sandwich section of the dielectric may be removed from the whole for measurement purposes; this done, the section removed may be bent into a doughnut shape, keeping the faces of the loading elements in the same planes as occupied in the whole. As the loaded section is now closed upon itself, all edge effects are eliminated and a tube of the material of length equal to the original depth is formed. This tube is now inserted in a short-circuited coaxial chamber for measurement,[8] as shown in Fig. 2 (following page). The spacing between the last metallic loading element and the short circuit is made $d/2$.

We now treat the metallic loaded section as if it were a sample of a homogeneous dielectric material. If a sample of homogeneous dielectric material is inserted

* Decimal classification: R282.9×R310. Original manuscript received by the Institute, August 11, 1950; revised manuscript received, February 23, 1951.

The material contained in this paper has been submitted to the University of Minnesota in partial fulfillment of the requirements for the degree of Professional Engineer.

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

[1] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, vol 27, pp. 58–82; January, 1948.

[2] S. B. Cohn, "Analysis of metal-strip delay structures for microwave lenses," *Jour. Appl. Phys.*, vol. 20, pp. 257–262; March, 1949.

[3] L. Brillouin, "Wave guides for slow waves," *Jour. Appl. Phys.*, vol. 19, pp. 1028–1041; November, 1948.

[4] M. A. Brown, "The design of metallic delay dielectrics," *Jour. IEE* (London), vol. 97, part III, pp. 45–47; January, 1950.

[5] B. A. Lengyel, "A Michelson type interferometer for microwave measurements," PROC. I.R.E., vol. 37, pp. 1242–1244; November, 1949.

[6] For a general background on the various methods used in measuring the dielectric properties of ordinary homogeneous dielectric materials, the reader is referred to a recent book edited by C. G. Montgomery, "Technique of Microwave Measurements," Radiation Laboratory Series 11, Chap. 10 (by R. M. Redheffer), McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

[7] Since the writing of this paper an article has appeared by S. B. Cohn, "Electrolytic tank measurements for microwave metallic delay-lens media," *Jour. Appl. Phys.*, vol. 21, pp. 674–680; July, 1950. This paper shows how the low-frequency index of refraction of such a medium may be calculated from electrolytic tank measurements on individual loading elements.

[8] The slight "curvature effect" introduced in place of the "edge effect" is comparatively small and calculations indicate that for the cases considered the dielectric constant is decreased only a few per cent by this transformation.

at the short-circuited end of a line, the standing-wave pattern will shift by an amount dependent on the length and dielectric constant of the sample inserted. The di-

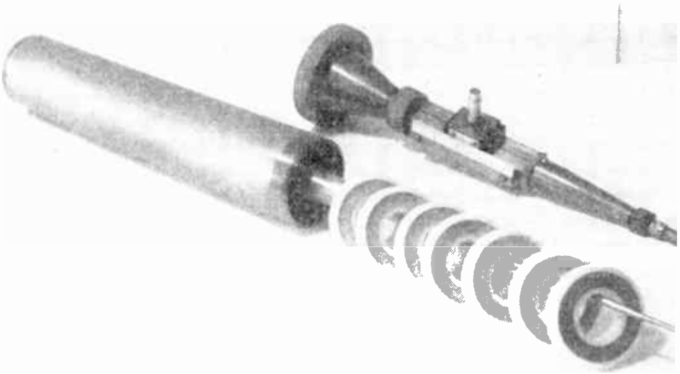


Fig. 2—Photograph of an elongated coaxial chamber with a few sections of the strip-type dielectric material.

electric constant of the material is calculated, as shown in Fig. 3, from the shift in the standing-wave pattern.

The loss in the sample of dielectric material is calculated from the change in the magnitude of the standing wave. With the dielectric sample inserted, the standing-wave ratio will be maximum when the electrical length

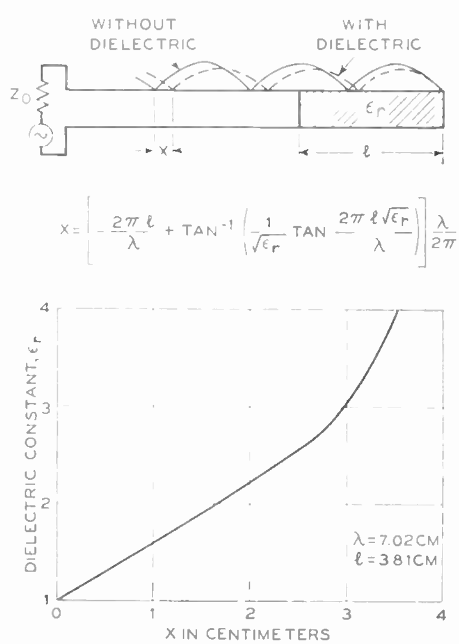


Fig. 3—Standing-wave relations used for determining the dielectric constant.

of the sample is a multiple of a half-wavelength, and minimum when the electrical length of the sample is an odd multiple of a quarter-wavelength. For low-loss materials, transmission-line theory gives

for $l = m\dfrac{\lambda}{2}$ where m is an integer

$$\text{loss} = \frac{8.7}{l}\left[\frac{\sqrt{\epsilon_r}}{S_T} - \frac{\sqrt{\epsilon_r}}{S_0}\right] \text{ db per meter,} \tag{1}$$

for $l = (2m-1)\dfrac{\lambda}{4}$

$$\text{loss} = \frac{8.7}{l}\left[\frac{1}{\sqrt{\epsilon_r}S_T} - \frac{\sqrt{\epsilon_r}}{S_0}\right] \text{ db per meter,} \tag{2}$$

where S_T = voltage standing-wave ratio with sample inserted, S_0 = voltage standing-wave ratio with sample removed, l = length of sample in meters, and ϵ_r = the dielectric constant relative to air.

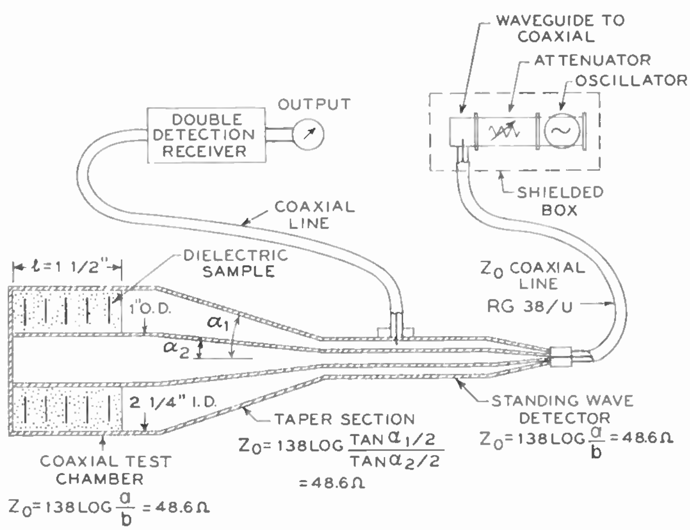


Fig. 4—Apparatus arrangement for measuring dielectric properties.

Apparatus

Fig. 4 is a sketch of the apparatus used in making the measurements. The coaxial-test chamber used for dielectric constant measurements is shown. This chamber allows for a representative sample of the dielectric, and is satisfactory for measurements on materials such as those shown in Fig. 1, where $a+t$ is equal to $\frac{5}{8}$ inch. The enlarged section is made short to prevent the supporting of higher order modes. The tapered section is made smooth and accurately coaxial to prevent any tendency to generate unwanted modes. By taking these precautions no trouble due to multiple moding was experienced with this short test section. The remaining equipment shown on the sketch is composed of standard components.

Theoretical Relations

The metal-loaded material shown in Fig. 1 can be represented by the strip-loaded transmission line shown in Fig. 5(a). The equivalent transmission line with lumped circuit elements is shown in Fig. 5(b). Using simple transmission-line equations, the impedance looking into the short-circuited line AB–GH (Fig. 5(a)) is

$$Z_{AB} = R_{AB} + j\omega L_{AB}$$
$$= R_{AB} + j120\pi \frac{d}{b} \tan\left(2\pi \frac{a/2}{\lambda}\right), \tag{3}$$

where b is the assumed width of the line (in the magnetic plane at right angles to the paper). Hence, the loading inductance in the equivalent loaded line (Fig. 5(b)) is

$$L_{AB} = 20 \frac{d\lambda}{b} \tan\left(\pi \frac{a}{\lambda}\right) 10^{-8}. \tag{4}$$

The total inductance per unit length of the equivalent line is the inductance of a tape line of spacing t plus the loading inductances,

$$L' = 40\pi \frac{t}{b} 10^{-8} + 2L_{AB}/d$$
$$= 40\pi \frac{t}{b}\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right) 10^{-8}. \tag{5}$$

When d/t is small, the capacity per unit length is

$$C' = \frac{1}{36\pi\ 10^9} \frac{b}{t}.$$

In general

$$C' = \frac{1}{36\pi\ 10^9} \frac{b}{t} K. \tag{6}$$

In a private communication, Schelkunoff gives the formula

$$K = \frac{\pi t}{2d \cosh^{-1} e^{\pi t/2d}} \tag{7}$$

where $d \leqq t$.

The velocity in the equivalent line is

$$V' = \frac{1}{\sqrt{L'C'}} = \frac{3 \times 10^8}{\sqrt{K\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right)}}. \tag{8}$$

The equivalent dielectric constant is $[3 \times 10^8/V']^2$; or

$$\epsilon_r = K\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right). \tag{9}$$

The impedance of the equivalent line is

$$Z_0' = \sqrt{\frac{L'}{C'}} = 120\pi \frac{t}{b} \sqrt{\frac{1 + \frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda}}{K}}. \tag{10}$$

By (9)

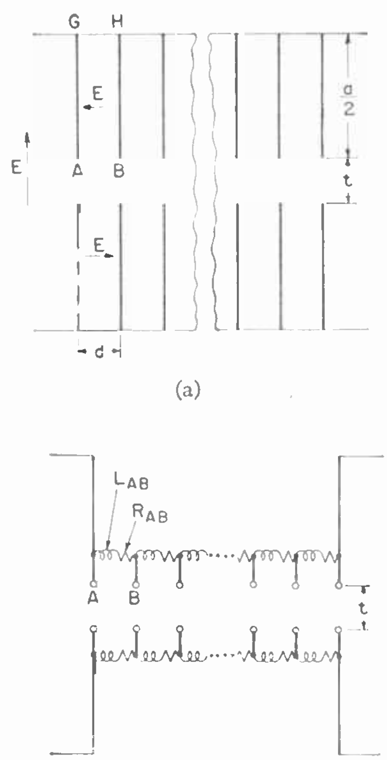
$$Z_0' = 120\pi \frac{t\sqrt{\epsilon_r}}{bK}. \tag{11}$$

The attenuation of the equivalent line is largely caused by the series resistance R_{AB} Fig. 5(b)), which is the resistive component of the input impedance to line AB–GH (Fig. 5(a)). For $a < \lambda/4$,

$$R_{AB} \approx \frac{a}{2} R,$$

where R is the resistance per unit length of line AB–GH. For copper ($g = 5.8 \times 10^7$),

$$R_{AB} \approx \frac{a}{2} \frac{9.06 \times 10^{-3}}{b\sqrt{\lambda}}.$$



(a)

(b)

Fig. 5—Equivalent transmission line with lumped circuit elements. (a) Cross section of strip-loaded material of width b. (b) Equivalent transmission line of width b.

The approximate attenuation of the equivalent line is

$$T' = 8.7 \frac{2R_{AB}}{d} \frac{1}{2Z_0'}$$
$$= \frac{aK}{td\sqrt{\epsilon_r}\sqrt{\lambda}} 1.05 \times 10^{-4} \text{ db per meter (for copper).} \tag{12}$$

The units used in these equations are mks, and thus all dimensions are in meters.

Results

Fig. 6 shows the coaxial measuring section together with several single-layer samples of loaded and unloaded dielectric materials in the foreground. On the extreme lower left is depicted the coaxial equivalent of a single section of a metallic strip dielectric of the type shown in Fig. 1. Low-loss styrofoam[9] is used as the spacer material.

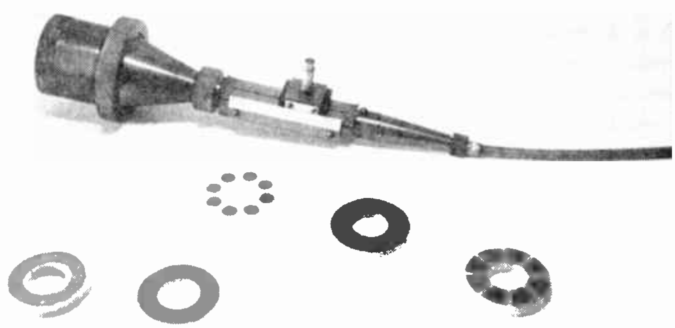


Fig. 6—Photograph showing coaxial measuring section together with several samples of dielectric materials.

Table I below lists the important dimensions, as well as the measured and calculated values for the dielectric constant ϵ_r of several different strip dielectrics. It is seen that the measured and calculated values of the dielectric constant agree in most cases to better than 10 per cent.

TABLE I

a (inch)	d (inch)	t (inch)	K (Equation (7))	λ (cm)	Dielectric Constant ϵ_r Calculated (Equation (9))	Dielectric Constant ϵ_r Measured (Equation (9))
$\frac{1}{2}$	$\frac{3}{32}$	$\frac{1}{8}$	0.755	3.18	8.38	
				7.00	4.13	4.32
0.365	$\frac{3}{16}$	0.26	0.766	3.18	2.39	2.10
				7.00	1.95	1.68
0.365	$\frac{3}{8}$	0.26	0.622	7.00	1.55	1.58

Loading arrangements other than the strip type have also been investigated. The solid curve of Fig. 7 shows the variation with wavelength of the dielectric constant ϵ_r of a dielectric material consisting of thin, isolated, $a=\frac{1}{2}$-inch square metal loading elements, spaced $\frac{1}{8}$ inch. The space between layers, $d=\frac{3}{8}$ inch, is again filled with low-loss styrofoam. The coaxial equivalent of a single section of this material is shown on the extreme right of Fig. 6. The dotted curve of Fig. 7 is calculated from (7) and (9) for a continuous strip dielectric with the same dimensions. The calculations for the strips gave, as expected, a slightly higher dielectric constant than the measured values for the closely spaced squares. According to (9), resonance will occur when $a=\lambda/2$. It is seen that this point is rapidly being approached at the left side of Fig. 7. To avoid anomalous dispersion regions, these particular materials should not be used for a wavelength below about 5 cm.

[9] Low-loss styrofoam has a measured dielectric constant of only 1.03 ± 0.01 in the microwave range, and thus forms an attractive light-weight spacer material.

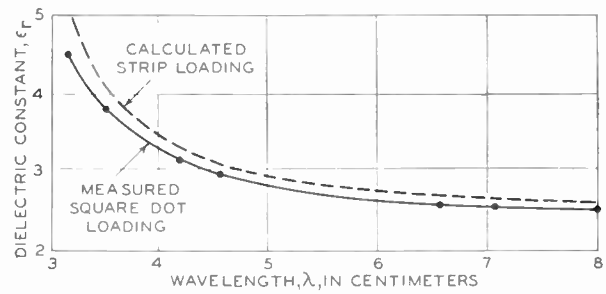


Fig. 7—Variation of the dielectric constant of two artificial dielectrics with wavelength. $a=1/2$ inch, $t=1/8$ inch, $d=3/8$ inch.

The transmission loss, as stated earlier, may be calculated from (1) or (2) if we know the magnitude of the standing-wave ratio in the chamber when the electrical length of the dielectric sample is either a multiple of a half wavelength or an odd multiple of a quarter wavelength. The standing-wave ratio for the air-filled chamber must also be known.

Fig. 8 shows a measured curve of the variation of the standing-wave ratio with sample length for a strip-type dielectric material made up in sections $d=\frac{3}{8}$ inch thick. This curve was taken at a frequency of 3,919 mc and the length was changed in steps equal to d; the circled points show the information so obtained. The exact maximum and minimum values on the curve, indicated by X's, were arrived at by slight frequency variations in the "peak" and "valley" regions. Only a single maximum or minimum need be known to solve either (1) or (2); however, when several cycles are taken, as in

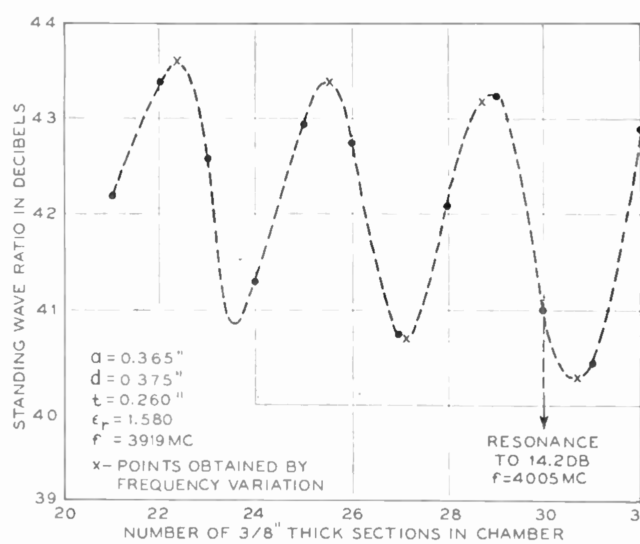


Fig. 8—Variation of the measured standing-wave ratio with the length of a dielectric sample.

Fig. 8, the multiple solutions possible improve the accuracy of the loss measurement. By the use of the longer chamber, shown in Fig. 2, it was possible to allow for the length variation needed in obtaining such a curve.

When extremely low-loss materials are being measured, as is the case for most metal-loaded dielectrics, the observed standing-wave ratios must be corrected by taking into account the losses in the line between the probe of the standing-wave detector and the face of the dielectric sample. This correction was calculated and applied to the measurements of Fig. 8 before using (1) and (2). Furthermore, care must be taken to avoid making measurements in regions where the chamber itself may become resonant at a higher mode. One such resonance was found in taking the curve of Fig. 8 when 30 sections were in the chamber. Such resonances might be eliminated by placing the proper "killers" in the enlarged chamber section; but in most cases this will not be necessary if sufficient points are taken on a curve for a resonance to be recognized when it occurs.

The measurements shown on Fig. 8 were taken with the dielectric material pictured in Fig. 2, which was an aluminum-strip type where $a=0.365$ inch, $t=0.26$ inch, and $d=0.375$ inch. The measured transmission loss was only 0.0535 db per meter, which is equivalent to a loss tangent of 0.00012. The calculated loss, as given by (12), is 0.0358 db per meter.[10] This agreement between the measured and the calculated losses is quite satisfactory.[11] Equation (12) applies for copper-loading elements. If material of higher resistivity is used for loading, the transmission loss will increase by the square root of the resitivity ratio. Equation (12), therefore, may be used for estimating the effect of changes in the design of strip-loaded dielectric material. For example, the equation shows that as the density of the loading elements is increased the transmission loss will also increase.

[10] This loss was calculated on the basis of the dc resistivity of aluminum, as given in handbooks. Actually, at microwaves, the loss will be a few percent higher because of surface roughness. See paper by A. C. Beck and R. W. Dawson, "Conductivity measurements at microwave frequencies," PROC. I.R.E., vol. 38, pp. 1181–1190; October, 1950.

CONCLUSIONS

The short-circuited coaxial-line method has been found useful for the measurement of the dielectric constant of metal-loaded artificial dielectric material in the microwave region. Samples must be prepared in a particular manner to permit measurements to be made in a coaxial cavity, but edge effects are eliminated in the process and only small-sized samples are required for the measurements. The loss factor of artificial dielectrics also may be measured by this method although considerable care is required to obtain reliable measurements on extremely low-loss materials. The measurements of dielectric constant and transmission loss are in good agreement with values calculated from approximate formulas based on simple transmission-line theory.

ACKNOWLEDGMENT

The author wishes to express his appreciation to members of the Holmdel Radio Laboratories, and especially to W. E. Legg, for assistance and co-operation during the course of this work.

[11] In the case of metal-loaded dielectrics, the so-called dielectric loss, for all practical purposes, is almost entirely in the form of power dissipated in the metallic loading elements.

A Precise Sweep-Frequency Method of Vector Impedance Measurement*

D. A. ALSBERG†, MEMBER, IRE

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

Summary—**The impedance of a two-terminal network is defined completely by the insertion loss and phase shift it produces when inserted between known sending and receiving impedances.**

Recent advances in precise wide-band phase and transmission measuring circuits have permitted practical use of this principle. Reactive and resistive impedance components are read directly from a simple graphical chart in which frequency is not a parameter. The basic principle described promises attractive possibilities in many cases of impedance measurements where present methods are inadequate.

* Decimal classification: R244. Original manuscript received by the Institute, March 26, 1951; revised manuscript received April 10, 1951. Presented at the second joint AIEE-IRE-NBS Conference on High-Frequency Measurements, Washington, D. C., January 11, 1951.

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

INTRODUCTION

IMPEDANCE is a fundamental property of any transmission facility; thus, from the earliest beginning of electrical transmission impedance has been a parameter of importance. As the number of links constituting a transmission facility increases, the properties of its component elements must be measured with increasing precision. Extreme examples are the transcontinental television transmission systems now nearing completion. A representative planned coaxial-cable system is made up of about 1,000 repeaters in tandem from New York to Los Angeles; a corresponding microwave radio-relay system consists of about 116 repeater stations in tandem. The most precise impedance requirements are assigned to terminations used to interconnect networks and other transmission elements and to components. Often even small impedance irregularities in their summed effect result in severe impairment of the transmission characteristics. In many cases it is inefficient to measure impedances at discrete frequencies, and sweep-frequency measurements must be made in order to detect significant irregularities. Often impedances must be measured in positions which do not permit short leads to the measuring circuit.

Impedance bridges do not lend themselves readily to sweep-frequency techniques, and must be located close to the impedance to be measured. Similarly, slotted lines and the three-voltmeter method do not adapt themselves to sweep methods.

Numerous devices, such as hybrids and directional couplers, have been developed which use variations of the reflected-energy principle. These adapt themselves to sweep-frequency techniques. However, they all require some form of mutual coupling, which restricts the bandwidth over which they are usable for sweep-frequency methods. The internal phase-shift changes and attendant calibration corrections of these devices are usually so severe that sweep-frequency measurements have only been used to measure the magnitude but not the phase, of an impedance.

Impedances which have to be measured with the greatest precision are those in the vicinity of transmission-line impedances, terminations, and connectors. Commercially available bridges do not possess adequate accuracy in many applications.

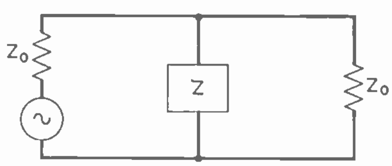


Fig. 1—Impedance shunt insertion.

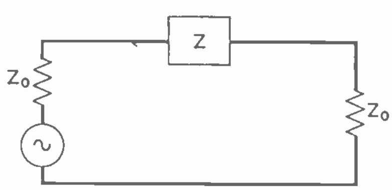


Fig. 2—Impedance series insertion.

General Principle

Since the development of highly precise phase- and transmission-measurement methods, a principle of impedance measurement has become practical, which, heretofore, has had little attention despite its many inherent advantages. Let us assume, in the ideal case, a transmission line terminated on both the sending and receiving ends with its characteristic impedance. If we now insert an impedance Z in shunt (Fig. 1) or in series (Fig. 2) with this line, the insertion loss and phase shift thus produced are directly a measure of the impedance Z in terms of the characteristic impedance of the line. A measuring circuit using this principle consists essentially of an insertion loss and phase-shift measuring circuit (Fig. 3) incorporating a fixed reference transmission line and a measuring transmission line having some provision for inserting the two-terminal impedance Z into the line.

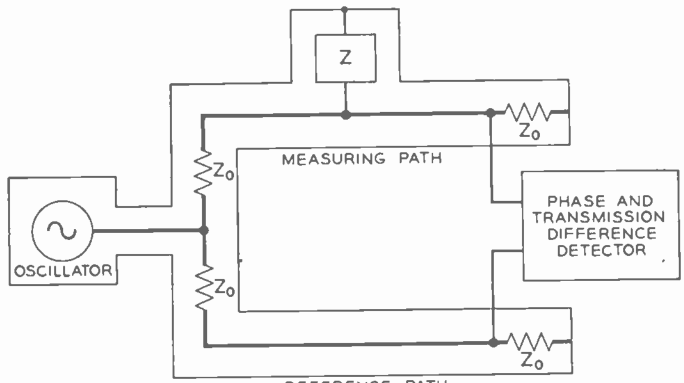


Fig. 3—Basic measuring circuit.

Some advantages of the insertion loss and phase principle are immediately evident.

1. The perturbation created in the transmission line by insertion of the impedance is directly the measure of the impedance. No mutual coupling or probes are required as an intervening link.
2. Impedance is measured as a transmission property, thus constant impedance corresponds to constant transmission, independent of frequency.

These two properties in combination permit precise sweep-frequency measurements over a much wider band of frequencies than are possible with any other known method.

3. As the transmission line is terminated in its characteristic impedance, no restriction is placed upon its length. Thus the junction point for insertion of the unknown impedance can be as remote from the measuring circuit as desired, without any impairment of accuracy.

Definitions and the exact equations for the relationship of insertion loss and phase and impedance or admittance are given in the Appendix.

In practice it would be cumbersome to solve these equations for every measurement. To avoid this, convenient charts have been prepared which are entered with the measured insertion loss and phase so as to read directly resistive and reactive impedance components.

The first chart (Fig. 4) covers all impedances containing positive resistances, and is drawn on a normalized impedance basis. The impedance grid is similar to the familiar Smith chart. Insertion loss and phase are represented by circles and radii centered on the point of zero impedance. The chart yields answers for either shunt impedance or series admittance to two significant figures. It is used for solving problems where this ac-

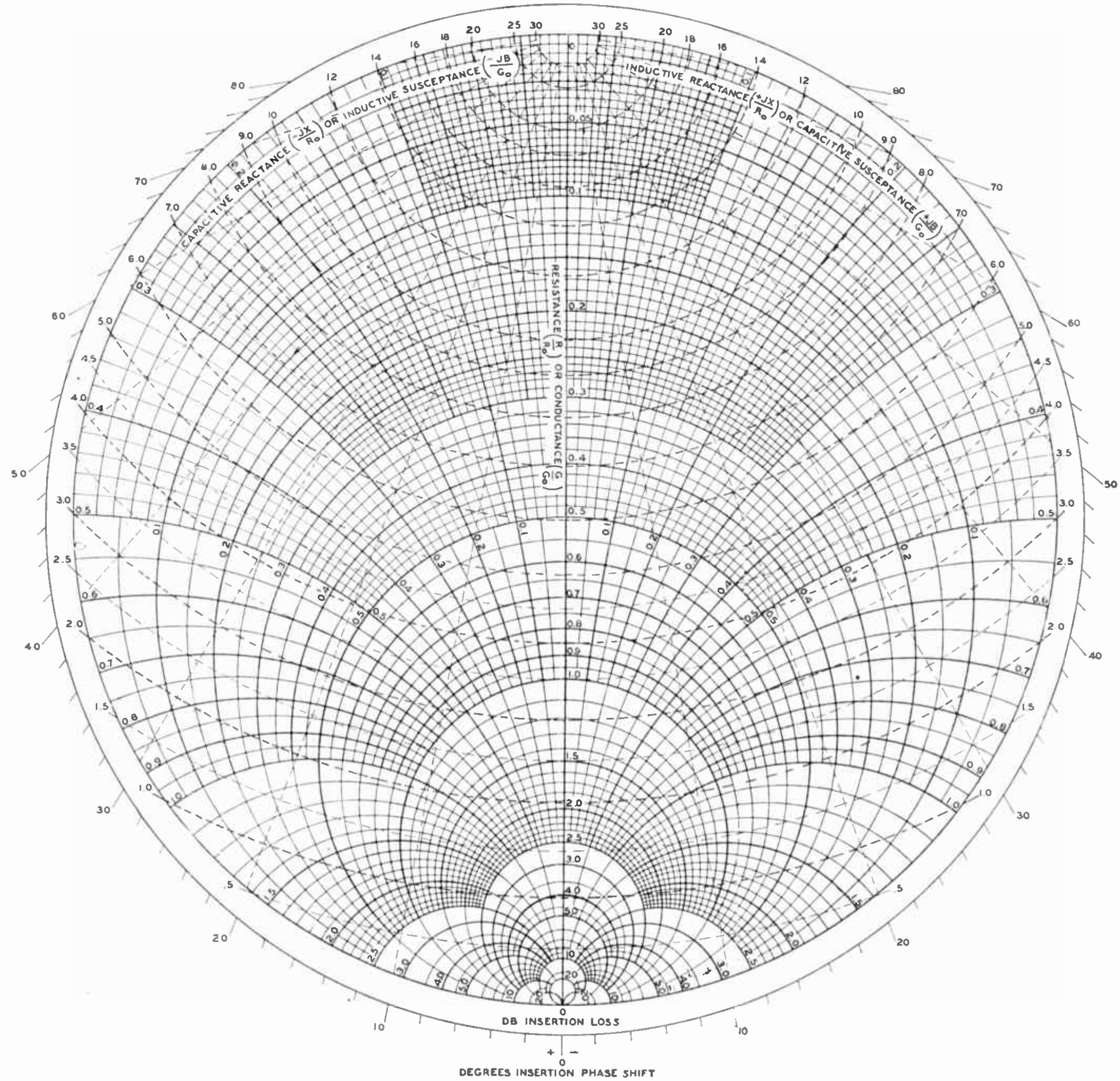


Fig. 4—Universal impedance and admittance chart. Insertion and phase shift produced by normalized shunt impedance (Z/R_0) or series admittance (Y/G_0) between terminations R_0 or G_0.

curacy is adequate, and it is used also for evaluating the range of loss and phase shift a measuring set must cover in order to measure a predetermined range of impedances.

The following chart (Fig. 5) was drawn specifically for 75- and 50-ohm circuits, to cover a range of impedance of ± 20 per cent about the circuit impedance. It is in this region where the majority of the most precise impedance measurements have to be made. It may be read to an accuracy of about 0.1 per cent. Resistance and reactance are read directly in ohms. It should be noted that in order to cover an impedance variation of ± 20 per cent, or 10 per cent reflection coefficient, about the circuit impedance, a loss range of only ± 0.5 db and a phase range of ± 3.5 degrees are required.

If other impedance ranges have to be read with higher precision than is possible with Fig. 4, existing published equalizer charts[1] may be used, or, when warranted, special charts may be readily constructed.

Limitations and Errors of Measurement

As may be seen from inspection of the impedance charts, accurate measurement of impedances relies on the ability to measure with high accuracy small losses and phase shifts. The alternative reflection loss and phase method relies on measuring larger losses and phase shifts with less accuracy.

[1] "Equalizer Charts," Bell Telephone System Monograph B-1643.

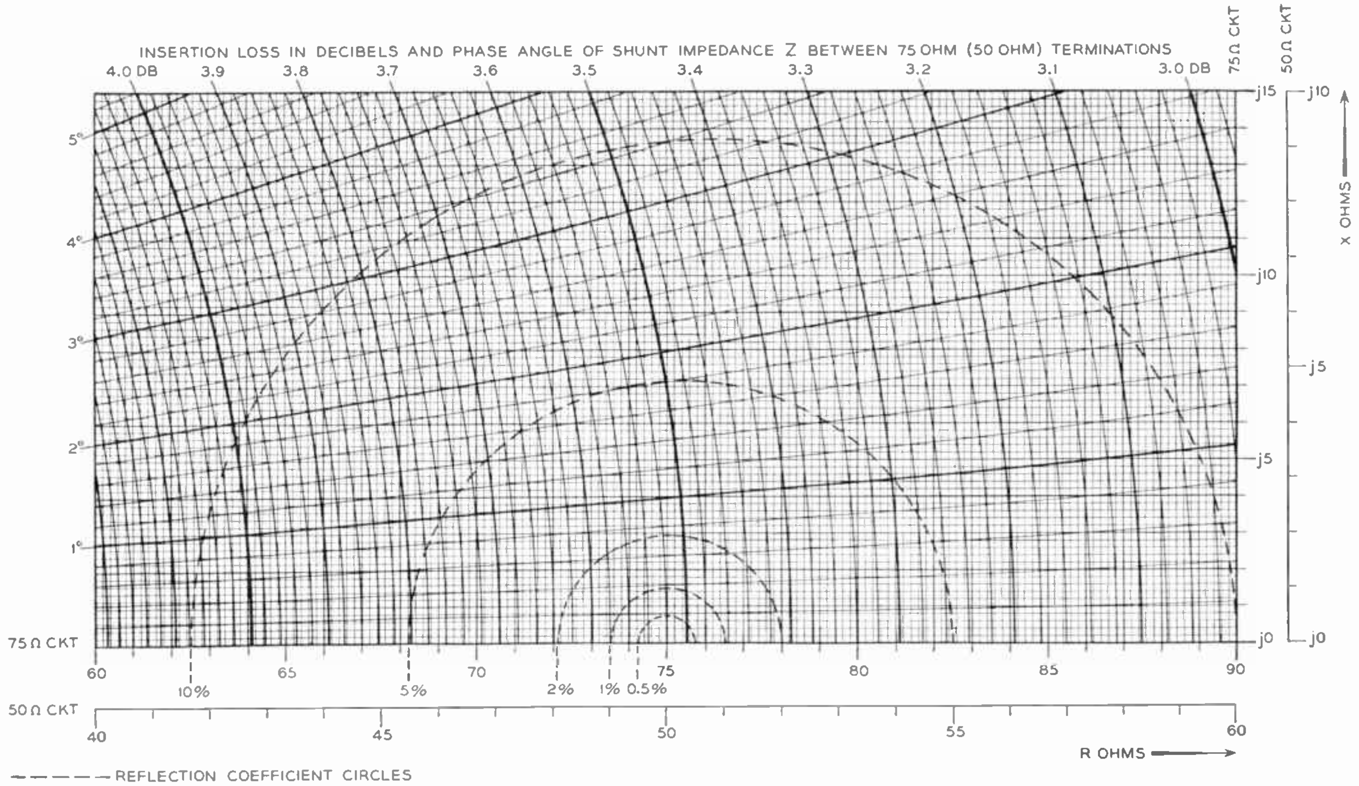


Fig. 5—Impedance chart for impedances in the vicinity of 50 and 75 ohms.

At high frequencies the effects of residual capacitances and inductances are often sufficiently large to make it necessary to use shields. Such shields can be grounded conveniently when shunt insertion is used. When using series insertion for measuring large impedances, the uncontrolled stray impedances to ground limit the accuracy of measurement, and yet many satisfactory results have thus been obtained in measurements on spurious resonances of piezoelectric crystals.

The effect of the fixture used to insert the unknown impedance into the phase- and loss-measuring circuit may be difficult to evaluate, but it may be eliminated by a simple procedure. In a coaxial circuit, a tee is inserted in the line connecting the test circuit generator to the detector. To establish the reference zero of the circuit, the open end of the tee is terminated with a known impedance standard and the test circuit is adjusted to read the nominal loss and phase shift associated with the impedance standard. For instance, we find from Fig. 5 that a 75.00-ohm impedance shunted across a 75.00-ohm circuit produces a 0.0-degree insertion phase shift and a 3.52-db loss. This setting is then as accurate as the impedance standard is known. Using coaxial-line techniques and thin film resistors, it is possible to design standards for frequencies up to 80 mc, whose reactance component is less than 0.1 per cent of the nominal impedance and whose resistance component can be determined by dc measurement.

As the coaxial cables used to connect the measuring circuit have a surge impedance which is a function of frequency, and which is subject to manufacturing tolerances, the impedance Z_0 is not constant over a wide frequency band and is dependent on the specific cable used. If the direct-reading chart (Fig. 5), which assumes a constant R_0 of 75 ohms or 50 ohms, is used, an error then results. To determine if this error is negligible, the formulas derived in the Appendix may be used.

The impedance measured is the one at the junction point of the unknown impedance with the transmission line. Often the impedance to be measured cannot be inserted physically at the junction point, and a coaxial line is used between the coaxial tee and the unknown impedance. If the surge impedance of this line and the unknown impedance are significantly different, the measured impedance value must be corrected for the effects of the line by use of conventional transmission-line theory, most conveniently by use of the Smith Chart.[2] To avoid this correction, it is advisable to make the connection between unknown impedance and the junction point as short as possible, and to lengthen instead the cables connecting the junction point to the generator and detector. The impedance of these connecting cables can be measured precisely, however; if it is found to be substantially different from the design impedance R_0, significant errors can be computed from (20) in the appendix. Thus, the unknown impedance can be measured as remotely from the insertion-phase and -loss measuring circuit as desired without impairment of accuracy.

[2] P. H. Smith, "An improved transmission-line calculator," *Electronics*, vol. 12, pp. 29–30; January, 1939; also, *Electronics*, vol. 17 pp. 130–133; January, 1944.

Application

The method described was demonstrated first by using an improved version of a phase and transmission measuring circuit, previously described.[3] The improved version (Fig. 6) has the following characteristics:

Frequency range	0.05 to 20 mc (9 octaves)
Loss range	0 to 70 db
Accuracy (absolute, recording)	±0.02 db (up to 30-db loss)
	±0.1 degree
(differential, recording)	±0.01 db
	±0.05 degree
(differential, with special technique)	±0.002 db
	±0.01 degree
Zero characteristic less than	±0.1 degree and ±0.02 db from 0.05 to 20 mc.

Circuit impedance 75.0 ohms.

The circuit is self-tuning and the test-frequency oscillator is motor driven at the rate of 1 mc for sweep measurement from 0.05 to 20 mc. The insertion phase shift and loss may be read directly on the indicating meter, or may be automatically recorded.

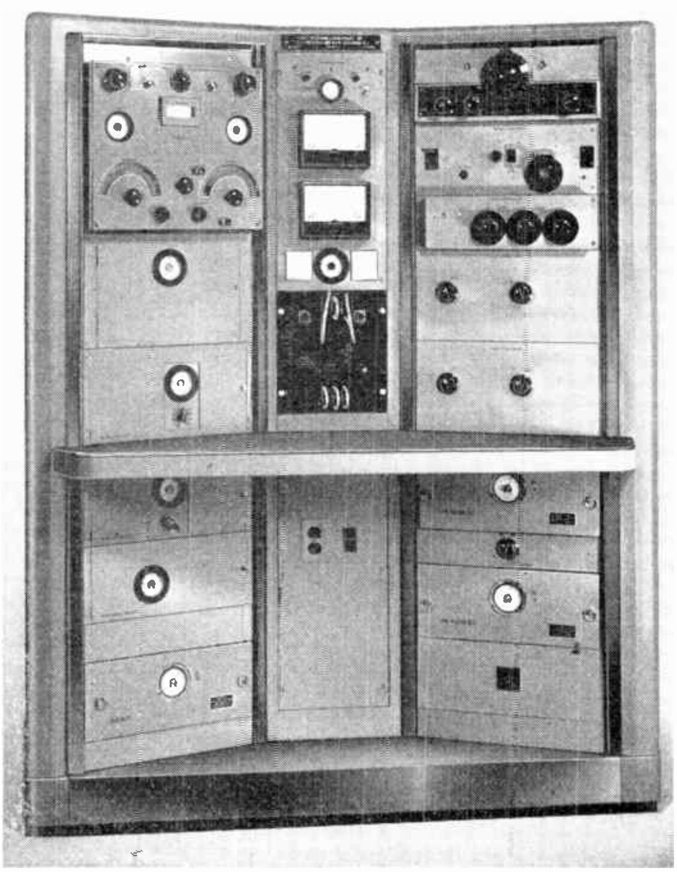


Fig. 6—50-kc to 20-mc phase, transmission and delay measuring set.

From Fig. 5 it is evident that, with a differential accuracy of 0.01 db and 0.05 degree, impedance accuracies of ±0.25 per cent are attained in the vicinity of 75 ohms. The 75-ohm coaxial tee (Fig. 7) is inserted into the test branch of the measuring circuit and the unknown impedance is inserted in the open arm of the tee. The motor drive of the test oscillator is started and the frequency is swept from 0.05 to 20 mc, or over any portion of the band. Deflections of the indicating meters reveal any impedance variations immediately.

Inspection of chart Fig. 4 shows that as the impedance increases, accuracy decreases, when shunt insertion is used. For instance, from Fig. 4 for an impedance of 750 ohms ($R/R_0=10$) a resolution of 0.01 db corresponds to a 20-ohm or 2.5-per cent uncertainty. However, high accuracy is attained for low impedances

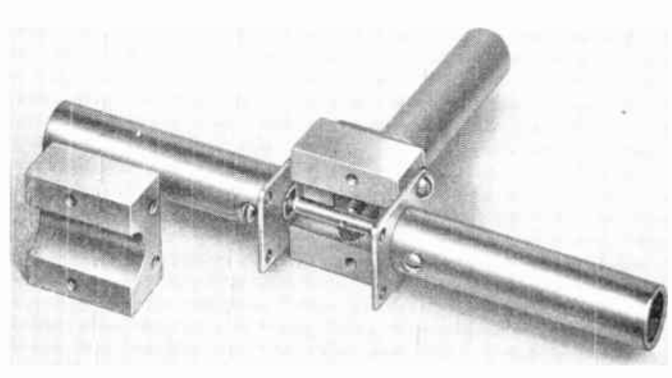


Fig. 7—75-ohm coaxial tee.

using shunt insertion. For example, from Fig. 4, an impedance of 1.5 ohms ($R/R_0=0.02$) corresponds to a loss of about 28 db; an accuracy of 0.02 db then yields an impedance uncertainty of about 0.003 ohms, or 0.2 per cent.

The differential accuracy limit of 0.01 db and 0.05 degree is set by the circuit stability. By increasing the indicating-meter sensitivity and averaging a number of consecutive measurements, it is possible to attain a statistical differential accuracy of 0.002 db and 0.01 degree, corresponding to impedance accuracies of ±0.05 per cent at 75 ohms.

The validity of the general impedance-measurement method was verified at frequencies up to 80 mc. The method has found practical use in the measurement of coaxial repeaters. quartz crystals, cable impedances, and precise terminations.

Examples

An example of a recording sweep measurement is shown in Fig. 8 on page 1398: 158 feet of RG 6/U cable were terminated with a resistance standard of 75.10 ohms. Both the transmission and phase records have two traces, one the reference zero trace, and one the measurement trace, the net measurement value being the difference between the two traces. As the deviation of the measured impedance from 75 ohms was small, phase and loss departure readings could be labeled directly in ohms without the need for conversion by use of graphical tables. The irregularity of the measurement trace shows clearly the frequency dependence and irregularity of the characteristic impedance of RG 6/U cable, a common fault of all flexible coaxial cables.

[3] D. A. Alsberg and D. Leed, "A precise direct reading phase and transmission measuring system for video frequencies," *Bell Sys. Tech. Jour.*, vol. 28, pp. 221–238; April, 1949.

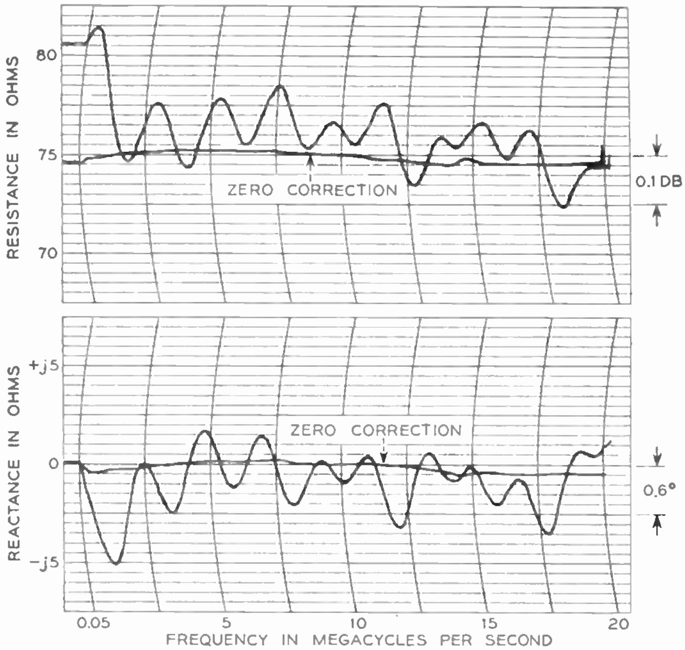


Fig. 8—Recording of input impedance of 158 ft of RG 6/U coaxial cable terminated in 75.10 ohms.

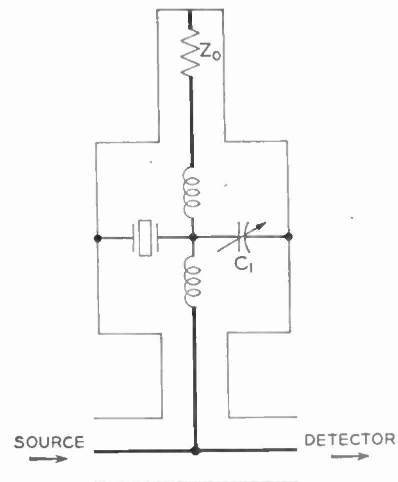


Fig. 10—Adjustment of low pass filter to absorb static crystal capacitance C_0.

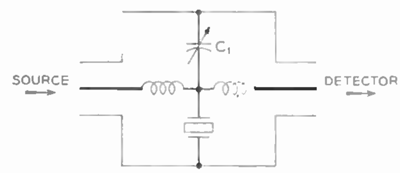


Fig. 11—Insertion of crystal and low pass filter in impedance measuring circuit.

Another important application has been the measurement of quartz-crystal primary parameters. Fig. 9 shows the conventional equivalent circuit of a quartz crystal. The measurement procedure is as follows in a 75 ohm circuit:

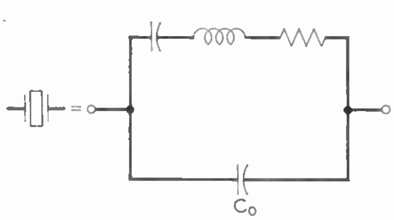


Fig. 9—Equivalent circuit for piezoelectric crystal.

The low-pass filter (Fig. 10) is terminated with a 75-ohm resistance standard. The crystal is inserted, as shown, and at a frequency remote from resonance the capacitor C_1 is adjusted until the filter is transparent to 75 ohms. Thus, the static crystal capacity C_0 is absorbed into the filter. After this initial adjustment the filter and crystal are inserted in the measuring circuit, as shown in Fig. 11. The resulting measurement then contains only the effective impedance of the series-resonance elements. As the static capacitance C_0 has been absorbed into the filter, zero phase shift occurs exactly at resonance. With this method the residual resistance, or Q, of high-frequency crystals, has been determined with higher accuracy than heretofore possible.

A typical low-pass filter structure for crystal measurements is shwno in Fig. 12.

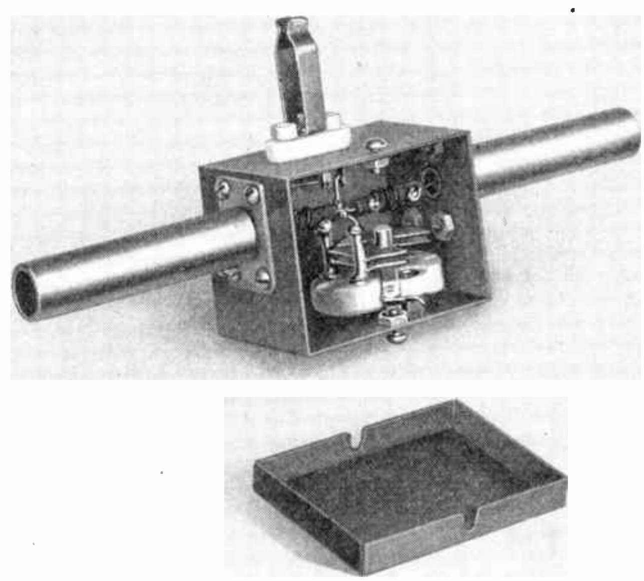


Fig. 12—Typical low-pass filter for piezoelectric crystal impedance measurements.

Conclusion

The examples used were cited as experimental verification of the possibilities of the basic method. It is the primary purpose of this paper to stimulate others to take advantage of the properties of the insertion phase and loss principle, which is a general principle of measurement not restricted to the coaxial measurement circuits. The fundamental properties are as follows:

1. The impedance is measured as a transmission property; thus, constant impedance corresponds to constant transmission, and frequency is not a parameter. This method may be adapted ideally to sweep-frequency measurement techniques.
2. The perturbation created in the transmission line by insertion of the impedance is the quantity directly of interest; in contrast to methods like slotted lines, where the disturbance created by probes is a source of error and limitation, no mutual impedance is required.
3. The measurement relies on the ability to measure precisely small phase shift and potential differences in contrast to methods using the reflected energy principle, such as hybrids, slotted lines, and directional couplers which require measuring large phase shifts and potential differences, though with less accuracy.
4. Measurement may be made at a remote distance from the measuring circuit.

Where an accurate impedance-measurement device for a limited impedance range is required, greatly simplified but very precise phase- and transmission-measuring circuits can be built. Where the general transmission measurement of a communications device requires more elaborate phase- and transmission-measuring facilities, the mere addition of a line-bridging device converts these facilities, at practically no added expense, to measure impedance as well.

In addition to enabling sweep measurements, the application of the insertion-phase and -loss principle has resulted in more precise vector-impedance measurements at high frequencies than have been obtained by other available methods. The method has not yet been fully exploited, and it offers the possibility of precision measurements at very high frequencies, now only attained in the best bridges at much lower frequencies.

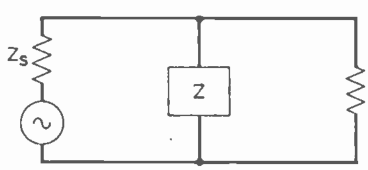


Fig. 13—Impedance shunt insertion definitions.

Appendix

Theory

The insertion loss α and phase shift β, produced by insertion of a two-terminal impedance between known sending and receiving impedances, is given by the following equations (see bibl. ref. 1):

For shunt impedance, (Fig. 13).

$$Z_0 = 2\frac{Z_r Z_s}{Z_r + Z_s}, \tag{1}$$

then

$$e^{\alpha + j\beta} = 1 + \frac{1}{2Z/Z_0} \tag{2}$$

if Z_0 is a pure resistance R_0.

$$e^{\alpha + j\beta} = 1 + \frac{1}{2Z/R_0} = 1 + \frac{1}{2(R/R_0 + jX/R_0)} \cdot \tag{3}$$

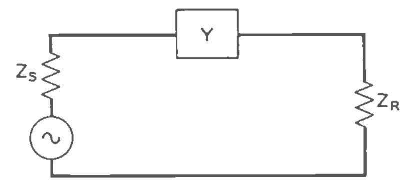


Fig. 14—Admittance series insertion definitions.

For series admittance (Fig. 14) let

$$Y_0 = \frac{2}{Z_s + Z_r}, \tag{4}$$

then

$$e^{\alpha + j\beta} = 1 + \frac{1}{2Y/Y_0} \tag{5}$$

if Y_0 is a pure conductance G_0.

$$e^{\alpha + j\beta} = 1 + \frac{1}{2Y/G_0} = 1 + \frac{1}{2(G/G_0 + jB/G_0)} \cdot \tag{6}$$

Both (3) and (6) are of the general form

$$e^{\alpha + j\beta} = 1 + \frac{1}{u + jv} \cdot \tag{7}$$

In (7) the loci of constant loss and phase shift are a family of orthogonal circles if v and u are plotted in rectangular co-ordinates.[1]

If we apply the complex transformation

$$u + jv = \frac{1 + \zeta}{1 - \zeta}, \tag{8}$$

where ζ is a complex number, to (7), it will result in a family of orthogonal circles representing u and v (Fig. 4). These are identical to the familiar Smith chart,[2] except for a scale factor of $\frac{1}{2}$ arising from the definition of Z_0 and Y_0. The outside rim of the chart $u=0$ is the unit circle in the ζ plane. The lines of constant insertion loss are now circles centered on $u=0$, $v=0$, the locus of which is defined by

$$\alpha = -20 \log \left| \frac{1+\zeta}{2} \right| \text{ decibels} \tag{9}$$

Lines of constant inertion phase shift are the radii of the loss circles in the ζ plane.

Equations (3) and (6) could also be written in terms of shunt admittance and series impedance, resulting in a chart similar to Fig. 4. This subject will be discussed more extensively by the author in a forthcoming article on equalizer charts.

Error Computation

Error occurs when actual Z_0 is different from R_0 assumed in direct reading charts.

From (1) let us define

$$Z_0 = R_0 + \delta_z. \tag{10}$$

The unknown impedance is defined as

$$Z = R_0 + \epsilon_z. \tag{11}$$

The measurement procedure fixes the loss and phase readings of the measuring set to the values prescribed by the impedance chart for the design impedance R_0 when the impedance standard is inserted, even though Z_0 differs from R_0 by δ_z. This introduces an error Δ in loss and phase readings. From (2) we can write the following:

The nominal value of the table for zero reference is

$$e^{\alpha+j\beta} = 1 + \frac{1}{2R_0/R_0} = 1.5. \tag{12}$$

The actual value measured for zero reference is

$$e^{\alpha'+j\beta'} = 1 + \frac{1}{2R_0/(R_0+\delta_z)} = 1.5 + \frac{\delta z}{2R_0}. \tag{13}$$

The nominal value of the table for Z is

$$e^{\alpha_1+j\beta_1} = 1 + \frac{1}{2Z/R_0}. \tag{14}$$

The actual value measured for Z is

$$e^{\alpha_1'+j\beta_1'} = 1 + \frac{1}{2Z/(R_0+\delta_z)}. \tag{15}$$

Therefore,

$$\Delta = -(\alpha+j\beta)+(\alpha'+j\beta')+(\alpha_1+j\beta_1)-(\alpha_1'+j\beta_1'). \tag{16}$$

Substituting (12), (13), (14), and (15) in (16), we find as a complete expression for Δ

$$\Delta = lg\left[1 + \frac{\frac{\delta_z}{3R_0}\left(1+\frac{R_0}{2Z}\right) - \frac{\delta_z}{2Z}}{\left(1+\frac{R_0}{2Z}\right)+\frac{\delta_z}{2Z}}\right]. \tag{17}$$

If δ_z is small, this yields

$$\Delta = lg\left(1+\frac{\delta_z}{3R_0} - \frac{\delta_z}{2Z+R_0}\right)$$

$$= lg\left(1+\frac{\delta_z}{3R_0} - \frac{\delta_z}{3R_0+2\epsilon_z}\right). \tag{18}$$

If the deviation ϵ_z of Z from R_0 is small, (13) approximates to

$$\Delta = lg\left(1 - \frac{\delta_z\epsilon_z}{4.5R_0^2}\right), \tag{19}$$

or as δ_z and ϵ_z are small,

$$\Delta = -\frac{\delta_z\epsilon_z}{4.5R_0^2} \text{ (nepers, radians).} \tag{20}$$

For example, let

$$\delta_z = \delta_R + j\delta_X \text{ ohms} \tag{21}$$

and

$$\epsilon_z = \epsilon_R + j\epsilon_X \text{ ohms.} \tag{22}$$

If (16) and (17) are inserted into (15), assuming $R_0 = 75\omega$ and converting to db and degree,

$$\Delta \text{ loss} = 3.4 \times 10^{-4}(\delta_R\epsilon_R - \delta_X\epsilon_X) \text{ decibels} \tag{23}$$

$$\Delta \text{ phase} = 2.26 \times 10^{-3}(\delta_X\epsilon_R + \delta_R\epsilon_X) \text{ degree.} \tag{24}$$

In practice Δ can be neglected in most applications. For a given δ_z the error Δ decreases linearly as the unknown impedance Z approaches R_0.

When the error, as given by (20), becomes sufficiently large to warrant correction of the chart readings, at times it will be found more convenient to use the actual Z_0 and Y_0, as defined in (1) and (4), directly in reading the charts rather than to use the arbitrary chart-design impedance plus correction from (20).

Bibliography

1. K. S. Johnson, "Transmission Circuits for Telephone Communications," D. Van Nostrand Co., Inc., New York, N. Y. Chap. 9, pp. 89–91; 1925.
2. O. J. Zobel, "Distortion correction in electrical circuits," *Bell Sys. Tech. Jour.*, vol. 7, pp. 438–534; July, 1928.
3. E. A. Guillemin, "Communications Networks," John Wiley and Sons, Inc., New York, N. Y. vol. 2, pp. 509–511; 1935.

Speech-Reinforcement System Evaluation*

L. L. BERANEK†, SENIOR MEMBER, IRE, W. H. RADFORD†, SENIOR MEMBER, IRE, J. A. KESSLER†, AND J. B. WIESNER†, SENIOR MEMBER, IRE

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

Summary—**Speech-reinforcement systems in six large auditoriums were evaluated, using subjective rating tests, word-articulation tests, and, in two cases, a new test method. This method, called the "terminal-word test," makes possible the quantitative measurement of speech intelligibility for a sound system in actual use. A graphical method is presented for calculating the performance of a sound system in which account is taken of the frequency response of the system, the reverberation time of the room, the directivity index of the loudspeaker, and the room noise. Test results indicate that a flat frequency response in the range between 400 and 4,000 cps is required for good intelligibility. The graphical method indicates that little further increase in intelligibility would result from extending this range upward or downward. If the loudspeaker system is sufficiently directive in this frequency range and properly located in the room, room reverberation has little effect on speech intelligibility.**

I. INTRODUCTION

EARLY IN 1949, in connection with the Mid-Century Convocation at the Massachusetts Institute of Technology, the authors designed three sound systems for use in auditoriums, ranging in size from 237,000- to 5.5-million cubic feet. These auditoriums were highly reverberant when empty, and considerable difficulty in obtaining satisfactory speech intelligibility was anticipated.

After preliminary study, it was decided to restrict the frequency responses to the interval between 200 and 7,000 cps, which is the range known to be of principal importance to speech intelligibility.[1,2] Two purposes were served by this decision: excitation of the low-frequency room resonances was avoided, and the cost of the loudspeaker installations was reduced. Although it was expected that a loss of naturalness would be observed because of the restricted low-frequency response, we were surprised to learn that many people judged these systems superior to average installations. Our interest was aroused in performing a series of experiments to determine some of the psychological factors involved in sound-system design for speech. No effort has been made to extend the studies to music.

The experimental program was planned to yield answers to the following questions: (a) Should the bass response of the system be restricted to preserve intelligibility in reverberant space? (b) If bass response is restricted, to what extent is naturalness impaired? (c) Does the articulation index theory apply to sound systems in reverberant surroundings? (d) What is the best location for the loudspeakers? (e) Can a suitable test be devised for measuring the performance of a sound system under normal operating conditions?

II. DESCRIPTION OF SYSTEMS

Our studies eventually included a cathedral, a memorial auditorium, and a highly reverberant industrial building, in addition to the three auditoriums mentioned above. In every case the loudspeakers consisted of one or more multicellular high-frequency units selected and installed so that from any seat the listener could see the throat of one of the horn cells. In Walker Memorial (Morss Hall) at M.I.T. (volume 237,000 cubic feet), a horn two cells high and five cells wide was used. In the Rockwell Cage at M.I.T. (volume 1,100,000 cubic feet), two horns, each two cells high and five cells wide, and one horn two cells high and four cells wide, were used. In the Boston Garden (volume 5,500,000 cubic feet), two horns, each three cells high and four cells wide, and one horn three cells high and six cells

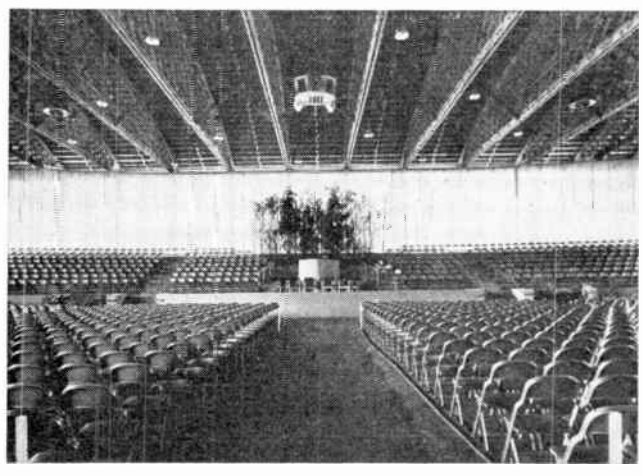


Fig. 1—View of Rockwell Cage showing loudspeaker installation. The speakers are mounted approximately 15 feet forward of and 20 feet higher than the microphone.

* Decimal Classification: R391×534. Original manuscript received by the Institute, March 1, 1951; revised manuscript received, June 18, 1951.

† Massachusetts Institute of Technology, Cambridge, Mass.

[1] L. L. Beranek, "Design of speech communication systems," PROC. I.R.E., vol. 35, pp. 880–890; September, 1947.

[2] N. R. French and J. C. Steinberg, "Factors governing speech intelligibility," *Jour. Acous. Soc. Amer.*, vol 19, pp. 90–119; January, 1947.

wide, were used. In the cathedral (volume 1,400,000 cubic feet), the equivalent of a two-cell high, five-cell wide horn was used. In the memorial auditorium (volume 1,160,000 cubic feet), a two-cell high, five-cell wide horn was used. In the industrial building (volume 990,000 cubic feet), two horns, one two cells high and six cells wide, and the other two cells high and four cells wide, were used. In addition, in the Rockwell Cage and in the memorial auditorium, two direct-radiator low-frequency units were used during portions of the experiments. A general view of the Rockwell Cage is shown in Fig. 1.

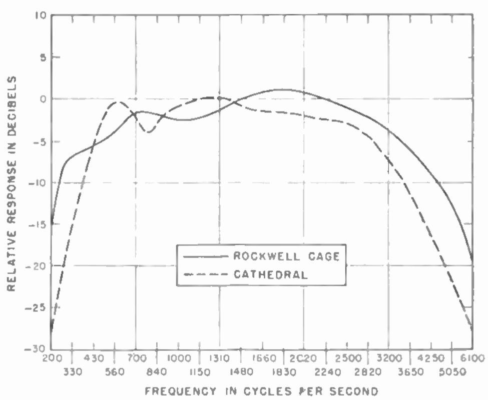


Fig. 2—Response of high-frequency channels of speech-reinforcement systems in Rockwell Cage and cathedral.

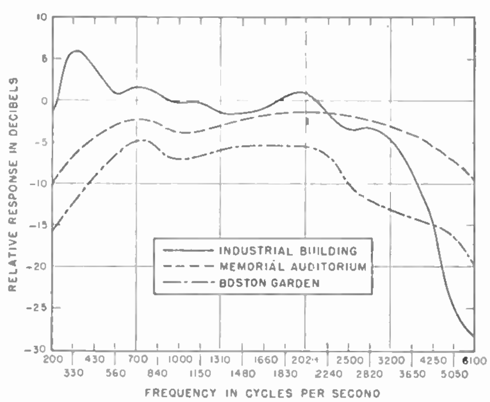


Fig. 3—Response of high-frequency channels of speech-reinforcement systems in industrial building, memorial auditorium, and Boston Garden.

In each auditorium, the sound pressure level produced by the sound system was measured as a function of frequency at several locations. In each test, the microphone was replaced by a warble-tone source having the same nominal internal impedance. The frequency response of the microphone was then added to that response curve to obtain an over-all response characteristic. Space averaging of the data was obtained by moving the microphone back and forth a few feet at each location. The resulting curves for the high-frequency channels alone, of five of the systems tested, are shown in Figs. 2 and 3. The abscissas of these figures show frequency in cycles per second plotted on a scale such that, in quiet, equal distances along the scale are of equal importance to speech intelligibility.[1,2] For example, the frequency range from 700 to 1,310 cps is as important to speech intelligibility as the range from 2,020 to 3,200 cps. Equipment by three different manufacturers was used. It should be noted that in nearly every case the system response falls off rapidly at high frequencies.

Distributions of relative sound levels over the main-floor seating area in four of the auditoriums are shown at two frequencies in Figs. 4 and 5. If 1-db contours were plotted on these figures, it would be seen that the sound levels varied over a range of as little as 5 db to as much as 12 db. Based on our observations, it is desirable to adjust the system for a maximum variation of about 6 db. However, this condition was achieved only in the memorial auditorium.

In two of the installations, considerable trouble from acoustic feedback was encountered when the systems were first tested. The feedback point was reduced 3 db in one case and seven in the other, by blocking off (with absorbent cotton) those horn cells which pointed directly at the walls on either side of the stage.

III. Acoustical Conditions

The reverberation times as a function of frequency were measured in five of the auditoriums with no audience present (see Fig. 6). In the cathedral alone, data also were taken with an audience. The reverberation times were then calculated for the number of people present during the tests[3] (see Fig. 7). Detailed reverberation data were not taken in the industrial building. At 500 cps, stop-watch observations indicated a reverberation time of about 5 seconds.

Although the desired reverberation time for amplified speech is believed to be less than 1 second,[4] it was found, by means of the techniques described in this paper, that a high level of intelligibility can be achieved, even when the reverberation time at 500 cps is as high as 2 to 4 seconds.

IV. Intelligibility Tests

Two different methods were used for measuring the speech intelligibility. One of these was the familiar word-articulation test in which one or more talkers and a number of listeners participated. The test words were monosyllabic and were used in phonetically balanced groups of fifty. Each word was read in one of the following carrier sentences: "You will write ______"; "Please write down ______"; "This time put ______"; and "I will read ______."

The other method of testing intelligibility was devised especially for these studies. It was performed at a public gathering while the scheduled speaker was talk-

[3] V. O. Knudsen and C. M. Harris, "Acoustical Designing in Architecture," John Wiley and Sons, Inc., New York, N. Y., chap. 8; 1950.

[4] R. H. Bolt and A. D. MacDonald, "Theory of speech masking by reverberation," *Jour. Acous. Soc. Amer.*, vol. 21, pp. 577–580; November, 1949.

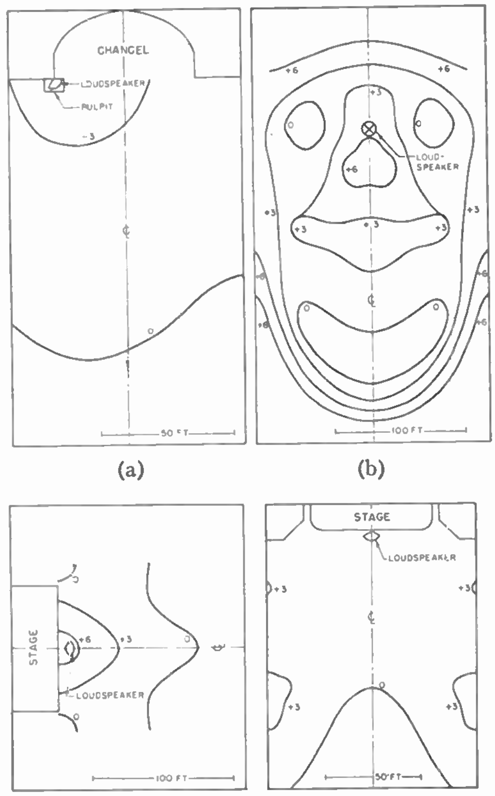


Fig. 4—Contours of constant sound pressure level at 500 cps for: (a) cathedral, (b) Boston Garden, (c) Rockwell Cage, and (d) memorial auditorium.

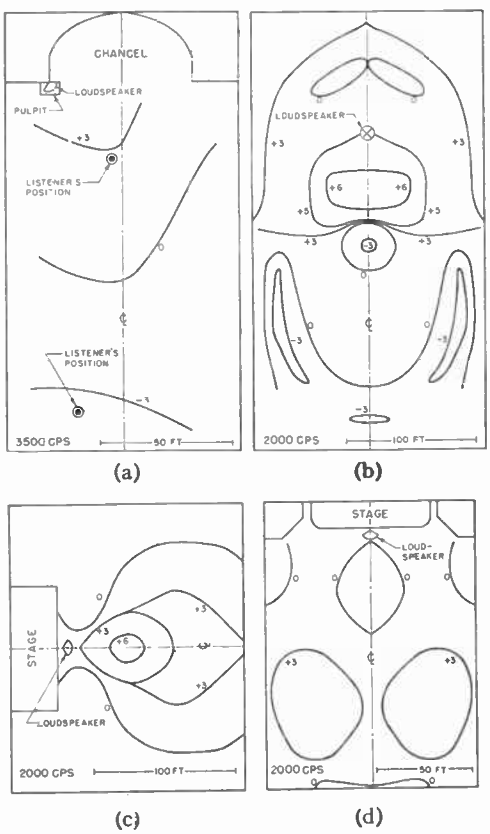


Fig. 5—Contours of constant sound pressure level at frequencies indicated for: (a) cathedral, (b) Boston Garden, (c) Rockwell Cage, and (d) memorial auditorium.

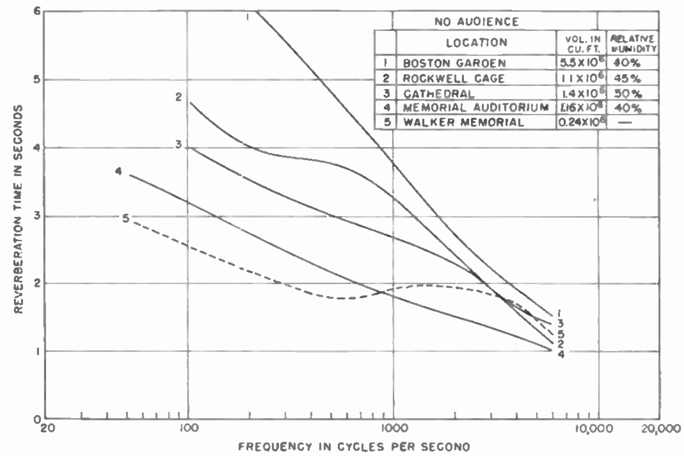


Fig. 6—Measured reverberation times for empty auditoriums. The volumes and relative humidities are shown in the table.

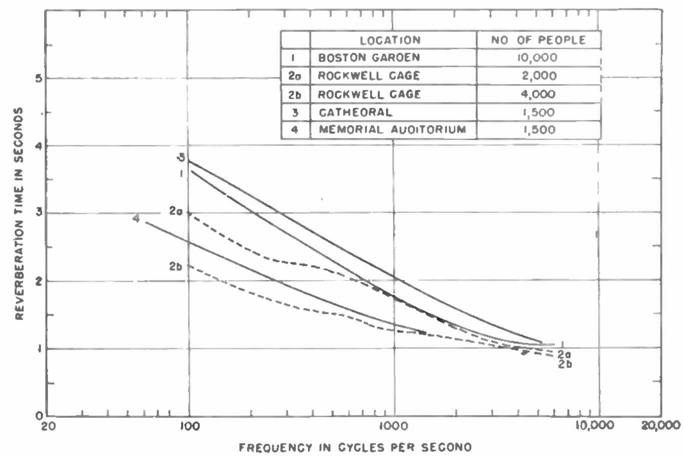


Fig. 7—Reverberation times calculated (except for the cathedral where the times were measured) assuming the occupancy given in the table.

ing. Certain members of the audience were asked to write down the terminal word of every sentence uttered by the speaker. Simultaneously, a recording of the speaker's voice was made with a disk or tape recorder connected to the microphone circuit. The word lists so obtained were graded with the aid of a master sheet prepared from the recording.

In the cathedral, word-articulation tests were performed on two occasions with no audience but with different observers. One group consisted of two talkers and five listeners, and the other of one talker and twenty listeners. The smaller group was composed of male college graduates. The larger group was composed of men and women whose educational backgrounds ranged from grammar school to college. For the smaller group, four locations were selected for the listeners: two on the main floor (see Fig. 5(a)) and two in the two galleries at the rear of the auditorium. Each listener sat in four different locations during the test, and was at each location once for each of the two talkers. In all, eight lists of fifty words each were read. For the larger crew, eight locations in the cathedral were selected, one for each of the eight word lists. The results of the tests, grouped so as to apply to four separated locations in the cathedral, are shown in Fig. 8. The average percentage of words correctly recorded by the small crew was 63, while that of the large crew was 41. However, the differences between the average score and the score at each of the four positions was very nearly the same for both crews. This illustrates an important consideration in articulation testing; namely, only comparisons among scores obtained by the same crew on the same day should be made because the composition of the crew, experience, and other factors, greatly influence the absolute value of the score.

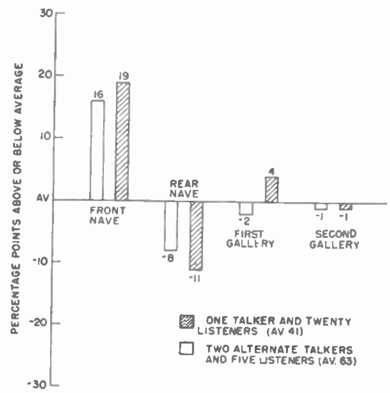


Fig. 8—Articulation scores obtained in the cathedral for each of two groups of observers plotted relative to the average score for the group.

In the front part of the nave, where the reverberation characteristics of the room are of least importance because of the high ratio of direct-to-reflected sound energy, the scores were 79 per cent for the small crew. Undoubtedly this score would have been higher if the sound system had had a greater frequency range.

From the frequency-response curve for the cathedral shown in Fig. 2, it can be seen that the response was uniform within ±3 db over only 65 per cent of the frequency range which is essential to good speech intelligibility. A restricted frequency range lowers the intelligibility of speech, and the score of 79 per cent observed is reasonable for the range used, as we shall see later. If the frequency response had been increased, the word intelligibility for both crews would have improved substantially, assuming low background noise level. A method for calculating the improvement is given in Section VI.

The lower scores at the rear of the nave and in the galleries are due, in part, to the reduced intensity of the sound (see Figs. 4 and 5) and, in part, to the reverberant conditions. The average sound levels were nearly the same in the galleries as at the rear of the nave. The poor acoustic conditions (reverberation and echoes) at the rear of the nave greatly lowered the scores. In this cathedral, the reverberation time was nearly independent of the number of people in the audience because of the high absorption of the seat cushions and the lower relative humidity when the cathedral was empty.

The results obtained from the terminal-word test described above are shown in the middle column of Table I. The word-articulation scores for comparable positions with the large group are also shown. The scores from the terminal-word test were expected to be higher than those from the word-articulation test because the listeners were aided by sentence context. It was anticipated that the scores would approach what is commonly called the "sentence intelligibility."

TABLE I

Comparison of Terminal-Word Scores Determined with Full Audience and Word-Articulation Scores Determined without Audience—Cathedral

Location	Terminal-Word Score	Word-Articulation Score
Front of nave	80	60
Center of nave	80	55
Rear of nave	69	35
Galleries	66	39

The data in Table I are plotted as crosses in Fig. 9 along with a curve of sentence intelligibility versus word articulation as determined by Egan from extensive experiments conducted at Harvard University.[5] These results show that the terminal-word test is approximately a measure of sentence intelligibility.

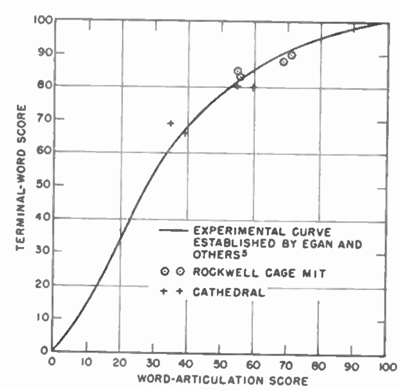


Fig. 9—Relation between terminal-word scores and word-articulation scores. An experimental curve of sentence intelligibility versus word-articulation scores is shown for comparison.

To determine the effect of restricting the bass response of a speech-amplifying system, tests were performed in the Rockwell Cage at M.I.T. using the loudspeaker arrangement of Fig. 1, and a cardioid microphone. From casual observation we decided that with the room empty, a flat frequency response down to very low frequencies produced undesirable reverbera-

[5] J. P. Egan; see L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., p. 628; 1949.

tion effects. Our first test was performed with only the moderate amount of bass shown by curve B (Fig. 10).

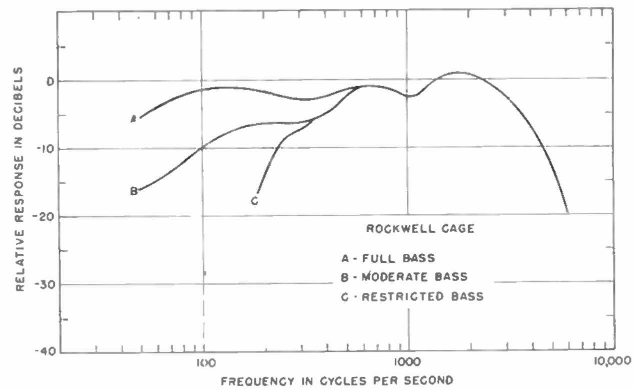


Fig. 10—Frequency-response characteristics of the Rockwell Cage sound-reinforcing system used during subjective tests.

Word articulation was measured with twenty-five listeners present for the two-system conditions shown by response curves B and C, Fig. 10. There were two talkers, each reading four hundred words, that is, two hundred words for each system. The twenty-five listeners were divided into four groups of about six persons each for the reading of a list of one hundred words. Then they were divided differently by a random-selection process to form another set of four groups for the reading of another list of one hundred words. By the time eight of the one hundred-word lists had been recorded, each person had been to each of four positions twice. For each of the one hundred-word lists, the two conditions of the system and the two voices had been employed. Results are shown in Table II.

TABLE II

Comparison of Sound System with Moderate and Restricted Bass (See Fig. 11)—Rockwell Cage

System	Condition	Word-Articulation Score
C	Restricted bass	62.8 per cent
B	Moderate bass	61.5 per cent

We see that the system performance with restricted bass is slightly superior to that with moderate bass. The difference is so small that either system is equally satisfactory. The low average scores of about 62 per cent were not surprising because of external traffic noise, the somewhat restricted high-frequency response of the system, the reverberant room conditions, and the variation of level in the room (see Figs. 4 and 5). There were only negligible differences among scores obtained with the two talkers.

In the second test, terminal-word articulation was measured with a full audience present, for the two system conditions shown by response curves A and C of Fig. 10. Thirty-nine observers recorded the last word of each sentence for 8 minutes without bass, then for 16 minutes with bass, and finally for 8 minutes without bass. Results are shown in Table III.

TABLE III

Comparison of Sound System with Full and Restricted Bass (Fig. 11)—Rockwell Cage

System	Condition	Terminal-Word Articulation Score
C	Restricted bass	86.6 per cent
A	Full bass	85.7 per cent

Our conclusion is the same as before, namely, that a slight advantage results from restricting the bass, but the difference between the performances of the two systems is negligibly small.

A further comparison of terminal-word intelligibility with sentence intelligibility is possible from this experiment. Data from four separated positions in the Rockwell Cage are shown in Table IV and are plotted as circles in Fig. 9. As was the case for the cathedral tests, the points fall near the Egan curve.

TABLE IV

Comparison of Terminal-Word Scores Determined with Full Audience, and Word-Articulation Scores Determined with Twenty-Five Listeners—Rockwell Cage

Location	Terminal-Word Score	Word-Articulation Score
Near center of room	90	71
Center edge	88	69
Rear corner	83	56
Center rear	85	55

Word-intelligibility tests were also performed in the industrial building. The results will be discussed in Section VI.

V. Preference, Naturalness, and Intelligibility Ratings

Manufacturers of sound equipment frequently state that it is true that only the frequencies from about 300 to 5,000 cps are needed for good speech intelligibility, but if naturalness is desired, the lower frequencies must also be reproduced. To test the validity of this statement, members of the audience were asked to make subjective judgments of the naturalness and intelligibility of speech in each auditorium studied. In the Boston Garden, where Winston Churchill and Harold E. Stassen spoke before near-capacity audiences, only enthusiastic appraisals of the system[6] were received. Favorable comments are still being received on the systems in the cathedral and in the memorial auditorium, despite the restricted bass, after two years of continuous use.

In general, qualitative comments are not wholly significant because the audience does not have specific

[6] This system was installed for the M.I.T. Mid-Century Convocation, and differs greatly from the one ordinarily used in the Boston Garden.

comparisons to make. To obtain more quantitative results, questionnaires were issued on each of three occasions in the Rockwell Cage at M.I.T. First, when three prominent Americans spoke; second, when a noted evangelist spoke; and third, when the word-articulation tests previously mentioned were being performed.

On the first occasion, with the bass condition of curve C of Fig. 10, twenty-seven observers judged the intelligibility, naturalness, and loudness on a rating scale of *Excellent*, *Good*, *Fair*, and *Poor*. The averages of their ratings were: Intelligibility—*Good to Excellent;* naturalness—*Good;* and loudness—*Good*. When the evangelist spoke, the system was operated part of the time with restricted bass, as shown by curve C of Fig. 10, and part of the time with full bass, as shown by curve A of Fig. 10. The three rating scales on which the 39 observers indicated their opinions, are reproduced in Fig. 11. The average results are indicated on each scale. These results show no discernible preference on the part of the listeners for one system over the other. On the third occasion, thirty observers, were asked to indicate

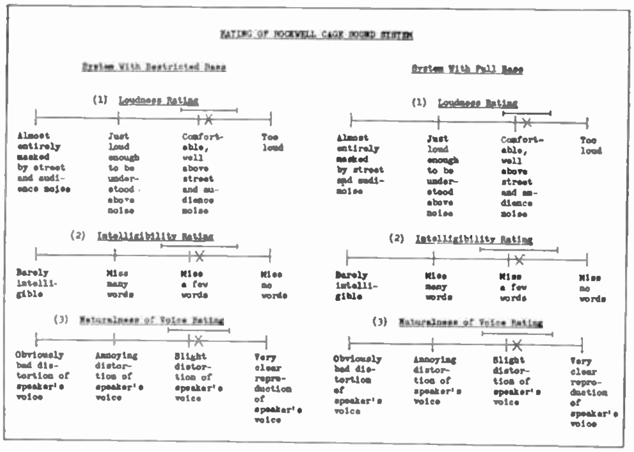


Fig. 11—Rating sheet used in Rockwell Cage test when a noted evangelist spoke. The crosses on each of the rating scales show the average of the ratings for 39 observers. The bars above the crosses are equal in length to twice the standard deviations.

a preference when the bass was switched in and out. Of these, thirty preferred the system with moderate bass (curve B of Fig. 10), and eighteen preferred it with no bass. On this occasion, the full bass condition was not compared.

VI. Calculation of Speech Intelligibility

A simple graphical method for calculating speech intelligibility was used to predict the articulation scores measured in the industrial building. By this method the articulation index area of Fig. 12 is combined with the over-all response curve of the sound system to produce the shaded area of Fig. 13,[1] bounded by the curves "amplified speech peaks" and "amplified speech minimums." This area was located vertically on the coordinate system by measuring the average sound pressure level in each of eight octave bands at several positions, while the talker was reading a word list. These eight octave band levels were then converted to spectrum levels (sound pressure level in a band 1 cps wide). Finally, 10 db were added to these spectrum levels to give a plot of the speech peaks as a function of frequency, because the average, as read on a vu meter, lies about 10 db below the speech peak level.[7] This

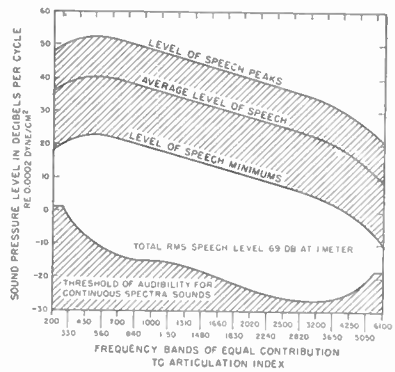


Fig. 12—Graph relating speech levels to frequency. The upper shaded area between the curves "level of speech peaks" and "level of speech minimums" is, by definition, the area corresponding to 100 per cent "articulation index." The speech levels were measured with a microphone placed 1 meter in front of a man speaking in a raised voice in an anechoic chamber.

plot, having only eight points, was used to locate vertically the more accurate plot of Fig. 13 by making it and the upper edge of the shaded region coincide as closely as possible.

The background noise in the room during the tests was also measured by the eight-band analyzer. After converting these sound levels to spectrum levels, they were plotted on the same graph (Fig. 13). The proportion of the shaded area lying above the background noise (or the threshold of hearing, whichever is greater) is known as the articulation index. For the example shown in Fig. 13, the articulation index is 0.51. This calculation was made for one particular position in the building, for one particular gain-control setting. The sound level varied by about 10 db over the floor area. Calculations were performed for three other positions, and the results are shown in Table V.

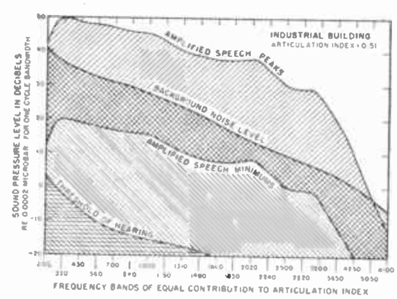


Fig. 13—Graph for determining an articulation index in the industrial building. The fraction of the articulation index area that lies above the "background noise level" is the articulation index, here equal to 0.51.

[7] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., pp. 702–703; 1949.

TABLE V

COMPARISON OF CALCULATED ARTICULATION INDEXES WITH MEASURED ARTICULATION SCORES—INDUSTRIAL BUILDING

Position	Calculated Articulation Index	Word-Articulation Score
A	83	90
B	71	76
C	62	65
D	51	59

The articulation index is approximately a measure of the percentage of the syllables correctly heard by a listener.[1] The word-articulation score is always higher; the exact relation depends upon the education of the listeners and their previous practice. The relations of Table V are typical of results obtained for college graduates without previous experience at this kind of test. A relation for college graduates with months of daily experience at this kind of test is shown in curve (a), Fig. 11 of the literature 1. Lower scores will be obtained for personnel whose reading habits are less well cultivated.

Bolt and MacDonald[4] have shown that the reverberation in a room, caused by the speech itself, is similar to background noise in reducing speech intelligibility. For each of the twenty bands of equal importance to speech intelligibility shown on the abscissas of Figs. 12 and 13, the number of decibels that the reverberant "speech" lies below the peaks of the direct speech is related to reverberation time, as shown in Table VI. For a multicellular horn with a mouth opening of 22 by 30 inches, the directivity index is approximately 8 db at 200 cps, 11 db at 300 cps, and 13 db for all frequencies above 500 cps. The reverberant "speech" level should be plotted as though it were background noise, and the higher of the two curves (room background noise or reverberant "speech" level) used as the background noise in determining the articulation index. It is seen that if the directivity index is high, speech intelligibility is affected little by reverberation. This agrees with the subjective impression of listeners. However, in a reverberant room, the background noise level will be higher than that in a nonreverberant room. Reduction of the reverberation time results in both a reduction of background noise and of the reverberant "speech," both of which increase the articulation index.

TABLE VI

RELATION BETWEEN REVERBERATION TIME IN AN AUDITORIUM AND THE DIFFERENCE IN DECIBELS BETWEEN THE DIRECT SPEECH PEAK LEVELS AND THE REVERBERANT "SPEECH" LEVELS. THE PARAMETER IS DIRECTIVITY INDEX OF THE LOUDSPEAKER[8]

Reverberation Time Seconds	Direct Speech Peak Level Minus Reverberant "Speech" Level—Decibels				
	Loudspeaker Directivity Index—Decibels				
	0	5	10	15	20
0.5	25	30	35	40	45
1	20	25	30	35	40
2	15	20	25	30	35
3	13	18	23	28	33
4	11	16	21	26	31

[8] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., pp. 668–684; 1949.

VII. OTHER OBSERVATIONS

An important reason for the satisfactory performance of these sound systems, even in rooms where the reverberation times at 500 cps are between two and four seconds, is the location of the loudspeakers. It was found that the best results were obtained when the loudspeaker was between 20 and 30 feet above the head of the talker, and the principal axis of the horn was pointed downward so as to keep as much of the sound off the side walls and ceiling as possible. Also, a more uniform distribution of sound over the seating area is obtained when the loudspeaker is high.

Experiments in Morss Hall at M.I.T., and at various other installations in Boston, showed that a single loudspeaker over the podium is superior to one or more loudspeakers on either side of the stage. With loudspeakers on opposite sides of the stage, distracting aural effects occur along and adjacent to the center line of the hall. With one loudspeaker, even though mounted high above the podium, the amplified speech appears to come from the talker's mouth because of the poor vertical directivity of the human ear.

The fact that so little difference in preference was found between systems with restricted bass and those with full bass, resulted partly from the satisfactory high-frequency response of all systems tested. If the high-frequency response is restricted, and if full bass is present, the quality of the reproduced speech is poor, especially in highly reverberant spaces. Situations such as this are common in the United States, for example, in large railway stations.

CONCLUSIONS

(1) The results reported herein are consistent with the prediction from articulation theory, that the over-all frequency response of a speech-reinforcing system should extend at least from 400 to 4,000 cps, and lie within about ± 3 db of the average value of the response through that region.

(2) The distribution of sound in an auditorium should be sufficiently uniform so that the requirements of the first conclusion are satisfied at all seat locations.

(3) Our results indicate that, contrary to general belief, the naturalness of the reinforced speech is not affected if frequencies below approximately 400 cps are attenuated. This conclusion is based on experiments where the response above 400 cps meets the requirements under the first conclusion above.

(4) The results show that the reinforcement of the low voice frequencies in highly reverberant auditoriums does not reduce the intelligibility of the amplified speech. Little, if any, harm results from the presence of

the full bass range of frequencies, provided the response at the high end is good out to at least 4,000 cps.

(5) The preceding two conclusions might no longer be valid if the loudspeaker location were to differ from those used in these experiments and if the loudspeaker had poor directivity. The locations used here were selected after observations of many existing systems, which indicated that the loudspeakers should be hung high above the podium (20 to 30 feet) and be directed downward, so that the sound is absorbed by the people and upholstered seats, and so that the sidewall reflections are minimized. A single loudspeaker system so located yields more natural reproduction than a split loudspeaker system because (a) the sound appears to come from the talker and because (b) no regions of the seating exist at which the sounds from two loudspeakers overlap, with resulting disturbing psychological effects.

(6) The terminal-word test described herein for measuring the sentence intelligibility of a sound system under normal operating conditions yields results which are consistent with results from conventional word-articulation tests, which can rarely be carried out under normal operating conditions.

(7) Previously reported theories for the calculation of speech intelligibility predict word-articulation scores within acceptable tolerances.

Radiation Resistance of a Two-Wire Line*

JAMES E. STORER† AND RONOLD KING‡

Summary—**A general formula for the radiation resistance of a two-wire line is derived by means of a Poynting vector integration over a large sphere. The result is shown to be in agreement with that computed by other techniques. Formulas for the useful special cases of a lossless system and a nonresonant line are presented.**

I. Introduction

THE PROBLEM of the power radiated from a two-wire line is one of considerable interest. Although numerous papers[1-5] have appeared on the subject, none, to the authors' knowledge, has contained a general formula for the radiation resistance. It is the purpose of this paper to derive such a formula and compare it with previous work.

There are three techniques available for calculating the radiation resistance. One is the integration of the normal component of the Poynting vector over a large sphere surrounding the line. A second, usually called the emf method, is equivalent to the integration of the normal component of the Poynting vector over the surface of the wires. The third and last possibility is to obtain the radiation resistance directly from the integral equation for the distribution of current along the line. It will be shown that these methods all yield consistent results.

II. The Distribution of Current Along the Line

Before it is possible to obtain a formula for the radiation resistance, it is necessary to know the distribution of current along the line. This has been done in standard texts on the subject.[6] For purposes of reference, however, it is worth while to write down the integral equation for the current and its solution. Hallén's equation[7] for such a two-wire-line circuit, as indicated in Fig. 1, is

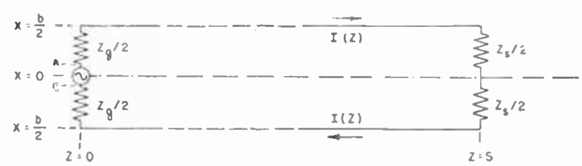


Fig. 1—Two-wire line circuit.

$$Z\delta(s) = z^i I(s) + \frac{j\zeta_0}{4\pi}\oint_{s'} I(s')\left[\beta_0^2 \hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right]\frac{e^{j\beta_0 R_{ss'}}}{\beta_0 R_{ss'}}\,ds', \quad (1)$$

where

z^i = the internal impedance per unit length

Z = the input impedance across the generator terminals AB

$\delta(s)$ = the Dirac delta function defined by $\int_{-x}^{x}\delta(x)dx = 1,\ x>0$

$I(s)$ = the current in the line at a point s^{-x}

ζ_0 = the impedance of free space ($=120\pi$ ohms)

$\beta_0 = 2\pi/\lambda$ where λ is the free-space wavelength

* Decimal classification: R117.1. Original manuscript received by the Institute, October 30, 1950; revised manuscript received, February 21, 1951.

The research reported in this document was made possible through support extended Cruft Laboratory, Harvard University, jointly by the Navy Department (Office of Naval Research), the Signal Corps of the U. S. Army, and the U. S. Air Force, under ONR Contract N5-ori-76, T. O. 1.

† Electronic Research Laboratory, Harvard University, Cambridge, Mass.

‡ Cruft Laboratory, Harvard University, Cambridge, Mass.

[1] C. Manneback, "Radiation from transmission lines," *Trans. AIEE*, pp. 289–300; February, 1923.

[2] A. A. Pistolkers, "The radiation resistance of beam antennas," Proc. I.R.E., vol. 17, pp. 562–579; March, 1929.

[3] S. A. Schelkunoff, "A general radiation formula," Proc. I.R.E., vol. 27, pp. 660–666; October, 1939.

[4] C. W. Harrison, "On the pickup of balanced four-wire line," Proc. I.R.E., vol. 30, pp. 517–518; November, 1942.

[5] E. J. Sterba and C. B. Feldman, "Transmission lines," Proc. I.R.E., vol. 20, pp. 1163–1202; July, 1932.

[6] R. W. P. King, "Electromagnetic Engineering," vol. 1, chap. VI, McGraw-Hill Book Co., New York, N. Y.; 1945.

[7] J. Aharoni, "Antennae," chap. II, Oxford University Press, Oxford, England; 1946.

$\hat{s}$ = a unit vector tangent to the wires at a point s, and in the direction of the current

$R_{ss}' = \{[\text{distance between } s \text{ and } s']^2 + a^2\}^{1/2}$

a = the radius of the wires, satisfying the inequality $\beta_0^2 a^2 \ll \beta_0^2 b^2 \ll 1$

$\oint_s' ds'$ indicates an integration around the contour of the circuit.

It will be assumed, although not explicitly written, that all electromagnetic quantities have a time dependence of the form $e^{j\omega t}$.

Subject to the five conditions

1) $\beta_0^2 b^2 \ll 1$,
2) $\beta_0^2 a^2 \ll \beta_0^2 b^2$,
3) $\alpha^2/\beta_0^2 \ll 1$, where α is the attenuation constant of the line,
4) The line is balanced, i.e., the currents in the two wires at the same position on the line are equal but opposite in direction, and
5) The current across the terminations is a constant, i.e., the terminations occupy regions of space that are small in comparison with the wavelength,

the solution of (1) can be written as a power series expansion in the parameter $\beta_0^2 b^2$ as follows:

$$I(z) = I_0(z) + \beta_0^2 b^2 I_1(z) + \beta_0^4 b^4 I_2(z) + \cdots. \quad (2a)$$

King[8] has demonstrated that $I_0(z)$ can be written in the form

$$I_0(z) = I_0(0) \frac{\cosh [(\alpha + j\beta)(s - z) + \rho + j\Phi']}{\cosh [(\alpha + j\beta)s + \rho + j\Phi']}. \quad (2b)$$

The quantities ρ and ϕ' are defined in terms of the characteristic impedance Z_c, the generator impedance Z_g, and the terminal impedance Z_s as follows:

$$\rho + j\Phi' = \tanh^{-1} Z_s/Z_c.$$

The input impedance Z is found to be

$$Z \doteq Z_g + Z_c \tanh [(\alpha + j\beta)s + \rho + j\Phi'] + \text{terms of order } \beta_0^2 b^2. \quad (3)$$

The $\beta_0^2 b^2$-term in the current-distribution expansion is impossible to evaluate unless we assume a specific wire configuration for the terminations. Even in the simplest case of wire bridge terminations for the line, this $\beta_0^2 b^2$ term appears to be so complicated that it would have to be evaluated numerically.

III. Calculation of the Radiation Resistance by the Integration of the Poynting Vector Over a Large Sphere

The integration of the normal component of the Poynting vector over the surface of a large sphere surrounding an antenna system is the usual technique to determine the power radiated from such a system.[9] Using the standard formula for the Poynting vector, and taking the limit as the spherical surface recedes to infinity, the following equation is obtained for the radiation resistance R:

Average power radiated

$$= \frac{1}{2} R I(0) I^*(0)$$

$$= \frac{1}{2} \frac{\zeta_0}{4^2\pi^2} \beta_0^2 \int_0^{\pi} \int_0^{2\pi} \sin\theta d\theta d\phi [\vec{\eta}\cdot\vec{\eta}^* - (\hat{r}\cdot\vec{\eta})(\hat{r}\cdot\vec{\eta}^*)] \quad (4)$$

$$\vec{\eta} = \int_v \vec{J}(x', y', z') e^{-j\beta_0(\hat{r}\cdot\vec{r}')} dx'dy'dz'$$

where

θ, ϕ = spherical co-ordinates describing points on the spherical surface

$\hat{r}$ = a unit vector directed outward from the surface

$*$ indicates a complex conjugate

$\vec{r} = x'\hat{x} + y'\hat{y} + z'\hat{z}$

$J(x', y', z')$ = the current density at a point x', y', z'.

If the previously given current distribution (2) is inserted, the integral for $\vec{\eta}$ may be evaluated. It consists of three parts, a contribution from the line, and one from each termination. If only the first-order terms $\beta_0 b$ are kept, the integral for $\vec{\eta}$ becomes independent of the physical structure of the terminations (wire bridge, coil, and the like). The contributions from the terminations are equivalent to that of a single current filament across the terminations. Inserting this expression for $\vec{\eta}$ into (4), the radiation resistance may be evaluated. The details of this process are left to the Appendix. The final result is

$$R = \frac{\zeta_0}{4\pi} \beta_0^2 b^2 \frac{\cosh (\alpha s + 2\rho)}{|\cosh [(\alpha + j\beta)s + \rho + j\Phi')]|^2} \cdot \left[\cosh \alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right] + \text{terms of order } \beta_0^3 b^3. \quad (5)$$

It is apparent from the technique used to obtain (5) that the $\beta_0^3 b^3$ terms would be difficult, if not impossible, to evaluate in a general way. It is worth while to see how formula (5) simplifies for certain special cases.

Case I: Line Approaching the Nonresonant State, i.e., $\rho > 3$.

$$R = \frac{\zeta_0}{2\pi} \beta_0^2 b^2 \left[\cosh \alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right] e^{-\alpha s} \quad (\pm 0.5 \text{ per cent}).$$

Then for a long line, $B_0 s > 50$, and good conductivity (α small)

$$R = \frac{\zeta_0}{2\pi} \beta_0^2 b^2 = 60 \beta_0^2 b^2 \text{ ohms} \quad (\pm 0.5 \text{ per cent}). \quad (6a)$$

[8] R. W. P. King, "Transmission-line theory and application," *Jour. Appl. Phys.*, vol. 14, pp. 577–600; November, 1943.

[9] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y.; 1941.

Case II: Small Losses in Termination and Line, i.e., $(\alpha s+2\rho)<0.1$.

There

$$R = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2\frac{1}{\cos^2(\beta_0 s+\Phi')}\left[1-\frac{\sin 2\beta_0 s}{2\beta_0 s}\right]$$

$(\pm 0.5$ per cent).

This resistance can be referred to the maximum current on the line, i.e., power radiated $=\frac{1}{2}R_{\max}|I_{\max}|^2$; then

$$R_{\max} = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2\left[1-\frac{\sin 2\beta_0 s}{2\beta_0 s}\right] \quad \text{(See Fig. 2).}$$

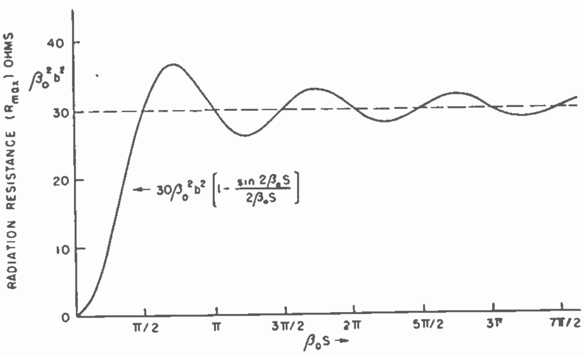


Fig. 2—Radiation resistance of a lossless line.

It is to be noted that this resistance is independent of the terminations and reasonably constant. In the case of wire bridge terminations ($\phi'\cong 0$) and a resonant line ($\beta_0 s=n\pi$), these formulas reduce to

$$R = R_{\max} = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2 = 30\beta_0{}^2b^2 \text{ ohms.} \tag{6b}$$

The formulas for these last special cases are in agreement with those of previous authors. The difference between the above nonresonant-line expression and that of Sterba and Feldman[5] is attributable to the fact that radiation from the terminations has not been included in their work.

Integration of the normal component of the Poynting vector over a large sphere has the advantage over other methods of obtaining the power radiated from a two-wire line in that the radiation resistance may be evaluated without reference to the physical construction of the terminations. It is interesting to see, however, that other techniques lead to the same results in a more restricted fashion.

IV. Evaluation of the Radiation Resistance by Other Techniques

In the Poynting vector method of evaluating the radiation resistance, it is necessary to know the distribution of current. This being the case, it is apparent that the fundamental method of getting at the radiation resistance is to obtain it directly from the integral equation (1) defining the current distribution. This was attempted by King.[10]

The method used resembles the procedure familiar in the approximate solution of the integral equation for the current in an antenna. The principle involved is to substitute for the current under the sign of integration a zero$^{\text{th}}$-order approximation of the current. In the case of the transmission line, this zero$^{\text{th}}$-order current is the well-known solution of the conventional transmission-line equations which neglect radiation. For a low-loss resonant line, the distribution is very nearly sinusoidal. In the case of the antenna, the zero$^{\text{th}}$-order current also is sinusoidal, but this is a very much poorer approximation of the actual current than is true for the transmission line. It might be supposed, therefore, that the evaluation of the radiation resistance of the transmission line should be correspondingly much more accurate than the corresponding calculation of the impedance of the antenna. But this is not necessarily true. If the radiation resistance of a resonant line were of the same order of magnitude as the radiation resistance of a resonant antenna, the more accurate zero$^{\text{th}}$-order current available for the transmission line certainly should lead to a correspondingly more accurate radiation resistance. However, the radiation resistance of a resonant line is of the order of magnitude of 30 $\beta_0{}^2b^2$ ohms, and hence is a very small fraction of an ohm for $\beta_0 b\ll 1$. For an antenna, the radiation resistance at resonance is of the order of magnitude of 70 ohms. Clearly, in order to determine accurately the very small value of radiation resistance for a transmission line, a very much more accurate zero$^{\text{th}}$-order distribution of current is necessary for the line than for the antenna. Actually, a distribution is required that is accurate to terms in $\beta_0{}^2b^2$. This can not be true of the conventional transmission-line distribution, since it is determined by ignoring radiation resistance which is known to be as large as 30 $\beta_0{}^2b^2$. Nevertheless, in the case of a resonant line an even number of half-wavelengths long, King arrived at essentially the correct result. But for other lengths, in particular, for a line an odd number of half-wavelengths long, an incorrect result was obtained as should, indeed, be expected.

This sensitivity to current distribution can be circumvented by the use of a variational principle. Then the eigen value so obtained, in this case Z, is correct to one order higher than the distribution function used to compute it. Hence, when the approximate current-distribution function (2b) is used, Z is correct to one higher order of $B_0{}^2b^2$ than (2b). Thus, Z so computed is correct to $B_0{}^2b^2$.

[10] R. W. P. King, "Electromagnetic Theory," Chap. VI, Sec. 25, pp. 483–485, eq. 1–12, vol. I, McGraw-Hill Book Co., New York, N. Y.; 1945.

The original equation (1), can be written as follows:

$$Z\delta(s) = \frac{1}{I(0)}\oint_{s'} I(s')G(s, s')ds', \tag{7}$$

where

$$G(s, s') = z^i\delta(s - s') + \frac{j\zeta_0}{4\pi}\left[\beta_0^2\hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right]\frac{e^{-j\beta_0 R_{ss'}}}{\beta_0 R_{ss'}} = G(s', s).$$

The variational expression for Z is

$$Z = \frac{1}{I^2(0)}\oint_s\oint_{s'} I(s)I(s')G(s, s')ds'ds.$$

This can be proved readily by performing the variation, i.e.,

$$\delta Z = \frac{1}{I^2(0)}\oint_s\oint_{s'}\left[I(s)\delta I(s') + I(s')\delta I(s)\right]G(s, s')ds'ds - 2\frac{1}{I^3(0)}\delta I(0)\oint_s\oint_{s'} I(s)I(s')G(s, s')dsds'.$$

Using the original equation (7), this becomes

$$\delta Z = \frac{1}{I(0)}\oint_{s'}\delta I(s')Z\delta(s')ds' + \frac{1}{I(0)}\oint_s \delta I(s)Z\delta(s)ds - 2\frac{\delta I(0)}{I(0)}Z = \frac{\delta I(0)}{I(0)}Z + \frac{\delta I(0)}{I(0)}Z - 2\frac{\delta I(0)}{I(0)}Z = 0.$$

A technique is accordingly made available for determining Z correct to the order $\beta_0^2b^2$. Assuming this, the next problem is to separate out the radiation resistance from the input impedance Z. When there are no losses in the line or terminations, this is simple, because then the real part of the input impedance must be the radiation resistance, since there are no other power losses in the system. In the case of no losses, $z^i = 0$ and $I(s)$ is real. Hence,

$$R = \text{real part}\frac{1}{I^2(0)}\oint_s\oint_{s'} I(s)I(s')G(s, s')dsds' = \frac{\zeta_0}{4\pi}\oint_s\oint_{s'} I(s)I(s')\left[\beta_0^2\hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right]\frac{\sin\beta_0 R_{ss'}}{\beta_0 R_{ss'}}dsds'.$$

After a rather tedious integration, it can be shown[11] that this integral yields a result identical to that obtained previously (Case II). It must be emphasized that this method yields the correct result even if the lossless assumption is not made.

[11] J. E. Storer, "Radiation Resistance of a Two Wire Line," Technical Report No. 69, Cruft Laboratory, Harvard University; March, 1949.

It is also possible to obtain the radiation resistance by integrating the normal component of the Poynting vector over the surface of the wires. This procedure is often referred to as the emf method. It can be shown[11] that this technique also yields results in agreement with those obtained in this paper.

V. Ratio of Ohmic Losses to Radiation Losses

Antenna systems that consist essentially of a wire close to and parallel to a metal surface, are of practical importance. If the metal surface is sufficiently large compared with the distance $b/2$ of the parallel antenna from it, the conducting surface may be replaced by an image of the wire. The wire, plus its image, is equivalent to a two-wire line; ohmic losses and radiation are one-half those of a two-wire line. It is apparent that the efficiency of such a system is determined by the ratio of ohmic losses to the radiation resistance. Restricting the work to the case of lossless terminations and a near lossless line, the radiation resistance of the two-wire line is

$$R = \frac{\zeta_0}{4\pi}\beta_0^2b^2\frac{1 - \dfrac{\sin 2\beta_0 s}{2\beta_0 s}}{\cos^2(\beta_0 s + \Phi')};$$

the factor $1 - \sin 2\beta_0 s/2\beta_0 s$ is approximately equal to 1, provided $\beta_0 s$ has any appreciable value (see Fig. 2). As only the order of magnitude is of interest in this case, this factor can be set equal to 1. Therefore

$$R \cong \frac{\dfrac{\zeta_0}{4\pi}\beta_0^2b^2}{\cos^2(\beta_0 s + \Phi')}.$$

As it has been assumed that there are no losses in the terminations, the total ohmic loss for the two-wire line is given by

$$\text{ohmic loss} = \frac{1}{2}R_{\text{ohmic}}|I(0)|^2 = \frac{1}{2}r_i\int_0^s |I(z)|^2dz + \frac{1}{2}r_i\int_s^0 |I(z)|^2dz,$$

where r_i is the resistance per unit length of the line wires. Therefore

$$R_{\text{ohmic}} = 2r_i\int_0^s\frac{|I(z)|^2}{|I(0)|^2}dz = 2r_i\int_0^s\frac{|\cosh[(\alpha + j\beta_0)(s - z) + j\Phi']|^2}{|\cosh[(\alpha + j\beta_0)s + j\Phi']|^2}dz$$

$$\cong 2r_i\int_0^s\frac{\cos^2[\beta_0(s - z) + \Phi']}{\cos^2[\beta_0 s + \Phi']}dz = r_i s\,\frac{1 + \dfrac{\cos(\beta_0 s + \Phi')\sin\beta_0 s}{\beta_0 s}}{\cos^2(\beta_0 s + \Phi')}.$$

For appreciable $\beta_0 s$, the factor

$$1 + \frac{\cos(\beta_0 s + \Phi') \sin \beta_0 s}{\beta_0 s}$$

is approximately equal to 1. Hence,

$$R_{\text{ohmic}}/R \cong \frac{s r_i}{\zeta_0/4\pi} \beta_0^2 b^2.$$

It is interesting to evaluate this for a particular case. Assuming the frequency is 300 mc and the wire is of copper and 1 mm thick, $r_i = 1.4$ ohms/meter, letting $s = 1$ meter. This becomes

$$R_{\text{ohmic}}/R = \frac{1 \times 1.4}{30} \cong \frac{1}{20}.$$

VI. Conclusions

It has been shown that various methods yield consistent results for the radiation resistance. A general formula was obtained for the radiation resistance, the two most useful special cases of this formula being

$$R = \zeta_0/2\pi(\beta_0^2 b^2)$$

for a long, near nonresonant line.

$$R_{\max} = \zeta_0/4\pi(\beta_0^2 b^2)\left[1 - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right]$$

for a near lossless system.

APPENDIX

Evaluation of the Radiation Resistance by the Poynting Vector Method

Using the previously given current distribution (2) and remembering that the radius of the wires satisfies the inequality $\beta_0^2 a^2 <<< 1$, $\vec{\eta}$ can be expressed as follows:

$$\vec{\eta} = \hat{z} \int_0^s I(z) e^{-j\beta_0 \cos\theta z'} \left[e^{-j(\beta_0 b \sin\theta \cos\phi)/2} - e^{j(\beta_0 b \sin\theta\cos\phi)/2}\right] dz'$$

$$+ \underset{\text{Generator Termination}}{I(0) \int \hat{s}' e^{-j\beta_0 \hat{r}\cdot\vec{r}'} ds'} + \underset{\text{Terminal}}{I(s) \int \hat{s}' e^{-j\beta_0 \hat{r}\cdot\vec{r}'} ds'}.$$

If terms only of the order $\beta_0 b$ are kept in the expression for $\vec{\eta}$, all three integrals simplify; the two involving the terminations become independent of the physical structure of the terminations (wire bridge, coil, etc.). Hence the contributions from the terminations to the radiation become equivalent to that of a single current filament across the terminations. Thus,

$$\vec{\eta} = \hat{z}[-j\beta_0 b \sin\theta\cos\phi] \int_0^s e^{-j\beta_0 z \cos\theta} I(z') dz'$$

$$+ I(0)\hat{x} \int_{-b/2}^{b/2} dx' + I(s)\hat{x} e^{-j\beta_0 s \cos\theta} \int_{-b/2}^{b/2} dx'$$

$$+ \text{terms of order } \beta_0^2 b^2.$$

It is now seen that terms of order $\beta_0^2 b^2$ in the current-distribution function can be dropped, since $\vec{\eta}$ is calculated correctly only to order $\beta_0 b$. The integrations necessary for $\vec{\eta}$ can now be readily performed, as the integrand is simply a product of an exponential and a trigonometric function. Hence

$$\vec{\eta} = bI(0)\left\{J_1(\cos\theta)\hat{x} + \frac{\cos\phi}{\sin\theta}\left[\cos\theta J_1(\cos\theta) - J_2(\cos\theta)\right]\hat{z}\right\},$$

where

$$J_1(\cos\theta) = 1 - \frac{\cosh(\rho + j\Phi')}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} e^{-j\beta_0 s \cos\theta}$$

$$J_2(\cos\theta) = \frac{\sinh[(\alpha + j\beta)s + \rho + j\Phi']}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} - \frac{\sinh(\rho + j\Phi')}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} e^{-j\beta_0 s \cos\theta}.$$

Now, $\hat{r}\cdot\hat{x} = \sin\theta\cos\phi$ and $\hat{r}\cdot\hat{z} = \cos\theta$. After substituting $\vec{\eta}$ into the formula for the radiation resistance, this becomes

$$R = \zeta_0 \frac{\beta_0^2 b^2}{4^2\pi^2} \int_0^\pi \int_0^{2\pi} \sin d\theta d\phi \left\{J_1 J_1^* + \frac{\cos^2\phi}{\sin^2\theta}[\cos\theta J_1 - J_2][\cos\theta J_1^* - J_2^*] - \left[J_1 \sin\theta\cos\phi + \frac{\cos\phi\cos\theta}{\sin\theta}(\cos\theta J_1 - J_2)\right] \cdot \left[J_1^* \sin\theta\cos\phi + \frac{\cos\phi\cos\theta}{\sin\theta}(\cos\theta J_1^* - J_2^*)\right]\right\}.$$

As J_1 and J_2 are not functions of ϕ, the ϕ part of this integration is readily performed, and the resulting terms simplify to

$$R = \zeta_0 \frac{\beta_0^2 b^2}{4^2\pi} \int_0^\pi \sin\theta d\theta [J_1(\cos\theta) J_1^*(\cos\theta) + J_2(\cos\theta) J_2^*(\cos\theta)]$$

$$= \zeta_0 \frac{\beta_0^2 b^2}{4^2\pi} \int_{-1}^{+1} [J_1(x) J_1^*(x) + J_2(x) J_2^*(x)] dx.$$

This last integration is just one of exponentials and is readily performed. The somewhat complex array of terms so obtained can be simplified by means of trigonometric identities to:

$$R = \frac{\zeta_0}{4\pi} \beta_0^2 b^2 \frac{\cosh(\alpha s + 2\rho)}{|\cosh[(\alpha + j\beta)s + \rho + j\Phi]|^2} \cdot \left[\cosh \alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right].$$

An Electrostatic-Tube Storage System*

A. J. LEPHAKIS†, MEMBER, IRE

Summary—A storage system which will store binary (on or off) pulses has been constructed, and should prove useful in laboratory studies of certain communication problems. The system comprises two storage channels, each utilizing an MIT electrostatic storage tube and switching circuits which route incoming pulses into one channel while stored pulses are being recovered from the other. Pulses are stored in each tube in a square array of discrete spots of charge; each spot may assume one of two possible potentials, corresponding to the two possible states of a binary pulse. The order of occurrence of incoming pulses is preserved during storage, but the time relationship is not; the time relationship of pulses recovered from storage is determined by an independent pulse source under control of the user. Consequently, the system may be used to compress, expand, or delay a group of pulses. The capacity of each storage channel is, at present, 256 pulses. The system operates reliably at all frequencies up to 33 kc when storing incoming pulses and up to 70 kc when supplying stored pulses.

One of the basic problems in communication is the obtaining of more efficient utilization of transmission channels, by reducing the necessary bandwidth, reducing the required average power, or improving the signal-to-noise ratio. This problem is most profitably attacked by applying information theory. The results obtained show that storage systems are necessary components of the information-processing equipment at both the sending and receiving ends of the transmission system; and furthermore, that each such storage system must be capable of absorbing alterations of the time scale because the instantaneous rate at which new information enters the system will, in general, be different from that at which stored information is recovered.[1] Other communication applications of storage systems are to obtain delays and to obtain uniform compression or expansion of the time scale of the input.

The storage system described in this paper was designed to serve as a general-purpose laboratory instrument, capable of providing a delay or uniform or arbitrary alterations of the time scale of the input. It stores binary pulses: pulses which have only the two states of on and off and which represent the binary digits of 1 and 0. The fact that the system stores only binary pulses does not limit its utility in communication because information can be represented arbitrarily closely by a sequence of binary numbers.

The characteristics of a storage system are primarily determined by the storage device which is used as the nucleus of the system. In the present system, the MIT electrostatic storage tube was selected as the storage device because it enabled the desired flexibility with respect to time-scale alterations to be obtained without the use of an excessively large quantity of equipment. The same flexibility characteristics, and a much greater operating speed, could have been obtained by using flip-flops (bistable multivibrators) instead of the storage tube. However, the size and cost of the equipment which would have been required more than offset the advantage of a higher operating speed. Other storage devices, such as magnetic drums and ultrasonic-line storage loops,[2] were considered unsuitable because these devices cannot be easily made to absorb time-scale alterations.

The storage system contains two storage channels, each consisting of an MIT tube and the circuits which are necessary to operate the tube. Incoming pulses are routed to one of the two storage channels while pulses which have previously been stored in the other channel are being recovered; these operations are reversed when the channel in which storage is taking place is completely filled.

A continuous flow of information can be maintained through the system since the two storage channels perform opposite functions at any given time; one receives incoming pulses while stored pulses are being recovered from the other. All circuits have been made insensitive to pulse-repetition-frequency effects. The order of stored pulses is preserved, but their time relationship is lost. The time relationship of pulses recovered from storage is determined by an external source. Consequently, neither the input pulses nor the output pulses need have any periodic relationship; a delay or an alteration of the time scale may be easily obtained.

The MIT electrostatic storage tube was developed for use in the Whirlwind computer, and has been described in the literature.[3] The tube comprises a target assembly and two electron guns, one of which produces and the other maintains the two stable target potentials used to represent the two states of binary pulses. The high-velocity gun emits a narrow beam which can be positioned to any point on the target and which is turned on only when it is desired to write or read a pulse. The holding-gun beam covers the entire target, and is turned on except when writing or reading is taking place.

To write a pulse, it is necessary to position the high-velocity beam, turn off the holding beam, apply an ap-

* Decimal classification: R339. Original manuscript received by the Institute, August 11, 1950; revised manuscript received, February 16, 1951.

The work described in this paper is discussed in more detail in Technical Report No. 154, Research Laboratory of Electronics, MIT, based on a thesis, "Storage of Pulse Coded Information," submitted in partial fulfillment of the requirements for the degree of Master of Science at the Massachusetts Institute of Technology, September, 1949. This work has been supported in part by the Signal Corps, the Air Materiel Command and the ONR.

† Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.

[1] A. J. Lephakis, "Storage devices for communication," *Electronics*, vol. 23, pp. 69–73; December, 1950.

[2] Proceedings, "Symposium on Large-Scale Digital Calculating Machinery," Harvard University Press, Cambridge, Mass.; 1948.

[3] S. H. Dodd, H. Klemperer, and P. Youtz, "Electrostatic storage tube," *Elec. Eng.*, vol. 69, pp. 990–995; November, 1950.

propriate potential to the signal plate of the target assembly, energize the high-velocity beam for a short time, and finally turn on the holding beam. It is not necessary to erase before writing. A stored pulse is read by performing essentially the same operations, except that the high-velocity beam is intensity-modulated with a 10-mc radio-frequency voltage. The state of the stored pulse is determined by comparing in a phase-sensitive detector this voltage with the signal-plate rf output.

The design of the storage-system circuits was based on tentative storage-tube data which were available in the early part of 1949. Provisions were made to change easily the pertinent operating characteristics of the circuits so that future storage tubes might be accommodated. Although circuit flexibility has been obtained at the expense of more complicated equipment, it is believed that the increased complexity is justified by the fact that the equipment is not likely to become obsolete.

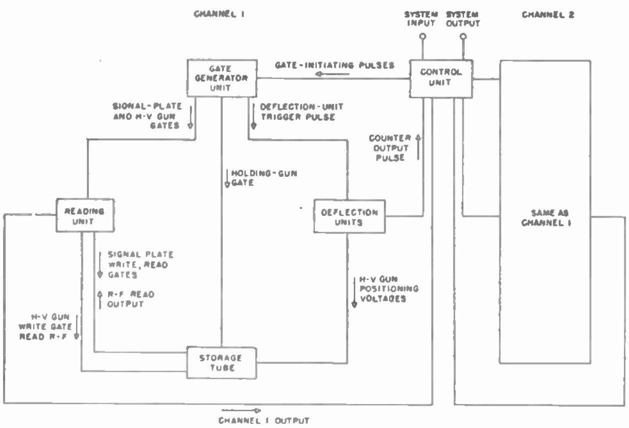


Fig. 1—Block diagram of the storage system.

A block diagram of the storage system is shown in Fig. 1. The circuits consist of four basic units. The deflection units generate the high-velocity-gun positioning voltages. The gate-generator unit generates the writing and reading gates which are required by the signal plate, the high-velocity-gun control grid, and the holding-gun control grid. Each gate is initiated by a trigger pulse. The gate-generator unit also provides the deflection-unit input pulses. The reading unit generates the pulsed oscillation which is applied to the high-velocity-gun control grid during reading, and detects the pulses which are read. The duration of the pulsed oscillation is determined by a gate from the gate-generator unit. The control unit provides the pulses which trigger the gate-generator units associated with the two storage tubes, and determines whether writing or reading is taking place in each storage tube. Switching of the writing and reading operations is caused by pulses which are produced by the deflection units whenever a storage tube is completely full or empty. When tube 1 is full, writing commences in tube 2 and reading commences in tube 1. Tube 1 must be emptied before tube 2 is filled. When tube 1 is empty, its circuits become quiescent and remain quiescent until tube 2 is filled, at which time writing commences in tube 1 and reading commences in tube 2. The circuits of tube 2 become quiescent as soon as this tube is empty, and when tube 1 is full, the cycle of operations is repeated.

It has been attempted to keep the current drain of the various circuits at a minimum, consistent with satisfactory operation, in order to simplify power-supply requirements. The low-impedance, high-current, circuits which are generally required to obtain extremely narrow pulses and extremely small rise and fall times of gates have been avoided. The widths of pulses generated by the equipment are approximately 0.2 μsec, and the rise and fall times of generated gates are approximately 0.2 μsec. Pulse transformers have been employed to invert the polarities of negative pulses and to obtain pulses at a low output-impedance level in order to prevent waste of current in inverter tubes and in cathode-follower output tubes. Pulse-repetition-frequency effects have been minimized by using clamping diodes in all capacitively coupled circuits.

The deflection units comprise two modified decoder circuits which supply 16 or 32 discrete voltage levels to the vertical and to the horizontal deflection plates of the high-velocity gun. Each such circuit consists of five pentode current sources, which supply weighted currents in the ratio 1:2:4:8:16, and are turned on and off by five flip-flops connected as a binary counter. The current-source outputs are added in a common load resistor, and the resulting voltage is applied to an amplifier which provides the balanced voltage necessary to drive the deflection plates. Each input trigger pulse shifts the output voltage to the next level. An established level is maintained constant until another input pulse is applied because direct coupling is used in the amplifier and between the current sources and the flip-flops. Thirty-two input pulses are required for a complete cycle of operation; at the end of each cycle, an output pulse is obtained from the counter. The output pulse of the horizontal counter is applied to the input of the vertical counter. In this manner, a square array of 32×32 spots, traced in sequence, is obtained on the storage-tube target. At the end of each frame an output pulse is obtained from the vertical counter and is applied to the control unit. A 16×16 array is obtained by by-passing the first stage of each counter.

The gate-generator unit supplies two sets of gates, one for reading and one for writing. All gates are generated by cathode-coupled monostable multivibrators. The gates are amplified and are applied through suitable circuits to the output lines. A typical output circuit is shown in Fig. 2. The gate is applied to the grid of a cathode follower, and the load is connected to the cathode of this tube. A discharge tube, which is normally not conducting, is connected across the load. Because of the load capacitance, the cathode potential of the cathode follower cannot change instantaneously. The positive-going edge of the gate, therefore, causes the grid-to-cathode potential of the cathode follower to increase by a considerable amount, and the resulting high cathode-follower current rapidly charges the capacitance. A large cathode resistor is used in order to allow most of the cathode-follower current to flow into the capacitance. The negative-going edge of the gate cuts off the cathode follower. A regenerative pulse amplifier, triggered by a pulse coincident with the negative-going edge of the gate, turns on the discharge tube which discharges the capacitance.

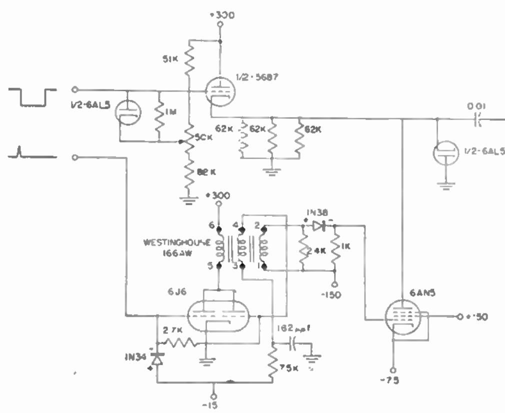


Fig. 2—Schematic diagram of the holding-gun-gate output circuit

The reading unit consists of a 10-mc pulsed oscillator which is energized by the high-velocity-gun read gate, an amplifier to which the signal-plate rf output is connected, and a phase-sensitive detector. The schematic diagram of the latter is shown in Fig. 3. Each of the

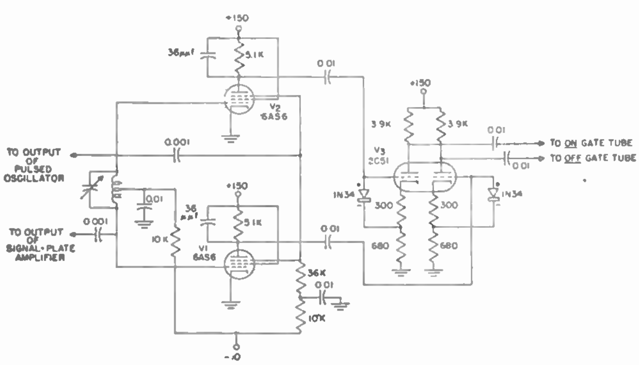


Fig. 3—The phase-sensitive detector used in the reading unit.

negative bias voltages which are applied to the suppressor grids and control grids of tubes V_1 and V_2 is sufficient to prevent plate-current flow; plate current will flow only if positive signals are simultaneously applied to both grids. The pulsed-oscillator output is applied to the suppressor grids of the tubes, and the signal-plate-amplifier output is applied to the control grids. The former signal has the same phase relationship at both tubes; the latter appears in phase opposition at the two control grids. Since the signal-plate output is either in phase or 180 degrees out of phase with the pulsed-oscillator signal, these signals coincide in one, and only one, of the two detector tubes. The output of the tube in which coincidence takes place is a negative gate. The positive gate obtained from the corresponding half of amplifier V_3 is applied to the suppressor grid of the *on* or *off* gate tube, which then passes a pulse supplied at the proper time by the control unit.

The major function of the control unit is to route the input pulses to the gate-generator units which operate the two storage tubes. Another function is to delay by appropriate amounts the pulses which initiate the signal-plate and the high-velocity-gun gates; at the start of a write or read cycle the holding beam is turned off first, then the signal plate is brought to the proper potential, and finally the high-velocity beam is turned on. Pulses are delayed by means of monostable multivibrators and pulse-forming tubes; the pulse applied to a multivibrator initiates a gate, and the pulse-forming tube generates a pulse at the trailing edge of this gate. The routing is accomplished by means of gate tubes which are opened and closed by flip-flops triggered by the deflection-unit-counter output pulses. Direct coupling is used between the gate tubes and the flip-flops.

The gate-generator units have been adjusted to allow a writing interval approximately 15 μsec and a reading interval of about 7 μsec. These units require 2 μsec to recover after a set of gates has terminated. The deflection units will provide a new spot location within 2 μsec after being triggered. On the basis of these figures, the maximum frequency limits for periodic operation of the storage system should be approximately 60 kc when writing and 110 kc when reading. It was found, however, that reliable operation did not occur under these conditions. Stored patterns degenerated, probably because of the low holding-gun duty cycle; the holding beam was on only 2/17 of the time during writing and 2/9 of the time during reading. The degeneration may have been aided by a slight defocussing of the high-velocity beam caused by stray magnetic fields.

Reliable periodic operation of the system, independent of frequency, was observed with holding-gun duty cycles of 1/2 or greater: writing frequencies of 33 kc or less, and reading frequencies of 70 kc or less. Tests involving nonperiodic operation have not been performed. However, data obtained from periodic operation indicate that the system will function properly if the minimum interval between adjacent pulses is at least 30 μsec during writing, and at least 15 μsec during reading.

Acknowledgment

The writer expresses his gratitude to J. B. Wiesner, H. E. Singleton, and S. H. Dodd and his colleagues of the Storage Tube Group of the MIT Servomechanisms Laboratory for their many valuable suggestions and advice, and to J. W. Forrester for extending the co-operation of the Servomechanisms Laboratory.

Determination of Aperture Parameters by Electrolytic-Tank Measurements*

SEYMOUR B. COHN†, MEMBER, IRE

Summary—**In this paper it is shown how the electric and magnetic polarizabilities of an aperture may be determined accurately by electrolytic-analog measurements. Measured magnetic-polarizability data are given for rectangular-, rounded-slot-, cross-, rosette-, dumbbell-, and H-shaped apertures.**

Introduction

THE ELECTROMAGNETIC problem of an aperture of any shape in an infinitely thin conducting wall between any two regions has been solved in a general manner by Bethe.[1,2] In order for the solution to be valid, it is necessary that the aperture be small compared to a wavelength and compared to the distance to the nearest sharp bend of the wall or other perturbation. Although these limitations appear rather severe, Bethe's method has proven very useful in many applications.

It was shown by Bethe that the field in the vicinity of an aperture may be represented approximately by the original field E_0, H_0 at the location of the aperture before the aperture is cut in the wall, plus the fields of an electric and magnetic dipole located at the center of the aperture. The electric dipole is oriented perpendicular to the aperture and the magnetic dipole is in the plane of the aperture. The strengths of the dipoles are related to their respective original fields by constants of proportionality that are functions only of the shape and size of the aperture. These constants are known as the electric and magnetic polarizabilities P and M. The latter is a dyadic quantity, and hence the magnetic dipole moment is in the same direction as H_0 only if H_0 coincides with a principal axis of the aperture. The two principal axes of an aperture are orthogonal, and may be easily determined from the symmetry that usually exists for a practical aperture shape. The magnetic polarizabilities along the two principal axes are scalar constants that are in general unequal. Coupling formulas containing P and M are given for many important aperture configurations in the literature.[1,3–5]

Formulas for the polarizabilities are given by Bethe only for circular and elliptical apertures, and for long slits where the magnetic field is transverse to the slit. The lack of exact formulas for the polarizabilities of other aperture shapes has been a handicap to the practical application of Bethe's method. Because of the tremendous mathematical difficulty that would be involved in obtaining precise formulas for shapes other than those previously analyzed, a method of accurate measurement is necessary. Microwave measurement of the polarizabilities immediately suggests itself, but the probable error would be of the order of 10 per cent. In this article an electrolytic-tank method capable of an accuracy of about 1 per cent is described, and extensive graphical data for M are presented for many practical aperture shapes.

Measurement Method for the Electric Polarizability

It will now be shown how the electric polarizability P of an aperture may be determined quantitatively from electrolytic-tank measurements. First, a formula will be derived that relates the electric polarizability of an aperture to the change in capacitance occurring when a magnetic-wall model of the aperture is inserted between a pair of parallel conducting plates. Then the electrolytic analog of this configuration will be presented, and the formula relating P to resistance measurements in an electrolytic cell will be given. CGS-Gaussian units are used unless otherwise indicated.

Assume a divided cell, as shown in Fig. 1(a). The cell consists of three equispaced horizontal electric walls and four vertical magnetic walls. The central electric wall, which is infinitely thin, contains an aperture of arbitrary shape whose electric polarizability is desired. With a voltage applied as shown, the field in region (2) is antisymmetrical to that in (1) about the central plane. Let the largest dimension l of the aperture be very small compared to the dimensions of the box. Then at distances $r \gg l$ from the aperture, the field in region (2) is the same as that existing if the aperture were not present plus the field of an electric dipole having the following dipole moment:

$$\Pi_e = \frac{P}{2\pi} E_0, \tag{1}$$

where P is the electric polarizability and E_0 the exciting field. This relation is (25) of Bethe's Radiation Laboratory Report.[1] (Bethe inserted the $1/2\pi$ factor in order to rationalize the electric-polarizability formula for a

* Decimal classification: R282. Original manuscript received by the Institute, October 4, 1950. The work described in this paper was accomplished for the Squier Signal Laboratory under Contract No. W36-039-sc-38246.

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

[1] H. A. Bethe, "Lumped constants for small irises," *MIT Rad. Lab. Rep.* 43-22; March 24, 1943.

[2] H. A. Bethe, "Theory of diffraction by small holes," *Phys. Rev.*, vol. 66, p. 163; 1944.

[3] C. G. Montgomery, R. H. Dicke, and E. M. Purcell, "Principles of Microwave Circuits," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 8; 1948.

[4] M. Surdin, "Directional couplers in waveguides," *Jour. IEE*, vol. 93, pt. IIIA, p. 725; 1946.

[5] C. G. Montgomery, "Technique of Microwave Measurements," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 11; 1947.

circular aperture.) The exciting field E_0 was defined verbally by Bethe on page 6 of his report, as follows:

$$E_0 = E_1 - E_2, \tag{2}$$

where E_1 and E_2 are the incident fields in regions (1) and (2). Since $E_2 = -E_1$, $E_0 = 2E_1$ and

$$\Pi_e = \frac{PE_1}{\pi}.$$

Similarly, the field in region (1) for $r \gg l$ is the original field plus that of a dipole oriented opposite to the previous one.

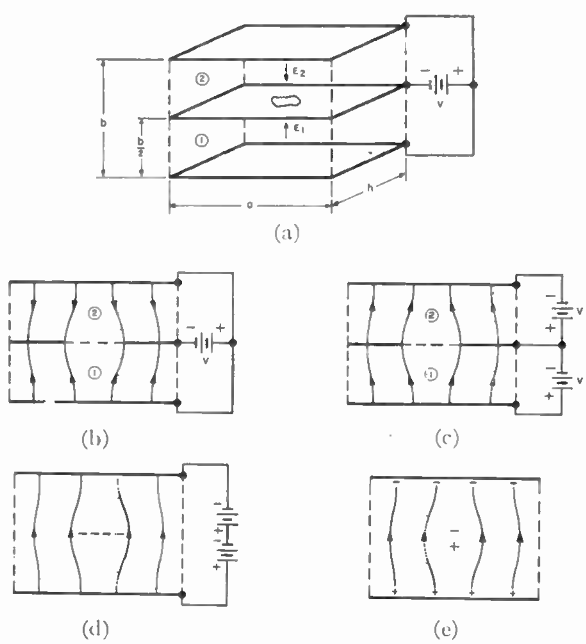


Fig. 1—Configuration for the electric-polarizability derivation.

The actual normal field in the plane of the aperture is zero in Fig. 1(a), and, therefore, one may close the aperture with a thin magnetic sheet without disturbing the field. This is shown in Fig. 1(b), where magnetic-wall surfaces are represented by dotted lines and electric-wall surfaces by solid lines. Because the two regions of Fig. 1(b) are isolated electrically, one may reverse the arrows on the field lines in region (2), as shown in Fig. 1(c), without otherwise affecting the field orientation. Now the electric field on the electric-wall portion of the partition is continuous across the partition, and hence the electric wall may be removed leaving only a thin magnetic-wall obstacle (Fig. 1(d)). The field in the entire cell of dimensions a, b, h is (for $r \gg l$) a superposition of the incident field E_1 plus the field of a single dipole replacing the obstacle of strength.

$$\Pi_e = -\frac{PE_1}{\pi}. \tag{3}$$

The minus sign takes account of the orientation of the dipole opposite to the impressed field.

Page and Adams give the following equation for the dielectric constant of a medium:[6]

$$\epsilon_r = 1 + 4\pi s_e, \tag{4}$$

where ϵ_r is the average dielectric constant relative to free space and s_e is the electric susceptibility of the medium, or the induced polarization per unit volume and unit impressed electric field. In the case of the cell of Fig. 1(e), the impressed field is E_1 and the volume is abh. The susceptibility, therefore, is

$$s_e = \frac{\Pi_e}{E_1 abh} = -\frac{P}{\pi abh} \tag{5}$$

and the dielectric constant is

$$\epsilon_r = 1 - \frac{4P}{abh}. \tag{6}$$

Let C_1 be the capacitance between the electric walls of the cell with the magnetic-wall obstacle inserted, and let C_2 be the capacitance with the obstacle removed. Then, since capacitance is proportional to dielectric constant,

$$\frac{C_1}{C_2} = \frac{\epsilon_r}{1} = 1 - \frac{4P}{abh}.$$

The electric polarizability is therefore given by the following expression:

$$P = \frac{abh}{4}\left(\frac{C_2 - C_1}{C_2}\right). \tag{7}$$

By means of this formula, the electric polarizability could be determined by capacitance measurements if the configurations were realizable. This is not the case, however, since not even a poor microwave approximation for a magnetic wall exists in nature. Examination of the analogy between conductance in the electrolytic tank and capacitance in free space shows, however, that the magnetic walls may be replaced by nonconductors in the tank, and hence the electrolytic-tank analog may be perfectly realized.[7]

In Fig. 2 (see page 1418) is shown the electrolytic cell which is the exact analog of the capacitance cell of Fig. 1(d). Two of the vertical boundaries are conductors (electric walls), the other two vertical boundaries, the bottom, and the surface of the liquid are nonconductors that simulate magnetic walls. The obstacle, which may be suspended by fine threads, is a thin, nonconducting model of the aperture under test. Let $G_1 = 1/R_1$ be the conductance that is analogous to C_1 and $G_2 = 1/R_2$ be the conductance analogous to C_2. Then in terms of the measurable resistances R_1 and R_2, P is given by

$$P = \frac{abh}{4}\left(\frac{R_1 - R_2}{R_1}\right). \tag{8}$$

[6] L. Page and N. I. Adams, "Principles of Electricity," D. Van Nostrand and Co., Inc., New York, N. Y., p. 44; 1931.
[7] S. B. Cohn, "Electrolytic-tank measurements for microwave metallic delay-lens media," *Jour. Appl. Phys.*, vol. 21, pp. 674–680; July, 1950.

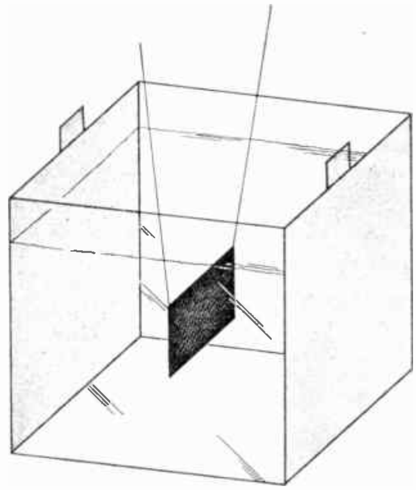


Fig. 2—Electrolytic cell containing a thin nonconducting obstacle.

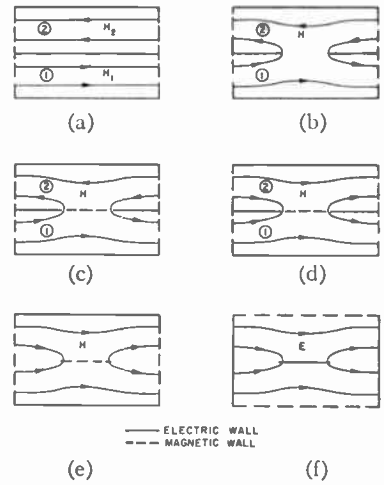


Fig. 3—Configuration for the magnetic-polarizability derivation.

Measurement Method for the Magnetic Polarizability

The derivation of the relationship for the magnetic polarizability of an aperture is similar to that for the electric polarizability. As a final step, however, an application of Babinet's electromagnetic duality principle is necessary in order to obtain a configuration suitable for electrolytic measurement. The derivation is as follows:

Assume the divided rectangular cell shown in Fig. 3(a). In this case the cell consists of three equispaced horizontal electric walls, two vertical electric walls, and two vertical magnetic walls. Let equal but oppositely-directed uniform fields H_1 and H_2 exist in regions (1) and (2) of Fig. 3(a). Since the field lines are parallel to the electric walls and perpendicular to the magnetic walls, the boundary conditions in the box are satisfied. Now assume a small aperture in the infinitely thin central wall, with one of its principal axes oriented parallel to the original field. Then the field adjusts itself as shown in Fig. 3(b). As in the electric polarizability case, the field in region (2) far from the aperture is equal to the initial field H_2, plus the field of a magnetic dipole having the magnetic dipole moment

$$H_m = \frac{M}{2\pi} H_0, \tag{9}$$

where the exciting field H_0 is given by

$$H_0 = H_1 - H_2 = 2H_1. \tag{10}$$

Because of the symmetry of the configuration, the aperture may be filled with an infinitely thin magnetic wall without disturbing the field (Fig. 3(c)). Since the two regions are now isolated, the field in region (2) may be reversed without affecting region (1) (Fig. 3(d)), and then the electric-wall portions of the central barrier may be eliminated without disturbing the field. These steps have produced the cell of Fig. 3(e) that contains a magnetic-wall model of the aperture. The average magnetic permeability of the cell containing the obstacle is given by

$$\mu_r = 1 + 4\pi s_m, \tag{11}$$

where μ_r is the average permeability relative to free space and s_m is the induced magnetic polarization per unit volume and unit impressed magnetic field. For the impressed field H_1 and cell volume abh, the permeability is

$$\mu_r = 1 + \frac{4M}{abh}. \tag{12}$$

Since, however, the cell has the field flowing in and out of its two magnetic wall boundaries, a direct electrolytic analog is not possible, and therefore the boundaries will be altered by application of Babinet's Principle. This principle states that if all the electric and magnetic walls of a nondissipative region are interchanged, and if the permeability and dielectric constant are interchanged, then E may be replaced by $-H$ and H may be replaced by E. These alterations transform Fig. 3(e) into 3(f), which has a direct electrolytic analog. The average dielectric constant of Fig. 3(f) is

$$\epsilon_r = 1 + \frac{4M}{abh}. \tag{13}$$

Let C_1 and C_2 be the respective capacitances between the electric walls of the cell with and without the obstacles present. Then

$$\frac{C_1}{C_2} = 1 + \frac{4M}{abh}. \tag{14}$$

The electrolytic analog of Fig. 3(f) is identical to that for the electric-polarizability measurement, except that a thin conducting model of the aperture parallel to the incident current flow is utilized instead of the nonconducting obstacle. In terms of the previously defined resistances, the magnetic polarizability is given by

$$M = \frac{abh}{4}\left(\frac{R_2 - R_1}{R_1}\right). \qquad (15)$$

Apparatus

A photograph of the electrolytic cell used for the magnetic-polarizability measurements is shown in Fig. 4. Two of the vertical walls inside the cell have a conducting surface of rhodium, while the other vertical walls and the bottom are nonconducting lucite. The internal dimensions of the cell are 6×6×6 inches. When in use, the cell was filled with a dilute aqueous solution of potassium chloride to a height of approximately $5\frac{1}{2}$ inches. The concentration was adjusted to give a cell resistance of about 7,000 to 10,000 ohms. The obstacles for the magnetic-polarizability measurements, which were supported in the cell by fine nylon threads, were rhodium-plated 0.0015-inch copper sheet.

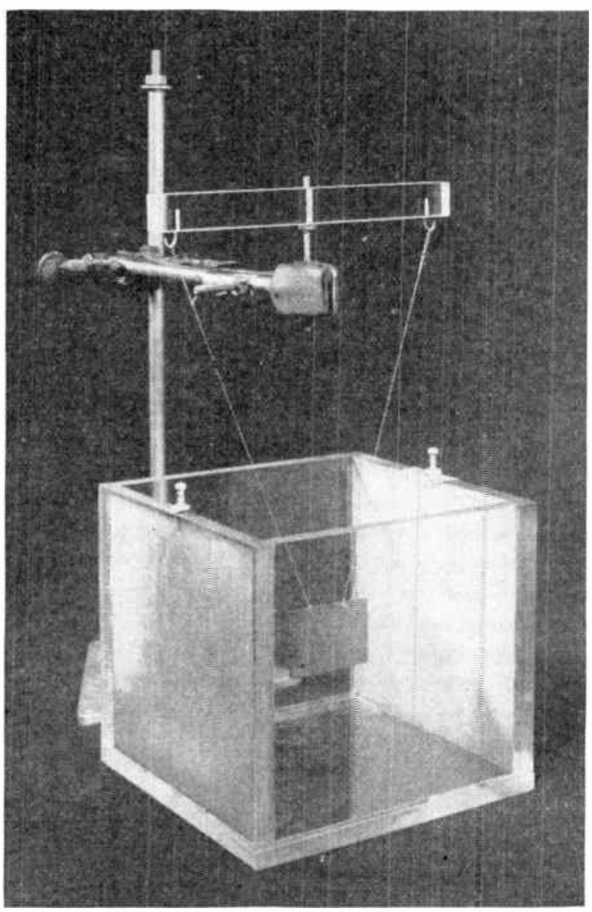


Fig. 4—Photograph of the cell and a suspended metallic obstacle.

The cell resistance was measured by a Wheatstone bridge utilizing a 1,000-cps generator and an oscilloscope detector. This equipment is discussed in more detail in a recent article on electrolytic-tank measurements for metallic delay lenses.[7] The bridge for the aperture measurements was made more sensitive, however, by the addition of 0.1-ohm steps to the variable arm. With this modification, a change of one part in 20,000 in the cell resistance was discernible by the detector.

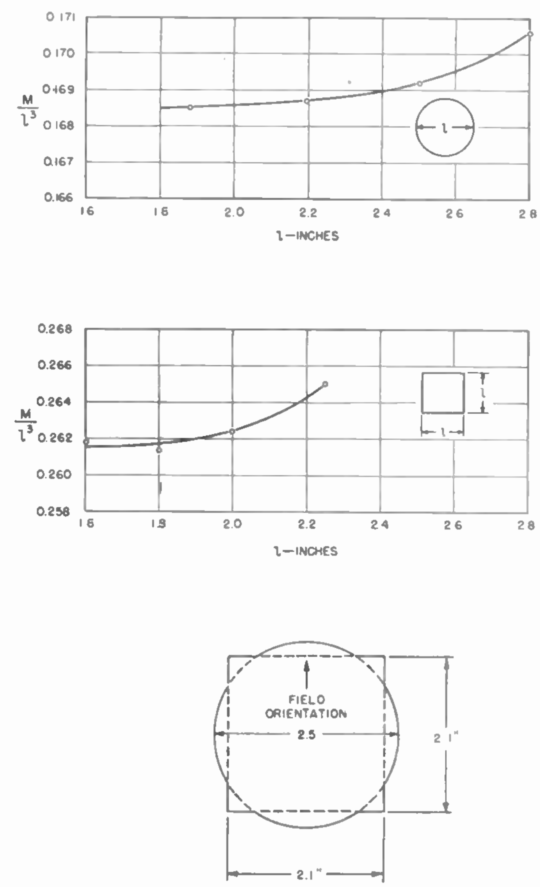


Fig. 5—Proximity effect and the criterion for maximum size.

Major Experimental Errors

In the derivation of (15), a metallic obstacle small compared to the cell was assumed. Above a certain size, the accuracy of the equation would be affected by the proximity of the cell walls. It is desirable to use the largest obstacle for which proximity effects may be neglected, since the larger the obstacle the greater the difference between R_1 and R_2, and hence the greater the precision of measurement. In order to determine the maximum permissible size, circular and square obstacles of various sizes were tested. The ratio M/l^3, which should be constant in the ideal case, is plotted in Fig. 5. It is seen that the curves are almost flat for the smaller obstacles, and bend upward for the larger obstacles. A diameter of 2.5 inches has been arbitrarily chosen as the maximum allowable for the circle and a length of 2.1 inches for the square. In a similar manner, the maximum dimensions of all other shapes might be determined, but this is not a practical procedure since it requires the

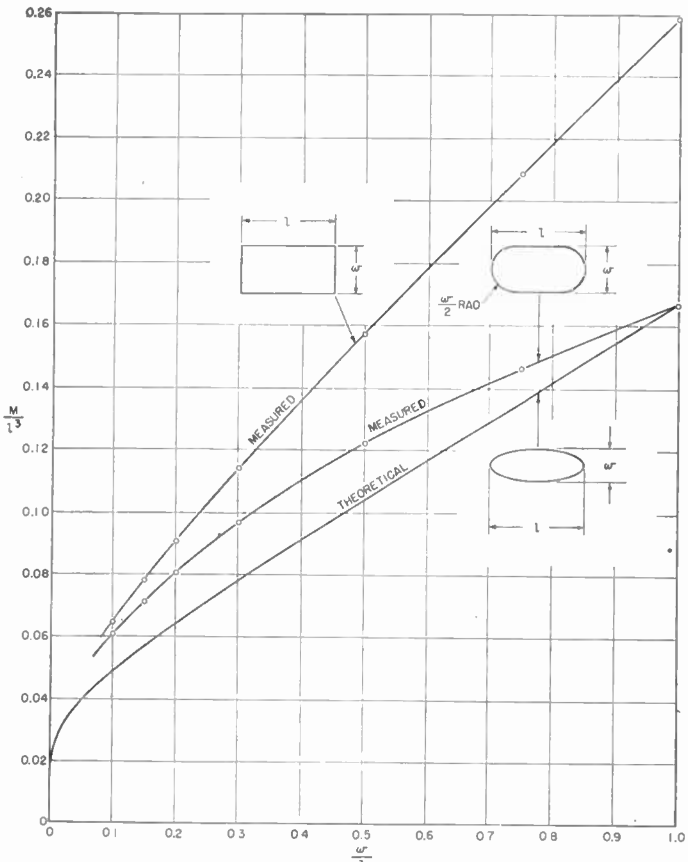


Fig. 6—Magnetic polarizabilities of rectangular, rounded-end, and elliptical slots, H parallel to l.

measurement of far too many obstacles. In order to reduce the number of measurements to one for each shape of obstacle, a criterion has been arbitrarily established that the proximity effect may be neglected for any obstacle that fits within the solid-line boundary shown in Fig. 5(c). This criterion is a reasonable one since the maximum allowable circle and square chosen from Figs. 5(a) and 5(b) fit the pattern.

The theoretical M/l^3 ratio for a circle is exactly 1/6 or 0.16666 . . . . In Fig. 5(a) it is seen that for the smallest circle the ratio was 1.1 per cent higher than theoretical, while for the 2.5-inch circle the ratio was 1.57 per cent above theoretical. This error is due not only to the proximity effect but also to the finite thickness of the obstacle and to equipment errors. In order to compensate as well as possible for the residual error due to proximity, thickness, and the like, all later M/l^3 data have been multiplied by 0.987 since this factor reduces the measured values of Fig. 5(a) in the range of permissible circle diameters to approximately the theoretical value. Although it is not rigorously justifiable to use this factor for shapes other than circles, it is believed that more accurate results are achieved thereby than if no correction factor were used at all.

As a further check on the accuracy of the measurement method, three elliptical obstacles having different eccentricities were tested with the field parallel to the major axes. In each case, the measured value of M and the theoretical value computed from Bethe's exact formula checked to 1 per cent or better. The excellent agreement verified the method and the 0.987 factor.

The Magnetic-Polarizability Data

The data for rectangular apertures and for slots with rounded ends are plotted in Fig. 6 for the case of the incident field oriented parallel to the long dimension of the apertures. As a comparison the theoretical curve for the elliptical shape is included in the graph. Fig. 7 shows

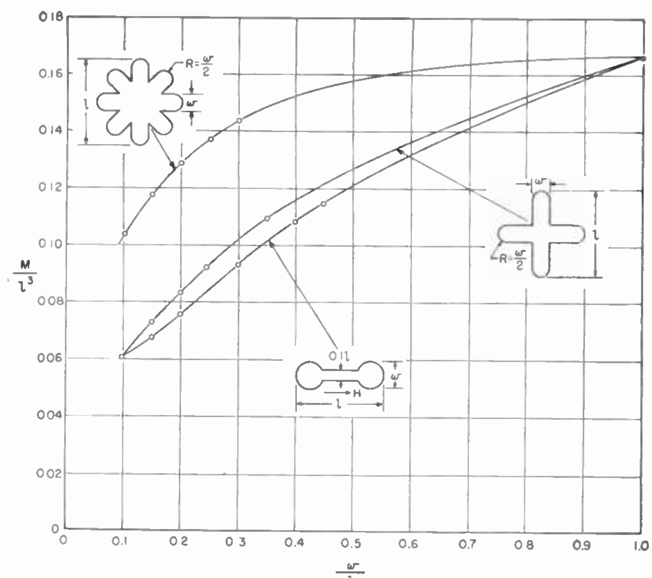


Fig. 7—Magnetic polarizabilities of rosette-, cross-, and dumbbell-shaped apertures.

the measured data for dumbbell-, cross-, and rosette-shaped apertures. For the first, the incident magnetic field is assumed in the direction of l. For the others M/l^3 is independent of the orientation of the field. Note that all three shapes reduced to a circular aperture for $w/l=1.0$. Also note that the dumbbell reduces to a rounded slot for $w/l=0.1$.

For w/l between 0.1 and 1.0 the M/l^3 curves for the dumbbell and cross are very nearly the same as that for the rounded slot. This indicates that the magnetic polarizability is determined almost entirely by the shape and size of the extremities of the aperture along the magnetic-field direction and that the effect of the intermediate portions of the aperture on M is very small. This observation should prove of value for applying the available data to other related aperture shapes. For example, although the connecting bar of the dumbbell was $0.1l$ for Fig. 7, the curve may be used with very good accuracy for any bar width less than w.

The data for an H-shaped aperture are shown in Fig. 8 for the magnetic field in the l direction. Note that for w/l between 0.1 and 0.3, M/l^3 is within 15 per cent of the value for a rectangular slot having $w/l = 0.5$. This is not surprising in view of the preceding discussion. For $w/l = 0.5$, the H-shaped aperture reduces to a rounded slot, and therefore the value for the latter is plotted at this point in Fig. 8.

Values of M/l^3 taken from the original graphs are given in Table I, below. These values are believed to have an accuracy of the order of 1 per cent.

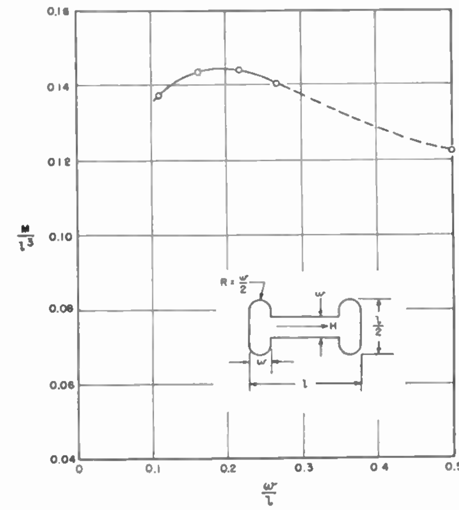


Fig. 8—Magnetic polarizability of an H-shaped aperture.

TABLE 1
Values of $M/2^3$

	$\frac{W}{2}=0.1$	0.15	0.2	0.25	0.3	0.35	0.5	0.75	1.0
Rectangle	0.0645	0.0780	0.0906		0.1139		0.1575	0.2096	0.2590
Rounded slot	0.0610	0.0711	0.0801		0.0964		0.1222	0.1455	0.1667
Cross	0.0611	0.0728	0.0832	0.0930		0.1093			0.1667
Rosette	0.1028	0.1172	0.1282	0.1368	0.1434		0.1208		0.1667
Dumbbell*	0.0610	0.0675	0.0757	0.0848	0.0932	0.1012	0.1222		0.1667
H-shape	0.1358	0.1426	0.1442	0.1418					

* Width of bar = 0.12.

Combination Open-Cycle Closed-Cycle Systems*

J. R. MOORE†

Summary—**The ancient idea of a combination coarse and fine adjustment is shown to be applicable to the design of precision automatic control systems. In the particular class of systems discussed, the coarse adjustment is taken to be a separate element operated by the input, but outside the feedback loop. Its position outside the feedback loop qualifies the coarse controller as an open-cycle system, and makes it possible to introduce such elements without affecting the system's transient response adversely. In this way, interference equalization of dynamical distortion errors is possible without such critical dependence being placed on a knowledge of series elements of the system as is required for interference equalization by a controller in the feedback loop.**

Three broad types of open-cycle systems are discussed: series, parallel, and partially parallel. Each of these may be "algebraic," "differential," or a combination. The algebraic controllers are useful when the average value of the input signal is predictable—particularly where a repetitive duty cycle is encountered.

The advantages of adding completely parallel open-cycle elements for improving speed range and reducing over-all cost are shown. Idea is also applicable to nonlinear and multiplicoupled systems.

Introduction

In the development of methods for designing automatic control systems, the effort has been concentrated almost exclusively on systems with feedback, called "closed-cycle systems," as ideally illustrated in Fig. 1(a). The reason for this is that, by the use of feedback, the designer's ignorance of the exact nature of his system elements can be rendered unimportant, and a performance which, for the great majority of applications is satisfactory, can be obtained with a minimum of effort.

Although it is theoretically possible to make a com-

* Decimal classification: 621.375.13×621.375.2. Original manuscript received by the Institute, December 9, 1949; revised manuscript received, December 1, 1950. Presented at IRE West Coast Convention, Los Angeles Section, Los Angeles, Calif., August, 1949.
† North American Aviation, Inc., Downey, Calif.

pletely open-cycle control system (Fig. 1(b)) as accurate as a completely closed-cycle system by knowing almost everything about the load and unalterable elements, the computer required for such a system would usually be impossibly complicated. Furthermore, the effort would probably be foredoomed to failure because the actuators and unalterable elements found in practice almost invariably integrate their inputs, thereby causing errors to accumulate.

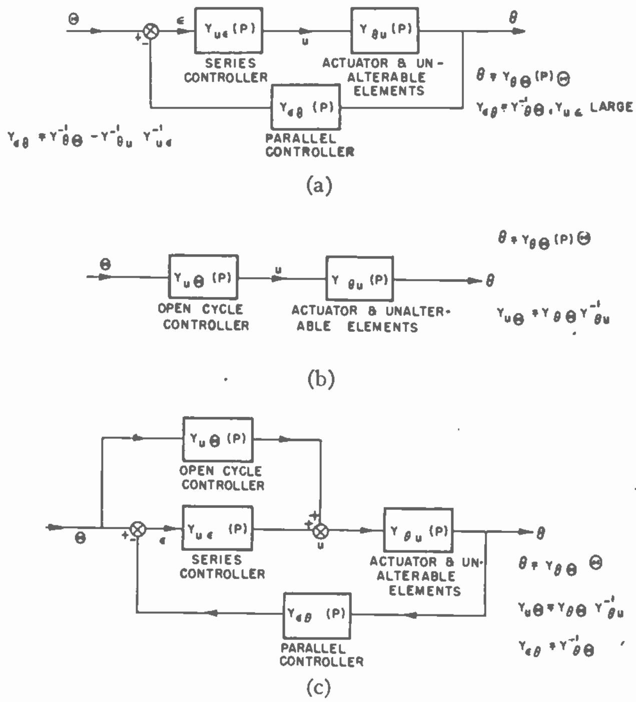


Fig. 1—(a) Idealized closed-cycle system. (b) Idealized open-cycle system. (c) Idealized combination (partially parallel) open-cycle closed-cycle system.

However, an open-cycle system may be combined with a closed-cycle system (as shown in Fig. 1(c)) to form an automatic control which is superior to either. Here the open-cycle system acts as the coarse controller providing the major portion of the output, while the closed-cycle system acts as the vernier.

The combination open-cycle closed-cycle system can be made to have the following advantages:

1. Small static errors.
2. Reduction of velocity, acceleration and higher derivative errors by factors of from two to more than 100 (compared with conventional closed-cycle systems).
3. Minimization of effects of disturbances not superposed on the desired input.
4. Maximum use of predictable input characteristics, particularly in applications with approximately periodic duty cycles.
5. Interference equalization without affecting the loop stability (contrary to the condition which exists with a parallel controller inside the feedback loop).
6. Possibility of making the closed-cycle system highly damped without reducing the velocity acceleration and higher derivative coefficients inordinately. (This overlaps advantages 2 and 3 above).
7. Possible reduction of required quality of large actuators and their controls without jeopardizing performance, thus leading to an over-all reduction in cost compared to completely closed-cycle systems of comparable performance.
8. Possibility of handling nonlinear and multiplicoupled systems with less apparatus than required for completely closed-cycle systems; furthermore, the form of this apparatus is often simpler and its perfection less critical than for corresponding parallel controllers.

The present paper attempts to illustrate these advantages and to document, formalize, and extend the basic idea of the combined coarse and fine adjustment (which many engineers and scientists have applied in some form or other to special problems of automatic control for several years[1]) via the use of open-cycle coarse control superposed on a closed-cycle vernier.

Combination Open-Cycle Closed-Cycle Systems

Despite the shortcomings of many completely open-cycle control systems, open-cycle elements are often useful for coarse controls. As indicated previously, these can, in many cases, be combined profitably with closed-cycle systems as fine adjustments to produce combination systems which are better than either a completely closed- or completely open-cycle system by itself.

Two basic ideas are involved: (a) If the input (s) is approximately predictable as a function of time—particularly if it is approximately periodic (as in the duty cycles of many processes)—this information may be used to construct an "algebraic"[2] type of controller which almost produces the correct output without ever comparing it with the desired input function. Such a controller may involve cams and computers producing a coarse output directly (in which case they are truly "algebraic") or the cams and computers may control the input to an actuator, thereby producing a "quasi-algebraic" controller. And (b) If the input is unpredictable, but the dynamic characteristics of actuators, unalterable elements, and the load are predictable in form (by "dynamic characteristics" is meant the equations relating inputs and outputs) a controller can be built which has approximately the inverse characteristics and which produces the input to the actuators.

These ideas are not confined to linear systems or linear elements. Indeed some of their greatest utility occurs

[1] One simple manifestation of the idea was mentioned by R. E. Graham as "feed forward" in his paper "Linear servo theory," *Bell Sys. Tech. Jour.*, vol. 25, pp. 616–651; October, 1946.

[2] An "algebraic" device is one whose output is (ideally) independent of the past history of its input.

where nonlinear elements are involved. They may, in general, take one or a combination of the three forms shown in Fig. 2.

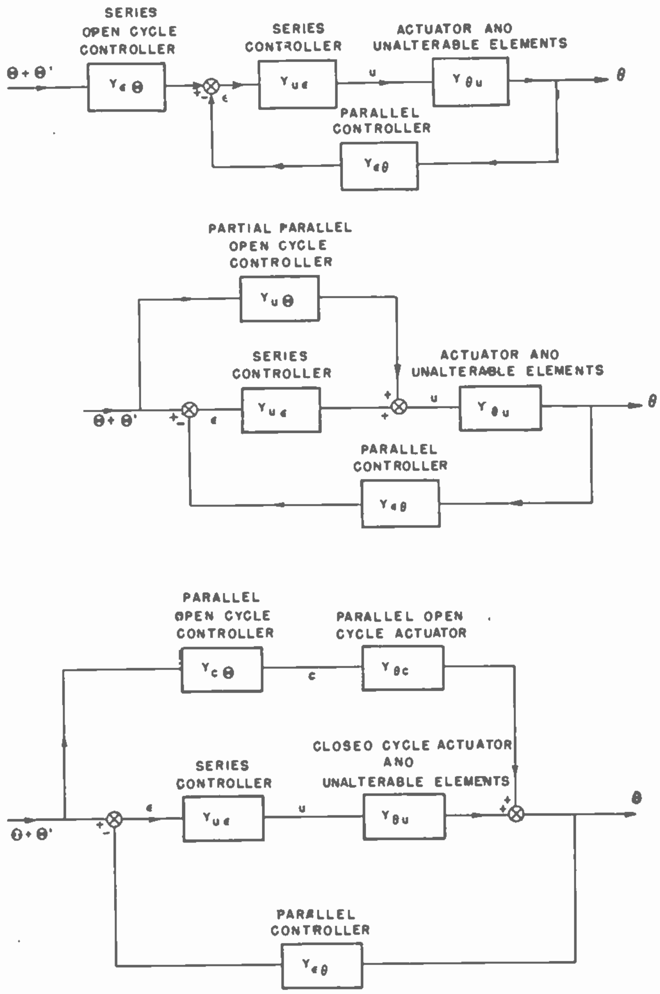


Fig. 2—Series showing partially parallel and completely parallel open-cycle elements.

Theoretical Comparison with Closed-Cycle System Synthesis

It will be seen that, in any type of combination open-cycle closed-cycle system, the open-cycle portions operate directly on the inputs and, therefore, unless unstable in themselves, do not affect the stability of the system nor its response to other desired or undesired inputs. In this way, they differ from parallel equalizers or minor loop controllers. Consequently, open-cycle controllers make it possible to design the system loops for maximum external noise suppression and low-frequency accuracy, while in themselves providing the necessary higher frequency response.

The full significance of this statement will be brought out in the discussion (to follow) on partially parallel open-cycle controllers versus parallel closed-cycle controllers. However, a brief consideration of the general problem of synthesis of closed-cycle systems will serve to introduce the idea. Referring to Fig. 1(a), the idealized closed-cycle two-variable system is shown. Here the important elements are indicated by diagram blocks.

If the primary controller is series, it reduces errors by compensating for actuator inadequacies and providing sufficient error magnification over the desired operating "frequency" range to insure a satisfactorily small error. This may be considered control by "division" or "inversion," since the error is minimized by dividing it by the series controller amplification factor at each frequency, or by inverting undesirable characteristics of the unalterable elements. Here the parallel controller is used merely to convert the output into a form suitable for comparison with the input.

If the primary controller is parallel, it reduces errors by subtraction and so may be thought of as controlling by "interference." Here the series controller may still be used to minimize the error by inversion and to "linearize," or otherwise clean up, the actuator and unalterable element dynamics.

These ideas may be illustrated for the broad problems of controller synthesis by considering the error equation of an idealized linear closed-cycle system in the presence of errors in the controller functions themselves. Thus, referring to Fig. 1(a), the output is

$$\theta = \frac{(\Theta + \Theta')}{Y_{\epsilon\theta} + Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}},$$

Here the transfer admittances Y_{ij} are assumed to be functions of the operator $P(=d/dt)$, and Θ' is the input noise. The equation is, of course, identical with the zero initial condition Laplace transform equation if P is replaced by the complex variable.

For purposes of the illustration, let us assume that the control system is a pure position servo. For this type of system, we wish to make $\theta = \Theta$. This will be written $\theta \doteq \Theta$, where $\doteq$ means "desired equal to."

As a result, the error is

$$\mathcal{E} = \theta - \Theta = \frac{[(1 - Y_{\epsilon\theta}) - Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}]\Theta + \Theta'}{Y_{\epsilon\theta} + Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}}. \quad (2)$$

We call the error in Θ, the "distortion error," and the error in Θ' the "input noise error."

If, now, the parallel controller is not to be used for interference equalization, we set

$$Y_{\epsilon\theta} = 1$$

giving

$$\mathcal{E} = \frac{-\Theta}{1 + Y_{\theta u} Y_{u\epsilon}} + \frac{Y_{\theta u} Y_{u\epsilon} \Theta'}{1 + Y_{\theta u} Y_{u\epsilon}}. \quad (3)$$

Obviously the only way in which $\partial\mathcal{E}/\partial\Theta$ can be made small is by making $Y_{u\epsilon}$ large. Evidently $\partial\mathcal{E}/\partial\Theta$ can theoretically be zero only if $Y_{u\epsilon}$ is infinite. This explains the term "synthesis by division." It is seen that if $Y_{u\epsilon}$ is large enough to make $\partial\mathcal{E}/\partial\Theta$ negligible, it is so large as to make $\partial\mathcal{E}/\partial\Theta'$ nearly unity.

Equation (2) shows that ideally $\partial\mathcal{E}/\partial\Theta$ may be made zero by use of a parallel controller if we make the coefficient of Θ zero. This requires that

$$Y_{\epsilon\theta} \doteq 1 - \frac{1}{Y_{\theta u}Y_{u\epsilon}}, \tag{4}$$

which gives

$$\frac{\partial\mathcal{E}}{\partial\Theta'} \doteq 1.$$

However, it is never possible to know any of the Y_{ij} exactly. Let us, therefore, indicate this ignorance as errors $\delta Y_{u\epsilon}$ and $\delta Y_{u\theta}$. This makes (3) more realistically

$$\frac{\partial\mathcal{E}}{\partial\Theta} = \frac{1 + \delta Y_{\epsilon\theta}Y_{\theta u}(Y_{u\epsilon} + \delta Y_{u\epsilon})}{1 + Y_{\theta u}(Y_{u\epsilon} + \delta Y_{u\epsilon})(1 + \delta Y_{\epsilon\theta})}, \tag{5}$$

while the error for the case of parallel equalization is (neglecting $\delta Y_{u\epsilon}$)

$$\mathcal{E} = \frac{-\delta Y_{\epsilon\theta}\Theta + \Theta'}{1 + \delta Y_{\epsilon\theta}}. \tag{6}$$

Evidently, with an unknown $\delta Y_{\epsilon\theta}$ the poles of the error function may have real parts not sufficiently negative to give a satisfactory degree of system stability. This is a limitation on error reduction in a closed-cycle system since it is not often possible to design for the average value of $Y_{\theta u}Y_{u\epsilon}$, but rather requires a complicated investigation of the worst combination of parameters and parameter errors. This precludes the possibility of using average values of system parameters.

By contrast, the partially parallel combination system of Fig. 1(c) relies on the open-cycle element for interference equalization. Here, however, the penalty for ignorance of $Y_{\theta u}$ is not nearly so great since $Y_{u\Theta}$ is outside the feedback loop. As will be shown later, the distortion error is, for a position servo,

$$\frac{\partial\mathcal{E}}{\partial\Theta} = \frac{Y_{\theta u}\delta Y_{u\Theta}}{1 + Y_{\theta u}Y_{u\epsilon}}. \tag{7}$$

Evidently, $\delta Y_{u\Theta}$ cannot affect the poles of $\partial\mathcal{E}/\partial\Theta$ except by adding poles of its own. Since these may always be made to have sufficiently negative real parts, errors in the open-cycle element (or ignorance of closed-cycle elements) cannot affect system stability adversely, and $\delta Y_{u\Theta}$ may be allowed to go as much negative as it can positive. This permits the design of the open cycle element for a suitably weighted average value of $Y_{\theta u}Y_{u\Theta}$, thereby permitting a splitting of uncertainty errors such as would seldom be feasible with a completely closed-cycle system.

Basic Block Diagrams and Equations of Linearized Systems

The ideas of the previous section may be understood by consideration of the basic block diagrams of three types of open-cycle closed-cycle systems. These are shown for a linearized single-output single-(nominally) input system in Figs. 3, 6, 7 and 8.

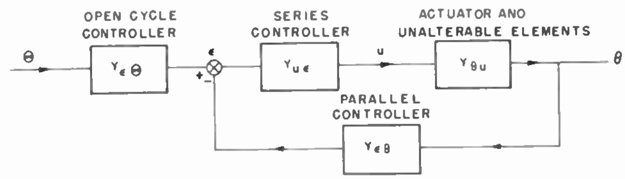


Fig. 3—The series open-cycle controller.

A. The Series Open-Cycle Controller

1. Compensation for Input Distortion Error

Considering first the system of Fig. 3, we note that the open cycle element $Y_{\epsilon\Theta}$ is placed in series with the input so that the system equation is

$$[1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}]\theta = Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\Theta}(\Theta + \Theta'). \tag{8}$$

Evidently this method is useful in the conventional sense if

$$\theta \doteq Y_{\theta\Theta}\Theta \doteq Y_{\epsilon\Theta}\Theta, \tag{9}$$

since this can be done without jeopardizing the stability or system response to disturbances entering elsewhere in the system. It is particularly desirable when the same result could not be conveniently accomplished by the more usual method of making

$$\Theta \doteq Y_{\epsilon\theta}\theta \tag{10}$$

because of the presence of an undesirable $Y_{\epsilon\theta}$ inside the feedback loop.

However, the series open-cycle controller may also be designed to minimize distortion error. Its greatest utility here comes when it is necessary to improve a system "as is" (without modifying it internally in any way, as would be required with a partially parallel controller), or when the inverse of the whole closed-loop system is simpler to mechanize with required accuracy than the inverse of some portion of the system containing the actuator and unalterable elements. As might be expected, the series open-cycle controller has a transfer function which is the inverse of the transfer function of the closed-loop part of the system multiplied by $Y_{\theta\Theta}$. Thus we start with the error equation

$$\mathcal{E} = \theta - Y_{\theta\Theta}\Theta$$
$$= \frac{[Y_{\epsilon\Theta} - Y_{\theta\Theta}(Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})]\Theta + Y_{\epsilon\Theta}\Theta'}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}. \tag{11}$$

To make $\partial\mathcal{E}/\partial\Theta$ zero, it is necessary that

$$Y_{\epsilon\Theta} \doteq (Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})Y_{\theta\Theta}. \tag{12}$$

This makes

$$\mathcal{E} \doteq Y_{\theta\Theta}\Theta'. \tag{13}$$

Actually, as pointed out in the previous section, it is not physically possible to satisfy (12) exactly. Furthermore, it is desirable to compromise between elimination of distortion error and elimination of input noise error. Thus, we assume the more realistic form

$$Y_{\epsilon\Theta} = (Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})Y_{\theta\Theta} + \delta Y_{\epsilon\Theta}, \tag{14}$$

where the desired form of $\delta Y_{\epsilon\Theta}$ is controlled by the character of the input noise, the physical realizability of the controller, the nature of the closed cycle distortion error (distribution among various derivatives or frequencies), and the economics of the situation.

The resulting error expression is

$$\mathcal{E} = \frac{\delta Y_{\epsilon\Theta}}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}\Theta + \left(Y_{\theta\Theta} + \frac{\delta Y_{\epsilon\Theta}}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}\right)\Theta'. \tag{15}$$

For physical realizability, $Y_{\epsilon\Theta}$ must never amplify high frequencies. This means that it must either be a constant or a rational fraction of powers of P whose denominator is at least of the same degree as the numerator. Thus

$$Y_{\epsilon\Theta} = \frac{K_{\epsilon\Theta}(1 + t_1P)(1 + t_2P) \cdots (1 + t_nP)}{(1 + T_1P)(1 + T_2P) \cdots (1 + T_{n+r}P)}, \tag{16}$$

where r is a positive number.

The synthesis procedure will be illustrated for an actual application of the series open-cycle controller to improving the performance of servos on an early model of the Reeves Electronic Analogue Computer (REAC). The purpose of these servos is to convert a voltage into a shaft rotation. Usually this voltage comes from one part of the computer and the servos are used to rotate resolvers or potentiometers. These early computer servos, as designed, emphasized good static accuracy, but were subject to relatively large velocity errors. In particular, when the servos were adjusted for satisfactory damping, the positional error of the output was approximately 1.2 per cent of the numerical value of the rotational equivalent of the input velocity (based on the desired relation between input voltage and output rotations), whereas the positional error of the output was 0.05 per cent of the numerical value of the rotational equivalent of the input acceleration (both being measured in "per second" time units). These give velocity and acceleration error coefficients of 83 and 2,000, respectively.

For the application being described, the acceleration and higher order errors were satisfactory considering possible acceleration of the servo, but the velocity error had to be markedly reduced. Furthermore, it was desirable to work only on the input circuits (explaining the use of a series open-cycle controller instead of the partially parallel open-cycle controller to be described later). In terms of the error expression of (11), if we denote the input voltage E by

$$E = K_{E\Theta}\Theta,$$

we may treat the servo as a pure position servo with

$$\theta \doteq \Theta,$$

so that

$$\frac{\partial \mathcal{E}}{\partial \Theta} = \frac{1}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}} - 1.$$

However, the error has already been stated in terms of $\dot{\Theta}$ and $\ddot{\Theta}$ as

$$\mathcal{E} = -\frac{\dot{\Theta}}{83} - \frac{\ddot{\Theta}}{2000} - \cdots. \tag{17}$$

This means that

$$\frac{1}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}} \approx 1 - \frac{P}{83} - \frac{P^2}{2000} - \cdots. \tag{18}$$

From which it can be inferred that

$$Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1} \approx 1 + \frac{P}{83} + \frac{P^2}{1550} + \cdots. \tag{19}$$

Since we desire only to reduce the P/83 term, it is satisfactory to strive for a

$$Y_{\epsilon\Theta} = 1 + \frac{P}{83} + \frac{\partial Y_{\epsilon\Theta}}{\partial(P^2)}P^2 + \cdots, \tag{20}$$

where the terms beyond $\partial Y_{\Theta}/\partial P$ are not to be determined, it being required only that they be negligible compared to, or opposite and nearly the same size as corresponding terms in (19). This can be done with a simple rc lead network having a transfer function of the form

$$Y_{\epsilon\Theta} = \frac{1 + \tau P}{1 + K\tau P} = 1 + \tau(1 - K)P - K(1 - K)\tau^2P^2 + \cdots. \tag{21}$$

Taking a reasonable value of K to be 1/10, we have

$$0.9\tau = 0.012, \quad \tau = 0.0135,$$

which makes the series open-cycle controller error

$$\delta Y_{\epsilon\Theta} = -\left(\frac{1}{1550} + \frac{1}{61,000}\right)P^2 + \cdots = -0.000661P^2 + \cdots. \tag{22}$$

Thus we expect the total distortion error to be, from equations (15), (19), and (22),

$$\frac{\partial \mathcal{E}}{\partial \Theta} \approx \frac{-[0.000661P^2 + \cdots]}{1 + \frac{P}{83} + \frac{P^2}{1550} + \cdots}. \tag{23}$$

The actual circuit used is shown schematically in Fig. 4. Here μ is so large that E_ϵ is negligible.

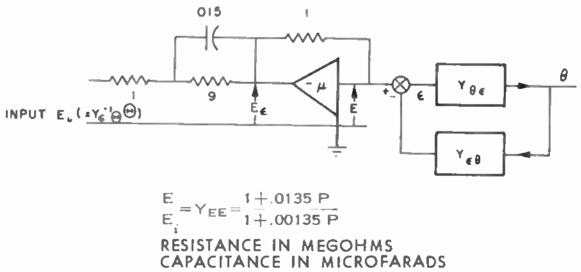


Fig. 4—REAC series open-cycle controller.

Fig. 5 below shows comparison photographs of input and error signals for the REAC servos with and without the series open-cycle controller for a step velocity, and sinusoids of three different frequencies. Evidently the velocity error has been reduced by a factor of about 15:1 for the amplitude of input signal used, whereas the sinusoidal error is improved by factors of 9.5:1, 8:1, and 4:1, respectively, for inputs of 0.5, 1, and 1.5 cps.

B. The Partial Parallel Open-Cycle Controller

Another type of open-cycle controller may be used in parallel with the series controller to produce an input directly to the actuator or actuator power amplifier. This is illustrated in Fig. 6.

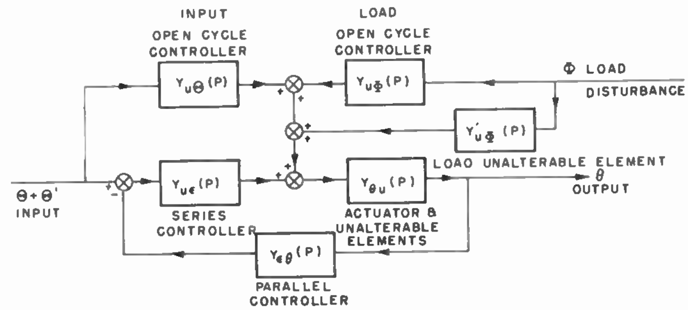


Fig. 6—Partially parallel open-cycle closed-cycle system with load compensation. The form shown here constitutes the generalized "feedforward" of R. E. Graham.

Here the linearized system equation is

$$(1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta})\theta = Y_{\theta u}(Y_{u\Theta} + Y_{u\epsilon})(\Theta + \Theta') + Y_{\theta u}(Y_{u\Phi} + Y_{u\Phi}')\Phi \tag{24}$$

where Θ' is the input noise and Φ is the load disturbance. If

$$\theta = Y_{\theta\Theta}\Theta, \tag{25}$$

the error is

$$\mathcal{E} = \theta - Y_{\theta\Theta}\Theta \tag{26}$$

$$= \frac{[Y_{\theta u}Y_{u\epsilon}(1 - Y_{\epsilon\theta}Y_{\theta\Theta}) + Y_{\theta u}Y_{u\Theta} - Y_{\theta\Theta}]\Theta + Y_{\theta u}(Y_{u\Theta} + Y_{u\epsilon})\Theta' + Y_{\theta u}(Y_{u\Phi} + Y_{u\Phi}')\Phi}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}}. \tag{27}$$

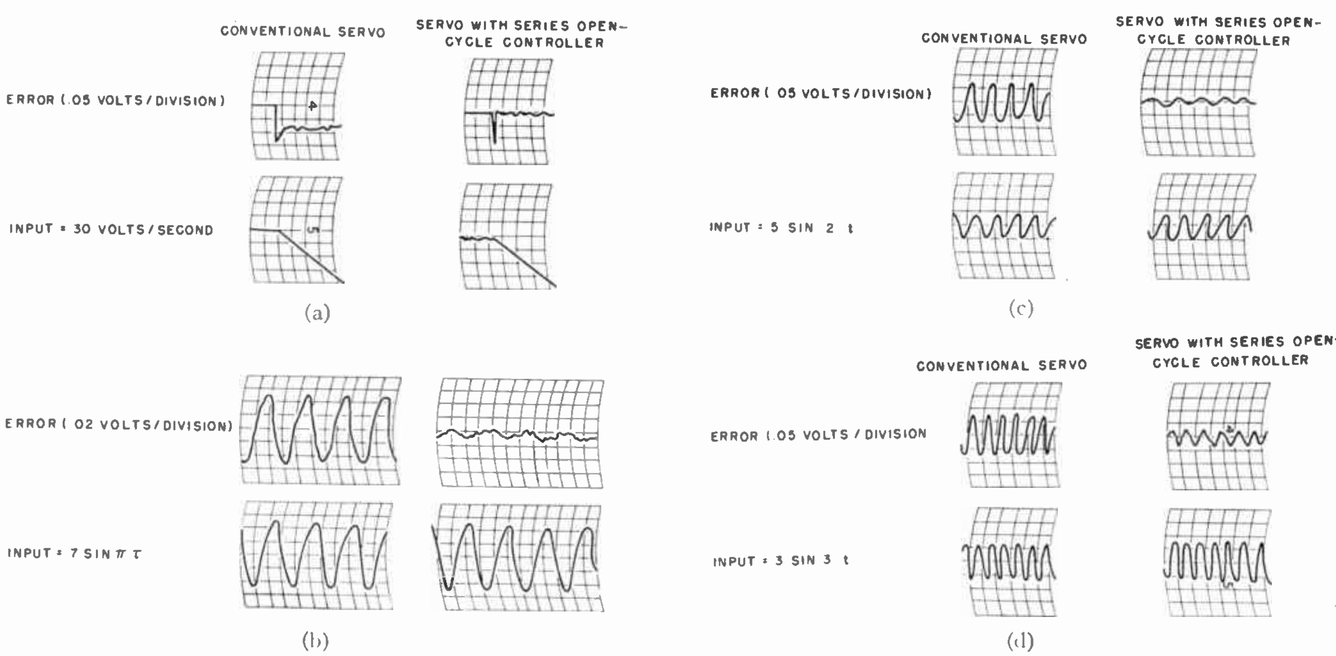


Fig. 5—The effect of open-cycle correction on a simulator servo. (a) Error, 0.05 volts per division; input = 30 volts per second. (b) Error, 0.02 volts per division; input = 7 sin $\pi\tau$. (c) Error, 0.05 volts per division; input = 5 sin 2 t. (d) Error, 0.05 volts per division; input = 3 sin $3t$.

Here it can be seen that if we choose

$$Y_{\epsilon\theta} \approx Y_{\theta\Theta}^{-1} \tag{28}$$

$$Y_{u\Theta} \approx Y_{\theta\Theta} Y_{\theta u}^{-1} \tag{29}$$

and

$$Y_{u\Phi} \approx -Y'_{u\Phi}, \tag{30}$$

(when the symbol $\approx$ means "equal over a satisfactory range"), the distortion and load errors can almost be "interfered out," just as with the series open-cycle system. The penalty for errors in $Y_{u\Theta}$ and $Y_{u\Phi}$, together with the input noise, Θ', is an expression of the form

$$\mathcal{E} = \frac{Y_{\theta u}\delta Y_{u\Theta}\Theta + [Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon} + Y_{\theta u}\delta Y_{u\Theta}]\Theta' + Y_{\theta u}\delta Y_{u\Phi}\Phi}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\theta\Theta}^{-1}}. \tag{31}$$

Again, since the open-cycle controller does not appear inside the feedback loop, uncertainties in it do not affect stability so that the closed loop may be designed for much higher attenuation than would otherwise have been possible, consistent with satisfactory response to high-frequency inputs.

However, addition of the open-cycle element makes the input noise error worse at high frequencies, illustrating the necessity for a compromise between distortion error and input noise error.

Finally, if Θ' and Φ are impulsive, the system should have good damping in all of its normal modes. In a completely closed-cycle system, good damping is obtained at the expense of gain and distortion error. The combination system permits good damping while, at the same time, keeping the distortion error low.

C. Completely Parallel Open-Cycle Controllers

The previous section considered the use of open-cycle controllers affecting the input to the system actuator or power amplifier. The method of the present section makes use of a separate actuator to feed the coarse correction into an adder where it is combined with the closed-cycle system output. Fig. 7 on page 1428 shows such an idealized servo with negligible load.

The equations of an idealized linearized system are

$$(1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta})\theta = (Y_{\theta u}Y_{u\epsilon} + Y_{\theta c}Y_{c\Theta})(\Theta + \Theta') \tag{32}$$

and

$$\mathcal{E} = \frac{[Y_{\theta c}Y_{c\Theta} - Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon}(1 - Y_{\theta\Theta}Y_{\epsilon\theta})]\Theta + (Y_{\theta c}Y_{c\Theta} + Y_{\theta u}Y_{u\epsilon})\Theta'}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}}. \tag{33}$$

As in the previous section, the best results are obtained by making

$$1 - Y_{\theta\Theta}Y_{\epsilon\theta} \approx 0 \tag{34}$$

and

$$Y_{c\Theta} \approx Y_{\theta\Theta}Y_{\theta c}^{-1}. \tag{35}$$

The practical or theoretical inability to satisfy (35) yields an error equation

$$\mathcal{E} = \frac{Y_{\theta c}\delta Y_{c\Theta}\Theta + (Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon})\Theta'}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\epsilon}}. \tag{36}$$

The greatest advantages of a completely parallel open-cycle system accrue when the closed-cycle actuator and unalterable elements are of limited capacity, or when practical and economic factors (such as will be considered in more detail later) intervene.

Practical Advantages of the Parallel Open-Cycle Closed-Cycle Systems

The purely theoretical advantages of idealized completely parallel open-cycle systems are complemented by the practical advantages of the actual devices used. Thus, any actuator has a very definite ratio of maximum to minimum speed (set by friction, slot locking, or other factors) over which it will run smoothly. This range may be increased by the introduction of "jitter" but is, nevertheless, often less than desired for wide range applications. As a result, when the servo must operate smoothly at speeds which are a small fraction of its maximum, a single actuator may be unable to accomplish the task, whether inside or outside the feedback loop. For such a situation, it is often better to resort to the completely parallel open-cycle closed-cycle system of Fig. 7 in which the open-cycle actuator is relied upon

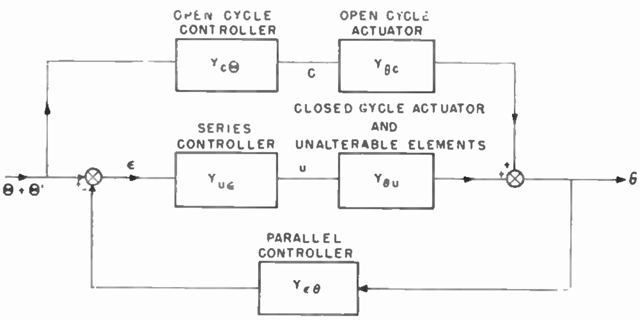


Fig. 7—Completely parallel combination system.

to produce approximately the required output speed, and the closed-cycle system serves to add or subtract small corrections to the open-cycle output. In this way, the maximum speed (referred to the output) required of the closed-cycle actuator and, correspondingly, its maximum power, may be held to, say, less than 10 per cent of what would have been required had no open-cycle element been used. Thus, if the closed-cycle actuator has only the same speed range as its open-cycle counterpart, it can operate at more than ten times the gear reduction, thereby increasing the speed range of the system by a similar factor. In such operation, the closed-cycle element adjusts the output at all speeds of operation and continues to drive the load at very low speeds after the open-cycle actuator has slipped to a halt.

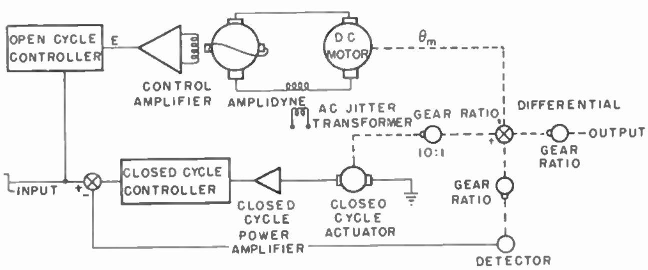


Fig. 8—Increased speed range with parallel open-cycle controller.

Such a system has other advantages besides the mere increase of operating speed range. A major one of these is economic. Thus, the cost of a closed-cycle system for fine control of a large power actuator is proportionally more than that for a small power actuator. Furthermore, for many operations in which a completely closed-cycle system could meet the specifications, the actuator and its controls would have to be of much higher quality than if it were to be part of a combination system. As a result, the closed-cycle vernier can often be added to a relatively cheap open-cycle actuator at much less cost than a single good actuator and closed-cycle system to control it.

As an example, consider the system of Fig. 8. Here the open-cycle controller is considered to be a large dc motor (shown here to be controlled by an amplidyne for purposes of illustration). The speed range of the motor is limited to, say, 50:1 even with jitter and an open-cycle controller.[3] Furthermore, the large motor can be relatively cheap for its power rating, and sluggish to respond, so that the speed and position may be instantaneously in error by as much as 10 per cent of the maximum. We now add a closed-cycle vernier such that ten revolutions of the closed-cycle actuator correspond to one revolution of the open-cycle actuator at the load. This actuator, however, is smaller and can have a much faster response than the open-cycle system actuator. Moreover, the speed range of the closed-cycle system can be made, say, 500:1. Since this 1:500 is a fraction of the 10 per cent error in the open-cycle system, we have succeeded in making the total operating speed range of the system 5,000:1. Furthermore, we have been able to use a high-power system much poorer in performance and therefore cheaper than would have been required to obtain even a 500:1 speed range with a completely closed-cycle system. As a result, the small closed-cycle system plus the large open-cycle system actually costs less than a large closed-cycle system having poorer performance.

Nonlinear Open-Cycle Controllers

A. Basic Theory

Although linearized approximations to actual transfer functions have been analyzed in examples of the previous sections, there is no such basic limitation on the combination open-cycle closed-cycle idea. Thus, if the actuator and unalterable elements have nonlinear characteristics, the partially parallel open-cycle controller must have the inverse characteristics. Referring to Fig. 7, suppose that, for

$$Y_{\theta\Theta}(P) = N_{\theta\Theta}(P)/D_{\theta\Theta}(P),$$
$$F_\theta(D_{\theta\Theta}\theta) = F_u(u) \tag{37}$$

where F_θ and F_u are any physically realizable functions. This requires that

$$F_\theta(N_{\theta\Theta}\Theta) \approx F_u(u) \tag{38}$$

be the equation of the partially parallel open-cycle controller, normally requiring a computer.

B. Example of a Hydraulic-Rate Servo

To illustrate, suppose that we wish to design a speed-controlled ("rate") servo. Here

$$\dot{\theta} = K_{\theta\Theta}\Theta. \tag{39}$$

[3] The reader will readily note that the open-cycle controller might be aided by a tachometer-accelerometer feedback around the large motor. This would correspond to an internal feedback loop in a closed-cycle system, and should be considered part of the open-cycle controller.

If the actuator is hydraulic and controlled by a valve, we consider the actual nonlinearities of the pressure drop across the valve. Referring to Fig. 9, a typical valve-controlled hydraulic ram is shown. Here we take the empirical formula for flow versus pressure

$$K_{\dot{\theta}\theta}\dot{\theta} = \delta(\Delta p)^n \tag{40}$$

where

$$\frac{\Delta p}{2} = p_1 - p = p' - p_0.$$

The load force is proportional to

$$p - p' = p_1 - p_0 - \Delta p$$

so that, if L is the load force

$$p - p' = K_{p\ddot{\theta}}\ddot{\theta} + K_{pL}L = p_1 - p_0 - \Delta p. \tag{41}$$

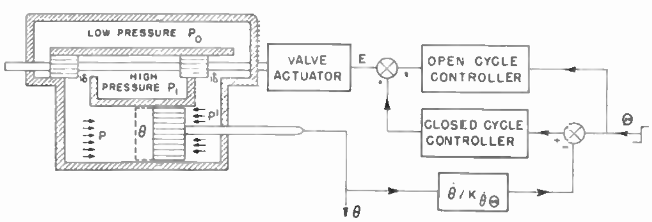


Fig. 9—Nonlinear open-cycle controller with hydraulic actuator.

The valve motion is related to the control voltage by the relation

$$(P^2 + 2\zeta_\delta\omega_\delta P + \omega_\delta^2)\delta = K_{\delta E}E. \tag{42}$$

We now seek the relation between Θ and E required for the open-cycle controller to make

$$\dot{\theta} = K_{\dot{\theta}\Theta}\Theta. \tag{43}$$

If we are successful in designing the open-cycle controller (and if (40) is sufficiently accurate), we may replace $\dot{\theta}$ by $K_{\dot{\theta}\Theta}\Theta$ in (40) and (41) to give

The reader will readily recognize that such a controller need not be restricted to open-cycle operation, but may also be used as a "linearizer" in series with the actuators and unalterable elements of a completely closed-cycle system for synthesis by "inversion." However, if used in the open-cycle controller, such a computer will have no effect on system stability.

Of course in any practical hydraulic system, the nonlinearity is masked by an output function feedback around the actuator-valve combination. This introduces problems of inner loop stability or sluggishness, but makes it possible to use a much more approximate open-cycle controller design.

C. Example of an Algebraic Cyclical Control

Where the desired output is approximately relatable to the time or an input variable by an algebraic (nondifferential) relationship, the open-cycle system may consist of a quasi-direct drive from input to output.

An example of such a condition is the flat card winding machine tension controller. Here, if a completely closed-cycle system (as shown in Fig. 10) is used, the tension actuator must be employed more to take up slack than to control tension. As a result, its ability to

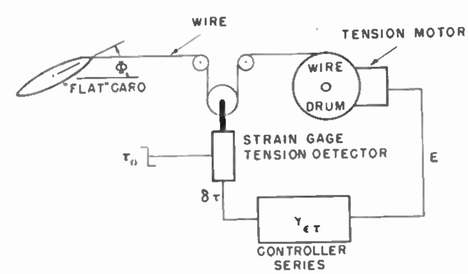


Fig. 10—Closed-cycle tension control system.

control tension is greatly impaired by the necessity for accelerating the mass of actuator and wire feed elements before the tension can even be affected. To illustrate this, we briefly analyze the problem of Fig. 10.

$$K_{\delta\theta}K_{\theta\Theta}(P^2 + 2\zeta_\delta\omega_\delta P + \omega_\delta^2)\left[\frac{\Theta}{(p_1 - p_0 - K_{p\ddot{\theta}}P\Theta - K_{pL}L)^n}\right] = K_{\delta E}E. \tag{44}$$

This shows what formula an idealized computer with Θ as an input and E as an output would take. Such a computer would, of course, be only approximate since perfect derivatives can never be mechanized. It would also be much more complicated than warranted for most applications, and is used here only as an example of a nonlinear open-cycle element design in all of which the dependent variable of the equations relating actuator output to input is replaced by the independent variable of the system.

Assuming that the friction and inertial torques of the rollers are negligible, the idealized tension equations are, for a separately excited dc or two-phase ac motor

$$\tau = -K_{\tau E}E + K_{\tau\dot{\theta}}P(1 + T_mP)\theta \tag{45}$$

$$-E = Y_{E\tau}(\tau_0 - \tau) \tag{46}$$

where τ_0 is the constant desired tension of the tension regulator.

It follows that

$$\tau = \frac{\tau_0 + K_{\tau\dot\theta}P(1 + T_mP)\theta}{1 + K_{\tau E}Y_{E\tau}}. \tag{47}$$

The tension error is

$$\delta\tau = \tau - \tau_0 = \frac{K_{\tau\dot\theta}P(1 + T_mP)\theta - K_{\tau E}Y_{E\tau}\tau_0}{1 + K_{\tau E}Y_{E\tau}}. \tag{48}$$

This shows $\delta\tau$ to depend on $\dot\theta$ and $\ddot\theta$. However, θ is determined by the position of the card given by the angle, ϕ. Thus, the tension regulation is evidently being greatly penalized by the full magnitudes of $\dot\theta$ and $\ddot\theta$ which may really be considered extraneous to the problem of tension control, since they depend on the size, shape and rotation speed of the card rather than on the tension.

Obviously, therefore, if we could minimize $\ddot\theta$ and modify E to eliminate the error chargeable to $\dot\theta$, the regulator could be made much more satisfactory. To do this, we add the open-cycle elements of Fig. 11. They have as their purpose the taking up of slack and the cyclical adjustment of winding speed so that $\dot\theta$ can be made nearly constant at the average rate of laying wire on the card.

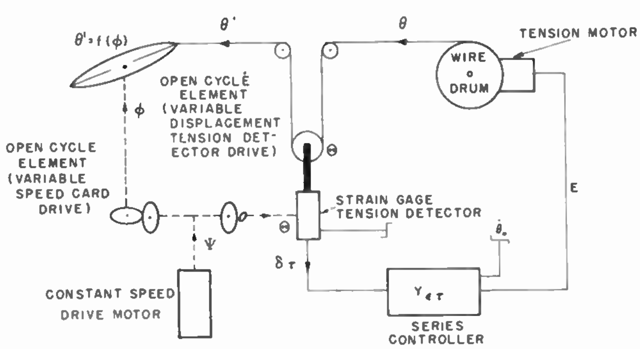


Fig. 11—Combination open-cycle closed-cycle tension regulator.

In the particular manifestation of the system shown in Fig. 11, the open cycle elements are a cam drive on the tension detector and a variable speed drive on the card such that, if $\dot\theta_0$ is the average speed of the wire drum,

$$2\Theta(\psi) = f[\phi(\psi)] - \dot\theta_0 t, \tag{49}$$

where $\Theta(\psi)$ is the cam drive and $\phi(\psi)$ is the card winding variable speed cam gear drive employed, if at all, to make the design of the $\Theta(\psi)$ cam more feasible mechanically.[4] In addition, setting of $\dot\theta_0$ sets a voltage E_0 which, for constant velocity conditions, maintains the required current in the motor to maintain τ equal to τ_0 without requiring a strain gage error to do so.

The improvement to be expected from such a system will now be demonstrated by an idealized analysis. We begin by assuming that τ is the tension throughout the whole length of wire from card to drum at any instant. If θ is the angular displacement of the wire drum with slack take-up operating, Θ is the displacement of the slack take-up device, and θ' is the displacement which the drum would have without a slack take-up, we have

$$\theta = \theta' - 2\Theta. \tag{50}$$

It follows, of course, that ideally we have

$$\dot\theta \doteq \dot\theta_0$$

so that

$$\ddot\theta \doteq 0.$$

If all of our idealized conditions can be met, there is no need for tension control other than calibration of the motor torque with supply current. However, in any practical case, imperfections will creep in all along the line and, indeed, it is the fact that a very considerable gain can be demonstrated despite these imperfections that makes the use of a combination open-cycle closed-cycle system desirable for this application.

Let the sum total imperfections be expressed as an equivalent error in $f[\phi(\psi)]$ so that substitution of (49) into (50) gives

$$\theta = \dot\theta_0 t - \epsilon \tag{51}$$

where

$$\epsilon = f[\phi(\psi)] - \theta'[\phi(\psi)]. \tag{52}$$

We now add the last open-cycle element by making

$$E = \frac{K_{\tau\dot\theta}\dot\theta_0}{K_{\tau E}} + Y_{E\tau}(\tau_0 - \tau). \tag{53}$$

Substituting (53) into (45) gives

$$\tau = \frac{\tau_0 + K_{\tau\dot\theta}P(1 + T_mP)\epsilon}{1 + K_{\tau E}Y_{E\tau}} \tag{54}$$

so that

$$\delta\tau = \frac{K_{\tau\dot\theta}P(1 + T_mP)\epsilon - K_{\tau E}Y_{E\tau}\tau_0}{1 + K_{\tau E}Y_{E\tau}}. \tag{55}$$

Comparing (55) with (48), it is observed that the error now involves ϵ where θ occurred before. Of course, $Y_{E\tau}$ can be selected to make the steady state error involving τ_0 negligible.

Evidently the tension error has been reduced by adding the open-cycle elements, an amount depending on the time variations of ϵ compared with the time variations of θ (in the system of Fig. 11). This permits an increase in winding speed which can greatly augment the output of the winding machine.

The reader will recognize that the specific example chosen does not represent the only means of accomplishing the desired objective of improved tension control. Thus there are immediately called to mind several improvements. In some of these, greater flexibility for use with cards of different shapes is attained by replacing

[4] Without such a cam gear drive, the required variations in speed of the $\Theta(\psi)$ cam follower would be prohibitive.

the cams by computer-operated servos, providing adjustment of the functions ϕ and Θ to correspond to variations in card depth. Indeed, the best mechanical design is obtained by putting the tension adjustment at Θ and the slack take-up as a ψ-controlled variation in angular displacement of the wire drum. However, the example has been given, not to describe an improved tension control, but rather to illustrate advantages of the addition of completely parallel, algebraic open-cycle elements to an otherwise closed-cycle system.

The same idea may obviously be applied anywhere that an approximate duty cycle can be completely established for the output of an automatic control system, requiring, of course, that the significant rates of the difference between actual and approximate duty cycles are suitably less than the corresponding rates of the true desired output. It is also useful for partially parallel open-cycle systems. Here the actuator voltage is continually adjusted by an algebraic open-cycle element in accordance with the predictable portion of the duty cycle, and corrected for the unpredictable portion by the closed-cycle vernier.

Multiple Variable Systems

The previous sections have demonstrated the use of combination open-cycle closed-cycle systems in a variety of linear and nonlinear single-input single-output combinations. However, many of the more advanced problems of precision automatic control system design involve systems with several inputs and several outputs. Examples include multiple variable process controls, automatic pilots, blind landing equipment, and gimbal system servos (to name only a few).

As with the two variable systems, series, partially parallel, completely parallel open-cycle controllers, or combinations thereof, may be employed. Here, however, the interaction is often nonlinear and the requirement for computers in the open-cycle elements is increased.

Suppose that a physical system contains n inputs E_i and m outputs θ_j related by the m equations

$$F_j(E_1, E_2, \cdots E_n, \theta_1, \theta_2, \cdots \theta_m) = 0. \tag{56}$$

Suppose, further, that it is desired to design a control system making use of the physical system by controlling its inputs or their effects in such a way that

$$\theta_j \doteq \theta_j(\Theta_1, \Theta_2 \cdots \Theta_n) \tag{57}$$

where the Θ_j are now considered the inputs of the whole automatic control system.

In a completely closed-cycle system, the scheme would be to control m of the E_i by errors

$$\mathcal{E}_j = \Theta_j - (\theta_j')^{-1}[\theta_1, \theta_2, \cdots, \theta_m]. \tag{58}$$

and the major problem of controller design would consist of finding a suitable set of relations among the $\mathcal{E}_j$ and E_j. The $(n-m)E_i$, which do not enter into the control, constitute unwanted inputs (disturbances or "noises") and their effects must be minimized.

In a partially parallel combination open-cycle closed-cycle system, the problem is to relate Θ_j to the m desired unalterable element inputs E_j. This requires that a computer be designed such that

$$\overline{F}_j(E_1', E_2', \cdots, E_m', E_{m+1}, \cdots, E_n, \theta_1', \theta_2', \cdots, \theta_m') = 0 \tag{59}$$

where the $\overline{F}_j$ represent practical approximations to the F_j. Here the θ_j' and $n-m$ of the E_i are considered controller inputs, and m of the E_i are considered controller outputs. The E_j' are added to the outputs E_k'' of a (by now greatly simplified) closed-cycle controller to establish the inputs E_j to the unalterable elements. Of course, it follows that

$$E_k'' = E_k''(\mathcal{E}_1, \mathcal{E}_2, \cdots, \mathcal{E}_n). \tag{60}$$

However, the E_k'' can each be made to depend primarily on one $\mathcal{E}_j$, since the open-cycle controller has insured that all $\mathcal{E}_j$ are much smaller than would be possible with a completely closed-cycle system.

Fig. 12 illustrates the system described by the foregoing equations. Here, however, we draw the diagram to

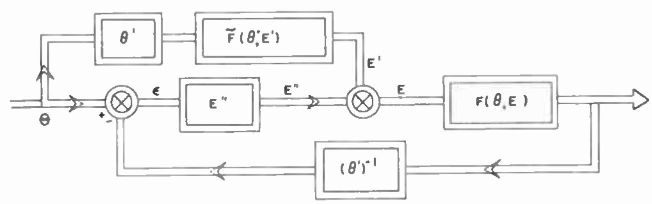


Fig. 12—Partially parallel open-cycle controllers in multiple variable systems.

represent one-way matrices defined by taking all similar quantities as a group. Thus letters without subscripts represent the matrices and

$E \equiv E_1, E_2, \cdots, E_n$	(unalterable element input matrix)
$E' \equiv E_1', E_2', \cdots, E_m'$	(open-cycle controller output matrix)
$\theta \equiv \theta_1, \theta_2, \cdots, \theta_m$	(actual output matrix)
$\theta' \equiv \theta_1', \theta_2', \cdots, \theta_m'$	(desired output matrix)
$F \equiv F_1, F_2, \cdots, F_m$	(unalterable element relations)
$\overline{F} \equiv F_1', F_2', \cdots, F_m'$	(open-cycle controller)
$\Theta \equiv \Theta_1, \Theta_2, \cdots, \Theta_m$	(desired input matrix)
$E'' \equiv E_1'', E_2'', \cdots, E_m''$	(closed-cycle series controller)
$\mathcal{E} \equiv \mathcal{E}_1, \mathcal{E}_2, \cdots, \mathcal{E}_m$	(closed-cycle controller inputs).

Such partially parallel open-cycle controllers are particularly useful in precision autopilot designs and systems involving appreciable transport lag. Again, the job may often be performed better by a series open-cycle element or a combination of series and partially parallel elements.

Conclusions

Although there are many applications where the addition of open-cycle elements carries the design beyond the point of diminishing returns (or, in some cases of bad input noise, may actually make the system performance poorer), the requirements for better and better automatic control systems reveal many other situations in which the use of a combination open-cycle closed-cycle system can be expected to "save the day." Thus, it is believed that the concepts and synthesis techniques involving combination open-cycle closed-cycle systems, although certainly no panacea, represent a powerful addition to the repertory of the precision automatic control system designer.

Notation and Nomenclature

Θ = system "desired" or "true" input
Θ' = input disturbance or "noise"
θ = system output
$\mathcal{E}$ = system error [difference between actual noise free) input and that function of output over which control is desired]
$\mathcal{E}_n$ = noise error
Φ = load disturbance—flux
$N_{\alpha\beta}(P)D_{\alpha\beta}(P)$ = numerator and denominator polynomials in P
$Y_{\alpha\beta}(P)$ = operator applied to element input, β, to get element output, α, usually a rational fraction
E = actuator input or control voltage
T_i, τ_i = system parameters
τ = wire tension
$K_{\alpha\beta}$ = constant
K = per unit gain uncertainty or variation
$P = d/dt$ (also taken as the complex variable of Laplace transforms)
p = hydraulic pressure
c_i = reciprocal error coefficient of the ith derivative
ω = input circular frequency
$\doteqdot$ = desired equal to
$\approxeq$ = equal over a satisfactory range
y = motor field flux.

Bibliography

1. W. R. Ahrendt and J. F. Taplin, "Automatic Regulation," vol. I, Ahrendt and Taplin, Washington, D.C.; 1947.
2. H. W. Bode, "Network Analysis and Feedback Amplifier Design," John Wiley and Sons, Inc., New York, N. Y.; 1948.
3. G. S. Brown and D. P. Campbell, "Principles of Servomechanisms," John Wiley and Sons, Inc., New York, N. Y.; 1948.
4. G. S. Brown and A. C. Hall, "Dynamic behavior and design of servomechanisms," *ASME Trans.*, vol. 68, pp. 503–52; July, 1946.
5. Caldwell, G. A., "Wide speed ranges with servomechanisms," *Westinghouse Eng.*, vol. 5, pp. 157–160; September, 1945.
6. C. S. Draper, Li, and Laning, "Measurement and control systems for engineers," *ASME Conference Paper*, San Francisco, Calif.; June 29, 1949.
7. M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," vol. I, John Wiley and Sons, Inc., New York, N. Y.; 1942.
8. R. E. Graham, "Linear servo theory," *Bell Sys. Tech. Jour.*, vol. 25, pp. 616–651; October, 1946.
9. H. Harris, Jr., "A comparison of two basic servomechanism types," *AIEE Trans.*, vol. 66, pp. 83–93; 1947.
10. H. Harris, Jr., "Analysis and Design of Servomechanisms," Fire Control Committee, Section D-2, NDRC, OJRD Report Number 454, 91 pp.; 1942.
11. H. L. Hazen, "Theory of servomechanisms," *Frank. Inst. Jour.*, vol. 218, pp. 279–331; September, 1934.
12. S. W. Herwald, "Forms and principles of servomechanisms," *Westinghouse Eng.*, vol. 6, pp. 149–154, September, 1946.
13. H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co. Inc., New York, N. Y.; 1947.
13. R. Latter, "Theory of Linear Servomechanisms," Naval Radiation Laboratory Report 2610, 105 pages; August, 1945.
14. H. Laner, R. Lesnick, and L. E. Matson, "Servomechanism Fundamentals," McGraw-Hill Book Co. Inc., New York, N. Y.; 1947.
15. H. T. Marcy, "Parallel circuits in servomechanisms," *AIEE Trans.*, vol. 65, pp. 521–529; August, September, 1946.
16. E. S. Smith, Jr., "Automatic Control Engineering," McGraw-Hill Book Co. Inc., New York, N. Y.; 1944.

CORRECTION

Abd El-Samie Mostafa, author of the paper, "Electron-Tube Performance with Large Applied Voltages," which appeared on pages 70–73 of the January, 1951, issue of the PROCEEDINGS OF THE I.R.E., has brought the following errors to the attention of the editors:

Page 71—Column 2, line 4
sin instead of rin

Equations (5), (7), and (8)
$-a_0E_c$ instead of $-a_0(E_c+B)$

Page 72—Column 2, line 1
$-E_c$ instead of E_c

Add to caption of Fig. 4
$E_b = 300$ volts, $E_c = 100$ volts, and $V_p = 130$ volts (R.M.S.)

The author's date of birth is April 27, 1913 instead of 1917.

Transient Response of a Narrow-Band Automatic Frequency-Control System*

THEODORE MILLER†, MEMBER, IRE

Summary—A method of analyzing the response of automatic frequency-control systems, operating in conjunction with a narrow band-pass filter, is presented in this paper. A time-lag, equivalent to the reciprocal of the bandwidth, is assigned to the filter. Circuit parameters are established for operation near the critically damped response condition. The equations of the system are derived with the aid of the Laplace transform. The method of residues is used to evaluate the transient response to a step-input frequency disturbance.

Introduction

A RECENT CIRCUIT PROBLEM required an automatic frequency-control system to operate in conjunction with a very narrow band-pass filter. The presence of such a filter introduces a time lag which complicates the behavior of the system, and makes its transient response difficult to predict. In a recent paper, van der Wyck[1] considers the dynamics of an afc system, and outlines conditions for a nonoscillatory response as derived from an unpublished work of de Cock Buning. However, the presentation is very brief, and no attempt is made to calculate the actual response time of the system.

This paper will present a method of analysis for narrow-band afc systems based on the Laplace transform. Conditions for a damped nonoscillatory transient response will be derived, and the response time of a system subjected to a step-input frequency disturbance will be calculated.

Description of the afc System

A block diagram of the afc system appears in Fig. 1. A portion of the output signal is amplified, limited, and applied to a frequency discriminator which is balanced for zero voltage output at the center frequency of the band-pass filter. A frequency variation in the output signal produces a discriminator output voltage which, acting through the integrating circuit and reactance tube, corrects the local oscillator frequency so that the output frequency is maintained within the band-pass region of the filter. The insertion of the integrating network permits adjustment of the transient response of the system.[2] The circuit parameters must be selected so that the steady-state output frequency deviation of the filter (resulting from an input frequency disturbance) will always remain in the band-pass region of the filter. This selection also must be consistent with any specified transient response.

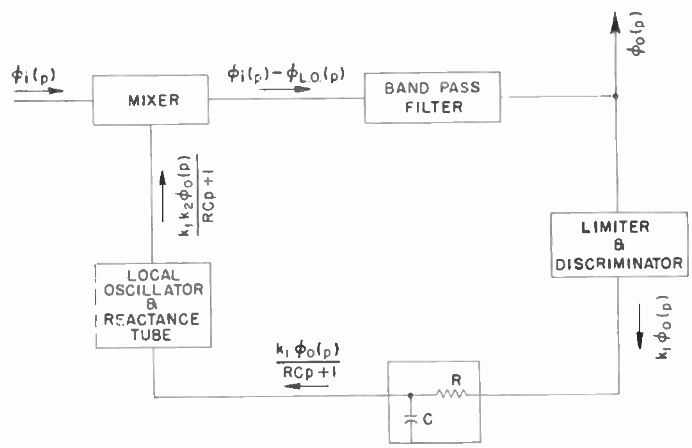


Fig. 1—Block diagram of the afc system. (The transforms of the variables are indicated as functions of P.)

Assumptions and Definitions

It will be assumed that at time $t=0$, an input frequency variation disturbs a previously attained steady-state condition in which the output frequency exactly coincided with the center frequency of the band-pass filter. The input disturbance will be designated as $\phi_i(t)$, and the resulting output frequency deviation as $\phi_0(t)$. The narrow band-pass filter will introduce a time delay τ, which is approximately equivalent to the reciprocal of its band-width at the one-half power points. It will also be assumed that the discriminator and reactance-tube responses are linearly proportional to frequency and voltage, respectively. From these assumptions the following definitions may be made:

$\phi_i(p)$ = Laplace transform of input frequency disturbance

$\phi_0(p)$ = Laplace transform of output frequency deviation

$\phi_{L.O.}(p)$ = Laplace transform of local oscillator frequency variation

$\phi_m(p)$ = Laplace transform of output of mixer

$e^{-\tau p}$ = Transfer function of band-pass filter[3]

* Decimal classification: R361.215. Original manuscript received by the Institute, July 14, 1950; revised manuscript received March 23, 1951.

† Westinghouse Research Laboratories, Westinghouse Electric Corp., East Pittsburgh, Pa.

[1] van der Wyck, "Netherlands PTT single-sideband equipment," Proc. I.R.E., vol. 36, pp. 970–980; August, 1948.

[2] Other kinds of control circuits, such as integral plus proportional control, may sometimes be employed advantageously. For an excellent discussion of this aspect of the automatic control problem, see Hall, "Analysis and Synthesis of Linear Servomechanisms," Technology Press, MIT, Cambridge, Mass.; May, 1943.

[3] The transfer function of a network is defined as the ratio of the Laplace transform of the output signal to the Laplace transform of the input signal; the Laplace transform of $F(t)$ is defined as $f(p)=\int_0^\infty F(t)e^{-pt}dt$. Because of amplitude limiting, the afc system is insensitive to amplitude variations, and therefore it is not necessary to include the amplitude versus frequency characteristics of the filter in the transfer function for the filter.

$\frac{1}{RCp+1}$ = Transfer function of integrating network

k_1 = Sensitivity of discriminator in volts per cycle

k_2 = Sensitivity of reactance tube and local oscillator in cycles per volt.

$k = k_1k_2$ = "Loop gain" of system.

System Equations

By referring back to Fig. 1, in which the transforms of the frequency deviations are designated at various parts of the system, and recalling the definition of transfer function, the following algebraic equations may be formulated:

$$\phi_m(p) = \phi_i(p) - \phi_{L.O.}(p)$$

$$\phi_0(p) = \phi_m(p)e^{-p\tau}$$

$$\phi_0(p) = [\phi_i(p) - \phi_{L.O.}(p)]e^{-p\tau}$$

$$\phi_{L.O.}(p) = \phi_0(p)\frac{k_1k_2}{RCp+1}.$$

Solving for $\phi_0(p)$ and writing K for k_1k_2,

$$\phi_0(p) = \phi_i(p)\left[\frac{e^{-p\tau}}{1+\frac{Ke^{-p\tau}}{RCp+1}}\right]. \tag{1}$$

Equation (1) relates the output frequency deviation of the system to any input frequency disturbance. To obtain an equivalent equation in the real time domain, some functional form for $\phi_i(p)$ is specified, and $\phi_0(p)$ may then be transformed back into the time domain by means of the inverse transform.[4] Thus, if at $t=0$ a step-input frequency variation of M cps is assumed, $\phi_0(p)$ becomes M/p and $\phi_0(t)$ may be written as

$$\phi_0(t) = \frac{1}{2\pi j}\int_{\alpha_1-j\beta_1}^{\alpha_1+j\beta_1}\frac{M}{p}\left(\frac{e^{-p\tau}}{1+\frac{Ke^{-p\tau}}{RCp+1}}\right)e^{pt}dp, \quad \beta_1\to\infty \tag{2}$$

where the path of integration includes all poles in the integrand.

Establishing the Transient Response Conditions

It is convenient to write (2) as

$$\phi_0(t) = \frac{1}{2\pi j}\int_{\alpha_1-j\beta_1}^{\alpha_1+j\beta_1}\frac{M(RCp+1)e^{p(t-\tau)}}{p(RCp+1+Ke^{-p\tau})}dp. \quad \beta_1\to\infty \tag{3}$$

The form of the transient response may be predicted from a knowledge of the roots of the denominator in (3). To find these roots let

$$p(RCp+1+Ke^{-p\tau}) = 0.$$

One root equals zero. The other roots must satisfy the following:

$$RCp + 1 + Ke^{-p\tau} = 0. \tag{4}$$

In general, p will be a complex number involving a real and an imaginary part. Positive real roots indicate an unstable response; negative real roots, a stable response degenerating as time increases. An imaginary root implies an oscillating response. If a nonoscillatory stable response is desired, it is necessary to establish circuit conditions which will either reduce the imaginary part to zero, or convert it to a real number. Let

$$p = \alpha + j\beta.$$

Substitute into (4):

$$RC\alpha + jRC\beta + 1 + Ke^{-\alpha\tau}(\cos\beta\tau - j\sin\beta\tau) = 0.$$

Separate real and imaginary terms, thus:

$$\cos\beta\tau = \frac{-(RC\alpha+1)}{Ke^{-\alpha\tau}} \tag{5}$$

and

$$\sin\beta\tau = \frac{RC\beta}{Ke^{-\alpha\tau}}. \tag{6}$$

Square, add (5) and (6), and solve for β, thus:

$$\beta = \sqrt{\frac{K^2e^{-2\alpha\tau} - (RC\alpha+1)^2}{(RC)^2}}. \tag{7}$$

To eliminate the imaginary roots of (4), β itself must be imaginary. Therefore, to yield a nonoscillatory response, the system parameters must satisfy the following inequality:

$$Ke^{-\alpha\tau} \leqq (RC\alpha + 1). \tag{8}$$

In principle, α can be determined by graphical solution of the above transcendental equations. However, such procedure is extremely tedious and time-consuming. Usually it is sufficient to solve the system for the critically damped condition for which β equals zero.[5] However, it may sometimes be desirable to operate the system with a very low-frequency, oscillating damped response, or sometimes with a slightly over-damped response. The former condition corresponds to a small positive number for β, the latter condition to a small imaginary number for β.

From (5) and (6),

$$\frac{\tan\beta\tau}{\beta\tau} = \frac{-RC}{(RC\alpha+1)\tau}. \tag{9}$$

In practice, τ will usually be much less than unity, and $\tan\beta\tau/\beta\tau$ may be replaced approximately by unity, and (9) solved for α if β is limited to small values. Thus

$$\alpha = -\left(\frac{\tau+RC}{RC\tau}\right). \tag{10}$$

[4] The inverse transform $F(t)$, of $f(p)$ is defined as

$$F(t) = \frac{1}{2\pi i}\int_{\alpha_1-i\beta_1}^{\alpha_1+i\beta_1}f(p)e^{pt}dt.$$

Alternatively, a table of transform pairs which relates $f(p)$ with its corresponding $F(t)$, may sometimes be employed.

[5] The term "critically damped" is used here to indicate the transition from an aperiodic response to an oscillatory response.

Equation (10) indicates that near critical damping, the roots of (4) will be real and negative in character, and hence the system response to a step-input disturbance will be stable. From (8),

$$K\exp\frac{\tau+RC}{RC} \leqq \frac{\tau+RC}{\tau} - 1.$$

Usually RC will be much greater than τ so that the non-oscillatory transient response condition may finally be formulated as[6]

$$Ke \leqq \frac{RC}{\tau}.$$

By setting $\beta=$ to zero in (7), the condition for the critically damped response may be established as

$$RC = \tau K\exp\frac{\tau+RC}{RC}. \tag{11}$$

CALCULATION OF $\phi_0(t)$

The steady-state frequency deviation resulting from a step-input frequency disturbance of M cps may be determined by evaluating the integral of (3) at the simple pole $p=0$. Let $\rho_1, \rho_2, \rho_3 \cdots \rho_m$ equal the roots of the quantity $(RCp+1+Ke^{-p\tau})$. Then, by the method of residues, $\phi_0(t)$ can be represented symbolically as

$$\phi_0(t)=\frac{M}{1+K}+M\sum_1^m \left.\frac{(RCp+1)e^{p(t-\tau)}}{\frac{d}{dp}(RCp+1+Ke^{-p\tau})}\right]_{p=\rho_1,\rho_2\cdots\rho_m}.$$

Since for a stable response the roots cannot assume positive real values, the steady-state deviation may be determined by allowing t to approach infinity.[7] Thus

$$\phi_0(t)_{t\to\infty} = \frac{M}{1+K}.$$

Imposing the restriction that the steady-state deviation must be less than half the bandwidth of the filter yields the following condition on K:

$$\frac{M}{1+K} \leqq \frac{BW}{2},$$

where $BW=$ the bandwidth at half-power points. Since it has been assumed that $BW=1/\tau$, this inequality will be satisfied if

$$1+K > 2\tau M. \tag{12}$$

The complete time response, including both the steady-state and transient deviations, may be determined by evaluating the integral of (2) at all its poles. This evaluation is most readily performed by first expanding the integrand into a power series.[8] Thus

$$\phi_0(t) = \frac{M}{2\pi i}\int_{\alpha_1-j\beta_1}^{\alpha_1+j\beta_1}\left[\frac{e^{-p\tau}}{p} - \frac{K}{RC}\frac{e^{-2p\tau}}{p\left(p+\frac{1}{RC}\right)} + \frac{K^2}{RC^2}\frac{e^{-3p\tau}}{p\left(p+\frac{1}{RC}\right)^2} - \frac{K^3}{RC^3}\frac{e^{-4p\tau}}{p\left(p+\frac{1}{RC}\right)^3} + \cdots\right]e^{pt}dp. \quad \beta_1\to\infty$$

Integrating each term by the method of residues, and applying the time restrictions involved in the translation theorem of operational calculus,[9]

$$\begin{aligned}\phi_0(t) = M &= 0 \quad \text{for} \quad t<\tau\\
- MK\left[1-e^{-(t-2\tau/RC)}\right] &= 0 \quad \text{for} \quad t<2\tau\\
+\frac{MK^2}{\overline{RC}^2}\left[\overline{RC}^2 - e^{-(t-3\tau/RC)}\left\{RC(t-3\tau)+\overline{RC}^2\right\}\right] &= 0 \quad \text{for} \quad t<3\tau\\
-\frac{MK^3}{\overline{RC}^3}\left[\overline{RC}^3 - e^{-(t-4\tau/RC)}\left\{\frac{RC}{2}(t-4\tau)^2+(t-4\tau)\overline{RC}^2+\overline{RC}^3\right\}\right] &= 0 \quad \text{for} \quad t<4\tau.\\
+\cdots.\end{aligned} \tag{13}$$

The number of terms which must be computed will depend on the rate of convergence of $\phi_0(t)$ to its final steady-state value. Calculations on practical afc systems (where τ is much less than one) indicate that usually only four or five terms will be required for an accurate description of the transient response.

[6] This condition is in agreement with that given by van der Wyck in the paper previously cited in footnote reference 1.

[7] The summation term in the equation for $\phi_0(t)$ is valid only for simple roots. For higher order roots, other techniques must be used to evaluate the residues. However, since only real negative roots are involved, $\phi_0(t)$ will always equal $M/1+K$ as t approaches infinity.

[8] This kind of expansion for evaluating residues is suggested in the following paper: L. A. Pipes, "The analysis of retarded control systems," *Jour. Appl. Phys.*, vol. 18, pp. 617–623; July, 1948. This paper considers servomechanisms with a time lag in the feedback loop.

[9] Churchill, "Modern Operational Mathematics in Engineering," p. 21, McGraw-Hill Book Co., New York, N. Y.; 1944.

Time Response of a Typical System

As a practical illustration of the foregoing principles, consider an afc system having a filter bandwidth of only 10 cycles at the half-power points, and a bandwidth of 200 cycles at the 60 db points. Assume that proper adjustment of the limiter and discriminator circuits will allow the system to correct for a maximum step-input frequency disturbance of 100 cycles. Setting $\tau = 0.1$ second and $M = 100$ cps, K (as determined from (12)) must equal at least 19 if the steady-state deviation is not to exceed 5 cycles. Let $K = 20$. For the critically damped transient response, the time constant of the integrating circuit must satisfy (11). Graphical solution of this transcendental equation yields a value of 5.539 seconds for RC.

The complete response time of the system may now be calculated from the infinite series solution of $\phi_0(t)$. The solid line curve, designated as $K = 20$ in Fig. 2, represents the critically damped response; the dashed line, the step-input disturbance. It is interesting to note that a time lag of τ seconds occurs before the output frequency responds to the input disturbance, and that another time lag of τ seconds must occur before correction of the output frequency is experienced.

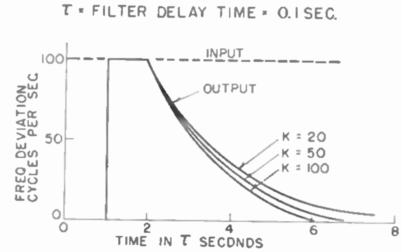


Fig. 2—The critically damped transient response to a step-input frequency disturbance for three values of "loop gain."

Conclusion

A method of calculating the transient response of a narrow-band afc system has been presented. Although the analysis was based on a step-input disturbance, any other kind of disturbance may be inserted into (2), and the resulting output-frequency deviation computed from a new series solution for $\phi_0(t)$.

Acknowledgments

Thanks are due P. Conley and C. E. Vogeley of the Westinghouse Research Laboratories, the former for helpful discussions during preparation of this paper, and the latter for proofreading the manuscript.

The Folded Fan as a Broad-Band Antenna*

R. L. LINTON, JR.†, MEMBER, IRE

Summary—**High-frequency model studies are reported which promise an antenna which operates over a 4-to-1 frequency range with a maximum power loss, due to mismatch, of 18.5 per cent. In the projected naval application, this loss is not unusual with currently used antenna systems. The antenna consists of an ordinary fan-type antenna folded back to ground, that is, electrically connected at the top to an identical fan grounded at its base.**

A means is described for matching the antenna to 52-ohm transmission line, although the folded structures center at about 160 ohms. The model described in detail is not necessarily of absolutely optimum proportions.

The maximum variation with azimuth of the radiation intensity of the antenna is less than 10 db. In the vertical plane, the antenna illuminates the horizon adequately, wasting no appreciable portion of the energy on the zenith.

Introduction

THE ANTENNA development described here was undertaken in an effort to solve partially the antenna problem aboard naval ships. In the past, the decks and superstructures of such vessels have become increasingly cluttered in a most confusing manner, with antennas of all types and sizes. In general, approximately one antenna is installed for each equipment: transmitter, receiver, or transceiver. As the most frequently used antennas in the communication band often have very poor impedance characteristics with several sharp resonances within the band, and deficient radiation patterns, often with severe nulls in the azimuth coverage, communication efficiency has not always been up to a practical minimum.

One mode of resolution of these difficulties proposes the wide use of multiplexing techniques. Each antenna would serve several units, either transmitters or receivers. We will concern ourselves with the transmitting problem. Multiplexing devices for the transmitting case generally require a broad-band antenna for a load.

It is further desired to use coaxial cable in place of the more bulky and bulkhead-weakening trunk, which leads from the transmitter to the antenna. The high standing-wave ratios and consequent power loss, or voltage breakdown and dielectric failure that would result with present antenna systems, prohibit the use of coaxial cable.

It was proposed that a broad-band type radiating structure to be fed by 52-ohm coaxial cable be developed for use in the 2- to 27-mc communication band.

* Decimal classification: R326.61. Original manuscript received by the Institute, August 8, 1950; revised manuscript received, February 1, 1951.

The work described here was done at the Antenna Laboratory, University of California, Berkeley, Calif., and was made possible through the support extended by the U. S. Navy Department, Bureau of Ships, under Contract NObsr-39401.

† Formerly, Antenna Laboratory, University of California, Berkeley, Calif.; now, Consolidated-Vultee Aircraft Corp., San Diego, Calif.

Model Studies

Impedance Measurements

A study of the impedance characteristics of various folded structures was carried out using ultra-high-frequency models. A scaling factor of 50 was used for much of the work, making the band of interest spread from 100 to 1,350 mc. Where other scaling factors were used, the data have been translated into terms of this frequency band.

The measurement equipment consisted of the following. A 20-foot-square simulated ground plane situated on the side of a building was used in conjunction with a Chipman measuring line.[1] The antenna models were excited by a small loop projecting above the surface of the ground plane and fed by a well-shielded transmitter. At the lower end of the frequency band, the General Radio model 757-A signal generator was used. At the high end, a local oscillator from an AN/SPR-2 receiver was used. The signal picked up by the sampling loop at the movable short circuit of the Chipman line was compared with the signal supplied from the Navy Type LAE-2 or LAF-3, and their associated direct-reading attenuators. The standing-wave ratio could be obtained directly in decibels from these attenuators, or the steepness of the current maxima could be obtained and translated to radius on the Smith chart. The AN/APR-1 receiver and its several radio-frequency heads were used in conjunction with an external microammeter. Except at the very lowest frequencies, where the receiver calibration was used, the frequencies were determined with the use of a series of transmission type cavity wavemeters.

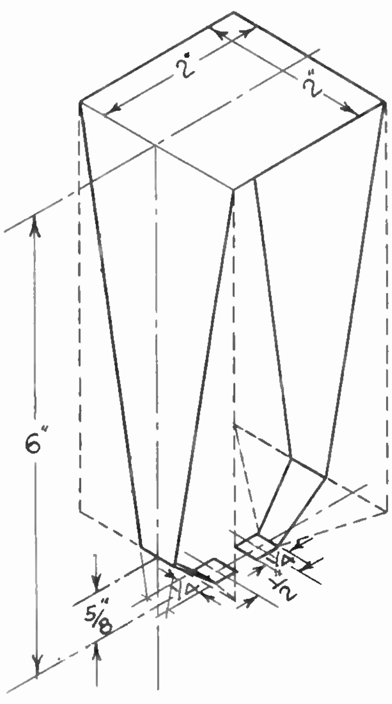


Fig. 1—Fat folded monopole.

The final design of antenna was not to be too large for installation on naval ships. A maximum height of 25 feet was selected as a nominal value to shoot for. This scaled to 6 inches. An early model investigated in these studies is illustrated in Fig. 1. One tab at the base was screwed to the inner conductor of the Chipman line; the other was soldered to the ground plane. It was desired to make the basic elements of the folded structure as broad-band as possible by themselves. The first model measured conformed to the dotted outline shown. Curve *B* in Fig. 3(a) represents the impedance characteristic of this model. The improvement noted in curve *A* was effected after trimming the model to conform to the solid lines in Fig. 1. Two effects are apparent here. First, trimming away the sides of the structure reduced the effective base capacitance, thus allowing the entire impedance characteristic to shift toward the inductive side of the chart. Second, the losses through radiation in the transmission-line mode were increased through a reduction in self-shielding, and the fatness in the monopole mode was reduced, bringing the Q's of the two modes more nearly into balance.

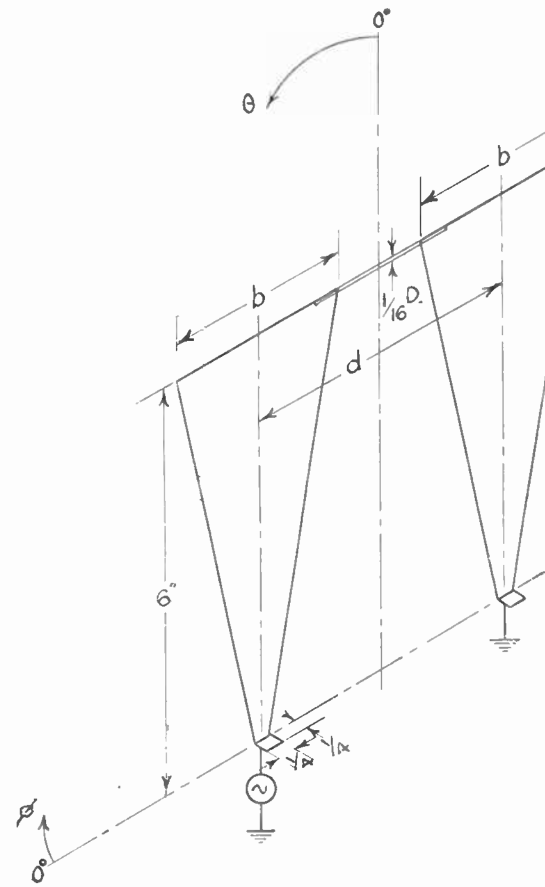


Fig. 2—Folded-fan antenna.

The effects of this trimming operation gave the clue to the folded-fan construction used in the remainder of the studies. In this device (see Fig. 2) the transmission

[1] J. T. Bolljahn: "Chipman Line Analysis and Development," Technical Report 132, Antenna Laboratory, University of California, Berkeley, Calif.; October 31, 1947.

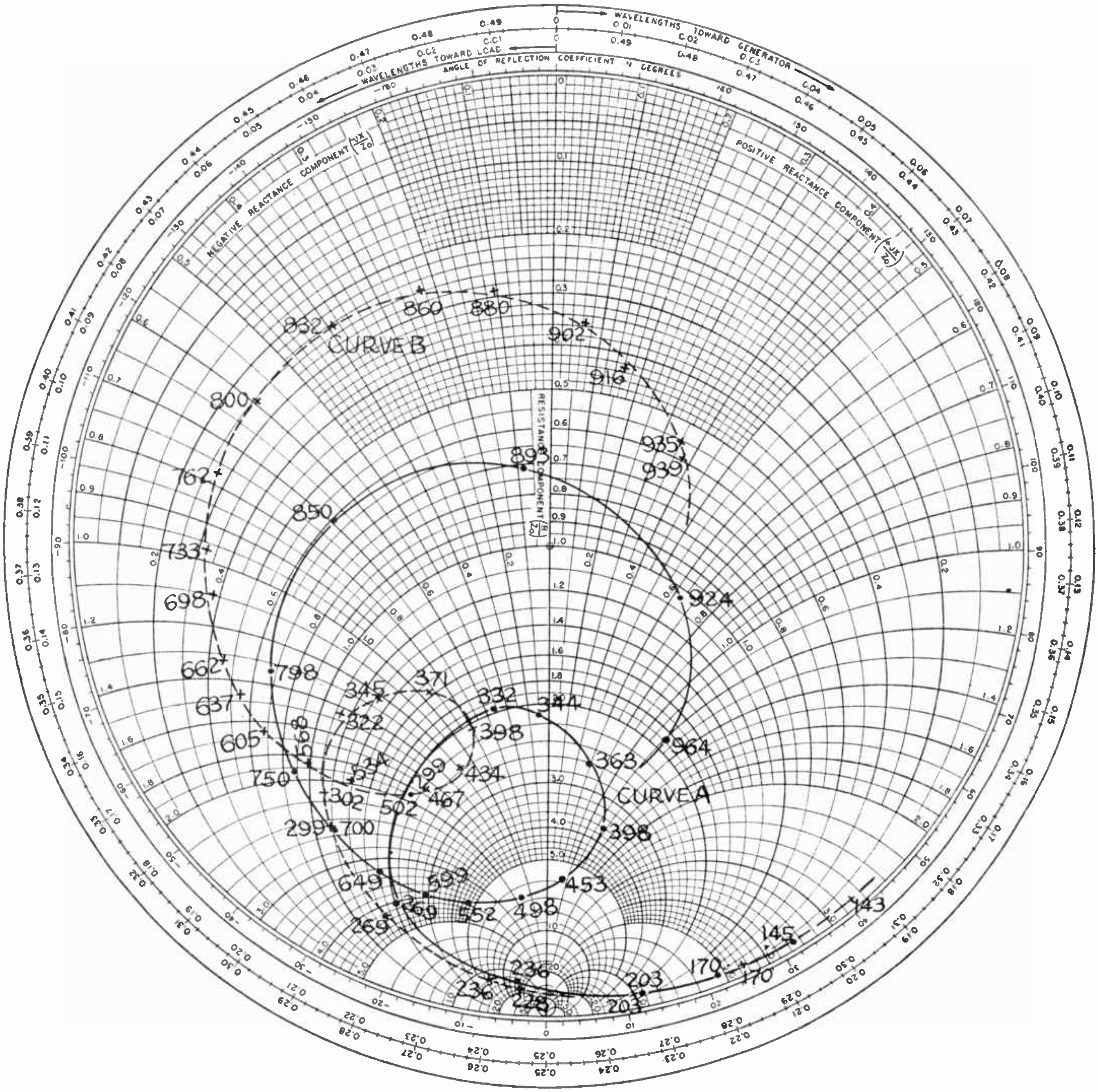


Fig. 3(a)—Smith chart plots of impedance referred to 52 ohms versus frequency in mc. (Chart from Philip H. Smith, "Transmission line calculator," *Electronics*, vol. 12, January, 1939; also vol. 17, p. 130, January, 1944.)

line has been opened up to the maximum extent possible, the entire structure lying in a plane. At the same time, the monopole is made fat by the use of the fan structure. A systematic survey of the impedance characteristics of this antenna for a wide array of different proportions was conducted. The dimensions d and b are identified in Fig. 2. Dimension d was varied from 2 to 6 inches, and b, from 1 to 5 inches. Some of the results are presented in Figs. 3(b) and 3(c). Fig. 3(b) shows the effect on the impedance locus of pulling two identical fans apart. Fig. 3(c) shows the effect of broadening the fans held at a constant distance on centers. In the latter, two isofrequency lines are drawn to show the manner in which the loci draw in toward a restricted region of the chart as the fans are enlarged.

Transmission-Line Matching

The impedance characteristic of Figs. 3(b) and 3(c) are compact but at too high an impedance level to be fed by 52-ohm cable. It was desired to match a representative model to 52 ohms by some convenient, practical method. The model of curve C, Fig. 3(c), was se-

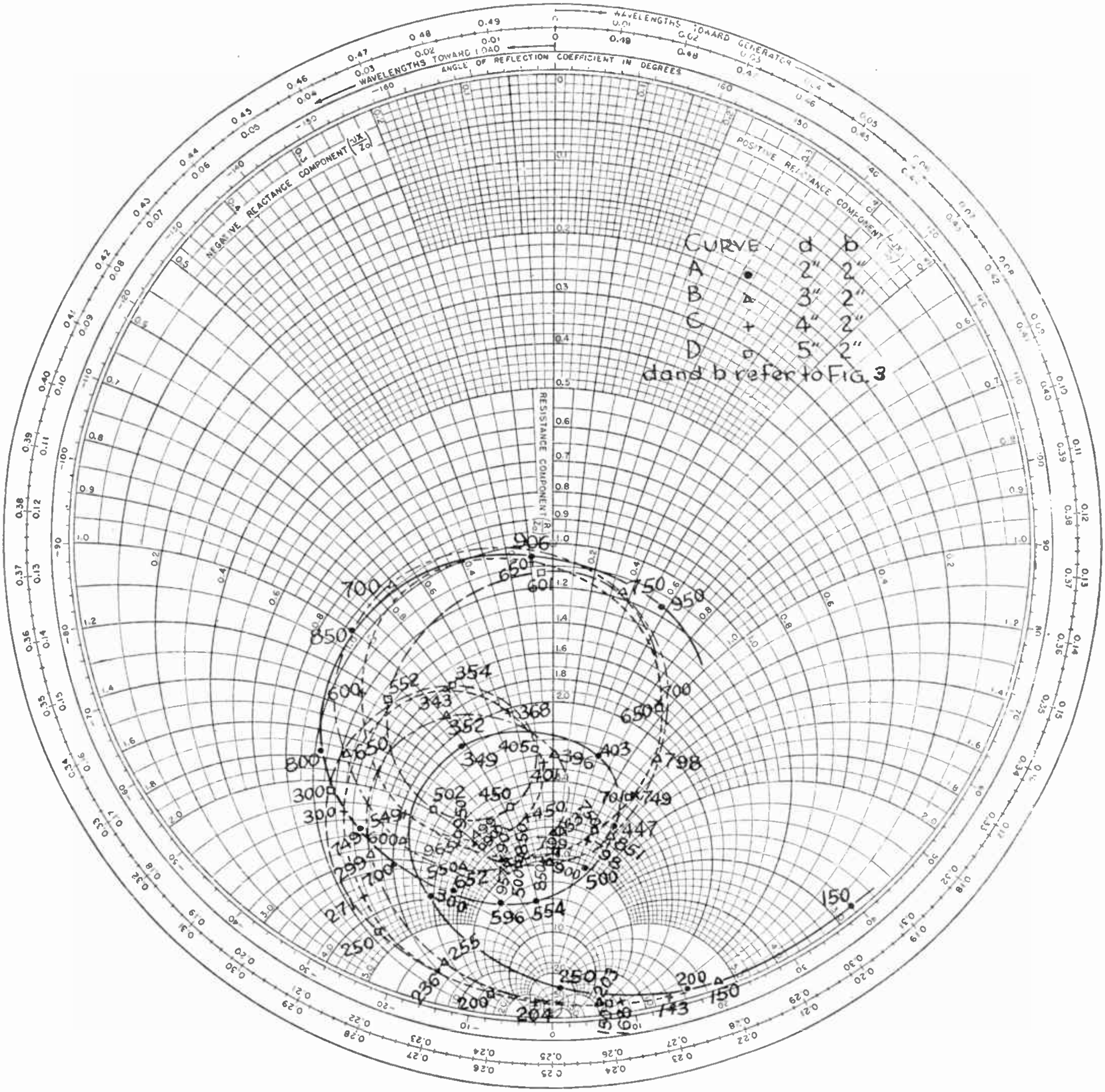


Fig. 3(b)—Smith chart plots of impedance referred to 52 ohms versus frequency in mc. B of Fig. 2 equal to 2 inches.

lected as having good balance of monopole and transmission-line Q's, and as having nominal dimensions. By a method already described[2] three sections of transmission line were selected to simulate an exponential taper. Values of characteristic impedance were selected which were available commercially, should a full-scale test be desired at a future time. The results of the analytical transformation were checked experimentally by means of an extention to the Chipman line.

[2] R. L. Linton, Jr., "Design charts for transmission line matching systems," *Tele-Tech.*, vol. 9, p. 19; January, 1950.

The inner conductor was stepped to produce the 125-ohm and the 93-ohm sections. The line itself has an impedance of 67.5 ohms. The results were transformed analytically through a portion of the line itself and then referred to 52 ohms. The data in final form appear in Fig. 3(d). To obtain information in the higher end of the frequency range of interest, a smaller scaling factor was used (35), but the results are referred to the 6-inch height for uniformity of presentation.

It is noted that the VSWR of this antenna and matching system is within 2.5 from about 285 to about

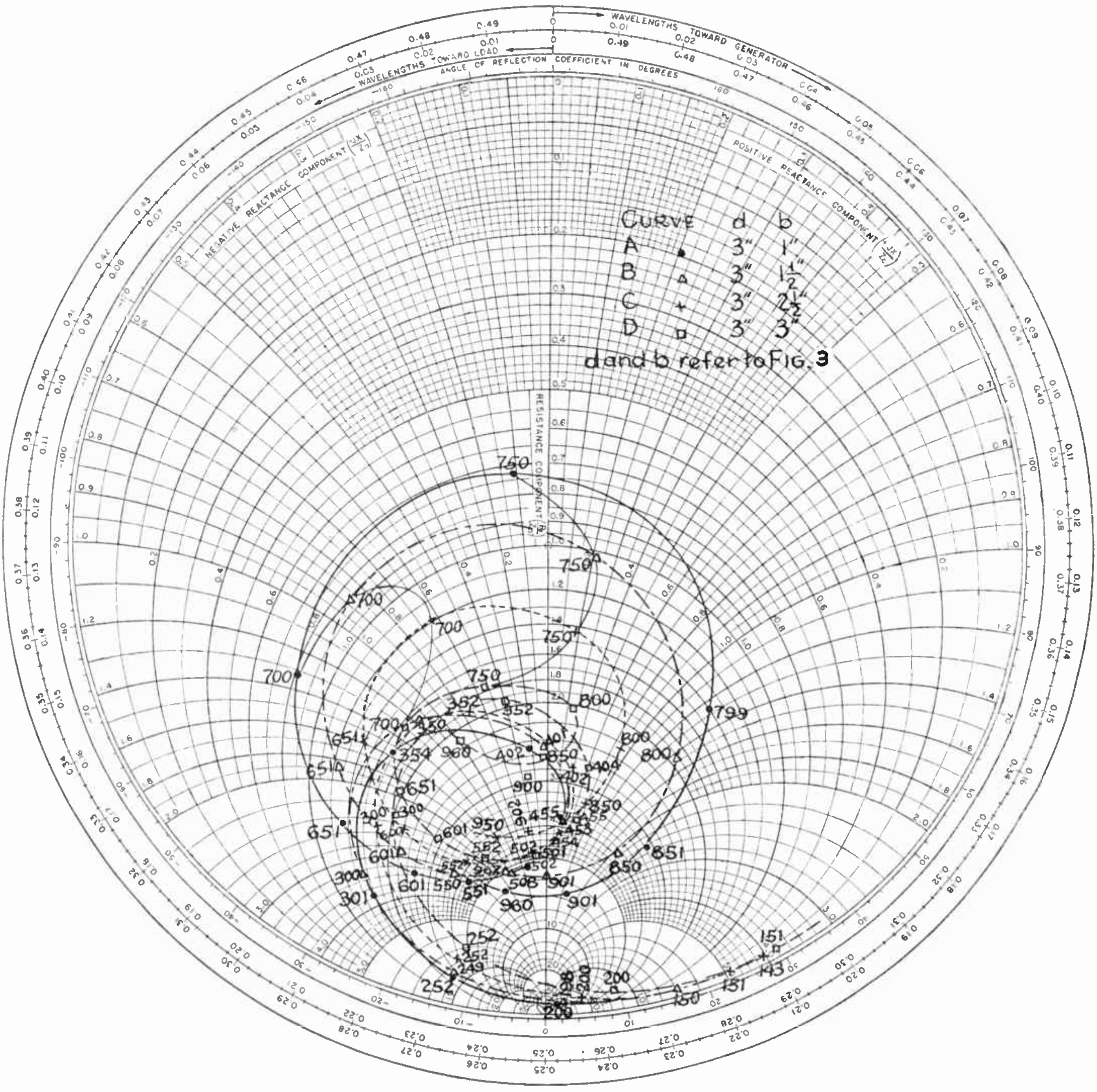


Fig. 3(c)—Smith chart plots of impedance referred to 52 ohms versus frequency mc. D of Fig. 2 equal to 3 inches.

1,160 mc, and within 4 from 265 to probably at least 1,350 mc.

Radiation Patterns

The radiation patterns of the model antenna in the horizontal and vertical planes were checked for acceptability for the projected application. For this purpose, a horizontal ground plane about 20 feet square with a turntable in the center was used. A lighthouse tube transmitter fed a one-inch stub antenna at the "focus" of a corner reflector which illuminated the ground plane. For the vertical pattern measurements, the model antenna and the corner reflector antenna were removed from the ground plane by about 10 feet. A Hoffman pattern recorder was used in conjunction with the AN/APR-1 receiver. The conventions for azimuth ϕ and zenith angle θ are indicated in Fig. 2. The patterns have been transcribed from the Hoffman pattern records and normalized to the outer edge of the charts.

Horizontal Patterns: The horizontal patterns are presented in Fig. 4. The ratio of maximum to minimum intensity throughout the useful band is well below 10 db.

As a rough check on the efficiency of the folded fan, the level of the pattern was compared, by direct substi-

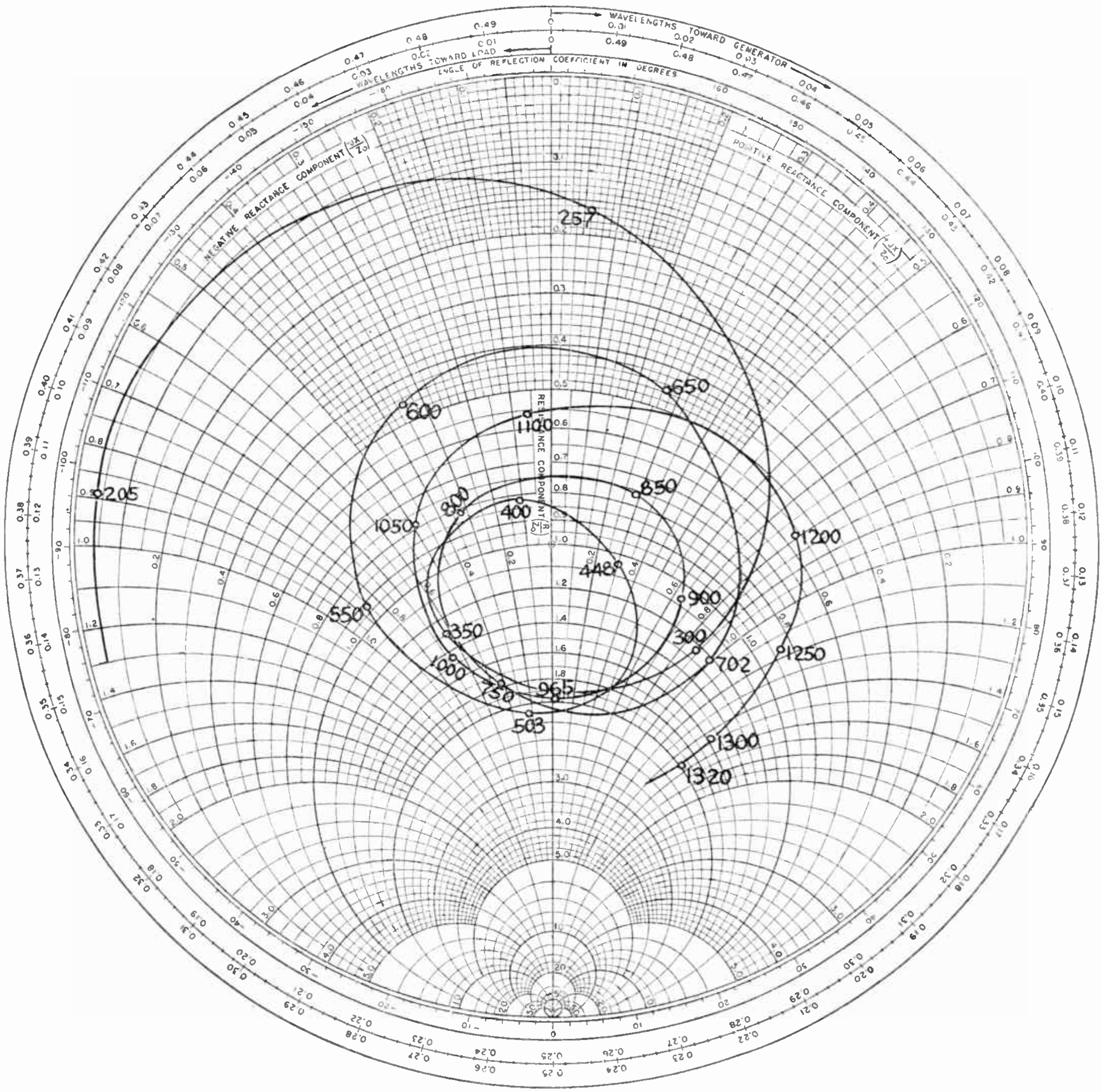


Fig. 3(d)—Smith chart plot of impedance referred to 52 ohms versus frequency in mc. Impedance transformed electrically through stepped matching transmission line. Characteristic impedance respectively 125 ohms, 93 ohms, and 67.5 ohms. Model 8 9/16 inches high. Frequencies for equivalent 6-inch antenna given.

tution, with that of a 1/16-inch diameter monopole of the same height at several frequencies. It is noted that the efficiencies of the two antennas are not radically different throughout the frequency range.

Vertical Patterns: For taking vertical radiation patterns, a balanced folded-fan dipole was constructed of the same proportions as those of the model previously selected. The balanced model was scaled to a factor of 74, but the frequencies given in Fig. 5 correspond to equivalent frequencies for the 6-inch unbalanced model. The model was mounted on a vertical phenolic rod and wooden tower. Patterns are given for three frequencies near the top of the band of interest in two vertical planes: the plane of the antenna and that perpendicular to it.

360-degree patterns are given so that the degree of symmetry displayed will indicate how well balanced the model was at the time the pattern was taken. The balance was adjusted by positioning the shorting bar of a balun incorporated at the feed of the balanced

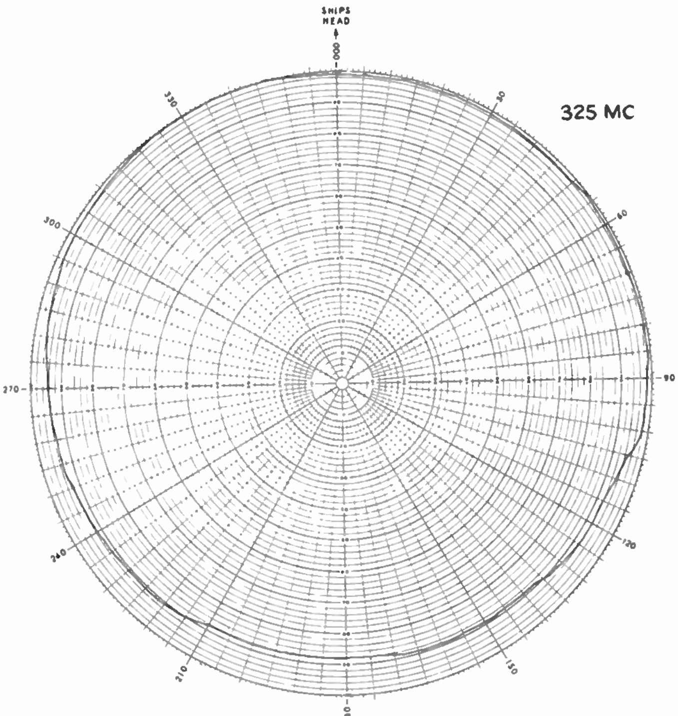


Fig. 4(a)—Horizontal radiation pattern of folded fan at 325 mc.

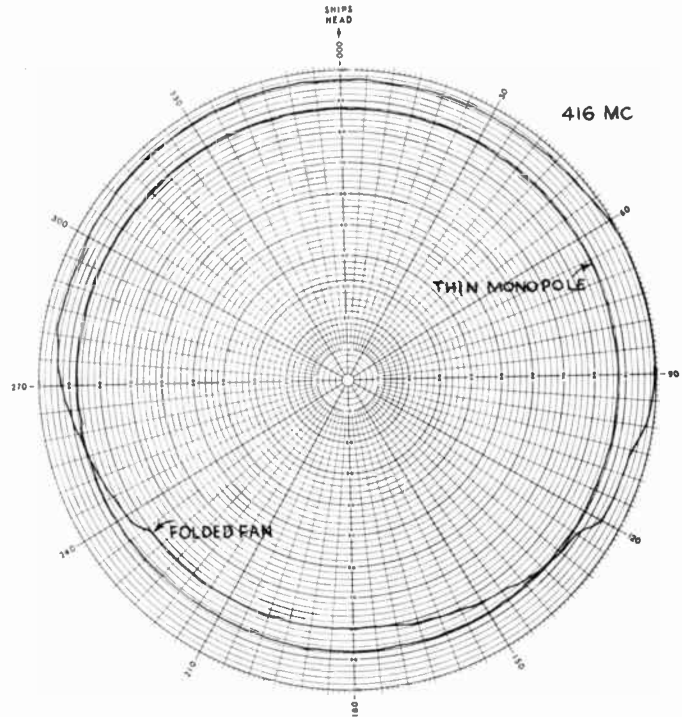


Fig. 4(b)—Horizontal radiation patterns of folded fan and comparison monopole at 416 mc.

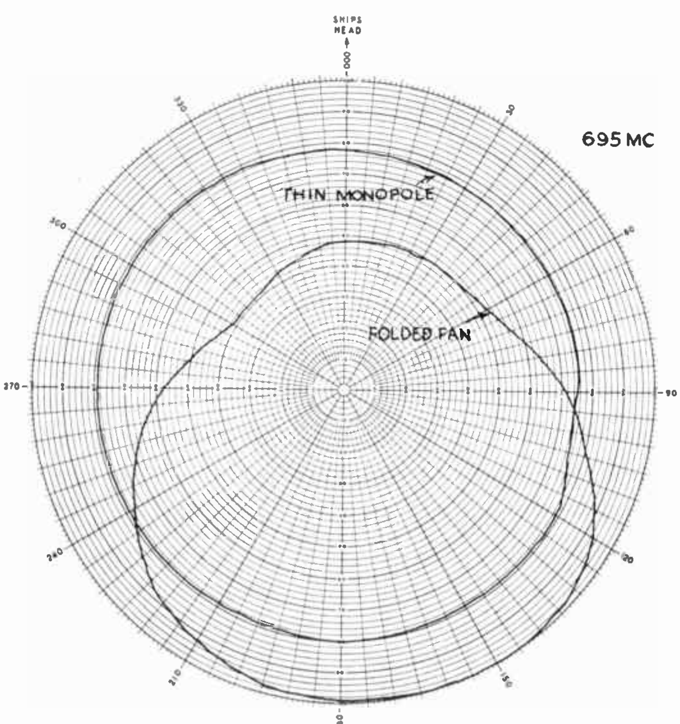


Fig. 4(c)—Horizontal radiation patterns of folded fan and comparison monopole at 695 mc.

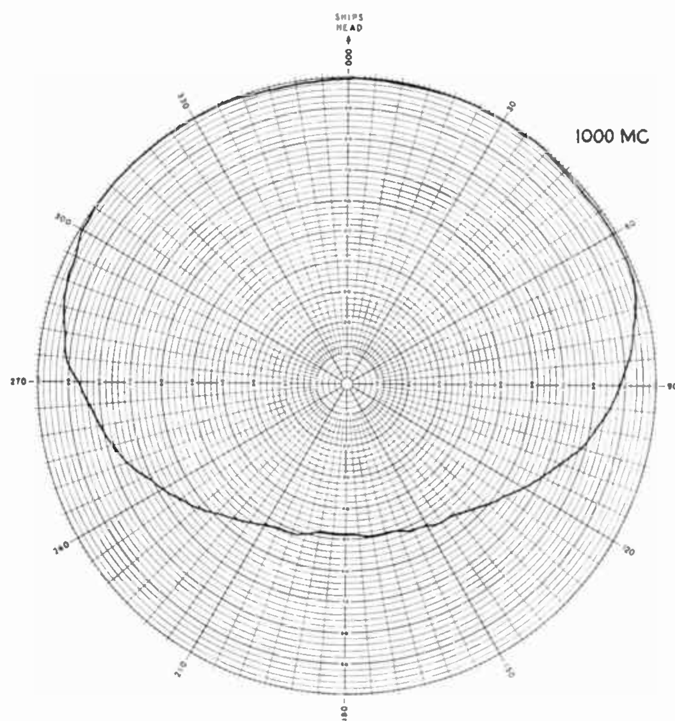


Fig. 4(d)—Horizontal radiation pattern at folded fan at 1,000 mc.

model. In all cases, the horizon receives a substantial illumination, and no major portion of the energy is diverted toward the zenith. In the figures, the zenith angle is shifted slightly from that indicated by the axis of symmetry due to inadvertent error in placing the graph sheets in the recorder. Time was not available for taking patterns at lower simulated frequencies; a still higher scaling factor would be needed to avoid the effects of reflections from the ground plane.

Conclusions

The results given here indicate further investigations that might prove fruitful. The impedance and radiation characteristics described should be checked in a full-scale model. Further model studies should be conducted to determine the feasibility of substituting available ship structures for the grounded leg of the folded fan. Many naval vessels possess structures of sufficient height. Where structures are larger than 25 feet, the

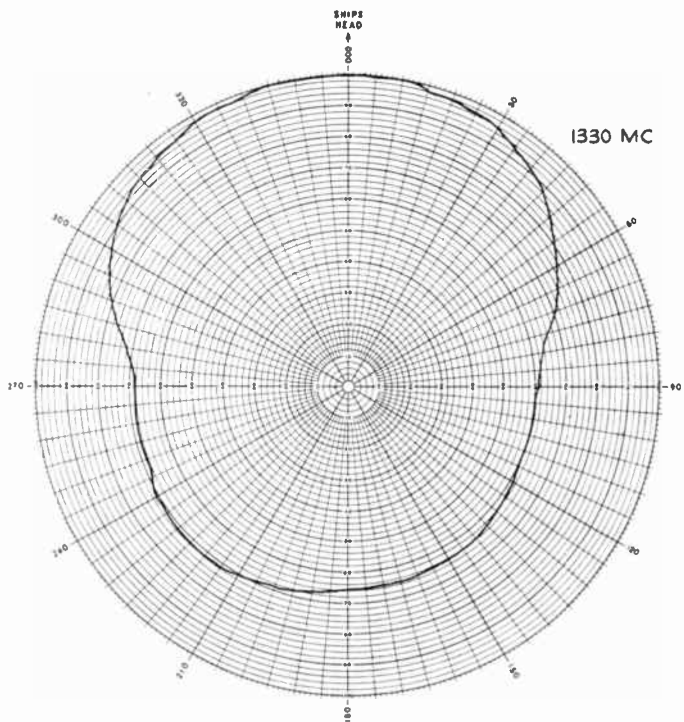


Fig. 4(e)—Horizontal radiation pattern of folded fan at 1,330 mc.

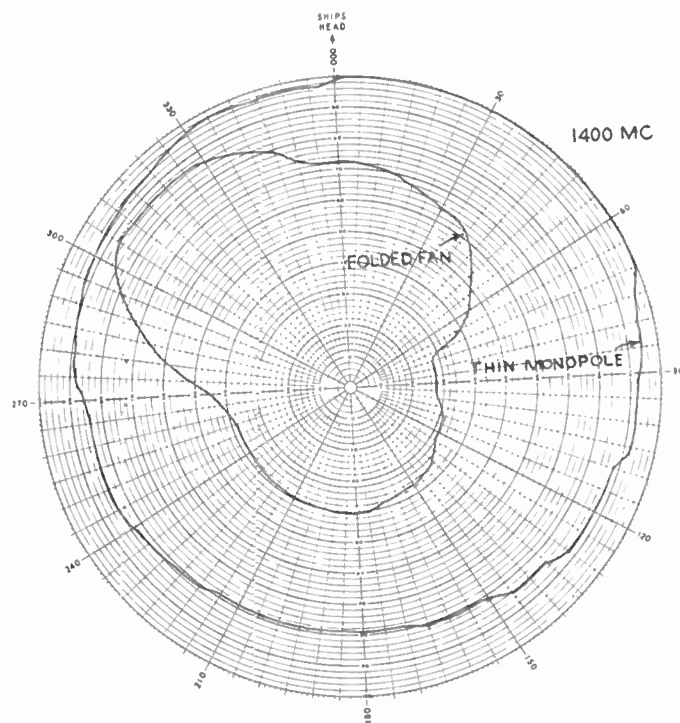


Fig. 4(f)—Horizontal radiation patterns of folded fan and comparison monopole at 1,400 mc.

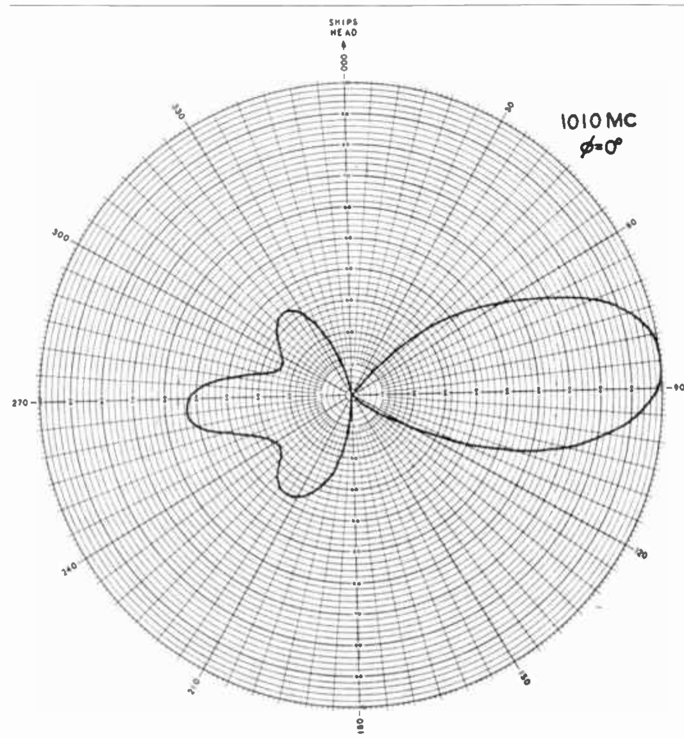


Fig. 5(a)—Vertical radiaton pattern of folded fan at equivalent frequency of 1,010 mc. Azimuth 0°.

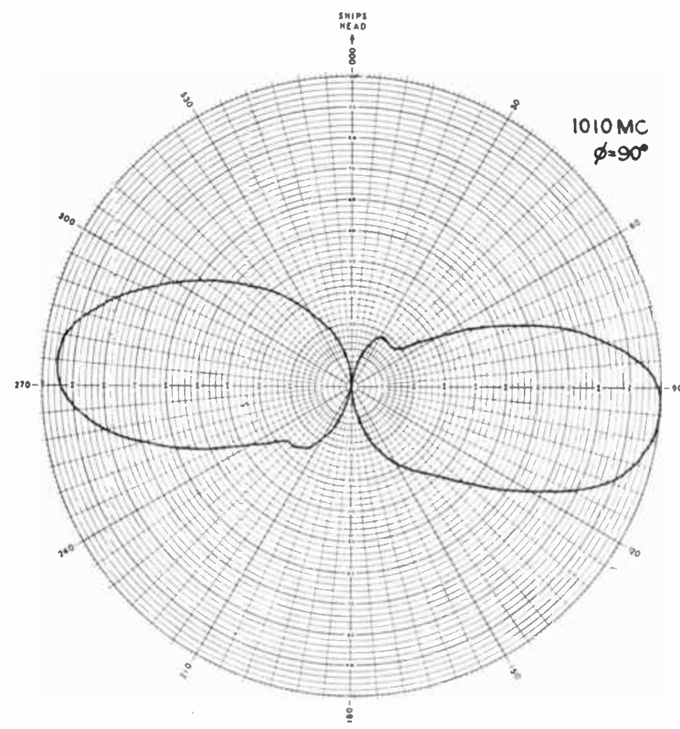


Fig. 5(b)—Vertical radiation pattern of folded fan at equivalent frequency of 1,010 mc. Azimuth 90°.

4-to-1 or 5-to-1 frequency range may be shifted to include the lower communication frequencies. For example, the stack of the *Iowa* class is three times as high as the full-scale antenna assumed in these studies. A folded fan of this height would have a 4-to-1 frequency range between 1.9 and 7.7 mc.

The present antenna seems to have possible application where a broad-band antenna is needed, either in the original projected application, or in other communication applications. At full-scale frequencies, the antenna would have a VSWR within 2.5 from 5.7 to 23.2 mc. Such a mismatch involves a power loss of only 18.5 per cent, which is not a serious loss compared to current naval practice in many instances. From 5.3 to 27 mc, the VSWR would be within 4, a power loss of 36 per cent. The matching cable needed to achieve these results

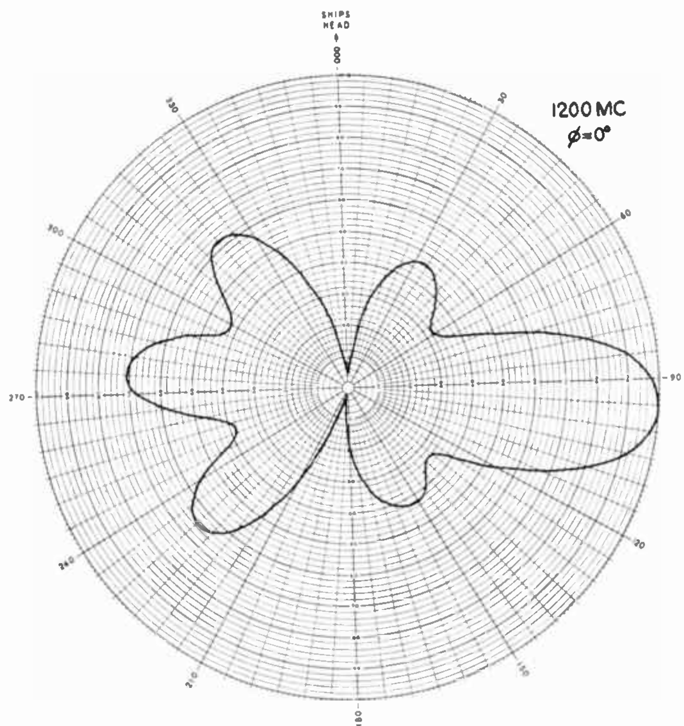


Fig. 5(c)—Vertical radiation pattern of folded fan at equivalent frequency of 1,200 mc. Azimuth 0°.

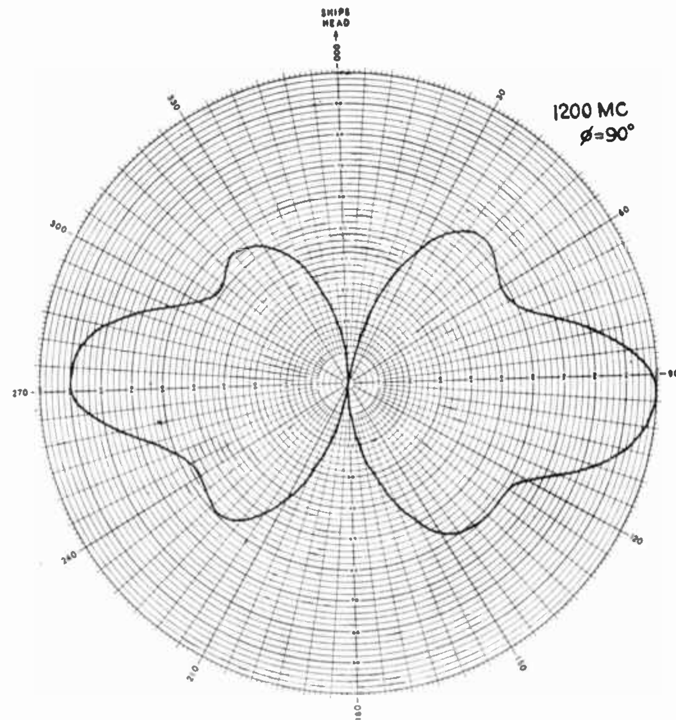


Fig. 5(d)—Vertical radiation pattern of folded fan at equivalent frequency of 1,200 mc. Azimuth 90°.

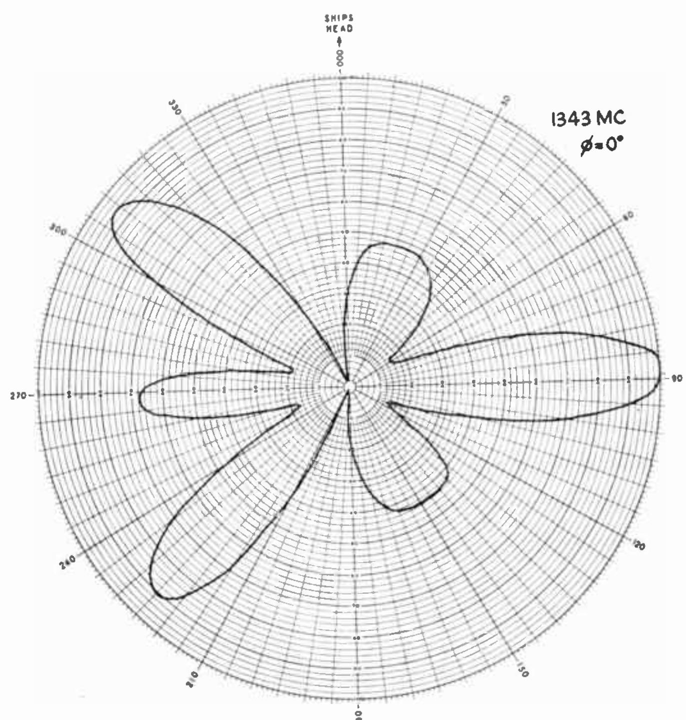


Fig. 5(e)—Vertical radiation pattern of folded fan at equivalent frequency of 1,343 mc. Azimuth 0°.

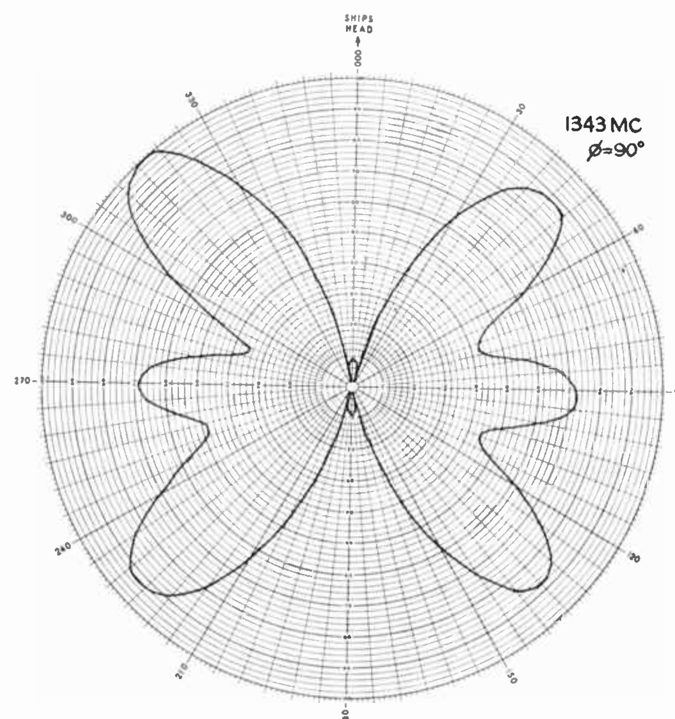


Fig. 5(f)—Vertical radiation pattern of folded fan at equivalent frequency of 1,343 mc. Azimuth 90°.

would be only 49.2 feet long, assuming air dielectric. The antenna would be 25 feet high, 23 feet in horizontal length, and of negligible width in the other dimension. The fans should be simulated with three or four wires. It should be pointed out that the particular proportions selected for detailed description are not necessarily optimum, and that even better results than those obtained may be possible with the folded-fan construction.

The balanced folded-fan dipole at ultra-high frequencies may have promise as a combined FM and television-receiving antenna.

Acknowledgment

The author acknowledges deep gratitude to Robert deLiban who directly supervised the work reported, and offered many helpful suggestions and much encouragement.

The Use of Complementary Slots in Aircraft Antenna Impedance Measurements*

J. T. BOLLJAHN†, ASSOCIATE, IRE, AND J. V. N. GRANGER†, SENIOR MEMBER, IRE

Summary—This paper describes a method for eliminating the feed-cable effect in the measurement of aircraft wing-cap and tail-cap antenna impedances with the aid of models. The procedure employed allows greater accuracy in measurement than that obtainable with conventional techniques, but its application is restricted to use use on simplified models. The advantages and limitations of the method are discussed, and typical experimental results are presented.

I. Introduction

Modeling techniques have been used successfully for a number of years in the measurement of antenna characteristics. These techniques are particularly well suited for measurements on antennas which are mounted on complex structures such as ships or aircraft, where adequate full-scale measurements are costly and difficult to accomplish.

Although the principles of modeling apply to both radiation patterns and impedance characteristics, their use in impedance measurements has been limited considerably by practical difficulties. In measuring the impedance of a model aircraft antenna in or below the frequency range where strong resonances may occur in the aircraft structure, for example, one must either use a model large enough to contain the measuring equipment, or make the measurements through a length of cable. The first alternative is impractical in most cases because of the model sizes required, while the second alternative leads to errors due both to the inaccuracy involved in transforming the measured impedance through a length of cable to determine the terminal impedance, and to the perturbation of the fields outside the model caused by the presence of the cable.

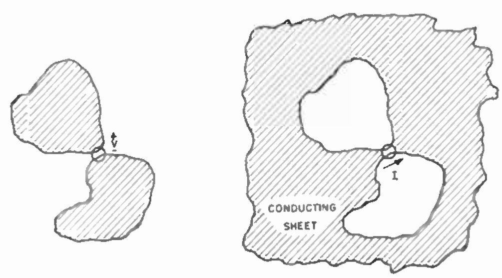
Measurements in this frequency range are of interest at present in connection with studies of wing-cap and tail-cap antennas. Since these antennas are formed by separating and insulating extremities of the aircraft from the main structure, it is seen that a systematic study of their impedance properties on full-scale aircraft would be completely impractical, and a consideration of new modeling techniques is in order.

This report describes a technique suitable for such work which employs a simplified aircraft model consisting of two or more strip conductors lying in a plane and arranged to simulate as closely as possible the shape of the aircraft. Impedance measurements are actually made on a slot which is complementary to this system of plane conductors and which is cut in a large conducting sheet. The impedance of a required configuration in the simplified aircraft model may be calculated from the corresponding measured value for the complementary case with the aid of Babinet's Principle.

Use of the slot analog rather than the model itself makes it possible to embed the feed cable in the ground plane, and thus minimize its effect on the external fields. In some cases the symmetry is such that an image plane may be used, and the feed-cable effect may be eliminated entirely by locating the measuring equipment behind the image plane. The accuracy of results obtained with this technique is limited, of course, by the approximations inherent in the model employed, but the experimental errors are considerably smaller than those which would occur in direct measurements on conventional models. The results of considerable experience lead to the conclusion that the net result of these two conflicting factors favors the slot technique for over-all accuracy.

II. Babinet's Principle

The basis for the slot analog is the electromagnetic Babinet Principle, and in particular the impedance result first pointed out by Booker.[1] This result relates the impedances seen by the two generators in the two problems pictured in Fig. 1. The configuration in the antenna problem consists of an arbitrary arrangement of plane conductors lying in a common plane and driven



(a) (b)

Fig. 1—Complementary antennas.

at one or more points by generators. In the slot problem, a corresponding arrangement of openings has been cut from an infinitely large, plane, perfectly conducting surface, and generators corresponding to those in the antenna problem are present as indicated. Designating the antenna and slot properties by the subscripts a and s, respectively, we may write the expression of interest as follows:

$$Z_a Z_s = \frac{\eta^2}{4} \tag{1}$$

* Decimal classification: R221×R326.21. Original manuscript received by the Institute, May 16, 1950; revised manuscript received, February 24, 1951.

† Stanford Research Institute, Stanford, Calif.

[1] H. G. Booker, "Slot aerials and their relation to complementary wire aerials," *Jour. IEE*, part IIIA, vol. 93, pp. 620 ff.; March, 1946.

where

Z_a = antenna impedance, ohms

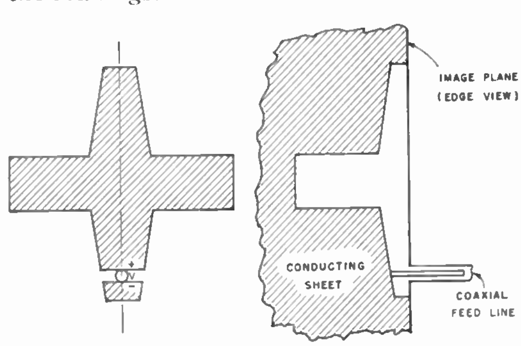
Z_s = slot impedance, ohms

$\eta = \mu_0/\epsilon_0 = 120\pi$ ohms.

The more general statement of Babinet's Principle from which this result is developed relates the electromagnetic fields in the two problems. Specifically, it states that the electric field in the antenna problem is of the same form as the magnetic field in the slot problem, and the magnetic field in the antenna problem is of the same form as the electric field in the slot problem. It is necessary to qualify this statement only to the extent that the field directions on one side of the plane containing the conductors must be reversed in one of the problems. The reason for this reversal may be seen by considering the field discontinuities involved. In either problem there must be discontinuities in the normal component of E and the tangential component of H at the conducting portion of the surface, corresponding to the charge and current distributions on the conductors. The reversal of fields on one side of the sheet in one of the problems is necessary to move the region in which field discontinuities occur from one set of conductors in the antenna problem to the complementary set of conductors in the slot problem.

III. Application to Impedance Measurements

As mentioned above, the chief advantage of using the slot analog in impedance measurements lies in the fact that the feed cable from the measuring equipment may be embedded in the conducting sheet, thus virtually eliminating its effect on the external fields. The inaccuracies associated with measuring through a long length of line are still present, however, since the operator and measuring equipment must be located sufficiently far from the slot that their presence does not affect the readings.



(a) (b)

Fig. 2—Simulated wing-cap antenna.

In systems having image symmetry such as the one shown in Fig. 2(a), this difficulty may be avoided by making measurements on a half-slot and image plane as indicated in Fig. 2(b). In this case, a coaxial slotted line may be connected directly to the desired feed point. If the slotted line has a characteristic impedance Z_0, it is readily shown with the aid of (1) that the normalized admittance values of the half slot are the same as the corresponding impedance values for the antenna problem with the latter normalized to a characteristic impedance $Z_0' = \eta^2/8Z_0$.

It is of interest to note that if the antenna problem involves more than one feed point, and if the coupling between the various feedpoints is represented by constructing a suitable equivalent network, then the dual of the antenna network is an equivalent network for the slot problem, provided that the impedance elements in the two networks are normalized to Z_0' and Z_0, respectively.

IV. Simplified Aircraft Models

Since the fields in the slot problem are accurately related to those in the complementary plane-conductor problem by Babinet's Principle, it follows that the basic approximations involved in the equivalent slot technique are simply those involved in using a system of plane conductors to represent the aircraft structure.

The use of simplified models of planar form is based on the concept of equivalent cross sections in two-dimensional static problems. Consider two infinitely long conductors having different cross-sectional shapes. It is always possible, at least theoretically, to adjust the transverse dimensions of one of the conductors while retaining its cross-sectional shape, so that the two conductors have the same static capacitance per unit length. The two conductors adjusted in this fashion to have the same static capacitance per unit length are said to have equivalent cross sections. It is possible to show[2,3] that two linear antennas will have equivalent impedance characteristics if their cross sections are equivalent in this sense, and if their cross-sectional dimensions are small compared with the wavelength.

The application of equivalent cross sections to the construction of planar aircraft models involves the assumption that the approximations inherent in the equivalent cross-section concept remain valid for more complex structures, even when the cross-sectional dimensions of the elements are appreciable fractions of the wavelength. A discussion of these approximations and an experimental investigation of their validity in the case of aircraft structures are given in the literature.[4]

The use of the planar model is further supported by the agreement found in an experimental comparison of the electromagnetic resonance behavior of planar and conventional models of the same aircraft structure.[5] These results serve to resolve a difficulty which arises in the case of the tail-cap structure, namely that the driven

[2] F. Bloch and M. Hammermesh, "Equivalent Radius of Thin Cylindrical Antennas," Rpt. 411-TM125, Radio Research Laboratory, Harvard University; June, 1944.

[3] C. Flammer, "Equivalent Radii of Thin Cylindrical Antennas with Arbitrary Cross Sections," Tech. Rpt. No. 4, Contract AF 19(122)78. Aircraft Radiation Systems Laboratory, Stanford Research Institute; February, 1950.

[4] J. V. N. Granger and T. Morita, "R-f current distributions on aircraft structures," Proc. I.R.E., to be published.

[5] A. S. Dunbar, "Electromagnetic Resonance Phenomena in Aircraft Structures," Tech. Rpt. No. 8. Contract AF 19(122)78. Aircraft Radiation Systems Laboratory, Stanford Research Institute; May, 1950.

vertical stabilizer does not fit into the planar models. Impedance measurements were made on the slot system complementary to the structure shown in Fig. 3, in which the vertical stabilizer is represented by a coplanar extension of the fuselage beyond the horizontal stabilizer. The resonances apparent in the impedance data shown in this figure correspond to those observed by Dunbar in back-scattering studies of a conventional model of the same airframe. Adoption of the complementary slot model for the tail-cap structure permits the use of image-plane measurements which are not possible when a conventional model is employed. The impedance data of Fig. 3 correspond in their important features to the limited data which are available from measurements on full-scale structures.

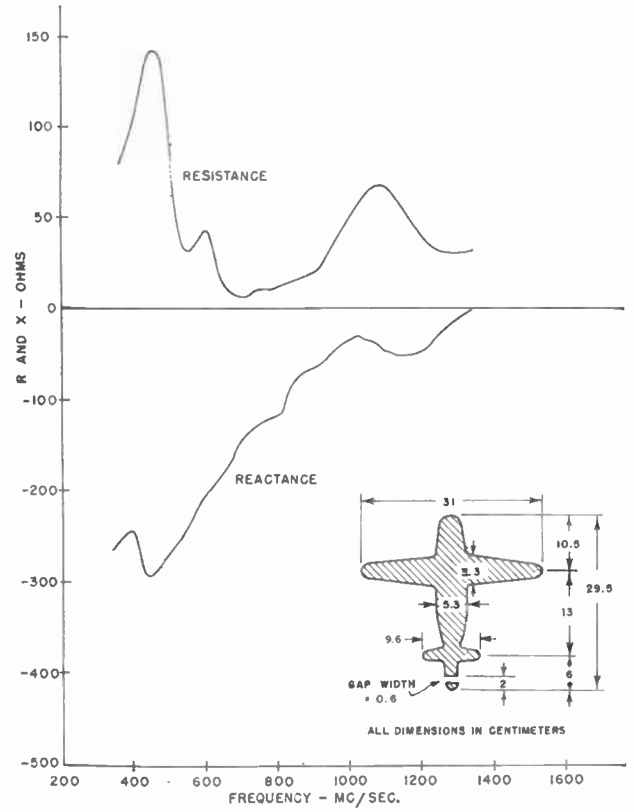


Fig. 3—Measured impedance characteristic of tail-cap antenna on simplified model

Measurements on simulated wing-cap antennas have been made with the slot arrangement shown in Fig. 2. The fore-and-aft symmetry of the slot representing the aircraft fuselage permits use of the image-plane technique which could not be used if a more accurate model were employed. Although this symmetry eliminates certain resonances associated with fuselage currents, measurements on such a model have yielded useful information on the effect of the length of the isolated section and the gap width on the impedance characteristics of the system.[6]

[6] J. V. N. Granger, "Wing-Cap and Tail-Cap Aircraft Antennas," Tech. Rpt. No. 6, Contract AF 19(122)78. Aircraft Radiation Systems Laboratory. Stanford Research Institute; March 1950.

V. The Feed System

It is of interest to consider briefly the details of the feed region in image-plane slot measurements. The inner-conductor extension which spans the slot in the slot problem is seen to correspond to the gap between the main structure and the isolated section in the planar model. If this inner-conductor extension is made of a flat strip lying in the plane containing the slot, its width corresponds directly to the gap width in the planar model. For extensions made with circular sections, the equivalent strip width may be calculated by employing again the theory of equivalent cross sections. The interpretation of the effect of nonplanar feeders in terms of the equivalent gap width is of particular interest in connection with problems which do not have image symmetry and which do not, therefore, permit use of the image-plane technique. In such a problem it would be desirable to provide the feed at the center of the isolating gap by some scheme such as that shown in Fig. 4. The use of equivalent cross-section theory for interpreting the planar-model gap width in terms of the feeder diameter in the slot problem has been verified experimentally for small diameter feeders by comparison checks using image-plane measurements.

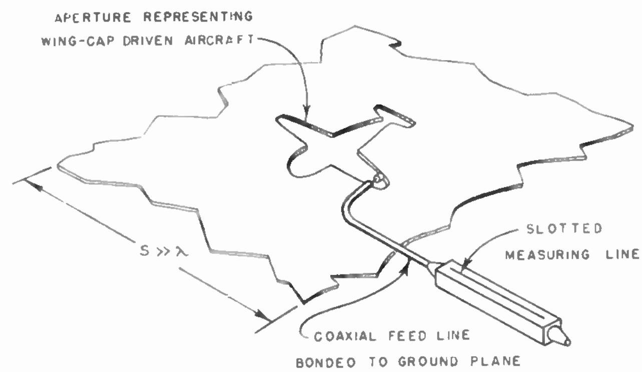


Fig. 4—Full-slot analogue of wing-cap antenna.

A final consideration of importance in connection with the feed configuration concerns the base capacitance. In an actual wing-cap or tail-cap antenna the structure has a finite thickness at the gap, and fields will be set up across the gap in the region within the contours of the structure. In the planar model, there is no counterpart to this region, and hence, the effect of the internal fields is not accounted for. Impedance values measured with the planar model may be corrected for this effect by adding a base-shunting capacitance correction term. The required value of base capacitance may be estimated from the dimensions of gap region.

VI. Experimental Work

All measurements to date have been made in the frequency range 400–1,600 mc using the image-plane technique.

Initial measurements were made on a relatively narrow, linear, center-fed slot so that the equivalent dipole impedance thus determined could be compared with available data on dipole impedances. The measured impedance values and the corresponding slot dimensions are shown in Fig. 5. Taking the equivalent diameter of the complementary strip dipole to be half the strip width, this dipole has a ratio of total length to equivalent diameter of 34.7. The resonant resistance is seen to have a reasonable value of about 60 ohms. The antiresonant resistance is 600 ohms, or about 11 per cent higher than the value of 540 ohms calculated from Schelkunoff's theory.[7] This check is considered to be adequate, in view of the general disagreement on the exact values of antiresonant resistance of such antennas.

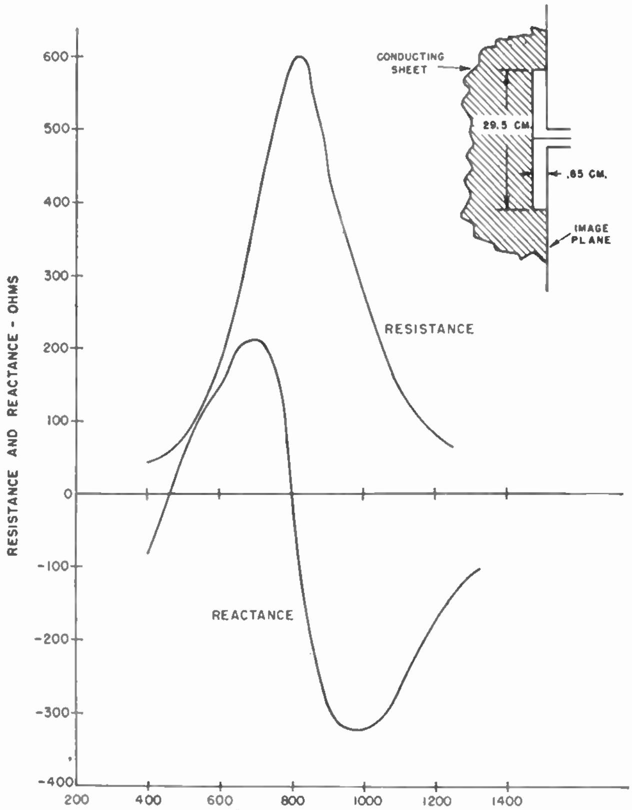


Fig. 5—Dipole impedance as determined from complementary slot measurements.

The complementary slot modelling technique has been employed in an extensive investigation of the impedance properties of wing-cap and tail-cap antennas. A typical measured impedance curve is shown in Fig. 3, in which the aircraft modelled was the B-29. An evaluation of the accuracy of the method is difficult because of the lack of adequate data on full-scale structures. Data obtained on the only comparable full-scale structure known to the writers are in reasonable agreement, when the effect of base capacitance is taken into account. These impedances are characterized by resistance values which vary between relatively narrow limits as compared with the corresponding variations for conventional wire antennas, and reactance curves which resemble that of a fixed capacitor in series with one or more low-Q, parallel resonant circuits.

Acknowledgments

The writers wish to acknowledge the contributions made to this work by C. T. Tai in many discussions of the problem. The experimental results reported herein were measured by Mrs. Robert F. Reese.

[7] S. A. Schelkunoff, "Theory of antennas of arbitrary size and shape," Proc. I.R.E., vol. 29, pp. 493–521; September, 1941.

Alignment and Adjustment of Synchronously Tuned Multiple-Resonant-Circuit Filters*

MILTON DISHAL†, SENIOR MEMBER, IRE

Summary—A simple method of "tuning up" a multiple-resonant-circuit filter quickly and exactly is demonstrated. The method may be summarized as follows: Very loosely couple a detector to the first resonator of the filter; then, proceeding in consecutive order, tune all odd-numbered resonators for maximum detector output, and all even-numbered resonators for minimum detector output (always making sure that the resonator immediately following the one to be resonated is completely detuned).

Also considered is the correct adjustment of the two other types of constants in a filter. Filter constants can always be reduced to only three fundamental types: f_0, $d_r(1/Q_r)$, and $K_{r(r+1)}$. This is true whether a lumped-element 100-kc filter or a distributed-element 5,000-mc unit is being considered. d_r is adjusted by considering the rth resonator as a single-tuned circuit (all other resonators completely detuned) and setting the bandwidth between the 3-db-down-points to the required value. $K_{r(r+1)}$ is adjusted by considering the rth and $(r+1)$th adjacent resonators as a double-tuned circuit (all other resonators completely detuned) and setting the bandwidth between the resulting response peaks to the required value.

Finally, all the required values for K and Q are given for an n-resonant-circuit filter that will produce the response $(V_p/V)^2=1+(\Delta f/\Delta f_{3db})^{2n}$.

* Decimal classification: R143.2×R386. Original manuscript received by the Institute, May 1, 1950; revised manuscript received, January 12, 1951. This paper was written in connection with work done on a contract for the Department of Air Force—AMC, Wright-Paterson Air Force Base.

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

I. Introduction

THIS PAPER attempts to answer two questions: "Exactly how can one 'tune up' a synchronously tuned multiple-resonant-circuit filter and be sure the tuning is correct?" and "Exactly how can one make sure that the mechanical design is actually supplying the required circuit constants?"

It should be noted that, for brevity, the paper will refer only to band-pass filters; the reader should realize that the discussion also applies similarly to the alignment and adjustment of low-pass, high-pass, and band-rejection filters when analogous frequency-variables and circuit constants are used.

The physical embodiment of a constant-K or equivalent type of filter, i.e., a filter having n complex frequency roots and no finite frequencies of infinite attenuation (all zeros at infinity), must exactly supply the numerical values of three kinds of quantities:[1] (1) resonant frequency f_0, (2) coefficients of coupling between adjacent resonators $K_{r(r+1)}$, and (3) resonator decrements $d_r(Q_r=1/d_r)$. It may be helpful to note that the above is true whether the elements of the unit are lumped, quasi-lumped, or distributed, so long as the percentage bandwidth is less than approximately 10 per cent for the latter two cases, as is most always true.

It should be realized that in the literature concerning filters a number of seemingly different types of constants have been used to describe the same or exactly equivalent networks. For example, classical filter theory which gives only approximate design data, usually produces the required values for L, C, M, and R; late in the 1930's, a number of papers described circuits by the so-called "ladder-network coefficients" for each arm; and at present many papers speak of the doubly or singly loaded Q of each resonator. In every case, the many different types of constants are all equivalent, but it has been the experience of the author that the constants f_0, $K_{r(r+1)}$, and Q_r are the "best" to use, particularly when dealing with *dissipative* filters.

For practical reasons usually involving mechanical tolerances, most selective-circuit designs incorporate a trimming adjustment for setting the resonant frequency of each resonator. After the filter is mechanically finished, the unit is aligned, i.e., all resonant frequencies are somehow correctly adjusted. Section III describes a method of alignment for multiple-resonant-circuit filters that is precise, requires no sweep-frequency generator, and can be performed by unskilled personnel.

The coefficient of coupling between adjacent resonators is usually not made variable as this adjustment requires a person "skilled in the art"; each K is carefully set by the designer as part of the mechanical design, which must be sufficiently stable to maintain it at the required value. Section IV describes an easy method for experimentally adjusting each coefficient of coupling to the exact desired value.

For the sake of completeness, a few comments are made about measuring Q, the third of these constants, in Section V.

Section VI presents some useful design data on what values of K and Q are required in a multiple-resonator-circuit filter.

[1] M. Dishal, "Design of dissipative band-pass filters producing desired exact amplitude-frequency characteristics," Proc. I.R.E., vol. 37, pp. 1050–1069; September, 1949. Also *Elec. Commun.*, vol. 27, pp. 56–81; March, 1950.

II. Symbols

n = total number of resonators used in a filter.

r = resonator number in filter chain. The resonator at the input end is numbered 1.

f_0 = resonant frequency of each resonator; this must include all coupling reactances. See Section IIIC.

$K_{r(r+1)}$ = coefficient of coupling between the rth and $(r+1)$th adjacent resonators. This may be defined fundamentally as the fractional bandwidth between the resulting response peaks that exist when each *pair* of adjacent resonators is considered separately (and the resonator Q's are infinite).

d_r = decrement of the resonator. This may be defined fundamentally as the fractional bandwidth between the 3-db-down points when each resonator is considered separately.

$\Delta f_{3\text{db}}$ = total bandwidth between 3-db-down response points.

$V_{1,r}$ = voltage across *first* resonator when all following resonators up to the rth resonator have been correctly tuned.

$p_{AB}=K_{AB}(Q_AQ_B)^{1/2}$ = fraction of "critical coupling" in a double-tuned circuit made up of resonators A and B.

$(\Delta f_p)_A$ = total bandwidth between response peaks in resonator A, to which a generator is coupled, when A is coupled *only* to an adjacent resonator B.

$(\Delta f_p)_B$ = total bandwidth between the response peaks in the above-mentioned resonator B.

$F_p=(\Delta f_p/f_0)$ = fractional bandwidth between response peaks.

$t=Q_A/Q_B$.

Δf_β = total bandwidth between response points that are V_p/V_β down from the peak response.

V_p = voltage output at peak of response curve.

V_β = voltage output at point of response curve where the bandwidth is Δf_β.

III. Alignment of Multiple-Resonator Filters

A. General Principle

This paper will refer mainly to small-percentage-bandwidth node networks. The reader should realize that in accordance with the principle of duality and with the following substitutions of words: mesh for node, current for voltage, voltage for current, open circuit for short circuit, and the like, the alignment procedure applies similarly to mesh networks. The procedure applies also to the large-percentage-bandwidth constant-K configuration discussed in Section IIIC.

The fundamental principle proposed in this section is that alignment can best be done by *completely assembling the filter and then concentrating on the amplitude phenomena occurring in the first resonator of the filter chain at* *the desired resonant frequency*. In Section IIIC, it is shown that if all the resonators are first completely detuned and if they are resonated in numerical order, calling the input resonator 1, then all odd-numbered resonators place an open circuit (high resistance) and all even-numbered resonators place a short circuit (low resistance) across the input terminals when correctly tuned.

B. Alignment Procedure

The alignment procedure will be described using the quadruple-tuned node-type band-pass filter shown in Fig. 1 as an example. Fig. 2 shows a practical physical embodiment of the circuit of Fig. 1.

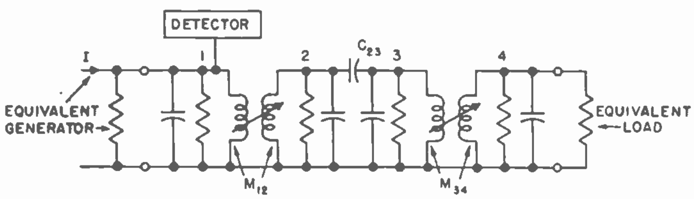


Fig. 1—Quadruple-tuned filter used to demonstrate alignment procedure of Section III. It should be realized that the alignment procedure applies to all the different types of synchronously tuned constant-K and coupled-resonant-circuit filters.

The procedure is applicable to all coupled-resonant-circuit filters, whether they be low-frequency constant-K configurations, medium-frequency coupled circuits, microwave quarter-wave-coupled waveguide filters, or the like.

1. Connect the generator to the first resonator of the filter and the load to the last resonator of the filter in exactly the same manner as they will be connected in actual use.

2. Couple a nonresonant detector directly and very loosely[2] to either the electric (voltage) or magnetic (current) field of the *first* resonator of the filter chain.

3. Completely detune[3] all resonators.

4. Set the generator frequency to the desired midfrequency of the filter.

5. Tune resonator 1 for *maximum output* indication on the detector. Lock the tuning adjustment.

6. Tune resonator 2 for *minimum output* indication on detector. Lock the tuning adjustment.

7. Tune resonator 3 for maximum output and lock the tuning adjustment.

8. Tune resonator 4 for minimum output and lock the tuning adjustment. The alignment of the filter shown in Fig. 1 is now complete.

If it is impracticable completely to detune all the resonators in a node network, a *single* device may be used to short-circuit *the resonator immediately following the one being tuned* since this will remove the effect of all the following resonators. It is important to make sure that this short circuit is fully effective at the measurement frequency.

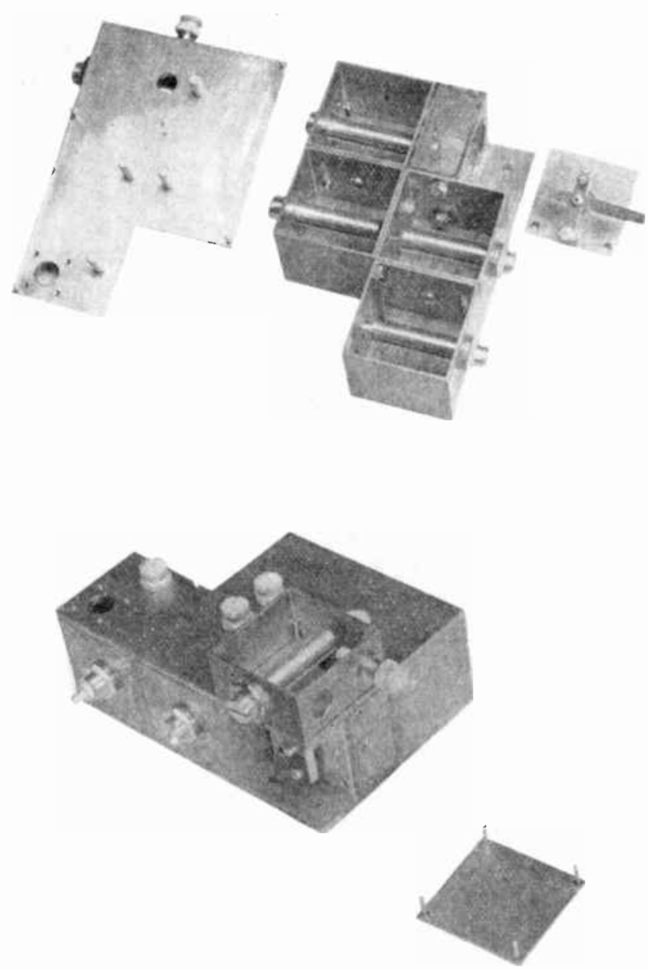


Fig. 2—A quadruple-tuned μhf filter embodying the circuit of Fig. 1. The midfrequency is 1,400 mc and the 3-db bandwidth is 6 mc. Note the smaller inductive coupling slot between resonators 1 and 2, the capacitive coupling hole between resonators 2 and 3, and the larger inductive coupling slot between resonators 3 and 4. The small plate with the "cross" mounted on it is the crystal mixer unit; when it is mounted in the last small cavity, the crystal is correctly coupled (capacitively) to both the fourth resonator and to the local-oscillator resonator.

Fig. 3 clearly demonstrates the amplitude-frequency phenomena that occur in each step of the alignment procedure. A sweep-frequency generator was used, and attention is called to the resonant-frequency marker. It should be clearly realized that since the alignment adjustments depend exclusively on the amplitude of the response at the resonant frequency f_0, a sweep-frequency generator is *not* required and all adjustments can be made with a single-frequency input f_0.

These oscillograms were obtained in aligning the quadruple-tuned filter shown in Fig. 1. It was designed to produce a Chebishev transfer shape[1] having a 1/2-db peak-to-valley ratio when loaded on one side only; i.e., it was fed by a high-impedance generator.

[2] A nonresonant detector (or generator) may be said to be "very loosely" coupled when it lowers the unloaded Q of the resonator by less than 5 per cent (say).

[3] A resonator is sufficiently detuned when its resonant frequency is at least 10 pass-band-widths away from the pass-band midfrequency.

Efficient filters with low internal losses, i.e., those using resonators having unloaded Q's very much greater than the fractional midfrequency ($f_0/\Delta f_{3db}$), produce deep and broad minimums when the even-numbered resonators are properly tuned, as may be seen in Fig.

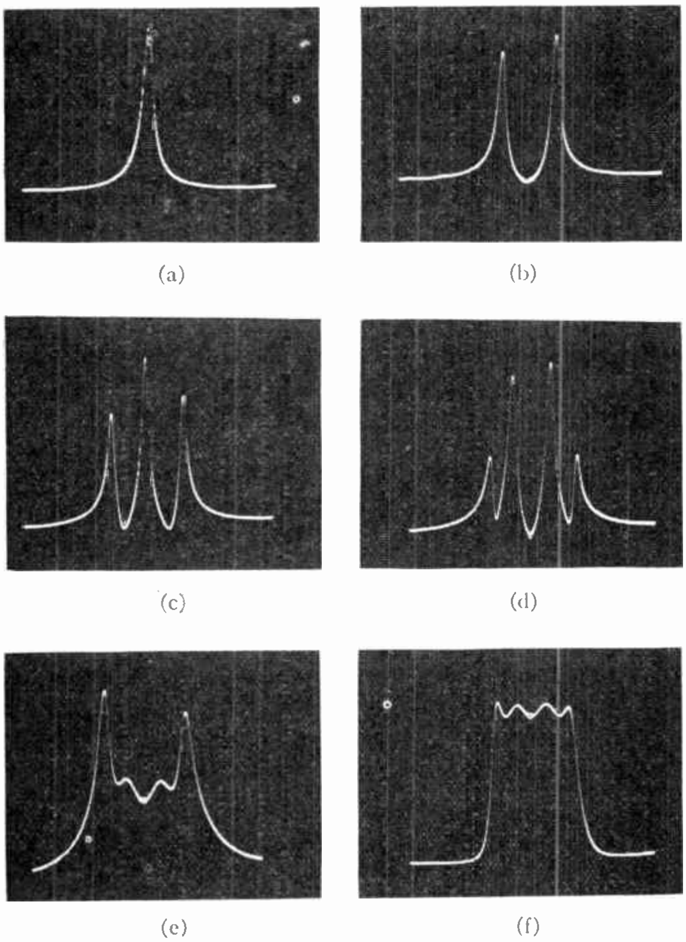


Fig. 3—Oscillograms of the amplitude-frequency phenomenon occurring in resonator 1 as the alignment steps of Section III are performed. (a) Resonator 2 is short-circuited or detuned and the input resonator 1 (odd-numbered) is adjusted for maximum marker signal of f_0. (b) Resonator 3 is detuned and the second resonator (even-numbered) is tuned for minimum amplitude of f_0. Oscillograms (c) and (d) show the continuation of the procedure of tuning odd-numbered resonators to produce maximum and even-numbered resonators to produce minimum values of f_0 response in resonator 1. It can be seen that as the rth resonator is tuned, there will be r peaks and $r-1$ valleys produced in resonator 1; this is a simple restatement of Foster's reactance theorem. Oscillogram (e) shows the voltage across resonator 1 as, with correct applied loading, the last resonator (even-numbered) of Fig. 1 is tuned for minimum output at f_0. Oscillogram (f) shows the resulting Chebishev transfer response shape (no tuning adjustments were retouched).

3(b). Therefore, it is important to use a large-amplitude signal input and high detector gain so that the middle of the minimum can be tuned accurately to the midfrequency. If the maximum generator input and detector gain still produce a broad null, the tuning adjustments should be set midway between two points of equal output.

C. Simple Theory of Alignment Procedure

Perhaps the simplest way of showing that the alignment procedure is correct is to consider the large-percentage-bandwidth constant-K filter chain shown in Fig. 4(a), to which all small-percentage-bandwidth coupled-resonant-circuit filters are exactly equivalent no matter what type of coupling is used between adjacent resonant circuits; and then to consider as a further example the small-percentage-bandwidth node circuit shown in Fig. 4(b).

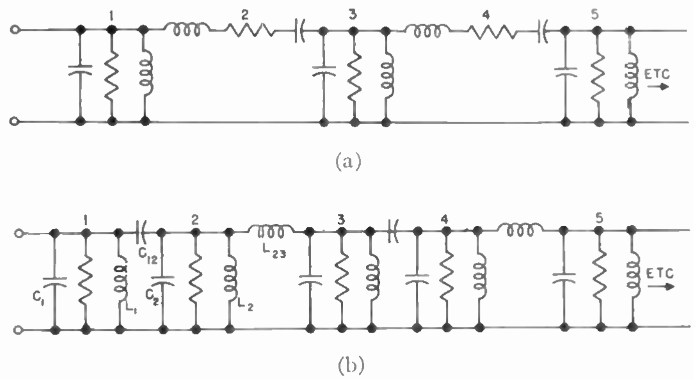


Fig. 4—(a) and (b) are, respectively, large- and small-percentage-bandwidth networks referred to in the explanations given in Section IIIC. It should be clearly realized that Fig. 3(b) shows alternate C and L couplings purely as an example; the explanation is true no matter what type of reactive coupling is used.

The reasoning applicable to the constant-K configuration of Fig. 4(a) requires previous knowledge of two simple facts.

First, complete detuning of all the resonators means that all the *series resonators* are effectively *open-circuited* and all the shunt resonators are effectively *short-circuited*.

Second, when correctly aligned, the resonant frequency f_0 of each separate resonator is identical.

Thus, when resonator 1 is tuned to f_0 (with resonator 2 open-circuited), maximum voltage will appear across the high parallel-resonant resistance of resonator 1. When resonator 2 is tuned to f_0 (with resonator 3 short-circuited), the low series-resonant resistance of 2 will shunt the terminals of 1, thus dropping the voltage across resonator 1 to a minimum. On tuning 3 to f_0 (with resonator 4 open-circuited), the high parallel-resonant resistance of 3 will remove the low series-resonant resistance of 2 from across the terminals of 1 so that the voltage across 1 will again rise to a maximum. Thus, starting at the front end of the filter, all odd-numbered resonators will produce a maximum voltage and all even-numbered resonators will produce a minimum voltage across resonator 1.

The reasoning applicable to the small-percentage-bandwidth node network of Fig. 4(b) requires previous knowledge of three simple facts.

First, complete detuning of a resonator means that the node involved is effectively short-circuited.

Second, when correctly aligned, the resonant frequency of each node is identical and the elements that resonate a node consist of every susceptance that touches the node, e.g., node 2 of Fig. 4(b) is resonated by adjusting C_2 (or L_2) to resonate with parallel combination of C_{12}, C_2, L_2, and L_{23}.

Third, if a group of reactances parallel resonate together, then any one of the reactances also series resonates with all the others, e.g., C_{12} series resonates with the parallel resultant of C_2, L_2, and L_{23}.

Thus, where node 1 is tuned to f_0 (with node 2 short-circuited), the high parallel-resonant resistance of C_1, L_1, and C_{12} will produce a voltage maximum at f_0. When node 2 is tuned to f_0 (with node 3 short-circuited), C_{12} will series resonate with the parallel resultant of C_2, L_2, and L_{23}, thus placing a short circuit across node 1 and producing a voltage minimum across node 1. The process repeats as alignment proceeds, producing maximums for alignment of odd-numbered and minimums for even-numbered resonators.

It will be shown in Section IVD that if we know the Q's of each resonator being used, then the ratios of maxima and minima occurring in resonator 1 can be used to set or check all the $(n-1)$ coefficients of coupling in a filter.

IV. Exact Adjustment of Coupling Between Resonators

A. General Principle

Before this section can be applied to the mechanical design and adjustment of a filter, it is, of course, necessary to determine by some synthesis procedure just what adjacent-resonant-circuit coefficients of coupling the mechanical embodiment must produce. As mentioned in the Introduction, no matter what type of constants are used to describe the synthesized network, they can always be transformed into f_0, $K_{r(r+1)}$, and Q_r.

The fundamental procedure being proposed in this section is to consider every pair of adjacent resonators as a double-tuned, i.e., two-pole, circuit (with all the other resonators completely detuned), and so be able to use the exactly known relation between the circuit constants and the resulting amplitude-frequency characteristic of a double-tuned circuit.

In a double-tuned circuit with Q_A and Q_B equal to infinity, the fractional bandwidth $(\Delta f_p/f_0)_A$ between primary response peaks is exactly equal to the coefficient of coupling between resonators A and B. (*In fact, this may be used as the basic definition of the constant that is commonly called the coefficient of coupling.*) When the resonators do not have infinite Q, the above equality is not exactly true, and Figs. 5 and 6 together with the described procedures supply two ways of finding the exact coefficient of coupling between adjacent resonators.

B. Adjustment Procedure

The procedure for measuring adjacent-resonator coupling, which is applicable to all coupled-resonant-circuit filters whether they be called low-frequency constant-K configurations, medium-frequency coupled circuits, or microwave quarter-wave coupled-waveguide filters, is as follows:

1. Designate as A and B the two adjacent resonators between which the coefficient of coupling is to be adjusted.

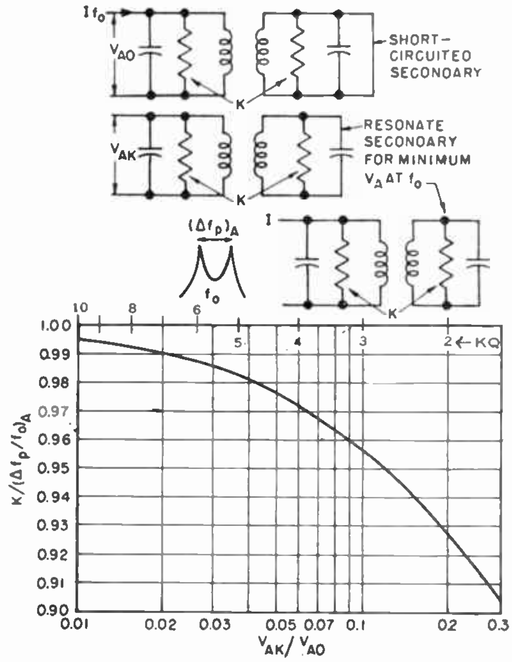


Fig. 5—Method of obtaining exact coefficient of coupling K between two resonators by measurements on only the primary circuit.

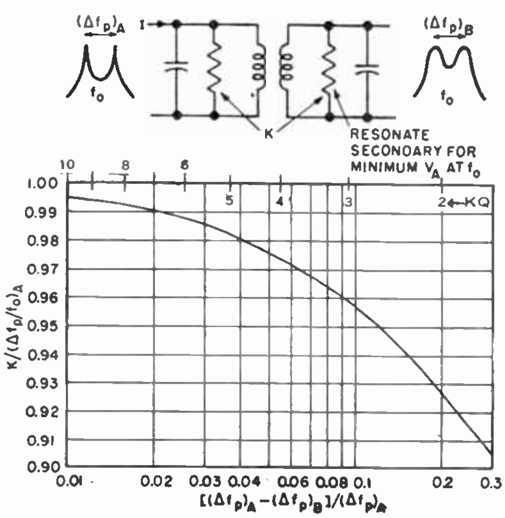


Fig. 6—Method of obtaining exact coefficient of coupling K between the two resonators by frequency measurements only.

2. Couple a nonresonant signal generator directly and very loosely to either the electric (voltage) or magnetic (current) field of resonator A.

3. Couple a nonresonant detector directly and very loosely to either the electric (voltage) or magnetic (current) field of resonator A.

4. Completely detune all the resonators in the filter chain.

5. Tune resonator A for maximum output from the detector. Record the signal-generator input and detector output.

6. Tune resonator B for minimum output from the detector (as in alignment procedure, Section III). Increase the signal-generator input to produce the same output obtained in step 5.

7. The ratio of the signal-generator input in step 5 to that in step 6 is the abscissa of the graph of Fig. 5. From the ordinate of this graph, read the ratio of the coefficient of coupling K between resonators A and B, to the percentage bandwidth $(\Delta f_p/f_0)_A$ between the response peaks that are now present across resonator A.

8. Carefully measure the bandwidth $(\Delta f_p)_A$ between the response peaks of resonator A.

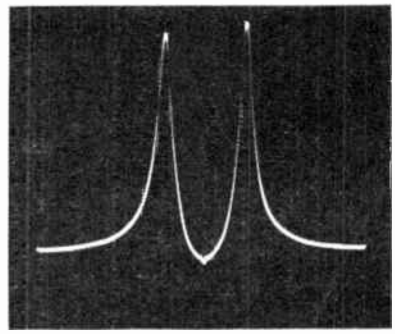
9. The exact coefficient of coupling is equal to the fractional bandwidth between these peaks times the ordinate obtained in step 7.

If it is not convenient to measure the amplitude ratio described in step 7, the following procedure involving frequency measurements only can be used. Omit the amplitude measurements from the above procedure and after step 6 carefully measure, by means of a nonresonant detector loosely coupled to resonator B, the bandwidth $(\Delta f_p)_B$ between the response peaks appearing on the secondary side.

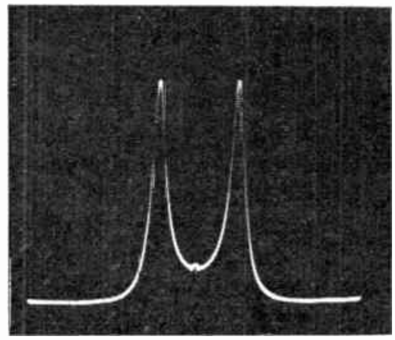
The fractional difference in peak bandwidth $[(\Delta f_p)_A-(\Delta f_p)_B]/(\Delta f_p)_A$ is the abscissa of the graph of Fig. 6. The exact coefficient of coupling is equal to the fractional bandwidth $(\Delta f_p/f_0)_A$ between *primary* peaks times the corresponding ordinate given in Fig. 6.

Examination of the ordinate values of Figs. 5 and 6 shows that even with a $(V_{A.A}/V_{A.B})$ ratio as small as 12 db, or a $[(\Delta f_p)_A-(\Delta f_p)_B]/(\Delta f_p)_A$ ratio as large as 25 per cent, the existing coefficient of coupling is only 8.5 per cent less than the percentage bandwidth between the primary peaks; therefore, in many cases it may be permissible simply to measure the bandwidth $(\Delta f_p)_A$ between primary response peaks and make the approximation that the coefficient of coupling is equal to about 0.96 times the measured fractional bandwidth.

Figs. 7(a) and 7(b) show the frequency-amplitude relations on the primary and secondary sides when the above procedure is used.



(a)



(b)

Fig. 7—Response peaks of a properly resonated pair of adjacent resonators with all other resonators completely detuned. By measuring the frequency bandwidth between the peaks occurring in the input resonator (a), the exact coefficient of coupling may be obtained. (See Fig. 4.) The corresponding peak bandwidth for the second resonator (b) differs slightly from (a) because the resonators have finite values of Q. (See Fig. 5.)

C. Quantitative Theory of Measuring Coefficient of Coupling

All the amplitude-frequency characteristics of a double-tuned circuit are most conveniently related to each other by means of two variables, the Q ratio of primary to secondary Q_A/Q_B and the $K_{AB}(Q_AQ_B)^{1/2}$ product. For the preparation of Figs. 5 and 6, we need the relations between the above variables and four quantities: (1) drop in primary voltage when the secondary is correctly resonated $V_{A.B}/V_{A.A}$; (2) bandwidth between response peaks in the primary side $(\Delta f_p)_A$; (3) bandwidth between peaks on the secondary side $(\Delta f_p)_B$; and (4) relation of bandwidth between primary peaks to the coefficient of coupling $K_{AB}/(\Delta f_p/f_0)_A$.

Straightforward analysis of a correctly resonated double-tuned circuit results in the following equations for these quantities in which the Q ratio will be denoted by $t=Q_A/Q_B$ and the $K_{AB}(Q_AQ_B)^{1/2}$ product by p.

$$\left(\frac{V_{A,A}}{V_{A,B}}\right) = 1 + p^2 \tag{1}$$

$$\left(\frac{\Delta f_p}{f_0}\right)_A = K\left\{\left[1 + 2(1+t)\frac{1}{p^2}\right]^{1/2} - t\frac{1}{p^2}\right\}^{1/2} \tag{2}$$

$$\left(\frac{\Delta f_p}{f_0}\right)_B = K\left[1 - 1/2\left(t + \frac{1}{t}\right)\frac{1}{p^2}\right]^{1/2}. \tag{3}$$

The fourth relation mentioned above is, of course, obtained from (2).

Equations (1), (2), and (3) are used straightforwardly to obtain the graphs of Figs. 5 and 6 for the case where the Q ratio is unity (most filters are built using resonators having the same unloaded Q).

D. Checking Coefficients of Coupling

During construction, or after a filter has been constructed, it is often desirable to be able to set or to double check each coefficient of coupling without going through the procedure of converting each pair of adjacent resonators into a double-tuned circuit as is required by Section IVB.

As shown below this can be accomplished by measurements made entirely at the input resonator. There are, in practice, two cases which have to be considered: In the first case (usually the large-percentage-bandwidth filter) the unloaded Q's of the resonators being used are very much greater than the fractional midfrequency $(f_0/\Delta f_{3db})$ being used; i.e., the unloaded individual Q's are essentially infinite. In the second case (usually the small-percentage-bandwidth filter), the unloaded Q's of the resonators are only 4 or 5 (say) times $(f_0/\Delta f_{3db})$.

For the first case above, the K's can be set, or measured, in consecutive order, by measuring the bandwidth between the various response peaks appearing in resonator 1, as each of the following resonators is resonated in consecutive order (see Fig. 3).

It will be remembered from Fig. 3 that there will be r response peaks occurring in the input resonator when the rth resonator is correctly tuned.

To calculate the peak bandwidths that should be measured when the rth resonator is tuned, straightforward analysis shows that we must solve the polynomial (4) for its roots

$$F_p{}^r - (\sum K^2)F_p{}^{(r-2)} + (\sum K^2K^2)F_p{}^{(r-4)} - (\sum K^2K^2K^2)F_p{}^{(r-6)} \cdots = 0. \quad (4)$$

The polynomial stops at the first- or zero-power term; i.e., no negative exponents are considered.

The coefficient of the $F_p{}^{(r-2)}$ term is the sum of all the products of the coefficients of coupling squared taken one at a time; the coefficient of the $F_p{}^{(r-4)}$ term is the sum of all the products of K^2 taken two at a time, *but in any pair a subscript number must not appear more than once*; the coefficient of the $F_p{}^{(r-6)}$ term is the sum of all the K^2 products taken three at a time, *but in any triplet a subscript number must not appear more than once*; and so forth.

As an example, as the first to fifth resonators are tuned consecutively, the following 5 polynomials must be solved consecutively to calculate the fractional bandwidth $(\Delta f_p/f_0)$ that should occur between response peaks.

$$F_p = 0 \quad (4a)$$

$$F_p{}^2 - K_{12}{}^2 = 0 \quad (4b)$$

$$F_p{}^3 - (K_{12}{}^2 + K_{23}{}^2)F_p = 0 \quad (4c)$$

$$F_p{}^4 - (K_{12}{}^2 + K_{23}{}^2 + K_{34}{}^2)F_p{}^2 + (K_{12}K_{34}) = 0 \quad (4d)$$

$$F_p{}^5 - (K_{12}{}^2 + K_{23}{}^2 + K_{34}{}^2 + K_{45}{}^2)F_p{}^3 + (K_{12}{}^2K_{34}{}^2 + K_{12}{}^2K_{45}{}^2 + K_{23}{}^2K_{45}{}^2)F_p = 0. \quad (4e)$$

These first five polynomials require simple linear- and quadratic-equation solutions, and, because the coefficients are known numerical values, the Graeffe root-squaring process can be used to solve accurately any of the polynomials.

For the second case described above, the coefficients of coupling should be set or measured in consecutive order as follows: Accurately measure the Q of each resonator in the filter, then proceed step by step through the alignment procedure of Section IIIB, accurately measuring (and recording) the magnitudes of the alternate maxima and minima produced. Straightforward analysis shows that the ratio of the detector output obtained when resonator 1 is alone resonated to that obtained when resonator r is resonated is given by the continued fraction of (5).

$$\left(\frac{V_{1,1}}{V_{1,r}}\right) = \left[1 + \cfrac{p_{12}{}^2}{1 + \cfrac{p_{23}{}^2}{1 + \cfrac{\cdots}{1 + p^2{}_{(r-1r)}}}}\right]. \quad (5)$$

It is important to realize that Q_1 is the Q of resonator number 1 with both the generator and detector coupled to it.

Since we know the desired value for each K and have measured each Q, we know the value of each $P_{12}{}^2 = K_{12}{}^2Q_1Q_2$, and so on in (5); and the measured value of the voltage ratios $(V_{1,1}/V_{1,r})$ should, of course, equal those calculated from (5).

V. Adjustment of Resonator Decrement $(1/Q)$

It has been the author's experience that any method of measuring Q that removes the resonator from the exact position it occupies in the filter chain is potentially inaccurate.

It has also been noted that measurements involving an amplitude-modulated oscillator can lead to erroneous results particularly in the uhf and microwave regions because of spurious frequency modulation. If an amplitude-modulated carrier is being used, an obvious check for appreciable spurious frequency modulation is to use an oscilloscope to examine the envelope of the wave form being measured and a narrow-band receiver to examine the frequency content of the supposedly purely amplitude-modulated carrier.

The most trustworthy method of making accurate unloaded or loaded Q measurements on a resonator that is part of a filter chain seems to be as follows:

1. Completely assemble the filter.
2. Completely detune all resonators except the one to be measured. Obviously, complete detuning of the resonator on each side of the one being measured should be satisfactory.
3. A nonresonant signal generator is coupled directly and very loosely to either the electric or magnetic field of the resonator.
4. A nonresonant detector is coupled very loosely and preferably to the field opposite to that being used for the generator; i.e., make sure that there is negligible direct coupling between generator and detector.
5. Using an unmodulated wave or an amplitude-modulated wave checked for negligible frequency modulation from the signal generator, measure the frequency difference Δf_β between the points that are V_p/V_β down from the peak response; the resonator Q is given by

$$Q = (f_0/\Delta f_\beta)\left[(V_p/V_\beta)^2 - 1\right]^{1/2}. \quad (6)$$

Obviously, when high Q's are to be measured, the apparatus must be capable of measuring very-small-percentage bandwidths. This may be accomplished by "beating down" the measurement frequency with a very stable local oscillator and a mixer and by making the measurements at the resulting difference frequency. By this means, the accuracy of the measuring apparatus is increased by the ratio of the original to the difference frequencies. The "cost" of this simplification is, of course, the necessity of using a very stable local oscillator.

VI. K's and Q's to Produce Response Shape $V_p/V = [1 + (\Delta f/\Delta f_{3db})^{2n}]^{1/2}$.

A straightforward synthesis procedure[1,4,5] shows that for the transfer response shape given just above, and if infinite Q resonators are used, i.e., in practice, resonators whose unloaded Q's are greater than $10/[\sin(90°/n)]$ times the fractional midfrequency ($f_0/\Delta f_{3db}$), it is possible to write very concisely the exact values of K and Q required for any number n of resonators for two important practical cases. The rate of cutoff obtained with the transfer response shape given in the title of this section is exactly $6n$ db per octave and the size of the octave is the bandwidth between the midfrequency and any frequency past the 3-db-down frequency.

For the case where *one end of the network can have only a pure reactance placed across it* (by either a reactive generator, e.g., the plate of a pentode tube, or by a reactive load), the required Q and K values are given exactly by (7(a)) and (7(b)).

$$\frac{Q_1}{(f_0/\Delta f_{3db})} = \sin\frac{90°}{n}, \quad Q_{2\to n} = \infty \tag{7a}$$

$$\frac{K_{r(r+1)}}{(\Delta f_{3db}/f_0)} = \frac{\cos(r\,90°/n)}{\{[\sin(2r-1)(90°/n)][\sin(2r+1)(90°/n)]\}^{1/2}}, \tag{7b}$$

where r is made equal to 1, 2, 3, and so forth, up to $(n-1)$; and n is the total number of resonators used. For the above design equations, *the end resonator that is loaded is called resonator* 1.

For the case where *both ends of the network must have resistances placed across them* (e.g., a 50-ohm generator and 50-ohm load are being used), then the required Q and K values are given exactly by (8(a)) and (8(b)).

$$\frac{Q_1(=Q_n)}{(f_0/\Delta f_{3db})} = 2\sin\frac{90°}{n}, \quad Q_{2\to(n-1)} = \infty \tag{8a}$$

$$\frac{K_{r(r+1)}}{(\Delta f_{3db}/f_0)} = \frac{0.5}{\{[\sin(2r-1)(90°/n)][\sin(2r+1)(90°/n)]\}^{1/2}}, \tag{8b}$$

where r is set equal to 1, 2, 3, and so forth, up to $(n-1)$; and n is the total number of resonators being used. Since the resulting circuit is symmetrical, it makes no difference which end resonator is called resonator 1

For the unfortunately practical case where the unloaded Q of the resonators being used is not infinite, it does not seem to be possible to obtain elegantly simple design equations like (7) and (8). Figs. 8 and 9 give the K and Q values required to produce exactly the transfer

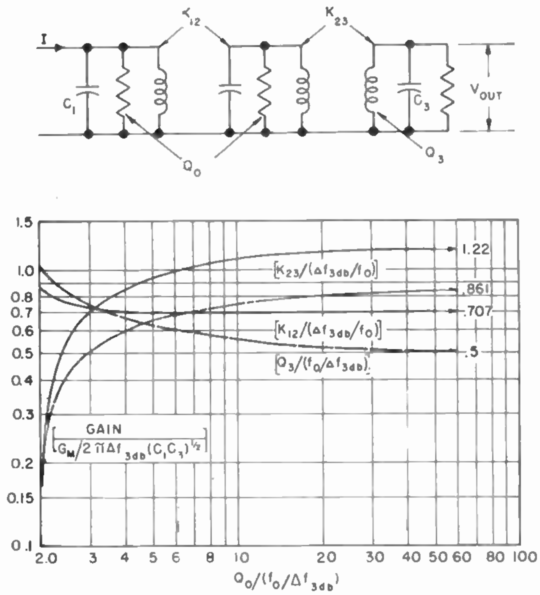


Fig. 8—Exact design for a finite-Q triple-tuned node circuit producing a Butterworth response shape when driven by an infinite-resistance generator.

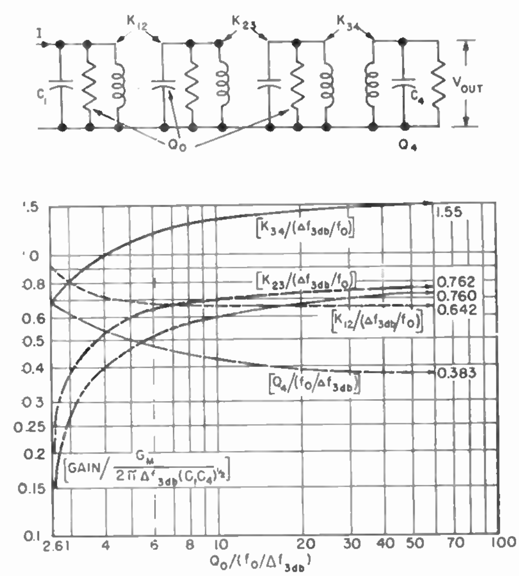


Fig. 9—Exact design for a finite-Q quadruple-tuned node circuit producing a Butterworth response when driven by an infinite-resistance generator.

response shape $(V_p/V)^2 = 1 + (\Delta f/\Delta f_{3db})^{2n}$ for triple- and quadruple-resonator filters, respectively, for the reactive-generator case. The abscissa of these graphs is the ratio of the unloaded $Q(Q_0)$ of the resonators being used to the fractional midfrequency ($f_0/\Delta f_{3db}$).

[4] E. L. Norton, U. S. Patent No. 1,788,538; January, 1931.
[5] W. R. Bennett, U. S. Patent No. 1,849,656; March, 1932.

The Multisection *RC* Filter Network Problem*

L. STORCH†

Discussion

In order to analyze the network of Fig. 1, the author of a recent paper[1] finds it necessary to postulate that equations (1) to (4) represent the desired voltage transfer ratio $(E_0/E_i)_n$ and input admittance G_n, which assumptions he then validates by an inductive proof:

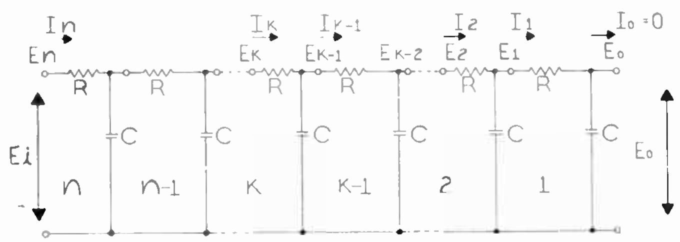


Fig. 1—(slightly modified). The n-section RC filter network.

$$\left(\frac{E_0}{E_i}\right)_n = \frac{1}{1 + a_1Tp + a_2T^2p^2 + \cdots + a_{n-2}T^{n-2}p^{n-2} + a_{n-1}T^{n-1}p^{n-1} + T^np^n} \tag{1}$$

$$G_n = \frac{Cp[n + b_1T_p + b_2T^2p^2 + \cdots + b_{n-2}T^{n-2}p^{n-2} + T^{n-1}p^{n-1}]}{1 + a_1T_p + a_2T^2p^2 + \cdots + a_{n-2}T^{n-2}p^{n-2} + a_{n-1}T^{n-1}p^{n-1} + T^np^n}, \tag{2}$$

where the coefficients of $(pT)^m$ in the case of n sections are:

$$a_{m,n} = \frac{(n+m)!}{(n-m)!(2m)!} \tag{3}$$

$$b_{m,n} = \frac{(n+m)!}{(n-m-1)!(2m+1)!}. \tag{4}$$

The ability to anticipate exactly the correct solution, which consists of four intricate equations, would seem to demand extraordinary intuition or prescience. It is the purpose of this note to show how the network characteristics can be obtained straightforwardly from the basic circuit equations.

This approach should be more meaningful to the engineer who is interested in the method of solving the actual problem, rather than in a formal proof of a set of elaborate postulates. In addition, it supplies him with a mode of attack which is valuable also when facing other circuit problems of the recurrent network type.

For any section k in Fig. 1 ($2 \leq k \leq n$), by the junction law of currents:

$$I_k = \frac{E_k - E_{k-1}}{R} = \frac{E_{k-1} - E_{k-2}}{R} + pCE_{k-1}.$$

Therefore ($T = RC$):

$$E_k = (2 + pT)E_{k-1} - E_{k-2} \tag{5}$$

with

$$E_0 = E_0 \quad \text{and} \quad E_1 = (1 + pT)E_0.$$

The single recursion process for E_k, carried on until $k = n$, is sufficient to produce the complete solution:

$$\left(\frac{E_0}{E_i}\right)_n = \frac{E_0}{E_n} \tag{6}$$

$$G_n = \frac{I_n}{E_n} = \frac{pC}{pT} \cdot \frac{E_n - E_{n-1}}{E_n}. \tag{7}$$

Furthermore, explicit expressions for the coefficients $a_{m,k}$ of the polynomial

$$\frac{E_k}{E_0} = \sum_{m=0}^{k} a_{m,k}(pT)^m, \tag{8}$$

which is generated by the recursion process, can also be derived from the fundamental equation (5).

Substituting (8) in (5) and collecting terms in $(pT)^m$:

$$a_{m,k} = 2a_{m,k-1} + a_{m-1,k-1} - a_{m,k-2} \tag{9}$$

with

$$a_{0,0} = 1, \qquad a_{0,1} = a_{1,1} = 1,$$

and by (8)

$$a_{m,k} = 0 \quad \text{when} \quad m > k, \text{ or } m < 0 .$$

Let us write (9) in the more symmetrical form:

$$\Delta_{m,k} = (a_{m,k} - a_{m,k-1}) = (a_{m,k-1} - a_{m,k-2}) + a_{m-1,k-1}, \tag{10}$$

and tabulate the first few terms[2] by forward differencing.

In view of the obvious nexus between the columns of Table I and sequences of binominal coefficients, these are listed in Table II. A comparison indicates that

* Decimal classification: R143.2. Original manuscript received by the Institute, April 19, 1950; revised manuscript received, December 13, 1950.

† Hughes Aircraft Company, Culver City, Calif.

[1] E. W. Tschudi, "Admittance and transfer function for an n-mesh RC filter network," Proc. I.R.E., vol. 38, pp. 309–310; March, 1950.

[2] This type of derivation is chosen since it is more graphic and also requires less space.

TABLE I. $a_{m,k}$ AND $\Delta_{m,k}$

k	$a_{0,k}$	$\Delta_{1,k}$	$a_{1,k}$	$\Delta_{2,k}$	$a_{2,k}$	$\Delta_{3,k}$	$a_{3,k}$
0	1	0	0	0	0	0	0
1	1	1	1	0	0	0	0
2	1	2	3	1	1	0	0
3	1	3	6	4	5	1	1
4	1	4	10	10	15	6	7
5	1	5	15	20	35	21	28
6	1	6	21	35	70	56	84
7	1	7	28	56	126	126	210
8	1	8	36	84	210	252	462

TABLE II. BINOMIAL COEFFICIENTS

k	$\binom{k}{0}$	$\binom{k}{1}$	$\binom{k}{2}$	$\binom{k}{3}$	$\binom{k}{4}$	$\binom{k}{5}$	$\binom{k}{6}$
0	1	0	0	0	0	0	0
1	1	1	0	0	0	0	0
2	1	2	1	0	0	0	0
3	1	3	3	1	0	0	0
4	1	4	6	4	1	0	0
5	1	5	10	10	5	1	0
6	1	6	15	20	15	6	1
7	1	7	21	35	35	21	7
8	1	8	28	56	70	56	28

$$a_{0,k} = \binom{k}{0}, \qquad a_{1,k} = \binom{k+1}{2},$$

$$a_{2,k} = \binom{k+2}{4}, \qquad a_{3,k} = \binom{k+3}{6}.$$

Apparently, the general solution is

$$a_{m,k} = \binom{k+m}{2m},$$

which can be verified by substitution in (9).

Factoring $pC\ E_0/E_n$ in (7):

$$\frac{E_n - E_{n-1}}{(pT)E_0} = \sum_{m=0}^{n-1} b_{m,n}(pT)^m.$$

Consequently:

$$b_{m,n} = a_{m+1,n} - a_{m+1,n-1}$$

$$= \binom{n+m+1}{2m+2} - \binom{n+m}{2m+2} = \binom{n+m}{2m+1}.$$

The explicit solution for the coefficients is, therefore:

$$a_{m,n} = \binom{n+m}{2m} = \frac{(n+m)!}{(n-m)!(2m)!} \tag{11}$$

$$b_{m,n} = \binom{n+m}{2m+1} = \frac{(n+m)!}{(n-m-1)!(2m+1)!}. \tag{12}$$

This completes the solution of the problem.

A quite different method of analysis, which may be of considerable interest, will now be outlined briefly.

A resistor $R/2$ connected in series with the upper output terminal in Fig. 1 does not alter the desired open-circuit characteristics. But the structure may now be considered as a chain of n symmetrical T sections, with a resistor $R/2$ in each series arm and a capacitor C in each shunt arm, which is fed through a source impedance $R/2$.

In terms of the image parameters θ and Z_0 of a single T section,[3]

$$\cosh\theta = 1 + \frac{pT}{2}, \quad Z_0 = \frac{\sinh\theta}{pC} = \frac{R}{2}\cdot\frac{\sinh\theta}{\cosh\theta - 1}, \tag{13}$$

the properties of the complete network are described in matrix form by

$$\begin{pmatrix} E_n \\ I_n \end{pmatrix} = \begin{bmatrix} 1, & \frac{R}{2} \\ 0, & 1 \end{bmatrix} \begin{bmatrix} \cosh\theta, & Z_0\sinh\theta \\ \frac{\sinh\theta}{Z_0}, & \cosh\theta \end{bmatrix}^n \begin{pmatrix} E_0 \\ I_0 \end{pmatrix}.$$

Substituting[4]

$$\begin{bmatrix} \cosh\theta, & Z_0\sinh\theta \\ \frac{\sinh\theta}{Z_0}, & \cosh\theta \end{bmatrix}^n = \begin{bmatrix} \cosh n\theta, & Z_0\sinh n\theta \\ \frac{\sinh n\theta}{Z_0}, & \cosh n\theta \end{bmatrix}$$

and performing the multiplication after setting $I_0 = 0$,

$$\frac{E_n}{E_0} = \cosh n\theta + \frac{R}{2Z_0}\sinh n\theta$$

$$\frac{I_n}{E_0} = \frac{\sinh n\theta}{Z_0}.$$

After substituting for Z_0 from (13) and simplifying:

$$\frac{E_n}{E_0} = \frac{\sinh (n+1)\theta}{\sinh\theta} - \frac{\sinh n\theta}{\sinh\theta} \tag{14}$$

$$G_n = \frac{I_n}{E_n} = \frac{pC\sinh n\theta}{\sinh (n+1)\theta - \sinh n\theta} \tag{15}$$

where

$$\theta = \cosh^{-1}\left(1 + \frac{pT}{2}\right).$$

This represents an alternate and compact solution[5] of the problem in terms of trigonometric functions of the iterative transfer constant of the RC section, which is also the image transfer constant of the full T section.

The same results can be obtained without the artifice of inserting a resistor $R/2$ in series with the output terminal. The transmission matrix of the RC section is expressed as a similarity transformation of the diagonal matrix containing its latent roots, in which form it is

[3] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., p. 194; 1943. Equations (75a) and (76); or directly from open circuit conditions, which determine the general circuit parameters A and C ($A = \cosh\theta$, $C = \sinh\theta/Z_0$) in terms of R and C.

[4] L. A. Pipes, "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 264–265; 1946.

[5] Reduces to the polynomial form, if desired, since

$$\frac{\sinh n\theta}{\sinh\theta} = \sum_n \binom{n+m}{2m+1}(pT)^m$$

by expanding in terms of

$$2\sinh\frac{\theta}{2} = \sqrt{pT}$$

according to #3.173 (modified), Smithsonian Mathematical Formulae, Smithsonian Institution, p. 67; 1939.

raised to the n'th power.[6] This process, however, involves more advanced concepts of the matrix calculus. Also, the solution in terms of hyperbolic functions can be obtained from the finite difference equation (5) by conventional methods, but requires more algebraic manipulation in that case.

Conclusions

It may be concluded that methods for the solution of the given problem are available which take nothing for granted but Ohm's and Kirchhoff's laws, and yet arrive at the goal with no more, or even less, algebraic manipulation than the "postulatory" method in the paper under discussion. Two of the possible methods have been presented in this note.

The finite difference method leads to the solution almost immediately. Its fundamental recursion formula (5) follows directly from the basic relation $\sum i = 0$ at a network junction, which can be formulated by inspection. A few more steps lead to explicit expressions for the coefficients $a_{m,n}$ and $b_{m,n}$, although, in general, solution by actual recursion according to (5) would be adequate. The second method, using image parameters suitable for cascade connection of symmetrical structures, may be even more attractive to the communications engineer.

[6] The first method is closely related, in mathematical terms, to a similarity transformation which equalizes the elements in the principal diagonal.

The Measurement of Antenna Impedence Using a Receiving Antenna*

E. O. HARTIG†, STUDENT, IRE, R. KING‡, SENIOR MEMBER, IRE, T. MORITA‡, ASSOCIATE, IRE, AND D. G. WILSON§, MEMBER, IRE

Summary—**Energy from a remote transmitter excites a receiving antenna that is erected vertically over a large conducting plane and base-loaded by a vertical slotted coaxial cavity of variable length. From measurements of the location and half-power width of resonance curves, the combined phase and damping factors for the two ends of the cavity are determined. By measuring these factors for the lower end of the cavity separately, those of the upper end are determined and used to calculate the impedance of the antenna. In effect, the receiving antenna is the generator driving the coaxial line, and it is the impedance of this generator that is measured. Curves of the measured impedance as a function of the electrical length of the antenna are given. Excellent agreement is obtained between impedances measured in this manner for the receiving antenna and corresponding impedances of the same antenna when base-driven through the slotted section. Both sets of measurements are in good agreement with the King-Middleton second-order theory.**

Is the impedance of an antenna that is loaded and used for reception equal to the impedance of the same antenna when used for transmission with the load replaced by a generator? The answer is simple. Yes, the impedances necessarily are equal if they are *defined to be the same*, and this is both useful and conventional. This is done in terms of an "equivalent" series circuit in which the receiving antenna is replaced by a concentrated emf V and a lumped internal impedance Z_0 in series with the lumped load Z_L. This circuit is rigorously equivalent to the antenna with load for the current I_0 entering and leaving the load, as is readily established using Thévenin's Theorem.[1] In the symmetrical, center-loaded antenna in Fig. 1(a) the open circuit voltage across AB as in Fig. 1(b), due to the action of the electric field, is $V = V_{AB}$ (open). Thévenin's Theorem states that I_0 in Fig. 1(a) is the same as I_0 in Fig. 1(c) if $V = V_{AB}$ (open) and Z_0 is the impedance looking into the terminals AB in Fig. 1(b) with the electric field E equal to zero. Hence,

$$I_0 = V/(Z_0 + Z_L). \tag{1}$$

Note that by definition Z_0 is the impedance of the antenna *as if driven by a potential difference across its terminals*. Hence, it is identically the transmitting impedance.

* Decimal classification: R221. Original manuscript received by the Institute, July 12, 1950; revised manuscript received, January 1, 1951.

The research reported in this document was made possible through support extended Cruft Laboratory, Harvard University, jointly by the Navy Department (Office of Naval Research), the Signal Corps of the U. S. Army, and the U. S. Air Force, under ONR Contract N5-ORI-76, T. O. 1. The original research was carried out by D. G. Wilson for his Ph.D. dissertation under the direction of R. King, and reported in Technical Report No. 43, Cruft Laboratory, May, 1948. Because of detector loading and an inadequate ground screen, the results obtained were not entirely satisfactory. Elimination of these sources of error and repetition of the measurements by E. O. Hartig and T. Morita culminated in Technical Report No. 94, Cruft Laboratory, December, 1949. This paper, combining parts of Wilson's work and the results of Hartig and Morita, has been prepared by R. King.

† Formerly Cruft Laboratory, Harvard University; now Goodyear Aircraft Corporation, Akron, Ohio.

‡ Cruft Laboratory, Harvard University, Cambridge 38, Mass.

§ Formerly Cruft Laboratory, Harvard University; now University of Kansas, Lawrence, Kan.

[1] Cruft Electronics Staff, "Electronic Circuits and Tubes," McGraw-Hill Book Co., New York, N. Y., p. 110; 1947.

V in (1) is that concentrated voltage which would maintain the same current I_0 in the load Z_L when connected in series with the lumped impedance Z_0 of the antenna as exists by action of the electric field along the loaded antenna. $V = V_{AB}$ (open) is defined in the literature[2,3] in terms of the electric field E and the dimensions and orientation of the antenna. Note that the simple circuit in Fig. 1(c) is equivalent to the actual

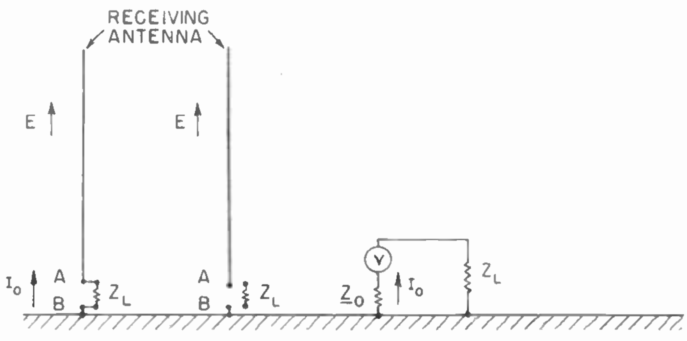


Fig. 1—Equivalent circuit for I_0 in receiving antenna.

antenna *only* in the determination of I_0. If the voltage V is connected in series with Z_L and the antenna, I_0 is the same as in the receiving antenna if V is properly defined, but I_z elsewhere on the antenna is not.

The impedance of the receiving antenna, defined as Z_0 in (1), may be measured with the apparatus shown in Fig. 2. Z_L is the input impedance of a vertical section of slotted line terminated in a sensitive detector and a piston, and the antenna is the vertical extension of the inner conductor above a large ground screen. V is the equivalent concentrated generator driving the line,

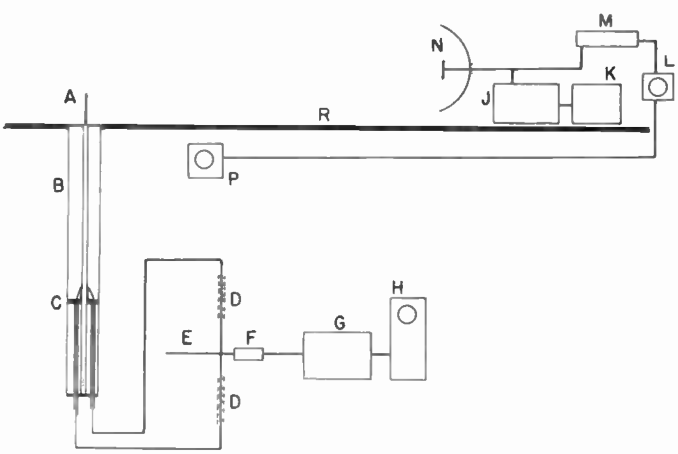


Fig. 2—Schematic diagram of complete equipment.

A = receiving antenna	J = transmitter
B = measuring line	K = modulator
C = movable piston	L = wavemeter crystal current
D = double line stretcher	M = wavemeter
E = shunt stub	N = transmitting antenna
F = bolometer mount	P = remote crystal current
G = tuned bolometer amplifier	R = ground plane.
H = Ballantine voltmeter	

and Z_0 is the internal impedance of the generator, i.e., the impedance of the antenna. Thus, the experimental problem is to measure the internal impedance Z_0 of the generator driving the line. This can *not* be done using either the conventional standing-wave-ratio method[4] or the distribution-curve method[4] since these involve only the load impedance. However, Z_0 is readily determined using the resonance-curve method.[4,5] It is necessary merely to determine the position and half-power width of the resonance curve obtained by moving the piston terminating the line for each length h of the antenna.

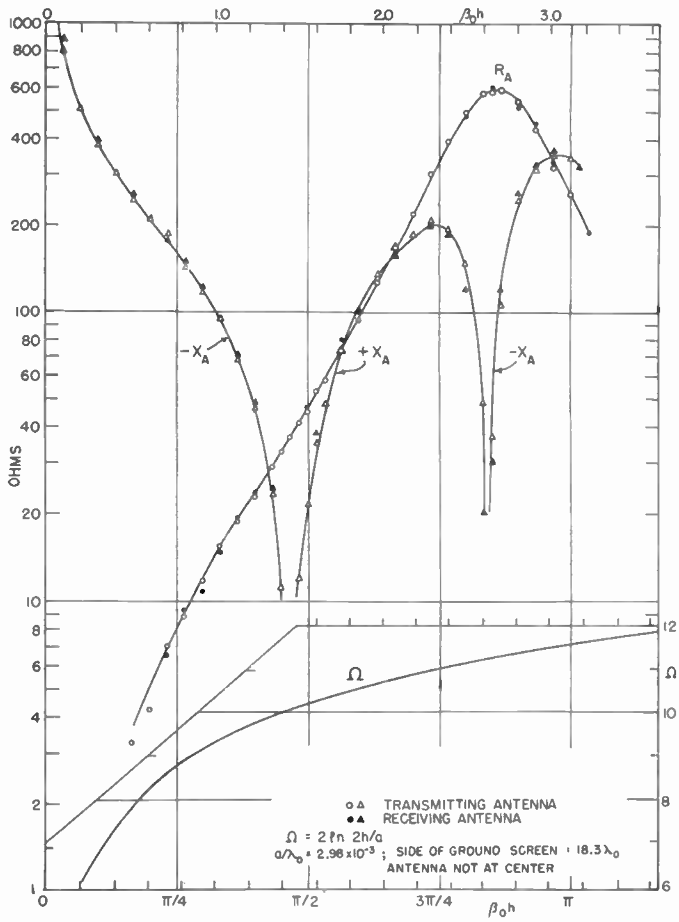


Fig. 3—Impedance of antenna by reception and transmission methods.

The resonance-curve method is expressed concisely using the terminal functions ρ_o and Φ_o of Z_0 at $z=0$, and ρ_s and Φ_s of Z_s at the end $z=s$ of the coaxial line with characteristic impedance $Z_c \doteq R_c$. These functions are defined by

$$Z_0 = Z_c \coth(\rho_o + j\Phi_o); \; Z_s = Z_c \coth(\rho_s + j\Phi_s). \quad (2)$$

The over-all phase function $(\beta s + \Phi_o + \Phi_s)$ is determined directly from the resonant length of the line; the over-all attenuation function $(\alpha s + \rho_o + \rho_s)$ from the half-power width, using

[2] R. King and C. W. Harrison, Jr., "The receiving antenna," Proc. I.R.E., vol. 32, pp. 18–49; January, 1944.

[3] R. W. P. King, H. R. Mimno, and A. H. Wing, "Transmission Lines, Antennas, and Wave Guides," McGraw-Hill Book Co., New York, N. Y., p. 164; 1945.

[4] D. D. King, "Impedance measurements on transmission lines," Proc. I.R.E., vol. 35, pp. 507–514; May, 1947.

[5] R. King, "Transmission-line theory and its application," *Jour Appl. Phys.*, vol. 14, pp. 577–600; November, 1943.

$$B = \beta s_n + \Phi_0 + \Phi_s = n\pi; \quad n = 1, 2, \cdots, \tag{3}$$

$$A = \alpha s_n + \rho_o + \rho_s = \beta W/2, \tag{4}$$

where α and $\beta = 2\pi/\lambda$ are, respectively, the attenuation and phase constants of the line, W is the full width of the resonance curve, and s_n is the length of the line at resonance. By predetermining Φ_s and ρ_s for the piston with its two detecting loops, measuring s directly, obtaining β from the measured wavelength, and computing the small quantity α from the dimensions and material of the line, ρ_o and Φ_o may be evaluated from (3) and (4).

$Z_0 = R_0 + jX_0$ is determined from curves of ρ_o and Φ_o, using

$$Z_0 = R_0 + jX_0 = \frac{R_c[\sinh 2\rho_o - j \sin 2\Phi_o]}{\cosh 2\rho_o - \cos 2\Phi_o}. \tag{5}$$

Since ρ_o and Φ_o are much more slowly varying than R_o and X_o, it is more accurate to draw smooth curves of ρ_o and Φ_o through the experimental points and use these to determine R_o and X_o than to substitute the experimental values of ρ_o and Φ_o directly in (5). Experimental curves of R_o and X_o are in Fig. 3. In Figs. 4 and 5 experimental points are compared with theoretical curves of the King-Middleton second-order theory.[6]

The vertical slotted line used in the measurements is that described by D. D. King.[7] It is provided with a long taper so that the gap may be kept sufficiently small to make terminal effects negligible. Styrofoam was used for the single insulator at the upper end of the line. The ground screen was a 36-foot square of sheet aluminum. The operating frequency was 500 mc.

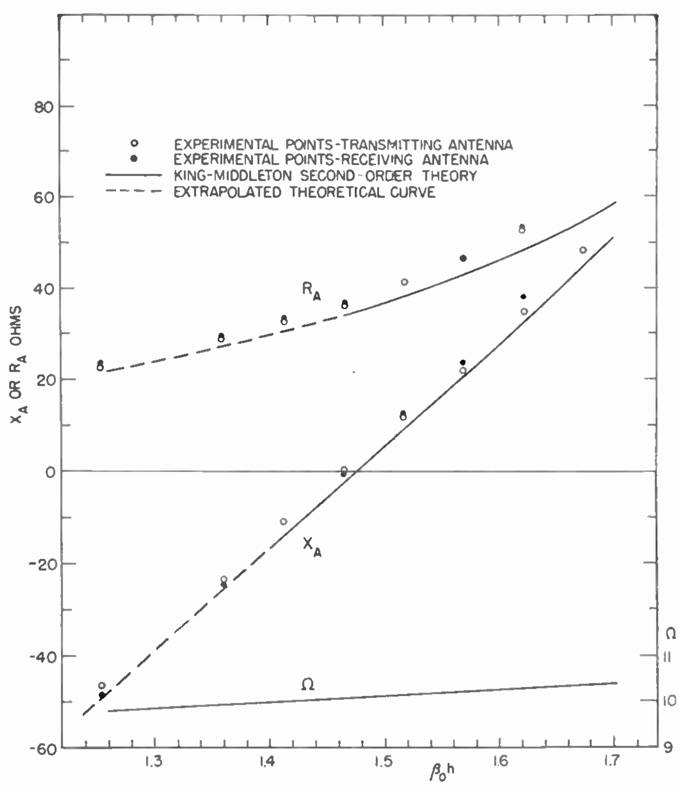


Fig. 4—Theoretical impedance near resonance with experimental points from Fig. 3.

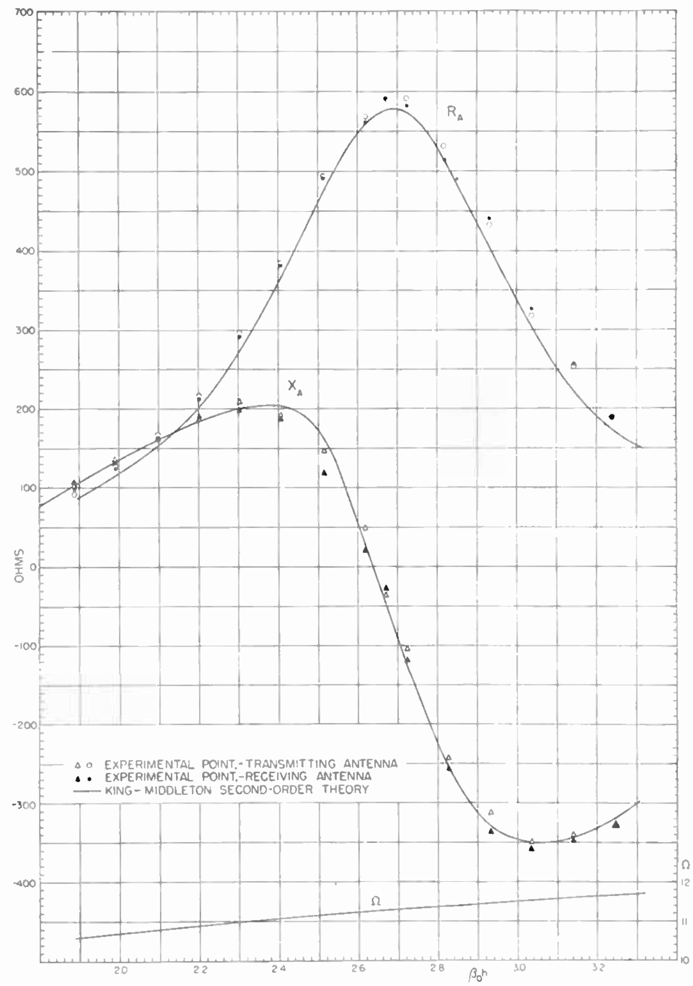


Fig. 5—Theoretical impedance near antiresonance with experimental points from Fig. 3.

The method used to determine ρ_s was to replace the receiving antenna by a metal cap, excite the line by a loosely coupled generator, and measure ρ_s with an auxiliary probe when $\rho_o = 0$. Φ_s, for the piston at the lower end of the line was determined by locating the minimum in the distribution curve using an auxiliary probe.

Although, in general, the measurement of the impedance of an antenna is more convenient when it is driven than when used for reception, since the standing-wave-ratio and distribution-curve methods are available, the receiving-antenna method has the advantage of requiring no transmitter at the location of the measurement. Thus, for example, only light test equipment is needed for measuring the impedance of an antenna in an aircraft or vehicle in motion, using a slotted line with predetermined ρ_s and Φ_s and a signal from a more or less distant high-powered transmitter.

[6] R. King and D. Middleton, "The cylindrical antenna, theory and experiment," *Quart. Appl. Math.*, vol. 3, pp. 302–335; January, 1946; also *Jour. Appl. Phys.*, vol. 17, pp. 273–284; April, 1946.

[7] D. D. King, "The measured impedance of cylindrical dipoles," *Jour. Appl. Phys.*, vol. 17, pp. 844–852; October, 1946.

Contributors to Proceedings of the I.R.E.

D. A. Alsberg (A'46–M'48) was born in Kassel, Germany, on June 5, 1917. He obtained his undergraduate training at the Technische Hochschule in Stuttgart, Germany, and completed his work there in 1938. After coming to the United States, he was engaged in graduate study at the Case School of Applied Science (now Case Institute of Technology) from 1939 to 1940. Following three years as development engineer with several companies in Ohio, he entered the United States Army and served at the Aberdeen Proving Ground, in Maryland and in Europe.

D. A. Alsberg

In 1945, Mr. Alsberg joined Bell Telephone Laboratories, Inc., where he has been concerned with phase, transmission, and related measurements problems in connection with coaxial-cable carrier and microwave radio-relay systems development.

❖

Leo L. Beranek (S'36–A'41–SM'45) was born in Solon, Iowa, on September 15, 1914. He received the B.A. degree, "with distinction," from Cornell College, Iowa, in 1936, the M.S. and D.Sc. degrees from Harvard University in 1937 and 1940, respectively, and an honorary D.Sc. degree from Cornell College in 1946.

Leo L. Beranek

Preceding his appointment to the Massachusetts Institute of Technology staff, as a physics research associate, in February, 1946, Dr. Beranek was associated professionally with Harvard University. From 1937 to 1946, he served successively as a research assistant, instructor, director of sound-control research, and director of Electroacoustic and Systems Research Laboratories. In February, 1947, Dr. Beranek was appointed associate professor of communications engineering in the Department of Electrical Engineering at M.I.T., and technical director of the Acoustics Laboratory.

Dr. Beranek is the author of two books and numerous papers and articles. In 1944, the Acoustical Society of America awarded him their biennial award for outstanding contributions to acoustics. In recognition of "outstanding services to his country," Dr. Beranek received the President's Certificate of Merit, in October, 1948.

Dr. Beranek is a fellow of the American Physical Society, a fellow and associate editor of the Acoustical Society of America, a member of the American Association for the Advancement of Science, and a member of Sigma Xi and Eta Kappa Nu. He has also served as national chairman of the Professional Group on Audio of the IRE, as vice-president of the Acoustical Society of America, and as chairman of the Acoustics Section Z-24 of the American Standards Association. He is now chairman of the Acoustics Panel of the Research and Development Board of the Department of Defense.

❖

J. T. Bolljahn (A'43) was born in Oakland, Calif., in 1918. He received the B.S. and Ph.D. degrees from the University of California in 1941 and 1950, respectively. From August, 1941 until January, 1946, he was employed by the Naval Research Laboratory in Washington, D. C. His work in this position was concerned with the development of aircraft and shipboard antennas. From February, 1946, until September, 1949, he was a member of the staff of the University of California Antenna Laboratory.

J. T. Bolljahn

Dr. Bolljahn joined the staff of the Stanford Research Institute as a senior research engineer in September, 1949. He has recently received a part-time appointment as acting associate professor in the electrical engineering department of Stanford University for the Summer Quarter, 1951. He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

Seymour B. Cohn (S'41–A'44–M'46) was born at Stamford, Conn., on October 21, 1920. He received the B.E. degree in electrical engineering from Yale University in 1942, also the M.S. degree in communication engineering in 1946, and the Ph.D. degree in engineering sciences and applied physics in February, 1948, from Harvard University. From 1942 through 1945, Dr. Cohn was employed as a special research associate by the Radio Research Laboratory of Harvard University. During part of this time, he represented that Laboratory as a Technical Observer with the United States Army Air Force in the Mediterranean Theater of Operations.

Seymour B. Cohn

Since March, 1948, Dr. Cohn has been employed by the Sperry Gyroscope Company as a project engineer in the microwave department. He is a member of Tau Beta Pi and an associate member of Sigma Xi. He is now serving on the Papers Review Committee of the IRE.

❖

Ramsay P. Decker was born on March 26, 1926, in Chicago, Illinois. He served in the United States Army in Italy from January, 1945, to June, 1946. In December, 1948, he received the B.S. degree in electrical engineering from the Technological Institute of Northwestern University.

Ramsay P. Decker

After graduation, Mr. Decker joined the Research Division of the Collins Radio Company, where he has been engaged in work on high-power vacuum-tube control circuitry, propagation-data analysis, and uhf airborne transmitter development.

❖

Milton Dishal (A'41–SM'46) was born on March 20, 1918, in Philadelphia, Pa. He received the B.S. degree from Temple University in 1939, and the M.A. degree in physics in 1941. From 1939 to 1941, Mr. Dishal was a teaching fellow in physics at Temple University.

Milton Dishal

In 1941 he entered the employ of Federal Telecommunications Laboratories, where he is now a senior project engineer, engaged in the development of radio receivers having special characteristics.

❖

Warren G. Findley was born in New York, New York, on September 23, 1906. He received the A.B. degree, with high honors in mathematics, from Princeton University in 1927. In 1929 and 1933, respectively, he received the degrees of M.A. and Ph.D. in educational psychology and statistics from Columbia University.

W. G. Findley

From 1927 through 1938, Dr. Findley served as College Personnel

Contributors to Proceedings of the I.R.E.

Officer at Cooper Union Institute of Technology, during which period he also developed aptitude tests for the College Entrance Examination Board. From 1938 through 1946, he was Assistant Director of Examinations and Testing in the New York State Education Department. He interrupted this work for 3 months in 1944 to supervise the construction of tests for the U. S. Armed Forces Institute at the University of Minnesota. In the capacity of Chief of the Evaluation Branch in the Educational Advisory Staff, he guided the development of tests at the Air University, Maxwell Air Force Base, Alabama, from 1946 through 1948. Since 1948 he has been Director of Test Development for the Educational Testing Service in Princeton, N. J., where a number of the tests and testing programs mentioned in his article, "Using Tests to Select Engineers," are built.

Dr. Findley is a fellow of the American Psychological Association and the American Association for the Advancement of Science, and a member of the American Educational Research Association, the Psychometric Society, and Phi Beta Kappa. He has contributed articles on testing to a number of professional journals and yearbooks.

❖

Irvin H. Gerks (A'32–M'41–SM'43) was born in New London, Wis., in 1905. He received the B.S. degree from the University of Wisconsin in 1927, and the M.S. degree from the Georgia School of Technology in 1932, both in electrical engineering.

Irvin H. Gerks

From 1927 to 1929, Mr. Gerks was employed by Bell Telephone Laboratories, where he developed and tested various special equipment for telephone centrals, such as tone signaling and centralized time announcing. Following this period, he became an instructor in and, later, an assistant professor of, communication and electronic engineering at the Georgia School of Technology, where he remained until 1940.

During the war years, Mr. Gerks was an officer in the Army Signal Corps. He was stationed first at Wright Field and later transferred to the Pacific Theater. At Wright Field he was in charge of the Communication and Navigation Division of the Aircraft Radio Laboratory.

At present Mr. Gerks is employed by the Collins Radio Company, where he is engaged in radio wave-propagation studies.

❖

John Van Nuys Granger (S'42–A'45–M'46) was born in Iowa, in 1918. He received the A.B. degree from Cornell College in 1941, and the M.S. and Ph.D. degrees from Harvard University in 1942 and 1948, respectively. During a part of 1942, he was a member of the staff of the Pre-Radar School at Cruft Laboratory, Harvard University, Cambridge, Mass. In November, 1942, he joined the Radio Research Laboratory of Division 15, OSRD, where he remained until 1945. During that interval he served with the American British Laboratory in Great Malvern, Worcestershire, England, and as a technical observer with the U. S. Air Forces in France and the Low Countries.

J. V. N. Granger

In 1945 he joined the staff of the Central Communications Research Laboratories of Division 13, OSRD, at Harvard, leaving in 1946 to resume his studies. In 1947 he became group leader of the Antenna Group at the Cruft Laboratory. In May, 1949, he was named supervisor of the Aircraft Radiation Systems Laboratory, Stanford Research Institute, Stanford, Calif. He has recently received a part-time appointment as acting associate professor in the electrical engineering department of Stanford University for the Summer Quarter, 1951.

Dr. Granger is a member of the Panel on Antennas and Propagation of the Research and Development Board, the American Physical Society, Sigma Xi, and also of Commission 6 of the U.S.A. National Committee of the URSI.

❖

Elmer O. Hartig (S'47) was born in Evansville, Ind., on January 28, 1923. He received the B.S. degree in electrical engineering from the University of New Hampshire in 1946. He attended Harvard University in the Department of Engineering Sciences and Applied Physics, where he received the S.M. degree in 1947, and the Ph.D. in 1950.

Elmer O. Hartig

From 1944 to 1946, while on active duty in the U. S. Army, Dr. Hartig was associated with the Manhattan Project at Columbia University and Los Alamos, N. M. From 1948 to 1950, while at Harvard's Cruft Laboratory, he served as a research assistant, doing research in antennas and pulsed circuits. Dr. Hartig has been associated with the aerophysics department of the Goodyear Aircraft Corporation since July, 1950, where he is heading the microwave and antenna group. He is a member of Sigma Xi.

John Alexander Kessler was born in Buffalo, N. Y., on December 19, 1920. He received the A.B. degree (cum laude) from Harvard College in 1942, in the Department of Music, and the S.M. degree from Harvard University in 1947, in the Department of Engineering Sciences and Applied Physics (acoustics and communications engineering).

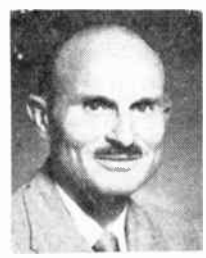
J. A. Kessler

Mr. Kessler has been affiliated with the research staff at the Massachusetts Institute of Technology, and with the Electroacoustic Laboratory at Harvard University. Mr. Kessler is now a staff member of the Division of Industrial Co-operation, and liaison officer of the Acoustics Laboratory at M.I.T.

Mr. Kessler is a member of the Acoustical Society of America, serving on the Music Committee and subcommittee Z 24-X-5 (Measurements of Acoustic Properties of Materials), and is a member of the Society of Sigma Xi.

❖

Ronold King (A'30–SM'43) was born on September 19, 1905, at Williamstown, Mass. He received the B.A. degree in 1927, and the M.S. degree in 1929 from the University of Rochester, and the Ph.D. degree from the University of Wisconsin in 1932. He was an American-German exchange student at Munich from 1928 to 1929, a White Fellow in physics at Cornell University from 1929 to 1930, and a Fellow in electrical engineering at the University of Wisconsin from 1930 to 1932. He continued at Wisconsin as a research assistant from 1932 to 1934. From 1934 to 1936, he was an instructor in physics at Lafayette College, serving as an assistant professor in 1937.



Ronold King

During 1937 and 1938, Dr. King was a Guggenheim Fellow at Berlin. In 1938 he became instructor in physics and communication engineering at Harvard University, advancing to assistant professor in 1939 and to associate professor in 1942. He was appointed Gordon McKay professor of applied physics at Harvard University in 1946.

Dr. King is a Fellow in the American Physical Society, the American Association for the Advancement of Science, and the American Academy of Arts and Sciences. He is a member of Phi Beta Kappa and also of Sigma Xi.

Contributors to Proceedings of the I.R.E.

A. J. Lephakis (A'44–M'50) was born in Glen Cove, L. I., N. Y., on May 14, 1921. Before the war, he attended the Massachusetts Institute of Technology. From 1943 to 1946, he served with the United States Army Air Force. In 1946, he returned to M.I.T., and received the S.B. and S.M. degrees in electrical engineering in 1948 and 1949, respectively. He remained at M.I.T. as a research assistant, and has been associated with the communication group at the Research Laboratory of Electronics since that time.

A. J. Lephakis

Mr. Lephakis is a member of Sigma Xi and an associate of the American Institute of Electrical Engineers.

❖

R. L. Linton, Jr. (S'41–A'44–M'46) was born on February 10, 1921, in Clemenceau, Ariz. He received the B.S. degree in electrical engineering at the University of California in May, 1942. He was employed by the Navy Department from then until April of 1944. For the Navy Department he was engaged in ultra-high-frequency antenna development at the Naval Research Laboratory, and in microwave airborne-radar and test-equipment design at the Bureau of Ships. Mr. Linton then became connected with the University of California at Berkeley. Here he worked at the Radiation Laboratory until June, 1945, when, at the formation of the Antenna Laboratory, he joined that group to be occupied with instrument and antenna development. Concurrently with his activity at the Antenna Laboratory, Mr. Linton was also a part-time lecturer for the Electrical Engineering Division from the fall of 1946 till mid-1949.

R. L. Linton, Jr.

Mr. Linton is now an electronics engineer in the Electronics and Guidance Section of Consolidated-Vultee Aircraft Corporation at San Diego, Calif. He is a member of the American Association for the Advancement of Science and of the Tau Beta Pi Association, and an associate of the Society of Sigma Xi.

❖

John F. Marshall was born on December 21, 1918, in New York City. He received the A.B. degree from Swarthmore College in 1941, and the Ph.D. degree from the University of Notre Dame in 1949.

Dr. Marshall joined the staff of the Bartol Research Foundation as a physicist in 1942. As a research assistant from 1946 to 1949, and an instructor from 1949 to 1950 at the University of Notre Dame, he specialized in theoretical nuclear physics.

J. F. Marshall

At present Dr. Marshall is engaged in the study of the theory of secondary electron emission at the Bartol Research Foundation.

❖

Theodore Miller (S'40–A'42–M'48) was born in Austria, on February 3, 1917. He received the B.S. degree in electrical engineering from the Carnegie Institute of Technology in 1946 and the M.S. degree in mathematics from the University of Pittsburgh in 1948. From 1941 to 1944 he was employed by the C. L. Hofmann Corporation of Pittsburgh as an electronics engineer, in which capacity he did development work on hearing-aid devices. Since 1944 he has been associated with the Westinghouse Research Laboratories, where he has been concerned with microwave research, radar and sonar engineering, and, at present, color-television research. He is the author of several technical papers on microwave apparatus and low-frequency instruments.

Theodore Miller

Mr. Miller is a member of the American Physical Society.

❖

J. R. Moore was born on July 5, 1916, in Saint Louis, Mo. After receiving the B.S. degree in mechanical engineering from Washington University in Saint Louis, in 1937, Mr. Moore joined the General Electric Company, completing the three-year course in advanced engineering in 1940. Continuing in the employ of General Electric until 1946, he became assistant head of the engineering mechanics group of the aeronautics and marine engineering division, and head of the theoretical section, project Hermes, respectively.

J. R. Moore

Mr. Moore was responsible for the development of wartime airborne and naval-gun fire-control equipment and of automatic production machinery, and worked on simulation problems for guided missiles. He was associate professor of mechanics, and director of the Dynamical Control Laboratory at Washington University from 1946 until February, 1948, when he moved to California to join the North American Aviation, Inc., Aerophysics Laboratory staff, where he is now assistant chief of guidance.

Mr. Moore is a member of two RDB groups. Most of his work has been in the fields of computing mechanisms, three-dimensional dynamics, and automatic control. Since the fall of 1948, he has also been a visiting associate professor of engineering at the University of California at Los Angeles, Calif., where he teaches a graduate course in advanced servomechanism theory.

❖

Tetsu Morita (S'44–A'49) was born in Seattle, Wash., on February 5, 1923. He attended the University of Washington from 1940 to 1942, and received the B.Sc. degree in electrical engineering from the University of Nebraska in 1944. He received the M.S. and Ph.D. degrees from Harvard University in 1945 and 1949, respectively.

T. Morita

During 1944 Dr. Morita assisted in the Army Specialized Training Program at the University of Nebraska. From 1946 to 1947 he was a teaching fellow at Harvard University, where he was a research assistant during the period from 1947 to 1949. Dr. Morita has been a research fellow in electronics at Harvard since 1949, and he is, at present, group leader of the Antenna Group at Cruft Laboratory.

❖

Martin A. Pomerantz was born on December 17, 1916, in New York City. He received the A.B. degree from Syracuse University in 1937, the M.S. degree from the University of Pennsylvania in 1938, and the Ph.D. degree from Temple University in 1951.

M. A. Pomerantz

Dr. Pomerantz joined the staff of the Bartol Research Foundation as research assistant in 1938. In 1941, he was promoted to Research Fellow. At present he is engaged, as a physicist, in the field of cosmic radiation, as well as that of secondary electron emission.

Contributors to Proceedings of the I.R.E.

As a leader of a number of National Geographic Society cosmic-ray expeditions, including several to Hudson Bay, Dr. Pomerantz has conducted cosmic-ray experiments at very high altitudes with instruments carried aloft by free balloons.

❖

W. H. Radford

William H. Radford (A'41–SM'45) was born in Philadelphia, Pennsylvania, on May 20, 1909. He received the B.S. degree in electrical engineering from the Drexel Institute of Technology in 1931, and the M.S. degree from the Massachusetts Institute of Technology in 1932. He has been a member of the staff of the department of electrical engineering at M.I.T. since 1932, and is now Professor of Electrical Communications there.

Professor Radford has been active as a consultant to government and industry on radio-communication facilities and specialized electronic applications since 1935. He served as a section member and consultant to the National Defense Research Committee, from 1940 to 1943. In 1941, he assisted in establishing the M.I.T. Radar School, which later became the principal center for training electronics specialist radar officers for the United States Navy. He was closely associated with the radar school throughout its period of operation, and became associate director in 1944. He is now devoting a large portion of his time to supervision of government-sponsored research programs in radio communications.

Professor Radford is a registered professional engineer in the Commonwealth of Massachusetts. He is a member of the American Institute of Electrical Engineers, the American Society of Engineering Education, the American Association for the Advancement of Science, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

W. M. Sharpless

William M. Sharpless (A'28–M'38–SM'43) was born on September 4, 1904, in Minneapolis, Minn. He received the B.S. degree in electrical engineering in 1928 and the E.E. degree in 1951, both from the University of Minnesota. As a member of the technical staff of the Bell Telephone Laboratories since 1928, he has been engaged in radio research projects. Mr. Sharpless' most recent work has had to do with the studies of the angle of arrival of microwaves, and the design of artificial dielectrics and microwave antennas.

Mr. Sharpless is a member of the American Physical Society, and is a licensed professional engineer in the State of New Jersey.

❖

Leo Storch

Leo Storch was born on March 3, 1921, in Vienna, Austria. In January, 1944, he received the B.E.E. degree, cum laude, from the School of Technology, College of the City of New York. He was awarded the M.S. degree by the Graduate School of Stevens Institute of Technology in June, 1947.

From 1944 to 1947, Mr. Storch worked as a development engineer with the Western Electric Company, Kearny, N. J. He was engaged in the circuit design for a wide variety of testing and measurement equipment. Subsequently, he transferred to RCA Victor, Camden, N. J., as advanced development engineer in the aviation radio department, where he was primarily concerned with the development of antenna impedance-matching networks and automatic tuning systems for wide-band aircraft communications transmitters.

Mr. Storch recently joined the Research and Development Laboratories of the Hughes Aircraft Company, Culver City, Calif., as research physicist in the communications group.

❖

James E. Storer

James E. Storer was born in Buffalo, N. Y., on October 26, 1927. He received the A.B. degree in physics in June, 1947, from Cornell University, followed by the M.S. and Ph.D. degrees in in 1948 and 1951, respectively, both from the Department of Engineering Sciences and Applied Physics at Harvard University. While at Harvard, Mr. Storer held an Atomic Energy Commission Fellowship. He is a member of the Sigma Xi Society.

Mr. Storer is at present engaged in research on thin wire antennas at the Electronics Research Laboratory of Harvard University.

J. B. Wiesner

Jerome Bert Wiesner (S'36–A'40–SM'48) was born on May 30, 1915, in Detroit, Michigan. He received his B.Sc. and M.S. degrees from the University of Michigan in 1937 and 1940, respectively. In 1950 he was granted his Doctor of Philosophy degree from the University of Michigan.

Dr. Wiesner was chief engineer of the Acoustical and Record Laboratory of the Library of Congress from 1940 to 1942, at which time he became a member of the staff at the Radiation Laboratory at the Massachusetts Institute of Technology. In 1945, he went to the Los Alamos Laboratory in New Mexico as a member of the staff, returning to M.I.T. in 1946. He has been assistant director of the Research Laboratory of Electronics at M.I.T. for the past three years, and is now its associate director. In addition to his work in the laboratory, he is professor of electrical engineering at M.I.T.

Dr. Wiesner is a member of the Acoustical Society of America, the Federation of American Scientists, and the American Association for the Advancement of Science. In addition, he is a member of Eta Kappa Nu, Sigma Xi, and Phi Kappa Phi.

❖

Donald G. Wilson

Donald G. Wilson (S'38–A'40–M'48) was born in Bridgeport, Conn., on September 20, 1917. He received the B.E.E. degree from Rensselaer Polytechnic Institute in 1938, and the S.M. degree in communication engineering from Harvard University in 1939. After a year in industry, he taught for two years at Rensselaer.

From 1941 to 1945, Mr. Wilson was engaged in microwave propagation research at the Radiation Laboratory of the Massachusetts Institute of Technology. In 1945 he returned to Harvard University for graduate study, and received the Ph.D degree in 1948.

He joined the department of electrical engineering at the University of Kansas in 1947 as an associate professor. In 1948 he was appointed chairman of the department of electrical engineering, and was advanced to the rank of professor in 1950.

Dr. Wilson is a member of the American Institute of Electrical Engineers, Sigma Xi, and Tau Beta Pi.

Institute News and Radio Notes

Calendar of COMING EVENTS

Third Annual Technical Conference, Kansas City Section, Hotel President, Kansas City, Mo., November 16, 17

First JETEC General Conference, Absecon, N. J., November 29–December 1

IRE Nuclear Symposium, Brookhaven National Laboratory, Upton, L. I., N. Y., December 3, 4

Joint IRE/AIEE Computer Conference, Benjamin Franklin Hotel, Philadelphia, Pa., December 10–12

Symposium on Williams Electrostatic Storage, National Bureau of Standards, Washington, D. C., December 13–14

IAS-ION-IRE-RTCA Conference on Air Traffic Control, Astor Hotel, New York, N. Y., January 30

1952 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6

IRE National Conference on Airborne Electronics, Hotel Biltmore, Dayton, Ohio, May 7–9

4th Southwestern IRE Conference and Radio Engineering Show, Rice Hotel, Houston, Tex., May 16–17

Technical Committee Notes

The August meeting of the **Standards Committee** was preceded by a meeting of the Administrative Committee. A. G. Jensen, presided as Chairman at both meetings. W. H. Pease was named as Chairman of the new **IRE Technical Committee on Servo-Systems**, recently organized, and W. D. Goodale, Jr., was appointed Chairman of the **IRE Technical Committee on Electroacoustics.** C. H. Page opened the question as to what the Committee's opinion was on the use of formulas in definitions. It was pointed out that some terms are inherently mathematical and could be defined best by giving a formula, although the tendency in the past has been to avoid this use. The Committee decided that it is better to use formulas, rather than using long, wordy, and often unclear, definitions. The Standards Manual will be revised shortly and this concept will be reflected in the Manual. Other additions to the Manual will be welcomed by Messrs. Weber, Baldwin, and Jensen.

The **Standards Committee** which convened on September 13, under the Chairmanship of A. G. Jensen, recommended for publication a paper on, "Fundamental Considerations Regarding the Use of Relative Magnitudes," by J. W. Horton. Dr. Horton is the Chief Consultant of the U. S. Navy Underwater Sound Laboratory, New London, Conn. His paper will be considered for an early publication in the PROCEEDINGS, by the editorial department.

The **Committee on Antennas and Waveguides** held a meeting on September 11, under the Chairmanship of Gardner Fox. This Committee is still engaged in the preparation of definitions which will be submitted to the Standards Committee for approval.

Consideration is being given to the necessity for broadening the scope of the **Mobile Communications Committee** and to the possibility of changing the name of the Committee to include land-mobile, air-mobile, and sea-mobile communications.

A Task Group of the **Receivers Committee** has prepared a paper on Methods of Calibration of Radiation.

Work is underway towards the preparation of the Report of the **Annual Review Committee.** Ralph Batcher, Chairman of the Annual Review Committee has requested that each Technical Committee Chairman appoint a member to the Annual Review Committee. The deadline for the receipt of material for the Annual Review Report is November 15.

A meeting of the **Measurements and Instrumentation Committee** was held on Sept. 14, F. J. Gaffney, Chairman. The Chairman announced that it would not be necessary to reactivate the **Subcommittee on Basic Techniques of Instrumentation** as the work of this Subcommittee has been taken over by the **Subcommittee on Basic Standards and Calibration Methods (25.1).** J. L. Dalke, Chairman of **Subcommittee 25.2, Dielectric Measurements,** reported that his Subcommittee is endeavoring to extend the frequency range presently defined and will redefine a series of terms on the low-frequency range. Definitions are needed that will extend over the entire frequency range, realizing of course that different methods of measurement will be required for the various frequency ranges. This Subcommittee 25.2 is also planning to review standards of other organizations and, if possible, incorporate them into IRE Standards. A report of the work in progress in **Subcommittee 25.3, Magnetic Measurements,** was presented by F. J. Gaffney in the absence of the Chairman, C. D. Owens.

Subcommittee 25.4, Audio-Frequency Measurements, under the Chairmanship of Dr. Peterson is collecting material for use in their standardizing work. **Subcommittee 25.11, Statistical Quality Control,** under the Chairmanship of Mr. Steen is endeavoring to prepare a bibliography of articles and information which will be helpful to the field of Quality Control. Another project of this Subcommittee is to make a survey of the field of other committees, subcommittees, and organizations toward the end of avoiding duplication of effort.

A. V. Loughren, Vice-President in Charge of Research, Hazeltine Electronics Corporation, has been appointed to replace Haraden Pratt on the **Joint Technical Advisory Committee.**

The IRE-IAS-ION-RTCA Symposium on Electronics in Aviation will be held on January 30, 1952 in New York City. Details will be announced as they are formulated.

Feedback Control Conference Planned

A special two-day conference on major developments in feedback control systems will be held December 6–7, at the Chalfonte-Haddon Hall Hotel, Atlantic City, N. J., it has been announced. The meeting will be sponsored by the American Institute of Electrical Engineers Committee on Feedback Control Systems.

Not only will developments of the past few years be reviewed during the conference but the program provides for covering many new, significant advances in design and components, according to Jerome Zauderer, chairman of the information committee.

Of special interest at the Atlantic City meeting will be a session on biological servo-mechanisms. Topics to be discussed also will include the role of statistics, discontinuous control systems, control systems operating from discrete data inputs, and a sampled data control system technique.

Scheduled to appear on the program are scientists and engineers from Massachusetts Institute of Technology, Columbia University, Minnesota, University of Connecticut, and representatives of several large industrial organizations.

The program and other information can be obtained by writing to Mr. Jerome Zauderer, American Measuring Instruments Corporation, 21–25 44th Avenue, Long Island City, N. Y.

Houston Section to Sponsor Southwestern IRE Conference

The Fourth Southwestern IRE Conference and Radio Engineering Show will be held in the Rice Hotel, at Houston, Texas, May 16 and 17, 1952. The Houston Section will act as host and sponsor.

Provisions have been made for the holding of all meetings, technical sessions, and equipment exhibits on one floor. Space will be available also in case of simultaneous technical sessions. Equipment exhibitors desiring to demonstrate sound producing apparatus will have isolated rooms on the same floor at their disposal.

The Houston Section's publication *Scope* plans two special Conference issues for the months of March and May, to be given region distribution.

Professional Group Notes

The **IRE Professional Group on Antennas and Propagation** held a joint meeting with the U. S. National Committee, URSI, on October 8, 9, and 10, at Cornell University, Ithaca, N. Y. C. R. Burrows of the School of Electrical Engineering, Cornell, and A. H. Waynick of Pennsylvania State College took charge of the program which included subjects on Radio Standards, Measurements, Tropospheric Propagation, Astronomy, and Ionospheric Propagation.

The **IRE Professional Group on Broadcast and Television Receivers** held an Administrative Committee meeting during the Radio Fall Meeting at the King Edward Hotel in Toronto, Ontario, October 29, 30, and 31.

The petition for the formation of an **IRE Professional Group on Electronic Computers** was approved by the Committee on Professional Groups at the last meeting. This Group is being promoted on the West coast by H. T. Larson of Hughes Aircraft Company, Culver City, Calif., and on the East coast by M. M. Astrahan of International Business Machines, Poughkeepsie, N. Y. Through the combined efforts of both promoters, the Group has a nationwide representation on its Administrative Committee.

Nathan Marchand, Chairman of the **IRE Professional Group on Information Theory,** has announced the appointment of D. L. Trautman of the University of California, as a member of the Administrative Committee of the National Professional Group. Professor Trautman is Chairman of the Los Angeles Group Chapter of the Information Theory Professional Group.

Eugene Mittelmann, Chairman of the **IRE Professional Group on Industrial Electronics** has reported that the recent membership drive conducted in the Sections of Akron, Chicago, Cleveland, Los Angeles, Milwaukee, New York, Pittsburgh, and Schenectady, has increased the membership of the Group to 848, as of September 15, 1951.

The petition for the formation of an **IRE Professional Group on Electron Devices** was approved by the Committee on Professional Groups at its last meeting. The Administrative Committee of this Group is being formulated; G. D. O'Neill, of Sylvania Electric Products Company, Bayside, L. I., is acting Chairman.

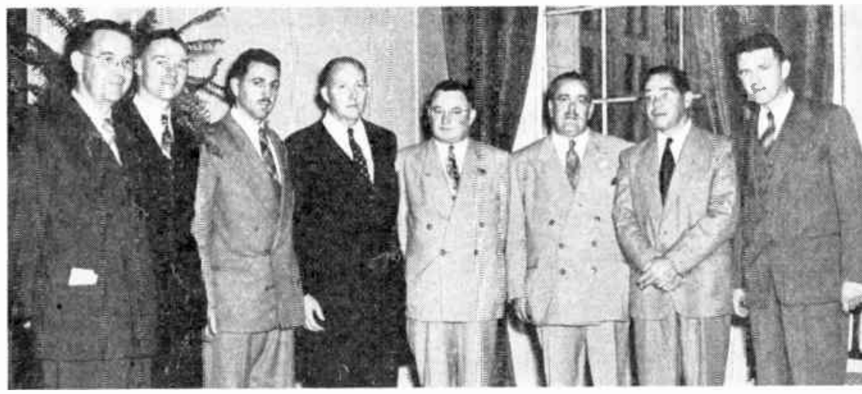
An IRE Symposium at Brookhaven National Laboratory, December 3 and 4, is being sponsored by the **IRE Professional Group on Nuclear Science** in co-operation with the Atomic Energy Commission. The afternoon and evening of December 3, will be composed of a program of papers on Nuclear Developments, with a social hour at 5:30, followed by dinner at 7:00, and resuming at 8:15 with papers on Atomic Energy, plus a panel discussion. The second session on the morning of December 4, consists of papers on Scintillation Counters, Functions of Research Reactors, Nuclear Reactors, and Fission Products in Industry.

A new **IRE Professional Group on Microwave Electronics** is being formulated. Those interested in becoming members of this Group are requested to contact Ben Warriner, IV, 151 Jefferson Avenue, Thornwood, N. Y. The scope of this Group is to encompass microwave theory, circuitry and technique, microwave measurements, tubes.

A national meeting on Land Mobile Communications was held at the Sheraton Hotel, Chicago, Ill., October 25 and 26, sponsored by the **IRE Professional Group on Vehicular Communications.** The welcoming address was given R. V. Dondanville, Chairman of the Chicago Group Chapter. Austin Bailey, Chairman of the National Group, and F. T. Rudelman, Vice-Chairman, acted as Moderators for the two-day session. Guest speakers at the dinner on October 25, and luncheon, October 26, were E. M. Webster of the Federal Communication Commission and J. D. O'Connell, Office of the Chief Signal Office, U. S. Army, respectively. Interesting papers on different phases of mobile communication were presented by: G. C. Terrell, L. P. Morris, M. E. Bond, J. F. Byrne, R. H. McRoberts, A. A. Macdonald, H. H. Davids, M. R. Friedberg, C. P. Williams, D. W. Bodle, J. S Brown, H. K. Lawson.

The **IRE Professional Group on Audio** was sponsor to the Audio Session at the National Electronics Conference held last month in Chicago. The Audio Session of the Radio Fall Meeting in Toronto, Canada was also sponsored by the Audio Group, under the Chairmanship of Frank Slaymaker of the Stromberg-Carlson Company. Plans are being formulated to transmit the papers presented at these Audio Sessions to all members of the IRE/PGA who have paid their assessments. A meeting of the Administrative Committee of this Group was held October 22, 1951, in Chicago. The September, 1951, issue of the NEWSLETTER of the IRE/PGA contains a technical editorial by H. F. Olson, entitled "Selecting a Loudspeaker." An announcement has been made in the NEWSLETTER of the arrangements made by the Group with Shure Brothers, Incorporated, to send a Shure Reactance Slide Rule to all members of the Group who send in the coupon contained in the NEWSLETTER. A similar arrangement has been made with the Ohmite Manufacturing Company by which they will send an Ohmite Ohm's Law Calculator. Previous plans had been made with the Mark Simpson Manufacturing Company, to send a Masco Sound Surveyor Slide Rule to all members of the Group on the same no-charge basis.

President Coggeshall Visits Vancouver Section



A most cordial reception was given President I. S. Coggeshall by the Vancouver Section of the IRE, during a recent trip to the West Coast. Section officers meeting with President Coggeshall are (left to right): Lorne Kersey, Student Branch Co-ordinator; J. S. Gray, Membership Committee; A. H. Gregory, Vice-Chairman; President I. S Coggeshall; G. C. Chandler, Chairman; D. D. Carpenter, Secy-Treas.; B. R. Tupper, Junior Past Chairman; and Miles Green, Asst. Secy-Treas.

IRE-AIEE Computer Conference Program Announced

The Joint IRE-AIEE Computer Conference to be held December 10, 11, and 12, in Philadelphia, Pa., will present papers from some 12 computing groups or manufacturing organizations which have succeeded in obtaining useful results from present large-scale digital computers. The following program lists the machines which will be discussed.

Monday, December 10, Benjamin Franklin Hotel:

1101, Engineering Research Associates; TESTRAC, Burroughs Adding Machine Corporation; UNIVAC, Remington Rand, Eckert Mauchly Division.

Tuesday, December 11, Edison Building:

Card-Programmed Electronic Calculator, IBM Corporation; Institute for Advanced Study Machine, Institute for Advanced Study, Princeton, N. J.; ORDVAC, University of Illinois; SWAC, Institute for Numerical Analysis, Los Angeles, Calif.; Harvard MARK III, Dahlgren Proving Grounds; University of Manchester Electronic Computer, Ferranti Limited, England.

Wednesday, December 12, Benjamin Franklin Hotel:

Whirlwind I, Massachusetts Institute of Technology; EDSAC, Cambridge University, Eng.; SEAC, Bureau of Standards, Washington, D. C.

A Keynote address by W. H. MacWilliams Jr., of the Bell Telephone Laboratories will precede the formal presentation of papers. J. W. Forester will summarize the Conference on Wednesday afternoon. Inspection trips have been arranged for visiting the Bureau of Census UNIVAC at Remington-Rand, Eckert-Mauchly Division, Burroughs Research Division, Moore School of Electrical Engineering, University of Pennsylvania, and Technitrol Engineering Company. Reservations for trips will be made at the Conference due to the space limitations of these groups.

Notice

IRE Members

Rates of the Proc. of the IEE

There has been a slight increase due to publication costs in the reduced rates given to those members of the IRE who subscribe to the *Proceedings of the Institution of Electrical Engineers.* These rates, effective January 1, 1952, are as follows:

	Sterling
Part I (General)	12s. 6d.
Part II (Power Engineering)	17s. 6d.
Part III (Radio and Tele-communications Engineering)	17s. 6d.
Part IV (Collected Monographs)	7s. 6d.
All Four Parts Together	£2.10s. 0d.

Student Awards Announced

The IRE Board of Directors established a plan last year whereby students in the different Student Branches may be given an award by the local Section. The Annual IRE Student Branch Awards for 1951, as well as the name of the Student Branch in which the student winner was enrolled, are listed as follows:

Student Branch Award Winner	*Student Branch*
R. D. Wengenroth	Rensselaer Polytechnic Inst.
G. M. Badoyannis	Rutgers Univ.
C. C. Townsend	Princeton Univ.
J. E. Shea	Univ. of Connecticut
T. G. Lynch	Univ. of British Columbia
R. D. Gloor	Univ. of Louisville
E. J. Breiding	Univ. of Kentucky
S. G. Bogan	Yale Univ.
C. G. Blanyer	Univ. of Illinois
J. R. Wood	Utah State Agricultural College
D. J. Groszewski	Univ. of Dayton
R. H. Wilcox	Lafayette College
H. D. Ruhl, Jr.	Michigan State College
R. O. Rheaume	Univ. of Detroit
F. H. Tendick, Jr.	Univ. of Michigan
H. J. Hummel	Illinois Inst. of Technology
R. A. Huwe	Univ. of Minnesota
J. R. Hall	Seattle Univ.
O. W. Fix	Univ. of Washington
D. O. Martin	Southern Methodist Univ.
V. C. Hathaway	Northwestern Univ.
E. C. Pauly	Univ. of Florida
D. M. Culler	Carnegie Inst. of Technology
H. R. Stillwell	Univ. of Pittsburgh
W. J. Huhtala	Michigan College of Mining and Technology
R. C. Ritchart	Univ. of Wisconsin
A. F. Petrie	Marquette Univ.
W. F. Kyle	California State Polytechnic College
F. R. Goodman	California Inst. of Technology
J. E. Niebuhr	Univ. of Southern California
P. H. McBride	Univ. of Miami
K. O. Timothy	Univ. of Utah

UHF Symposium Held

The IRE Professional Group on Broadcast Transmission Systems sponsored an UHF Symposium, the first of its kind, which was held at the Franklin Institute, Philadelphia, Pa., Sept. 17. Over 150 persons from 20 states heard eight leading specialists present outstanding papers on this all-important phase of television. The talks, well illustrated and supported by exhibits, were presented in the following order.

"Some Experiments with 850-Mc Television Transmissions in the Bridgeport, Connecticut, Area," G. H. Brown, RCA Laboratories Division; "DuMont 700-Mc UHF Installation," William Sayer, Jr., and Elliot Mehrbach, Allen B. DuMont Laboratories, Inc.; "Impedance and Frequency Measurements at UHF," R. A. Soderman and F. D. Lewis, General Radio Company; "Side Fire Helix UHF Television Transmitting Antenna," L. O. Krause, General Electric Company; "A Fundamental Approach to UHF Television Receivers," W. B. Whalley, Sylvania Physics Laboratories; "Progress Report on the RCA-NBC UHF-Project at Bridgeport, Connecticut," Raymond Guy, National Broadcasting Company; "Transmission Line Problems in the UHF Television Band," J. M. DeBell, Jr., Allen B. DuMont Laboratories; "An Electronic Radio Field Strength Analyzer for Use in Television Station Field Surveys," F. W. Smith, National Broadcasting Company.

Lewis Winner of *TeleVision Engineering* and Chairman of the IRE Professional Group on Broadcast Transmission Systems, opened the symposium with introductory remarks, and an "UHF Information Please Roundtable," completed the day.

IEE TV Convention Slated

The committee of the radio section, acting on behalf of the Council of the Institution of Electrical Engineers, are arranging a convention to be known as "The British Contribution to Television," to be held in London, April 28 to May 3, 1952, and they cordially invite IRE members to attend.

The convention will be organized in nine sessions covering the complete field of television. Each session will be devoted to the presentation and discussion of technical papers, and will be supported by demonstrations where applicable. Visits of inspection to organizations concerned with every aspect of television will be included in the program.

The sessions into which the convention will be divided are as follows: Program origination; Point-to-point transmission; Broadcasting stations; Propagation; Receiving equipment (2 sessions); Nonbroadcasting applications; and System aspects. An historical paper and a broad survey paper to act as an introduction to the whole convention will also be presented.

It is expected that proofs of all papers will be available in advance. A special issue of *The Proceedings of the Institution of Electrical Engineers* will be published containing all proceedings of the convention.

Full particulars, and a form of application for registration of those members wishing to attend the convention, will be issued shortly to all the regular recipients of Part III (Radio and Communication Engineering) of *The Proceedings of the Institution of Electrical Engineers.* Members who do not subscribe to this publication and who wish to take advantage of this invitation should notify the Executive Secretary of the Institute of Radio Engineers, 1 E. 79 St., New York 21, N. Y., so arrangements can be made for the supply of registration forms.

UHF Symposium Speakers

Speakers at the UHF Symposium of the IRE Professional Group on Broadcast Transmission Systems, held September 17, at the Franklin Institute in Philadelphia, are (left to right): front row; F. W. Smith, NBC; L. O. Krause, General Electric; R. A. Soderman and F. D. Lewis, General Radio; W. B. Whalley, Sylvania Electric; G. H. Brown, RCA Laboratories Division, RCA; William Sayer, Jr. and Elliot Mehrbach, DuMont; J. M. DeBell, Jr., DuMont; and Raymond Guy, NBC. In the rear appear the two Moderators for the session (left to right): D. D. Israel and S. L. Bailey.

IRE People

Bruce Williams (S'47–A'50) has been appointed as sales engineer for the John A. Green Company, Dallas, Tex. In his position he will call on industrial accounts, jobbers, research laboratories, and manufacturers in the states of Texas, Oklahoma, Louisiana, Arkansas, and New Mexico.

Bruce Williams

Mr. Williams was born on October 20, 1919, in Maryland, and attended the University of Cincinnati and the Oklahoma A and M College, Stillwater. He received his B.S. and M.S. degrees in electrical engineering at Oklahoma A and M College, in 1947 and 1949, respectively. He was also an instructor in electrical engineering from 1947–1948. Mr. Williams was a research engineer at the Field Research Laboratory of the Magnolia Petroleum Company, Dallas, until he joined the staff of the John A. Green Company.

Mr. Williams recently was Exhibits Chairman for the 1951 Southwestern IRE Conference which was held in Dallas. He is a member of Delta Tau Delta fraternity, and a member of the AIEE and the ARRL.

❖

O. L. Angevine, Jr. (S'36–A'37–SM'44), formerly chief engineer of the sound equipment division, Stromberg-Carlson Company, recently accepted an appointment as chief engineer of the Caledonia Electronics and Transformer Corporation, it has been learned.

Born in Rochester, N. Y., in 1914, Mr. Angevine received the B.S. degree in electrical engineering in 1936 from the Massachusetts Institute of Technology. Upon graduation he joined the staff of the Stromberg-Carlson Company as an engineer in the telephone laboratory. He was appointed staff engineer for the vice-president in charge of engineering in 1941, and chief engineer of the sound equipment division in 1946.

A former Chairman of the IRE Professional Group on Audio and of the Rochester Section, Mr. Angevine is active on the Video and Audio Techniques Committee. He belongs to the American Institute of Electrical Engineers, to the Acoustic Society of America, and is currently serving as chairman of the sound equipment section, engineering department, of the Radio-Television Manufacturers Association

❖

Marvin Hobbs (A'35–M'41–SM'43) has been named adviser to the Chairman of the Munitions Board. In this capacity he will co-ordinate all phases of the Defense Department's planning to meet the requirements for military electronics production. He will assist the Vice-Chairman for Production and Requirements, and the Military Director for Production.

Mr. Hobbs was born in Kyana, Ind., in 1912. He received the degree of B.S.E.E. at Tri-State College, Ind., in 1930. His engineering and production experience in the radio electronics and television industry extends over a period of 20 years. During World War II, he was associated with the War Production Board and with the Army Air Forces in the Pacific area. After the war he worked as a consulting engineer in Chicago for a number of radio and television manufacturers, including RCA and the Scott Laboratories. In 1950 he was appointed Deputy Executive Director of the Electronics Division of the Munitions Board in the Department of Defense, and has worked in this capacity until now.

❖

Quincy A. Brackett (M'41–SM'43), an assistant to Dr. Lee DeForest, inventor of the radio tube, died recently at the St. Andrew's Hospital, Boothbay, Me., after a long illness. His age was 66.

Mr. Brackett was graduated from Harvard University in 1907 and worked in the New York City laboratories of the Western Electric Company with Dr. DeForest for three years. After being associated with radio station KDKA, he was employed by the Westinghouse Electric Corporation during World War I, supervising production of radio equipment for the Army and Navy.

From 1921 to 1935, he worked with Westinghouse's East Springfield, Mass. plant, where radio transmitters, receivers, and other equipment were produced. In 1935, he helped found radio station WSPR in Springfield, and was president of that station from then until last spring, when he went into semi-retirement as vice president.

E. U. Condon (M'42–SM'43), Director of the National Bureau of Standards and noted nuclear physicist, has been appointed as director of research and development of the Corning Glass Works, Corning, N. Y. Dr. Condon has resigned as Director of the National Bureau of Standards, effective September 30, 1951.

For a photograph and biography of E. U. Condon, see page 707 of the June, 1951, issue of Proceedings of the I.R.E.

Norman L. Winter (A'47–M'47), chief sales engineer for Sperry Gyroscope Company, Great Neck, L. I., N. Y., has been appointed Chairman of the Navigation Technical Group of the Research and Development Board, Department of Defense, it was announced. This group was recently established to advise the Board on the integration and consolidation of air, land, and sea navigation research and development projects.

N. L. Winter

Mr. Winter has been associated with the Board since its establishment. He was Executive Director of the Committee on Electronics from 1946 until 1949, when he joined the staff of Sperry. Since that time he served the Board in a part-time capacity as a consultant on electronic, aircraft, and navigation problems.

In 1942 Mr. Winter was called to active duty with the Army in the Office of the Chief Signal Officer, in command of the Electronics Branch of the Engineering and Technical Service. For his services in this and other capacities during World War II, he was awarded the Legion of Merit.

From 1929 to 1941 he was employed by the General Electric Company, successively, as meter design engineer, motor design engineer, and motor application engineer.

Mr. Winter was graduated from Purdue University with the B.S. degree in electrical engineering, and has done work at the Massachusetts Institute of Technology, the University of Indiana, and Harvard.

He is a member of the American Institute of Electrical Engineers, the Indiana Engineering Society, the Air Force Association, and the Army Signal Association.

❖

Leo G. Sands (A'44–M'45–SM'50), formerly staff assistant to the general sales manager at Bendix Radio, was named director of public relations and advertising at that organization, it has been learned. A specialist in the design of control amplifier circuits, Mr. Sands has been actively concerned with the entire field of radio and electronic circuits. His career has included the following positions: Chief Inspector of Airborne Radar Equipment for the U. S. Army at the Sacramento Air Depot; service manager for various radio stores on the west coast, installations engineer of the Remler Company; and design engineer of the Coast Radio Company, San Jose, Calif.

Cyrus D. Backus (A'19–M'26–SM'43), former examiner in the radio division of the Patent Office, Washington, D. C., and an authority in the communications field, died recently at his home in Silver Spring, Md.

A native of Groton, N. Y., Mr. Backus was graduated in law and philosophy from Cornell University in 1896. In 1903 he received the degree of M.E.E. from George Washington University and later was named head of its electrical communication department. Mr. Backus was chief of Division 51, the principal radio division of the Patent Office, from the early years when that division was formed. He was associated with the Patent Office for 40 years, retiring in 1943, and for four years was patent law consultant of the International Telephone and Telegraph Company in New York, N. Y. He then returned to private law practice. A member of the District Bar in Washington, D. C., he was admitted to practice before the Court of Customs and Patent Appeals and the Supreme Court.

Mr. Backus was a Fellow of the American Association for the Advancement of Science and belonged to the American Institute of Electrical Engineers.

J. W. Hines (S'46–A'48) has been appointed as the new sales engineer of Magnecord Incorporated, Chicago, Ill. Mr. Hines will do liaison work between the engineering and sales department of Magnecord Incorporated, and will also handle technical service and sales problems.

J. W. Hines

Mr. Hines was born in Wilkinsburg, Pa., on September 26, 1923, and received his degree of B.S.E.E. at the Carnegie Institute of Technology in 1947. He has been active in the electronics field for the past 8 years, and served as a newsreel cameraman for the Signal Corps in Europe, during World War II.

Previous to his staff appointment with Magnecord, Mr. Hines was chief engineer for radio station WBVP, Beaver Falls, Pa.

Mr. Hines is a member of the Chicago section of the IRE.

❖

William J. Warren (SM'46) has joined the staff of Shell Development Company in Emeryville, California, and has been assigned to the Associate Directors' staff.

Mr. Warren was born in 1910 at Eureka, Calif. He received the B.S. degree in electrical engineering in 1931, from Santa Clara University and the Ph.D. degree in 1936, from the University of Illinois, where he taught from 1934 to 1937 and from 1938 to 1941. He was employed from 1937 to 1938 by the General Electric Company at Schenectady, N.Y., as a test engineer.

Before joining the Shell Development Company staff, Mr. Warren was affiliated with the University of Santa Clara where he was Professor of electrical engineering and part-time consultant for several industrial firms, since 1941.

Mr. Warren is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, Phi Kappa Phi, and a member of the AIEE.

❖

Loyd C. Sigmon (A'29–SM'46) has been awarded an honorary degree of electrical engineer by the Milwaukee School of Engineering during their June, 1951 commencement exercises, at which he was their featured speaker. Mr. Sigmon, who is chief engineer, vice president, and assistant general manager of radio station KMPC, Hollywood, Calif. advised the graduates that they have a knowledge which is needed now as in no other time in American history.

In prior years Mr. Sigmon attended the school of electrical engineering, Milwaukee, Wis., and from 1935 to 1940, was chief engineer for radio station KCMO, Kansas City, Mo. He then held the position of director of engineering for KMPC, Los Angeles, until his entry into the Armed Services, during World War II, where he served with the Army Signal Corps, with the rank of Lieutenant Colonel, as radio officer in the Signal Corps Communications Division of the European Theatre of Operations.

Mr. Sigmon was born on May 5, 1909, in Stigler, Okla. He is the holder of the Legion of Merit and the Order of the British Empire. He is an honorary member of the French Signal Corps.

❖

H. I. Romnes (SM'46), formerly general manager of the long lines department of American Telephone and Telegraph Company, has recently been appointed director of operations.

Mr. Romnes was born in Wisconsin, in 1907, and received the B.S. degree in electrical engineering from the University of Wisconsin in 1928. His professional experience includes service with the Wisconsin Telephone Company, and membership on the technical staff of the Bell Telephone Laboratories, Inc. In January, 1945, he joined the American Telephone and Telegraph Company, in charge of the toll transmission troup in the operation and engineering department, where he was responsible for long-distance transmission facilities of all types.

In 1950 Mr. Romnes became associated with the Illinois Bell Telephone Company, Chicago, Ill., as chief engineer, but returned to AT&T in December of that year.

Ernest R. Cram, a charter member of the IRE and pioneer in wireless telegraphy, died recently in the Long Island College Hospital, Brooklyn, N. Y. He was 70 years of age.

A native of Boston, Mass., Mr. Cram received his education at Harvard University and was first associated with the Stone Wireless Telegraph Company of Boston, for which he took out several patents on tuning circuits. Later, he joined the U. S. Signal Corps, Washington, D. C., as civilian engineer, and after 25 years of service in this line became associated with the Radio Corporation of America, with respect to patent and legal matters.

In the Signal Corps, his chief duties were the development of mica transmitting condensers, which led to their later commercial manufacture by the Dubilier Condenser Company, and as assistant to (then) Major George Owen Squier, U. S. Army Signal Corps, in experimenting with wired wireless, forerunner of the coaxial cable. He also assisted, for a time, John Hays Hammond, Jr., in the development of radio-controlled torpedoes.

Mr. Cram was founder of the Society of Wireless Telegraph Engineers, which later became part of the IRE.

J. W. Nelson, Jr. (A'46–SM'47) has been appointed as the manager of the newly formed application engineering section within the General Electric Government Sales Department, Syracuse, N. Y. His primary association with the new section will be to work closely in the field with all branches of the armed services, assisting them in the use of present electronic equipment, and locating possible application for which new electronic devices might be developed.

J. W. Nelson, Jr.

A native of Berkeley, Calif., Mr. Nelson received the degree of B.S. in electrical engineering from the University of California in 1941. During 1941–1942, he served as a research associate in the radiation laboratory at Massachusetts Institute of Technology, Cambridge, Mass., engaged in microwave research. Mr. Nelson served the next four years with the Air Force in World War II, as a development engineering officer, and later joined General Electric as a development engineer in their Government Division at Syracuse. He became sales engineer in 1947, and in 1949 he was named sales manager for the Air Force equipment section of General Electric. This was the position Mr. Nelson held prior to this announcement.

He is a member of Sigma Xi and RESA Tau Beta Pi, and Eta Kappa Nu.

Books

Propagation of Short Waves Edited by Donald E. Kerr

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 692 pages+22-page index+14-page appendix+xvii pages. 299 figures. 6×9¼. $10.00.

The next-to-last volume to be issued as part of the Radiation Laboratory Series will be a most welcome addition to the libraries of those working in the field of tropospheric radio-wave propagation. One cannot help but be immediately impressed by the fact that here, at hand, is a single volume containing an excellent summary of hundreds of wartime reports on vhf and uhf transmission. In spite of the many problems involved in the selection and review of these reports, the finished text is well constructed and concise, without too much loss of pertinent detail.

Naturally, it was impossible to incorporate all of the various phases of uhf propagation phenomena into this one volume. Very little material appears relating to diffraction by obstacles or forecasting techniques; however, in both instances, the subjects were either not studied thoroughly by Group 42, or overlapping material has already appeared in print. This is not mentioned as detracting from the usefulness of the volume, but only to illustrate its limitations.

The coverage of the book may best be summarized in a simple listing of the chapters. These are: "Elements of the Problem," "Theory of Propagation in a Horizontally Stratified Atmosphere," "Meteorology of the Refraction Problem," "Experimental Studies of Refraction," "Reflections from the Earth's Surface," "Radar Targets and Echoes," "Meteorological Echoes," and "Atmospheric Attenuation." The volume also contains an Appendix dealing with two additional views on the mathematical theory of scattering. These are the Lorentz Reciprocity Theory and Coherent vs. Incoherent Scattering.

This reviewer was particularly interested in the sections of the book which developed the radio-meteorology relationship. This is a subject that has sorely needed a comprehensive background, understandable to the radio engineer. Chapter 3 accomplishes this purpose and should find considerable use as a standard reference in this field. The qualitative aspects of the numerous experiments performed in order to correlate radio-meteorological pattern are extremely interesting. The treatment of this material here is much more descriptive and illustrative than it is in any comparable volume published to date. Accent is primarily placed upon modified-index distribution (M-profile) which renders a good agreement over medium-length paths. The higher fields over longer pathlengths may probably be accounted for by theories developed after the dissolution of Group 42.

A fair section of this book is devoted to the problems of scattering and target echoes. The theories and mathematical formulas relative to these problems are very well developed, and the authors are the first to admit that, at present, the cases of most practical interest are unfortunately beyond the scope of existing methods. Good discussions are noted in this chapter on clutter echoes, isolated targets, and the sea echo.

The authors have successfully avoided the duplications evident in British and American reports on the subject. The gaps in the coverage of the material which have been mentioned above do not detract seriously from the very apparent usefulness of the book. The only objection which can be raised is the inaccessibility of the many reports used as source material for which the authors can hardly be blamed.

OLIVER P. FERRELL
RASO, Radio Magazines, Inc.
121 S. Broad Street
Philadelphia 7, Pa.

Review of Current Research and Directory of Member Institutions, 1951

Published (1951) by the Engineering College Research Council of the American Society for Engineering Education, Room 7-204, 77 Mass. Ave., Cambridge 39, Mass. 215 pages+28-page index+x pages. 6×9. $2.25.

For the benefit of readers unfamiliar with the previous (1949) edition of this volume, it is a valuable listing of educational institutions active in pure and applied research, with an abstract of their current activities. The present edition has 35 per cent more pages than its predecessor.

About 80 colleges comprising the active members of the Engineering College Research Council are listed. Several pages are devoted to each one with information on its research officers, policies, personnel, expenditures, and projects now active.

Typical of some twelve particularly active research schools is an annual expenditure of over a million dollars to support a research staff of more than a hundred professors, consultants, other scientists and advanced students engaged on these projects.

The projects themselves cover the entire range of scientific subjects with emphasis on applied science and engineering. Such emphasis is to be expected as most of the work is sponsored by government agencies for military objectives. It is presumptive, however, that one of the principal aims is the subsidizing of education for increasing the technological potential of the nation; one might even go so far as to say that the many useful results may be regarded as byproducts of a system geared to the development of trained and inspired scientists and engineers.

Well indexed by subjects, the volume is a convenient guide to those colleges that may be specializing in any particular field. It is a concise record of the current activities in one of the most ambitious educational experiments in American history.

HAROLD A. WHEELER
Wheeler Laboratories, Inc.
Great Neck, L. I., N. Y.

Servomechanisms and Regulating System Design, Vol. I by Harold Chestnut and Robert W. Mayer

Published (1951) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N.Y. 497 pages+7-page index +6-page bibliography + xiii pages. 304 figures. 5¾×9¼. $7.75.

This is a book of good quality which does not particularly expand the field of its title but which does add a well-rounded work to those already on the market. It is intended for the "training of design and application engineers" and the particular volume under review is "adapted to the needs of engineers and engineering students who have not had previous training or experience in the field of closed-loop control systems."

To put down some thoughts which, although not unimportant, can be disposed of by a brief listing: (1) The present volume is the first of two; since the second has not been published, it is necessary for Volume I to stand on its own feet here. (2) The book has a profusion of excellent illustrations and charts. (3) Worked-out examples illustrating text material are good. (4) Typography and general appearance are good and typographical errors are infrequent. (5) There is a good bibliography of more than 100 items, almost all in English. (6) Problems for students to solve, covering some 50 pages at the end of the text, are good. (7) The book, in the General Electric series, is meant not only for electrical engineers but apparently also for other engineers who take General Electric Company courses. As a result there is a considerable amount of elementary material in the first third of the book.

Omitting consideration of elementary matter, Chestnut and Mayer's book covers material which might be described as customary, with the exception that considerably more emphasis is placed on attenuation and phase characteristics than usual. Chapter subjects (not titles) include stability, transfer functions, system types, complex-plane plots, attenuation and phase characteristics and applications, integral-network relations, multiple-loop systems, and transient performance.

On summary glance the text would seem to be excellent. It does not, on full reading, live up to original expectations. There is no one major reason for this; instead the result arises from many small causes; the book is over-written and could be appreciably reduced while supplying the same technical information; a more than occasional lack of balance exists; empirical information is introduced without even qualitative justification or explanation; sometimes major jumps depend on slight or nonexistent bases; a large number of curves in the text do not have the co-ordinates labeled; the design objective of the text tends to be slightly emphasized, with the result that discussions of which charts will produce an answer the soonest seem to be over-emphasized; and so on, and so on. Altogether it comes down to the fact that so many disturbing simple items can be found so frequently throughout the text, that the book falls short of being excellent.

A notation developed by a subcommittee of the AIEE Committee on Feedback Control Systems is used. This notation, which does not have the blessing of the AIEE Standards Committee nor the IRE's, nor the ASA's, is not only used here but, in addition, will appear in another servomechanisms text to be published shortly. Thus there will be a popularization of a notation which has been developed on the basic premise that the field of feedback control is one to itself and that correlation with other fields of electrical engineering is not necessary. An example of the use of the notation in the book under review will show the difficulty which a student would have in passing from one class to another, or a listener in passing from one session to another, if the Chestnut and Mayer notation were used in one. On one figure (page 282) appear $E=I$ and $O=C$, in which no symbol refers to voltage, current, zero, or capacitance. R (for reference input) and R (resistance) appear on one diagram, and E (error) and E (voltage) likewise. Many common symbols (B, C, E, H, M, Q, R, V, Z) have meanings completely different from the usual and standardized ones. This constitutes a considerable problem for the IRE Symbols Committee and the newly created IRE Technical Committee on Servomechanisms.

From the point of view of the student who wants to learn the theoretical and fundamental background of the servomechanism field and who can pick up his design and empirical information later, it may not be fully desirable that new books cover to a large extent the field already covered by predecessors, even though a reasonable choice is valuable. In January, 1946, Professor Guillemin raised the question of why servomechanisms were not designed from the desired over-all transfer characteristic. Recent work in the theses of Clanton, Truxal, and Aaron, as well as some unpublished work, show the potentialities of this approach. It is true that quite substantial problems remain to be overcome, but none the less there is a good possibility that the next major advance in a textbook in this field will be achieved by the one which starts from an attempt to synthesize an over-all characteristic and considers the details in the present book as cases evolving from that general problem.

J. G. Brainerd
Moore School of Elec. Eng.
University of Pennsylvania
Philadelphia 34, Pa.

An Introduction to Electron Optics by L. Jacob

Published (1951) by John Wiley & Sons Inc., 440 Fourth Ave., New York 16, N. Y. 137 pages+4-page index+1-page references+6-page bibliography. 48 figures. 4×6½. $2.00.

The author is affiliated with the Universities of Manchester and Liverpool, England.

This little book covers the entire field of electron optics from fundamental principles to lens aberrations, phase focusing, beam properties, and beam deflection. Field mapping and ray tracing in electrostatic fields, as well as the properties and aberrations of electrostatic lenses are included in about half of the book; however, magnetic lens fields are considered in somewhat less detail than the former.

The title, "Introduction to Electron Optics," is a little misleading. It constitutes rather, a summary of a very small considerable fraction of existing electron-optical literature. The beginner would have difficulty in discriminating between the more important and the less important techniques and findings of electron optics as they are reported here. Some of the general laws of electron optics are not brought out clearly and, too frequently, the logical basis of mathematical developments is not indicated, and symbols are inadequately defined.

Nevertheless, the worker in the field of electron optics will appreciate the many references to papers which may, heretofore, have escaped his notice. As a "refresher," particularly with reference to recent British work in the field of electron optics, the book should all the more so be useful, since the remarkable amount of material which has been compressed into this small volume appears to be relatively free of errors.

Edward G. Ramberg
RCA Laboratories Division
Princeton, N. J.

Television and FM Antenna Guide by Edward M. Noll and Matthew Mandl

Published (1951) by the Macmillan Co., 60 Fifth Ave., New York 11, N. Y. 308 pages+3-page index +ix pages. 225 figures. $5.50.

With the advent of television, rooftop antennas have reappeared. They have sprouted in a way that is reminiscent of the appearance of outside antennas 30 years ago in the early days of broadcasting. The TV antenna, however, is a different species because television waves are short. This is both an advantage, because directional types are feasible, and a handicap, because the effective apertures are small. For example, a 1-megacycle broadcast antenna 50 feet long is but a small fraction of a wavelength, yet it may have an effective aperture of 50,000 square feet over which it can extract energy from a passing radio wave. On the other hand, a half-wavelength TV antenna operating at 100 megacycles has a maximum effective aperture of less than 15 square feet over which it can collect energy from a television wave. This handicap of small aperture may be compensated in part by the use of directional-antenna arrays, which is why we find a great variety of stacked, colinear, Yagi, and corner-reflector types where signals are weak. Some of these have been adapted from earlier amateur vhf practice, but many have been designed especially for the wide band requirements of television. Considering the many types in use, it is rather surprising that long-wire and rhombic types have not found wider application in rural areas where the necessary acreage is available.

Television-receiving antennas form an extensive subject. It is natural, therefore, that an entire book has been devoted to this topic. This book, "Television and FM Antenna Guide," by Edward M. Noll and Matthew Mandl, should be of assistance to all interested in TV and FM receiving antennas, and in particular to technicians concerned with TV-antenna installation. The treatment is elementary, easily followed, and intensely practical. It is essentially nonmathematical.

The first 26 pages give an introductory picture of wave propagation at very high frequency. Included are FCC charts for the calculation of signal range. The next 30 pages cover some pertinent properties of transmission lines. This is followed by 56 pages on important practical considerations regarding antennas with a discussion of feeds and patterns of a number of simple antennas and directive systems.

The remainder of the book (187 pages) is in the form of a guide arranged for easy reference, which takes up first such topics as antenna-site selection, installation tools and procedures, transmission-line installation methods, input systems, booster amplifiers, and so forth. This is followed by a series of short, well-illustrated descriptions of numerous commercial TV antenna types ranging from simple dipoles to multielement stacked arrays. Many patterns that apparently are power plots are presented. There are also sections on long wire types, indoor type, antenna rotators, and multiple-outlet systems for apartments.

This book should be of great interest and value to all concerned with TV receiving antennas.

John D. Kraus
Dept. of Elec. Eng.
Ohio State University
Columbus, Ohio

New Publications

TV Factbook No. 13, published July, 1951, by Television Digest, 1519 Connecticut Ave., Washington, D. C. Martin Codel, editor and publisher. This new, 96-page, $5 edition is the 13th in a semiannual series and includes such reference material as:

Personnel and facilities data, with digests of rate cards for all of the 107 television stations and four networks now operating, together with listings of actual and proposed TV stations in Latin America and Canada; complete tabulation of the now frozen 416 applications for new stations pending before FCC; official tables of present vhf and proposed new vhf- and uhf-channel allocations by states, cities, and channels indicating educational assignments, proposed station shifts, and so forth.

New features of this edition are a population-dwelling-sales analysis of the 162 most important markets of the United States; directories of engineers, attorneys, and related services specializing in TV; a 34×22″ wall map in color, showing present television areas and actual and projected coaxial-microwave network routes.

Brought up to date are directories of the 92 TV-receiver manufacturers in the United States, 13 in Canada, 38 picture-tube and 12 receiving-tube makers, 27 concerns manufacturing TV transmitting and associated equipment, 465 firms providing films and other programs to TV stations, lists of station sales representatives, labor unions, trade groups, research firms, and so on. In addition there are tabulations of television-radio receiver production by months since 1946 and the latest count of TV sets in use by areas.

Books

High-Frequency Measurements, Revised Second Edition by August Hund

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N.Y. 631 pages + 35-page index + 7-page appendix + xi pages. 417 figures. 6×9¼. $10.00.

This second edition of a pioneer book in high-frequency measurements has been revised extensively in order to include the developments of the last 19 years. "High frequency" now is interpreted to include all frequencies above 20 kilocycles per second up to super-high frequencies (microwaves). To increase the usefulness of the book, standard measurements and audio frequencies are also given. The chapter on Line and Antenna Determinations has been completely rewritten, the chapter on Modulation Measurement has been greatly expanded, the use of radio-frequency and very-high-frequency bridges has been added, and there is a discussion of signal-to-noise measurements. Because of space limitation, specific microwave techniques have not been included.

The book now covers in its authoritative and comprehensive manner introductory chapters on Fundamental Relations and Circuit Properties, High-Frequency Sources and useful laboratory apparatus, such as oscillographs, bridges, and attenuation networks. It then takes up the measurement of small high-frequency currents, and in individual chapters, the Measurement of Voltage, Frequency, Capacitance, Self-Inductance, Mutual Inductance, Effective Resistance, High-Frequency Power and Losses, Logarithmic Decrement, and related quantities. In each instance, basic relations are given, typical values mentioned, and possible errors and necessary precautions discussed. A brief chapter on Ferromagnetic Measurements follows; Tube Measurements are discussed more fully, with attention to transit-time effects. The chapter on Modulation Measurements is rather comprehensive with considerable stress on frequency and phase modulation. The chapter on Measurements on Lines and Aerial Systems briefly reviews transmission-line relations and discusses measurements on parallel-wire and coaxial lines; it does not treat measurements on waveguide systems.

Determinations on Wave Propagation include field-strength measurements and polarization effects in skywave and ionosphere propagation. A final chapter on miscellaneous matters includes measurements on Quartz Crystals, Standard Field Calibration, Signal-to-Noise Measurements, and a brief treatment of latest developments in bridge measurements.

Throughout the book there are frequent references to the companion volume, "Short-Wave Radiation Phenomena," published by the McGraw-Hill Book Co., in 1951, and to a previous book, "Phenomena in High-Frequency Systems," also published by McGraw-Hill, in 1936. Many of the illustrations have been rather drastically reduced in size from the original, probably in order to save space; in several places this will make the reading difficult.

A great improvement over the first edition is the consistent use of the now internationally adopted mks system. The author has continued the use of mathematical nomenclature and the procedures recommended by the Standards Committee of the IRE. With this timely revision, the book will retain its leadership in the frequency range of greatest interest to practical radio engineers.

ERNST WEBER
Polytechnic Institute of Brooklyn
55 Johnson Street
Brooklyn 1, N.Y.

Transient Analysis in Electrical Engineering by Sylvan Fich

Published (1951) by Prentice-Hall, Inc., 70 Fifth Ave., New York, N. Y. 291 pages + 6-page index + 8-page appendix + ix pages. 92 figures. 5½×8¼. $7.35.

This book is a new addition to the Prentice-Hall Electrical Engineering Series, edited by Professor W. L. Everitt. It was written with the aim of extending transient analysis in electrical engineering to include modern operational methods without eliminating the presentation of classical theory. The only prerequisites for understanding the subject matter are the conventional courses in calculus, college physics, and steady-state circuit theory. No previous knowledge of differential equations is assumed.

The first five chapters are devoted to the solution of linear differential equations and their application to the classical solution of electrical and mechanical transients. The concept of a complex frequency is introduced by the definition of a complex angle in the classical analysis of an oscillating circuit. The Laplace transform is developed in Chapter 6 and applied to networks in the following chapter. Methods for solving higher-degree algebraic equations are presented in Chapter 8 and applied to circuit problems in Chapter 9. The Fourier series, integral, and transforms are developed in Chapter 10. This is followed by an introduction to complex-variable theory and its application to the calculation of inverse Laplace transforms. Chapter 13 deals with distributed parameters. A resume of electrical analogues of engineering systems is given in the last chapter.

This book should prove to be an excellent text for an undergraduate course. As suggested by the author, the first nine chapters might be used for this purpose. The reviewer is less enthusiastic about the suggested use of the last nine chapters as the basis for a first graduate course. Although admirably presented, the ground covered in this section is so great that the treatment is necessarily abbreviated. To illustrate, it may be questioned whether the two-page derivation of Heaviside's expansion theorem or the 13-page chapter on systems having distributed parameters will meet the needs of the graduate students. And, although it is a matter of personal opinion, the reviewer feels that offering the concepts of the complex variable at a point so late in the book has weakened the presentation. But on the whole the book is very well written and the author is to be commended particularly for his valuable treatment of the physical basis of the mathematical results, as well as for the style, clear and readable throughout, and for his ability to lead the reader to an understanding of the subject.

R. R. LAW
RCA Laboratories Division
Princeton, N. J.

Radio Communication at Ultra High Frequency by J. Thomson

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 200 pages + 3-page index + xix pages. 85 figures. 5×8¼. $4.50.

According to the preface, the aim of the author in writing this book was "to provide an account of modern developments in telecommunications employing radio waves of lengths ranging from a few meters to a few millimeters." It is addressed to students, and practicing communications engineers, as well as to workers in other electronic fields who might profit from it. The title of the book, would lead the prospective reader to expect a fairly complete discussion of the system considerations involved in uhf communications.

The author has not completely succeeded in accomplishing his own aim or in fulfilling the expectation mentioned above. The discussion of system considerations is restricted to a small portion of the final chapter of the book, while the remainder of the book is devoted to a discussion of uhf apparatus.

In several places the author relegated various developments to the indefinite future which actually were in commercial use at the time the book was written. One example of such questionable statements is in the final paragraph where he states: "Up to the moment of writing, the life of a U.H.F. valve is not satisfactory for unattended operation." In view of the General Electric Company's New York-Schenectady microwave relay system and the Bell System's New York-Boston and New York-Chicago systems, all of which were in commercial operation several years before the copyright date, this statement is difficult to understand.

In addition, there are a few minor errors such as the assigning of units to the dimensionless relative permeability on page 1, the use of $\bar{e_1}$ for $\sqrt{\bar{e_1}^2}$ throughout Chapter 4, and a few obvious typographical inaccuracies which could have been eliminated by careful proofreading.

This reviewer found the discussion of velocity-modulation devices and the chapter on receiver-input circuits interesting and well presented.

In spite of the shortcomings mentioned above, the author has produced a very readable book which should provide a valuable introduction to the field for students and which should prove to be, in part, a useful reference for the practicing engineer.

JOHN G. KREER
Bell Telephone Labs., Inc.
Murray Hill, N. J.

Sections and Subsections*

Chairman	Section	Secretary
R. L. Buriner Goodyear Aircraft Corp. 1210 Massillon Rd. Akron, Ohio	AKRON (4)	R. M. Byrne 316 Melbourne Ave. Akron 3, Ohio
E. S. Lammers, Jr. Westinghouse Electric Corp. Box 4808 Atlanta 2, Ga.	ATLANTA (6)	D. B. Buchanan 5130 Powers Ferry Rd., N.W. Atlanta, Ga.
D. G. Little 408 Old Orchard Rd. Baltimore 29, Md.	BALTIMORE (3)	G. R. White Bendix Radio Div. Towson 4, Md.
E. D. Coburn Box 793 Nederland, Texas	BEAUMONT-PORT ARTHUR (6)	C. B. Trevey 2555 Pierce St. Beaumont, Texas
L. E. Packard Technology Instrument Corp. 1058 Main St. Waltham 54, Mass.	BOSTON (1)	Beverly Dudley The Technology Review Mass. Inst. of Tech. Cambridge 39, Mass.
G. J. Andrews Fernandez Spiro 286 Acassuso, F.C.N.G.B.M. Buenos Aires, Argentina, S.A.	BUENOS AIRES	I. C. Grant Paseo Colon 995 Buenos Aires, Argentina, S.A.
R. E. Frazier R.D. 1. Center St. East Aurora, N. Y.	BUFFALO-NIAGARA (4)	W. K. Squires 115 St. John's Ave. Kenmore 23, N. Y.
V. R. Hudek 137 29 St. Drive S.E. Cedar Rapids, Iowa	CEDAR RAPIDS (5)	R. M. Mitchell 357 Garden St., S.E. Cedar Rapids, Iowa
Leroy Clardy Research Labs. Swift & Co. U.S. Yards Chicago 9, Ill.	CHICAGO (5)	A. A. Gerlach 4020 Overhill Ave. Chicago 34, Ill.
Calvin Bopp 1114 Maycliff Pl. Cincinnati 30, Ohio	CINCINNATI (5)	R. A. Maher 1329 Arlington St. Cincinnati 25, Ohio
T. B. Friedman 1900 E. 30 St. Cleveland 14, Ohio	CLEVELAND (4)	A. L. Hammerschmidt 111 Westbridge Dr. Berea, Ohio
D. C. Cleckner 1376 Aberdeen Ave. Columbus 3, Ohio	COLUMBUS (4)	J. H. Jaeger Ohio Bell Telephone Co. 42 E. Gay St. Columbus 15, Ohio
L. E. Williams Dept. of Elec. Eng. University of Connecticut Storrs, Conn.	CONNECTICUT VALLEY (1)	H. E. Rohloff The Southern New England Tel. Co. 227 Church St. New Haven, Conn.
E. A. Hegar 802 Telephone Bldg. Dallas 2, Texas	DALLAS-FORT WORTH (6)	J. K. Godbey 3990 Davilla Dr. Dallas 9, Texas
Joseph General 1116 Lexington Ave. Dayton 7, Ohio	DAYTON (5)	A. H. Petit 159 Aberdeen Ave. Dayton 9, Ohio
P. M. La Hue 3215 E. 34 Ave. Denver 5, Colo.	DENVER (5)	D. I. Peterson 933 Pontiac St. Denver, Colo.
Otto A. Tennant 1408 Walnut St. Des Moines, Iowa	DES MOINES-AMES (5)	A. D. Parrott 1515—45 Des Moines 11, Iowa
K. R. Schmeisser 13117 La Salle Blvd. Detroit 6, Mich.	DETROIT (4)	F. W. Chapman 1756 Graefield Rd. Birmingham, Mich.
W. R. Rolf 364 E. 5th St. Emporium, Pa.	EMPORIUM (4)	L. R. Maguire 4 E. 6 St. Emporium, Pa.
J. F. Sears 1736 Washington Ave. Evansville 14, Ind.	EVANSVILLE-OWENSBORO (5)	F. A. Gehres 2232 E. Powell Evansville 14, Ind.
J. F. Conway, Jr. 4610 Plaza Dr. Fort Wayne, Ind.	FORT WAYNE (5)	L. F. Mayle Capehart-Farnsworth Corp. Fort Wayne 1, Ind.
J. S. Turner 3422 Alani Dr. Honolulu, T.H.	TERRITORY OF HAWAII (7)	S. H. Lewbel P.O. Box 3564 Honolulu, T.H.
S. A. Martin, Jr. Box 2180, Rm. 209 Houston 1, Texas	HOUSTON (6)	H. T. Wheeler 802 N. Avenue "A" Bellaire, Texas
F. O. Meadows 1449 N. Emerson Indianapolis, Ind.	INDIANAPOLIS (5)	R. T. Van Nimam 4441 Indianola Ave. Indianapolis 5, Ind.
J. H. Gregory 200-A Ellis China Lake, Calif.	INYOKERN (7)	W. E. Vore 420-A Nimitz Ave. China Lake, Calif.
C. L. Foster 1901 W. 73 Terrace Kansas City 5, Mo.	KANSAS CITY (5)	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
G. R. Hosker c/o Richards-Wilcox Canadian Ltd. London, Ont., Canada	LONDON, ONTARIO (8)	Glenn Robitaille 19 McKinnon Place London, Ont., Canada
J. J. Fiske 128 W. Las Flores Ave. Arcadia, Calif.	LOS ANGELES (7)	Ellis King 3171 Federal Ave. Los Angeles 34, Calif.
R. B. McGregor 2100 Confederate Pl. Louisville, Ky.	LOUISVILLE (5)	M. I. Schwalbe Nichols Veterans Hospital Louisville 2, Ky.
P. H. Craig 1042 Catalonia Ave. Coral Gables, Fla.	MIAMI (6)	F. B. Lucas 5340 Davis Rd., S.W. South Miami 43, Fla.
W. H. Elliot 4747 N. Larkin St. Milwaukee 11, Wis.	MILWAUKEE (5)	D. A. Weller c/o WISN 540 N. Plankinton Ave. Milwaukee 1, Wis.
N. R. Olding Canadian Broadcasting Corp. Box 6000 Montreal, P.Q., Canada	MONTREAL, QUEBEC (8)	R. W. Cooke Canadian Marconi Co. 2442 Trenton Rd. Montreal, P. Q., Canada
B. J. Bittner 3110—42 Pl., Sandia Base Albuquerque, N. M.	NEW MEXICO (7)	G. A. Arnot, Jr. 401 N. 6 St. Albuquerque, N. M.
J. H. Mulligan, Jr. 14 Mitchell Pl. East Orange, N. J.	NEW YORK (2)	H. S. Moncton Sylvania Elec. Prods. Inc. Bayside, L. I., N. Y.
C. E. Hastings 117 Hampton Rd. Hampton, Va.	NORTH CAROLINA-VIRGINIA (3)	P. F. Hedrick 2472 Maplewood Ave. Winston-Salem 7, N. C.
O. A. Nash, Acting Chairman c/o CAA, AC-636 P. O. Box 1082 Oklahoma City, Okla.	OKLAHOMA CITY (6)	R. R. McClung, Acting Secretary-Treasurer 1615 Joane Street Oklahoma City, Okla.
F. W. Norris 3303 Woods Ave. Lincoln, Neb.	OMAHA-LINCOLN (5)	C. W. Rook Dept. of Elec. Eng. University of Nebraska Lincoln 8, Neb.
W. R. Wilson c/o Royal Canadian Mounted Police Ottawa, Ont., Canada	OTTAWA, ONTARIO (8)	D. V. Carroll Box 527 Ottawa, Ont., Canada
L. M. Rodgers 400 Wellesley Rd. Philadelphia 19, Pa.	PHILADELPHIA (3)	J. G. Brainerd Moore School of Elec. Eng. Univ. of Pennsylvania Philadelphia 4, Pa.
G. C. Rich Box 737 Scottsdale, Ariz.	PHOENIX (7)	A. M. Creighton, Jr. 2221 E. Osborn Rd. Phoenix, Ariz.
W. P. Caywood, Jr. 23 Sandy Creek Rd. Pittsburgh 21, Pa.	PITTSBURGH (4)	V. G. Shaw 17 Frankwood Rd. Pittsburgh 35, Pa.
L. C. White 3236 N. E. 63 Ave. Portland 13, Ore.	PORTLAND (7)	G. C. Ellison 11310 S. E. Market St. Portland 16, Ore.
W. H. Bliss 300 Western Way Princeton, N. J.	PRINCETON (3)	J. S. Donal, Jr. RCA Labs. Princeton, N. J.
J. A. Rodgers Huntington Hills Rochester 9, N. Y.	ROCHESTER (4)	J. C. O'Brien 283 Malden St. Rochester 13, N. Y.
W. F. Koch 1340 33 St. Sacramento 16, Calif.	SACRAMENTO (7)	H. C. Slater 1945 Bidwell Way Sacramento 18, Calif.
H. G. Wise 1705 N. 48 St. East St. Louis, Ill.	ST. LOUIS (5)	E. F. O'Hare 8325 Delcrest Dr. University City 24, Mo.
L. S. Cole 696 E. 4 North St. Logan, Utah	SALT LAKE CITY (7)	E. C. Madsen 739 Scott Ave. Salt Lake City 6, Utah

* Numerals in parentheses following Section designate Region number.

Sections and Subsections

Chairman	Section	Secretary
W. J. Hamm St. Mary's University San Antonio 7, Texas	San Antonio (6)	Paul Tarrodaychik 215 Christine Dr. San Antonio 10, Texas
J. P. Day 3565 Ingraham St. San Diego 9, Calif.	San Diego (7)	I. L. McNally 2714 Azalea Dr. San Diego 6, Calif.
A. R. Ogilvie 51 Michael Lane Millbrae, Calif.	San Francisco (7)	W. R. Abbott 274 Grizzly Peak Blvd. Berkeley 8, Calif.
E. S. Sampson 1402 Union St. Schenectady 8, N. Y.	Schenectady (2)	D. E. Norgaard 1908 Townsend Rd. Schenectady, N. Y.
J. Emerson Hogg General Electric Co. 710 2 Ave. Seattle 11, Wash.	Seattle (7)	H. M. Swarm Dept. Elec. Eng. Univ. of Washington Seattle 5, Wash.
D. C. Pinkerton 312 Cherry Rd. Syracuse 9, N. Y.	Syracuse (4)	Samuel Seely College Applied Science Syracuse University Syrcause 10, N. Y.
H. N. Rowe Rowe Industries 1702 Wayne St. Toledo 9, Ohio	Toledo (4)	R. G. Larson 2647 Scottwood Ave. Toledo 10, Ohio
George Sinclair 7 Broadway Ave., Apt. 21 Toronto 12, Ont., Canada	Toronto, Ontario (8)	J. R. Bain Dominion Sound Equip., Ltd. 386 Victoria St. Toronto, Ont., Canada
Daniel Silverman Stanolind Oil & Gas Co., Res. Lib. Box 591 Tulsa 2, Okla.	Tulsa (6)	C. E. Buffum Box 591 Tulsa 2, Okla.
A. A. Cohen Eng. Research Assoc. 1902 W. Minnehaha Ave. St. Paul 4, Minn.	Twin Cities (5)	O. W. Muckenhirn Dept. of Elec. Eng. University of Minnesota Minneapolis 14, Minn.
G. C. Chandler 846 Howe St. Vancouver, B. C., Canada	Vancouver (8)	D. D. Carpenter 1689 W. 29 Ave. Vancouver, B. C., Canada
P. De Forrest McKeel 9203 Sligo Creek Pkwy. Silver Spring, Md.	Washington (3)	H. P. Meisinger Hull Rd. & Old Court-house Rd. Route 3 Vienna, Va.
F. H. Scheer College Park Lewisburg, Pa.	Williamsport (4)	B. H. Bueffel, Jr. Dept. of Elec. Eng. Bucknell University Lewisburg, Pa.
D. G. Harmon Dept. of Elec. Eng. Texas Tech. College Lubbock, Texas	Amarillo-Lubbock (6) (Dallas-Ft. Worth Subsection)	R. B. Spear 510 E. Hill St. Brownfield, Texas
L. H. Stantz 168 Moeller St. Binghamton, N. Y.	Binghamton (4) (Syracuse Subsection)	E. L. Pittsley Reynolds Rd. R. D. 1 Johnson City, N. Y.
W. R. Smith Dept. of Elec. Eng. Pennsylvania State College State College, Pa.	Centre County (4) (Emporium Subsection)	H. F. Wischnia Dept. of Elec. Eng. Pennsylvania State College State College, Pa.
J. H. Pickett Aerovox Canada Ltd. 1551 Barton St., E. Hamilton, Ont., Canada	Hamilton (8) (Toronto Subsection)	G. F. Beaumont 55½ Sherman Ave., S. Hamilton, Ont., Canada
E. E. Spitzer c/o RCA Victor Div. Radio Corp. of America Lancaster, Pa.	Lancaster (3) (Philadelphia Subsection)	Werner Rueggeberg Research Lab. Armstrong Cork Co. Lancaster, Pa.
H. E. Webber Sperry Gyroscope Co. Lake Success, L. I., N. Y.	Long Island (2) (New York Subsection)	J. F. Craib R. D. 1 Hicksville, L. I., N. Y.
A. L. Samuel Electronics Labs. I.B.M. Corp. Plant 2, Dept. 532 Poughkeepsie, N. Y.	Mid-Hudson (2) (New York Subsection)	A. V. Platter R. D. 1 Salt Point Rd. Poughkeepsie, N. Y.
B. V. Blom 38 Lafayette St. Rumson, N. J.	Monmouth (2) (New York Subsection)	H. S. Bennett 31 Cedar Ave., Apt. 23 Long Branch, N. J.
J. F. Morrison Bell Telephone Labs. Whippany, N. J.	Northern N. J. (2) (New York Subsection)	P. S. Christaldi Box 145 Clifton, N. J.
J. M. Pettit Dept. of Elec. Eng. Stanford Univ. Stanford, Calif.	Palo Alto (7) (San Francisco Subsection)	H. M. Stearns 99 Washington St. San Carlos, Calif.
George Weiler 1429 E. Monroe South Bend, Ind.	South Bend (5) (Chicago) Subsection)	A. R. O'Neil Apt. 302 6712 Lake St. Falls Church, Va.
R. M. Wainwright Dept. of Elec. Eng. University of Illinois Urbana, Ill.	Urbana (5) (Chicago Subsection)	M. H. Crothers Dept. of Elec. Eng. University of Illinois Urbana, Ill.
	Winnipeg (8) (Toronto Subsection)	W. F. Souch Canadian Marconi Co. 149 Portage Ave., E. Winnipeg, Canada

Professional Groups

	Chairman
Airborne Electronics	John E. Keto Wright Field Dayton, Ohio
Antennas and Propagation	George Sinclair Dept. of Elec. Eng. University of Toronto Toronto, Ont. Canada
Audio	B. B. Bauer Shure Brothers, Inc. 225 W. Huron St. Chicago 10, Ill.
Broadcast and Television Receivers	D. D. Israel Emerson Radio & Phonograph Co. 111 8 Ave. New York 11, N. Y.
Broadcast Transmission Systems	Lewis Winner *TeleVision Engineering* 52 Vanderbilt Ave. New York 17, N. Y.
Circuit Theory	J. G. Brainerd University of Pennsylvania Philadelphia, Pa.
Engineering Management	Ralph I. Cole Griffiss Air Force Base Rome, N. Y.
Industrial Electronics	Eugene Mittelmann 549 West Washington Blvd. Chicago 6, Ill.
Information Theory	Nathan Marchand Sylvania Electric Products Inc. Bayside, L. I., N. Y.
Instrumentation	Ernst Weber Polytechnic Institute of Brooklyn Brooklyn, N. Y.
Nuclear Science	M. M. Hubbard Massachusetts Institute of Technology Cambridge, Mass.
Quality Control	R. F. Rollman Allen B. DuMont Laboratories 1000 Main Ave. Clifton, N. J.
Radio Telemetry and Remote Control	W. J. Mayo-Wells The Johns Hopkins University Silver Spring, Md.
Vehicular Communications	Austin Bailey American Telephone and Telegraph Co. 195 Broadway New York, N. Y.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with That Department and the *Wireless Engineer*, London, England

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger(†) must be regarded as provisional.

ACOUSTICS AND AUDIO FREQUENCIES

016:534 2309
References to Contemporary Papers on Acoustics—A. Taber Jones and R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 377–385; May, 1951.) Continuation of 1812 of September.

534+621.395.61/.62†(083.71) 2310
Standards on Electroacoustics: Definitions of Terms, 1951—(PROC. I.R.E., vol. 39, pp. 509–532; May, 1951.) Reprints of this Standard, 51 IRE 6 S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $1.00 per copy.

534.213–14 2311
Random Noise in an Attenuating Fluid Medium—R. E. Roberson. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 353–358; May, 1951.) "It is assumed that acoustic background noise is caused by a distribution of random "white" noise sources whose physical mechanism is unspecified. A law expressing the amplitude-distance attenuation characteristic of the medium is also assumed. Several distributions of noise sources are considered: uniform volume distributions, uniform surface dipole distributions, and two mixed cases. The noise dropoff, with frequency at a point below the surface, is found for each case. For an infinite volume of noise sources, this dropoff is 6 db/octave at all frequencies. It is shown how this simple model can be generalized to other attenuation laws and other spatial and amplitude distributions of noise sources."

534.232 2312
The Emission of Sound by a Piston—D. N. Chetaev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 813–816; February 21, 1951. In Russian.) Integral (1) determining the amplitude of the velocity potential of steady-state sound oscillations is considered, and the sound field at an arbitrary point *P*, determined. A formula (5) determining the radiation resistance is also derived.

534.232:538.652:621.314.212 2313
Electroacoustic Transformation by Means of Magnetostriction, with Special Reference to Radiation from Transformers—H. H. Rust. (*Z. angew. Phys.*, vol. 2, pp. 487–491; December, 1950.) Magnetostriction curves, obtained with Ni, Fe, and Fe/Si oscillators, and taking account of magnetostrictive hysteresis, indicate complex oscillations rich in harmonics. These results apply directly to transformers, which in normal operation must incidentally radiate such magnetostrictively-excited mechanical waves with adverse effects on the properties of any oil used for insulation. Suggestions are made for eliminating these harmful effects.

534.24+534.373 2314
Scattering and Absorption by an Acoustic Strip—A. Levitas and M. Lax. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 316–322; May, 1951.) An analytical method is described for determining the scattering and absorption of sound for a nonuniform-boundary case represented by the application to an infinite wall of a finite-width strip of soundproofing material.

534.24 2315
Focusing of Sound Waves by a Parabolic Reflector—L. D. Rozenberg. (*Zh. Tekh. Fiz.*, vol. 20, pp. 385–396; April, 1950.) A mathematical investigation is presented of the sound field near the focus. Formulas are derived for determining the pressure at the focus, and the effects of increasing the aperture of the reflector on its focusing properties, while keeping the focal length constant, are examined. Methods are indicated for choosing the optimum focal length for a given aperture of the reflector, and for determining the radius of the diffraction circle at the focus. Cases when the source of sound is at a finite distance from the reflector, and not on the axis, are also considered. Some experimental results are included.

534.24 2316
On the Nonspecular Reflection of Sound from Planes with Absorbent Bosses—V. Twersky. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 336–338; May, 1951.) The analysis developed in 9 of February for nonabsorbent surfaces is extended to the case of absorbent surfaces. The results for the two cases are compared. The effect of the finite impedance may be either to decrease or increase the radiation reflected at the specular angle.

534.24 2317
Sound Scattering of a Plane Wave from a Nonabsorbing Sphere—R. W. Hart. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 323–329; May, 1951.) An analytical treatment is developed. Consideration is restricted to the case where the acoustic properties of the sphere are not very different from those of the surrounding medium. See also 2139 of October (Hart and Montroll).

534.24 2318
On the Reflection of a Spherical Sound Wave from an Infinite Plane—U. Ingard. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 329–335; May, 1951.) Boundary conditions are given in terms of a normal impedance independent of the angle of incidence. The integral for the reflected wave is expressed in a form such that the wave can be considered as originating from an image source having a certain amplitude and phase. Graphs for determining these values are given in terms of a "numerical distance" which depends on the normal impedance and the position of the field point.

534.321.9:537.228.1 2319
Design of Variable Resonant Frequency Crystal Transducers—W. L. Hall and W. J. Fry. (*Rev. Sci. Instr.*, vol. 22, pp. 155–161; March, 1951.) A description of a system employing liquid mercury as a backing of continuously variable dimensions. The important aspects, viz., tight coupling of the crystal and mercury backing, and decoupling of the crystal and mercury from the supporting structure, are considered in detail. Construction procedure on a unit to cover the frequency range 40 to 80 kc is indicated. Experimental results on the magnitude of the electrical input impedance as a function of frequency and mercury-column length are given. The unit is compared with transducers having fixed resonance frequency.

534.414 2320
The Wavelength of a Spherical Resonator with a Circular Aperture—H. Levine. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 307–311; May, 1951.) An expression for the wavelength is derived in the form of an expansion exact as far as terms in $(a/R)^2$, where a is the aperture radius and R the sphere radius. A procedure for determining the wavelength approximately for a resonator of arbitrary shape is also described.

534.414 2321
Variation of the Resistance in the Resonator Neck with Intensity of Incident Sound—R. K. Vepa. (*Science and Culture* (Calcutta), vol. 16, pp. 482–483; April, 1951.) Measurements were made on a resonator formed by a plate 0.65 cm thick with a 1.25-cm circular orifice, appropriately spaced from a rigid backing. Results are compared with earlier measurements using a thinner plate and smaller orifice.

534.771 2322
Upper Limit of Frequency for Human Hearing—J. H. Combridge, J. O. Ackroyd and R. J. Pumphrey. (*Nature* (London), vol. 167, pp. 438–439; March 17, 1951.) Comment on 2959 of 1950 (Pumphrey) and author's reply.

534.78 2323
Effect of Delay Distortion upon the Intelligibility and Quality of Speech—J. L. Flanagan. (*Jour. Acoust. Soc. Amer.*, vol. 23, pp. 303–307: May, 1951.) Speech articulation tests were made on an all-pass system capable of advancing or delaying one frequency band relative to the rest of the spectrum. Measurements were made at signal/noise ratios of 30 db and 0 db. The results indicate that maximum impairment of intelligibility occurs when the delays or advances are of the order of $\frac{1}{4}$ second, and when the band delayed or advanced is near the center of the speech spectrum.

534.79 2324
Calculation and Measurement of the Loudness of Sounds—L. L. Beranek, J. L. Marshall, A. L. Cudworth, and A. P. G. Peterson. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 261–269; May, 1951.) An equivalent-tone method is described, in which the spectrum of the sound is divided into frequency bands which are treated as pure tones in calculating their loudness. Calculations for bands of white noise and for complex tones are compared with subjectively obtained data. The agreement is good.

534.833.4 2325
Absorption of Sound by Resonant Panels—G. G. Sacerdote and A. Gigli. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 349–352; May, 1951.) The experimental determination of resonance frequency and absorption of resonators, formed by plywood plates with uniformly spaced circular holes placed at various distances from the wall, is described. Measurements were made at normal incidence and with diffuse sound in a reverberant room. Moderate agreement between theoretical and experimental results is noted.

621.395.61/.62 2326
The Piezoelectric Sound Detector and its Electrical and Acoustic Equivalent Circuits—F. A. Fischer. (*Arch. elekt. Übertragung*, vol. 4, pp. 435–436; October, 1950.) An equivalent electrical circuit is derived which is applicable for operation of the piezoelectric transducer as generator, or as detector of sound. The paper is complementary to that noted in 2966 of 1950.

621.395.61/.62 2327
Post-War Developments in Electroacoustics [by Telefunken]—F. Bergtold. (*Telefunken Ztg.*, vol. 23, pp. 106–110; September, 1950.) Apparatus described includes moving-coil microphone, pickup, sound-distribution system, loudspeaker and arrays, cinema installation, and house-communication system.

621.395.623.7 2328
Loudspeaker Damping—A. Preisman. (*Audio Eng.*, vol. 35, pp. 22–23, 38 and 24, 45; March and April, 1951.) Loudspeaker characteristics are discussed theoretically, and an experimental method is described for determining the constants of the unit. The Q is determined from the shape of the impedance/frequency curve, and the source resistance for critical damping is calculated from the voice coil and motional impedance at resonance. An alternative method, based on consideration from a mechanical viewpoint, is also presented.

621.395.625.2 2329
Gramophone Turntable Speeds—G. F. Dutton. (*Wireless World*, vol. 57, pp. 227–231; June, 1951.) Suitable speeds for microgroove recordings are considered in relation to public demand, record materials, groove spacing, needle size, and distortion with different tangential velocities. Results are presented graphically, and a summary gives optimum speeds for different record diameters.

621.395.625.3:538.221 2330
Mixed Ferrites for Recording Heads—Herr. (*See* 2445.)

621.396.645.37.029.4 2331
Independent Control of Selectivity and Bandwidth—Villard. (*See* 2395.)

ANTENNAS AND TRANSMISSION LINES

621.39.09 2332
A New Solution of the Fundamental Problem of the Propagation of Electromagnetic Processes in a Multi-Wire System—N. A. Brazma. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 41–44; January 1, 1951. In Russian.)

621.392.09 2333
Surface-Wave Transmission Line—R. H. Nelson. (*Wireless Eng.*, vol. 28, p. 162; May, 1951.) Comment on 1300 of July (Barlow) and 563 of March (Rust).

621.392.22 2334
The Behaviour of Electromagnetic Waves in Highly Nonuniform Lines—H. Meinke. (*Z. angew. Phys.*, vol. 2, pp. 473–478; December, 1950.) The significant characteristic of the wave field in the region of a nonuniformity is the appearance of a wedge-shaped intrusion, resulting from longitudinal field components, in the field pattern in the neighborhood of the point of zero transverse electric field strength. This intrusion is calculated for the case of a field with constant curvature, and examples are given of the effects due to irregularities of arbitrary form.

621.392.26†:538.561 2335
Theory of the Excitation of Oscillations in a Waveguide by Means of a Linear Aerial—A. I. Akhiezer and G. Ya. Lyubarski. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1049–1064; September, 1950.) One of the main problems in the theory of antennas is the determination of the current distribution in an antenna to which given electromotive forces are applied. It has been shown by Leontovich and Levin (2618 of 1945) that in the case of an antenna in unlimited space, the problem is reduced to the solution of a linear integro-differential equation. A study is here presented of the current distribution in a linear antenna mounted along the axis of a cylindrical waveguide. In this case it is necessary to solve an equation of the same type as for an antenna in unlimited space. No effective methods of solving this equation for an antenna of arbitrary dimensions are known. The discussion is therefore limited to the case of a sufficiently long and thin antenna and, using a method proposed by Leontovich and Levin (2618 of 1945), an approximate solution of the equation is found by expanding the current in a series of powers of the inverse logarithm of the ratio of the length to the radius of the antenna.

The following two cases are considered separately: (a) when the wavelength differs considerably from the critical wavelength of the waveguide; (b) when this difference is not great. The current distribution in a tuned antenna differs very much from that in an antenna in unlimited space. Simple formulas for determining the amplitudes of the waves excited in the waveguide are also derived.

621.392.26†:621.39.09 2336
A Study of Asymmetrical Electromagnetic Waves from the Open End of a Circular Waveguide—L. A. Vainshtein. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 485–488; September 21, 1950. In Russian.) The methods used in 1283 of 1949 are not applicable to the case of asymmetrical electromagnetic waves, since in this case, owing to diffraction, two longitudinal components (1) and (2) of the electric vector appear at the open end, and the problem therefore cannot be reduced to a single integral equation. By using a generalized method similar to that presented in an earlier paper on sound radiation (*Compt. Rend. Acad. Sci.* (URSS), vol. 58, p. 1957; 1947) an exact solution can nevertheless be found. A system of equations (7) to (10) is derived, and methods are indicated for solving it.

621.392.26†:621.39.09 2337
The Diffraction of Waves at the Open End of a Circular Waveguide with Diameter Greater than the Wavelength—L. A. Vainshtein. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 909–912; October 11, 1950. In Russian.) The physical meaning of the formulas (see 1283 of 1949 and 2336 above) determining the radiation field under the above conditions, is discussed. Starting with the case of symmetrical waves, formula (7) determining the radiation field in the back half-space ($0<\theta<\lambda/2$) is considered, θ being the angle between the Z axis and the radius vector of the field point. With increase of distance from the edge of the waveguide, the primary cylindrical waves gradually become spherical. From a corresponding formula (12) for the front half-space, ($\lambda/2<\theta<\lambda$), it follows that waves from different sections of the edge interfere with one another and produce a complex spherical wave. Similar results are also obtained in the case of asymmetrical waves, but the process of development is more complicated.

621.392.43 2338
The Exponential Line at Cut-off Wavelength and in the Stop Range—A. Ruhrmann. (*Arch. elekt. Übertragung*, vol. 4, pp. 401–412; October, 1950.) The theory of current and voltage distribution at cutoff wavelength is discussed. The definition of characteristic impedance used in the pass range may be retained, but the transmission equations assume an indefinite form and require transformation. The cases of termination by characteristic impedance, short-circuiting, and open-circuiting, are considered separately; understanding of the mode of operation with complex termination is facilitated by reference to a circle diagram. On the basis of quadripole theory, it is possible to define a characteristic impedance within the stop range, although no wave propagation is to be inferred from the equations or circle diagrams. Attenuation factor and transmission factor are defined, the concept of a wave propagation process analogous to that in the pass range being made possible by introducing complex parameters. See also 1583 of 1950.

621.392.5:681.142 2339
Magnetic Delay-Line Storage—An Wang. (PROC. I.R.E., vol. 39, pp. 401–407; April, 1951.) A number of magnetic cores are connected together to form a static magnetic delay line in which a series of binary digits can be stored and read out. The operation of this type of line is briefly described and carefully analyzed, and the optimum operating conditions are derived. The effects of eddy-current loss and leakage inductance are considered, and criteria for stability of the system are discussed.

621.396.67 2340
Wrotham Aerial System: Part 1—New Design of Slot-Radiator for V.H.F. Broadcasting—C. Gillam. (*Wireless World*, vol. 57, pp. 210–214; June, 1951.) An omnidirectional horizontal-polarization radiator with a gain of 9 db, is obtained with 32 folded slots arranged in 8 tiers of 4, spaced uniformly around a vertical cylinder. The slots are cophased and fed by a branched transmission line with impedance-matching transformers, and are designed to handle simultaneously either three 25-kw FM transmissions or one 25-kw FM and one 18-kw AM transmission in the frequency band 87.5 to 95 mc.

621.396.67 2341
The Aerial Installations for the [German] Post-1945 High Power Transmitters—W. Berndt. (*Telefunken Ztg.*, vol. 23, pp. 39–52;

September, 1950.) Descriptions are given of the antenna installations for the medium- and long-wave broadcast and telegraphy transmitters described in 2565 below. A certain amount of improvisation was necessary. A feature common to all the installations is the spatial separation of transmitter and antenna system, the two being connected by hf cable.

621.396.67 2342
A Helix Theorem—J. D. Kraus. (Proc. I.R.E., vol. 39, p. 563; May, 1951.) For helical antennas of at least a few turns, with pitch angles of 10° to 15°, it is postulated that "when the circumference of an axial or end-fire helix is about one wavelength, (T_1 transmission mode dominant), there is a band of frequencies over which the phase velocity of wave propagation on the helix tends toward a value that makes the directivity a maximum."

621.396.67 2343
Radiation Properties of Spherical Antennas as a Function of the Location of the Driving Force—P. R. Karr. (*Bur. Stand. Jour. Res.*, vol. 46, pp. 422–436; May, 1951.) A theoretical analysis is made of the radiation from a conducting sphere fed at a narrow nonequatorial zone. Variations of radiation pattern, current distribution, and input admittance with the latitude of the feed zone, are studied. As long as the radius a of the sphere does not exceed $\lambda/2\pi$, the radiation conductance varies approximately as $\sin^4\theta_0$, where θ_0 is the colatitude of the feed zone. For $a>\lambda/2\pi$, the maximum value of radiation conductance may occur at values of θ_0 other than 90°.

621.396.67 2344
Slot Radiators—N. A. Begovich. (Proc. I.R.E., vol. 39, p. 508; May, 1951.) Correction to paper abstracted in 2711 of 1950.

621.396.67 2345
Biconical Electromagnetic Horns—W. L. Barrow, L. J. Chu, and J. J. Jansen. (Proc. I.R.E., vol. 39, pp. 434–435; April, 1951.) Corrections to paper noted in 1404 of 1940.

621.396.67:538.566 2346
Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 1—Transmission between Elliptically Polarized Antennas—V. H. Rumsey. (Proc. I.R.E., vol 39, pp. 535–540; May, 1951.) A method of analysis is discussed which makes use of the impedance concept of transmission-line theory. It is shown that P, the ratio of two orthogonal tangential components of the electric field, is related to q, the ratio between the left- and right-handed circularly polarized components corresponding to the orthogonal components, in the same way as impedance is related to reflection coefficient. Representation of polarization on a transmission-line impedance chart is described, and solutions of various polarization problems in terms of impedance analogies are discussed.

621.396.67:538.566 2347
Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2—Geometrical Representation of the Polarization of a Plane Electromagnetic Wave—Deschamps. (*See* 2418.)

621.396.67:538.566 2348
Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 3—Elliptically Polarized Waves and Antennas—Kales. (*See* 2419.)

621.396.67:53°.566 2349
Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 4—Measurements on Elliptically Polarized Antennas—J. I. Bohnert. (Proc. I.R.E., vol. 39, pp. 549–552; May, 1951.) Two methods for measuring polarization characteristics are outlined. One uses a rotating linear-polarization antenna, the other uses two circular-polarization antennas.

621.396.671.012 2350
Pattern Calculator for A.M.—G. R. Mather. (*Electronics*, vol. 24, pp, 100–101; April, 1951.) A graphical method for calculation of the radiation pattern of two-tower directive arrays.

621.396.677 2351
The Electric and Magnetic Constants of Metallic Delay Media containing Obstacles of Arbitrary Shape and Thickness—S. B. Cohn. (*Jour. Appl. Phys.*, vol. 22, pp. 628–634; May, 1951.) Methods of deriving the dielectric constant and permeability of a metallic-obstacle medium are given. The equivalent shunt capacitive susceptance and series inductive reactance of the individual obstacles are also determined. By means of a correspondence established between an infinitely thin conducting obstacle and an aperture in an infinitely thin conducting wall, formulas are derived for different shapes of obstacle. The effect on the magnetic field of obstacles of moderate thickness is evaluated. For obstacles of arbitrary shape and thickness, the constants can be determined by the electrolyte-tank method; this is demonstrated for a particular example.

621.396.677 2352
Directional Aerials for Radio Stations—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 4, pp. 49–56; February, 1951.) Review and discussion of different types of radiators for use at wavelengths between 3 cm and 3 m.

621.396.677:621.396.11 2353
A Wide Band Aerial System for Circularly Polarized Waves, Suitable for Inospheric Research—G. J. Phillips. (Proc. I.R.E., vol. 98, pp. 237–239; Part III, May, 1951.) The system described can be used to select, without readjustment, one or other of two waves circularly polarized in opposite senses and incident vertically, within the frequency range 2 to 6 mc. Two mutually perpendicular horizontal dipoles are associated respectively with two phase-shifting LC lattice networks. In the antennas, emfs initially 90° out of phase, may be added in phase, thus giving selection of a circularly polarized component. A discrimination ratio of at least 12:1 between the components has been obtained.

621.396.677.001.4 2354
Reflecting Surface to Simulate an Infinite Conducting Plane—S. J. Raff. (*Jour. Appl. Phys.*, vol. 22, pp. 610–613; May, 1951.) A finite reflecting surface, which simulates an infinite plane, is required for calibrating measurements of reflections back to a microwave transmitting antenna. The Fresnel-zone method of physical optics is used for the design calculations. Variational calculus is used to determine the optimum reflector shape for a given antenna pattern, reflector size, and range of antenna-to-reflector distance. Theoretical values are compared with results obtained on an example constructed for use at 25λ from a dipole antenna.

621.392 2355
Transmission Lines and Networks [Book Review]—W. C. Johnson. Publishers: McGraw-Hill Book Co., New York, N. Y., 1950, 361 pp., $5.00. (*Electronics*, vol. 24, pp. 278–280; April, 1951.) "Although basically a textbook for undergraduate students, the material covered should be of considerable interest to practicing engineers in both the power and communication fields."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.6:621.317.726 2356
Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances—H. Elger. (*Arch. elekt. Übertragung*, vol. 4, pp. 413–426; October, 1950.) Methods are presented for calculating the buildup conditions in CR elements when either steady or time-varying voltage is applied. Consideration is given to potential-divider circuits. The effect of nonlinear resistances is investigated, as, e.g., in the demodulation of rectified modulated voltage with a square-law rectifier. Methods based on the theory are discussed for varying time constants within wide limits, and an example shows how the readings of a diode peak-voltage meter, operating at very high voltages, may be corrected for errors due to small discharges through the insulation between pulses.

621.3.015.7:621.387.4† 2357
Single-Channel Analyzer—J. E. Francis, Jr., P. R. Bell, and J. C. Gundlach. (*Rev. Sci. Instr.*, vol. 22, pp. 133–137; March, 1951.) An analyzer for proportional and scintillation counters counts the number of pulses the heights of which lie between E and $E+\Delta E$; ΔE is constant to within ±1.2 per cent for $E=0-90v$.

621.3.016.352 2358
Relation of Nyquist Diagram to Pole-Zero Plots—H. F. Spirer. (Proc. I.R.E., vol. 39, p. 562; May, 1951.) Comment on 1086 of June (Harman).

621.314.212:534.232:538.652 2359
Electroacoustic Transformation by means of Magnetostriction, with Special Reference to Radiation from Transformers—Rust. (*See* 2313.)

621.314.3† 2360
High-Gain Magnetic Amplifier—R. Feinberg. (*Wireless Eng.*, vol. 28, pp. 151–155; May, 1951.) Self-excitation is an effective method of obtaining feedback in a transductor. High values of current amplification can be obtained by making the number of turns of the self-excitation winding sufficiently large in relation to the number of turns of the load winding. Turns relations for stable operation are discussed. When operating unstably, the system may be used as an on-off trigger relay.

621.314.3†:621.318.42 2361
Design of [Magnetic-] Amplifier Inductors with Series-Connected Ohmic Loads—E. Helmes. (*Arch. elekt. Übertragung*, vol. 5, pp. 39–46; January, 1951.) A formula is derived expressing the effective permeability and self-inductance and the transfer impedance between the coil and the load in terms of the core cross section and the number of turns. Families of curves are plotted from measurements on (a) a two-element inductor with transformer-sheet core, (b) a three-element inductor with mumetal core. These curves can be applied to other core materials by changing the scale.

621.314.3†:[621.396.615.17+621.396.619.2 2362
The Use of Saturable Reactors as Discharge Devices for Pulse Generators—W. S. Melville. (*Proc. IEE*, vol. 98, pp. 185–204; Discussion, pp. 204–207; Part III, May, 1951.) The development of materials used for saturable reactors is outlined; a magnetic material with rectangular hysteresis loop is required. Magnetic discharge devices can often replace electronic discharge devices. The merits of the two types are compared. The design and operation of saturable-reactor circuits for radar pulse-modulation and ignitron ignition are described.

621.315.592†+621.314.632 2363
The Characteristics and some Applications of Varistors—F. R. Stansel. (Proc. I.R.E., vol. 39, pp. 342–358; April, 1951.) A tutorial paper reviewing the properties of semiconductor nonlinear circuit elements, with particular reference to those available commercially. The principles and limitations governing the use of these elements are summarized. The varying importance of the different parameters with different types of application is illustrated by short discussions of the design of voltage

limiters, power rectifiers, hf modulators, and compandors.

621.316.726:681.142 2364
Automatic Frequency Control—J. M. M. Pinkerton. (*Electronic Eng.*, vol. 23, pp. 142–143; April, 1951.) Description of a method of controlling the clock pulse frequency in the storage system of a digital computer. The phase of a pulse, which has traveled down an ultrasonic delay line, is compared with that of a later pulse of the same series which has not been delayed. The phase difference is used to derive a voltage for control of the frequency of the master oscillator producing the clock pulses, by means of a reactance tube.

621.316.86 2365
Pyrolytic Film Resistors: Carbon and Borocarbon—R. O. Grisdale, A. C. Pfister, and W. van Roosbroeck. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 271–314; April, 1951.) A description of the production and structure of thin carbon films, deposited on ceramics or fused silica by the pyrolysis of hydrocarbon vapors, which are capable of providing resistors of high stability with resistances from a few ohms to tens of megohms. The incorporation of boron in the film results in a smaller temperature coefficient than that of many wire-wound resistors, and the negligible skin effect permits advantageous use of these film resistors at high frequencies. Resistance values up to $10^{9}\Omega$ have been obtained in the borocarbon type. See also 583 of April.

621.317.353.2.012.3 2366
Mixer Harmonic Chart—T. T. Brown. (*Electronics*, vol. 24, pp. 132, 134; April, 1951.) The chart facilitates identification of spurious frequencies resulting from beating of various harmonics of two inputs, the frequency of one being variable.

621.318.572 2367
A Three-State Flip-Flop—A. D. Booth and J. Ringrose. (*Electronic Eng.*, vol. 23, p. 133; April, 1951.) With Type-6J6 and Type-6SN7 tubes, particular values of cathode resistor were found to give three stable states in flip-flop units. An explanation is given of the circuit operation.

621.319.53:621.396.9 2368
High-Voltage Pulse Modulators for Radar Pulse Transmitters—Tigler. (*See* 2431.)

621.385.3:546.289 2369
Duality as a Guide in Transistor Circuit Design—R. L. Wallace, Jr., and G. Raisbeck. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 381–417; April, 1951.) The properties of a transistor are compared with those of a vacuum-tube triode, and the relation between them is found to be such that, by interchanging current and voltage, a known vacuum-tube circuit can be transformed into one suitable for use with transistors. The necessary changes in the circuit elements are considered, and practical examples of these networks (or duals) are given. Circuits which permit the simultaneous use of vacuum tubes and transistors, such as the Doherty amplifier, are also discussed.

621.392 2370
The Potential Analogue Method of Network Synthesis—S. Darlington. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 315–365; April, 1951.) The method developed is based on the analogy between the gain and phase of linear networks and the two-dimensional potential and stream functions produced by charges corresponding to the network singularities.

621.392 2371
The Synthesis of RC Networks to Have Prescribed Transfer Functions—H. J. Orchard. (PROC. I.R.E., vol. 39, pp. 428–432; April, 1951.) A more general method than that of Guillemin (2462 of 1949) is described. The resulting network is in the form of a lattice, and is capable of providing any transfer function physically realizable by an RC network. The design procedure is simple. A numerical example is included.

621.392.4 2372
An Application of Equaliser Curves to the Design of Two-Terminal Networks—P. W. Seymour. (*P.O. Elect. Eng. Jour.*, vol. 44, Part I, pp. 31–32; April, 1951.) Description of a method for deriving the circuit constants of a 2-terminal network from available design data for a 4-terminal constant-impedance equalizer having an insertion-loss/frequency characteristic similar to the impedance/frequency characteristic required for the 2-terminal network.

621.392.5 2373
Fluctuations in a Linear System with Periodically Varying Parameters—S. I. Borovitski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 233–236; September 11, 1950. In Russian.) A system is considered with parameters varying in accordance with the equation (1). Statistically, the behavior of such a system can be represented by the Einstein-Fokker equation (2). With the aid of the main solution (3) of this equation, the case of a statistically stable system is investigated. The spectrum of a perturbation consists of discrete lines on a continuous background. The discussion is illustrated by experimental curves obtained in analyzing the output of a superregenerative receiver in the absence of a signal.

621.392.5 2374
Response Characteristics of Resistance-Reactance Ladder Networks—R. R. Kenyon. (PROC. I.R.E., vol. 39, pp. 557–559; May, 1951.) Generalized expressions for the transfer functions for resistance-reactance ladder networks are given and discussed in detail. The output voltage as a function of time is derived for the cases of unit-impulse and unit-step input voltage. Two methods are discussed for determining the output voltage for the case of an arbitrary input.

621.392.5 2375
Passive Pulse-Sharpening Circuits—L. Reiffel. (*Rev. Sci. Instr.*, vol. 22, pp. 214–216; March, 1951.) A circuit is described using a Ge damping diode in conjunction with a resonant circuit to shorten pulses obtained from a GM counter or other pulse detector.

621.392.52 2376
Relations between Signals and Spectra—K. Fränz. (*Arch. elekt. Übertragung*, vol. 5, pp. 10–14; January, 1951.) Theoretical proof of certain laws of filter theory. The major part of the energy of any signal of given duration is confined to a narrow spectrum whose limits are independent of the waveform. The connection between filter bandwidth and signal buildup is derived. The pass-band response curve of a filter with monotonic buildup must be roughly bell-shaped.

621.392.52:621.396.611.21 2377
Lattice-Type Crystal Filter—R. Lowrie. (*Electronics*, vol. 24, pp. 129–131; April, 1951.) Description of a 2-section filter incorporating eight crystals, with a pass band of width 3.9 kc centered at 2 mc, and a bandwidth of 12 kc at 60 db attenuation.

621.394/.396].6 2378
A New Colour-Coded Wiring System—N. G. Partridge. (*Electronic Eng.*, vol. 23, pp. 138–139; April, 1951.) The method is based on the consecutive numbering of all the items involved, whether wires, cableforms, chassis, or racks, using the international color-code system to enable the allotted number to be carried by each item in the form of colored bands or labels.

621.394/.396].6:621.392.012 2379
Correlation of Circuit Diagram and Wiring Development of Electronic Systems—A. W. Keen. (*Electronic Eng.*, vol. 23, pp. 144–145; April, 1951.)

621.396.6:061.4 2380
Trends in Components—(*Wireless World*, vol. 57, pp. 185–188; May, 1951.) A survey of the components and accessories shown at the annual private exhibition organized by the Radio and Electronic Component Manufacturers' Federation in London, April, 1951.

621.396.611+621.317.7].029.63 2381
Circuits and Measurement Apparatus for the 30-cm Band—Safa. (*See* 2479.)

621.396.611.21 2382
Some Notes on Overtone Crystals and Maintaining Oscillators operating in the Frequency Range of 33-55 Mc/s—J. B. Supper. (*Proc. IEE*, vol. 98, pp. 240–247; Part III, May, 1951.) The terms "minimum impedance" and "inductive impedance" are proposed for identifying two general forms of maintaining circuit investigated, and the concept of "impedance diameter" as a measure of crystal goodness is introduced. Measurements on different types of crystal and their behavior in the inductive-impedance oscillator, are recorded and discussed. The effect of plating area on crystals is considered, and attention is drawn to the improvement obtained by reducing the plating diameter below the present standard value.

621.396.611.21 2383
Amplitude of Vibration in Piezoelectric Crystals—E. A. Gerber. (*Electronics*, vol. 24, pp. 142, 216; April, 1951.) The amplitude of vibration of a crystal has some influence on the crystal parameters. Simple expressions are derived relating amplitude to the rf current through the crystal and to the voltage across it, and comparison is made with the results of more general theories of a vibrating piezoelectric plate. Only thickness modes of vibration are considered. Experiments verifying the formulas are described.

621.396.611.3 2384
The Calculation of the Input Impedance of Coupled Oscillatory Circuits—F. A. Fischer and U. John. (*Arch. elekt. Übertragung*, vol. 5, pp. 33–38; January, 1951.) Description of a method based on the representation of input impedance as a function of the difference between the damping and the resonance frequencies of the two circuits and the coupling factor. Curves and diagrams are plotted for a typical case. The basic formula is extended to the case of n coupled circuits.

621.396.611.4 2385
Application of the Method of Curvilinear Coordinates to the Calculation of a II-Type Cavity Resonator—V. L. Patrushev. (*Zh. Tekh. Fiz.*, vol. 20, pp. 727–734; June, 1950.) Krasnushkin has studied the propagation of waves in waveguides of rectangular cross section by using the method of normal waves. In a previous paper (*Bull. Acad. Sci.* (URSS), *Sér. phys.*, vol. 12, p. 684; 1948) the author applied this method to the investigation of the em fields and natural frequencies of cavity resonators having rotational symmetry. Under certain conditions, a II-type cavity resonator can be regarded approximately as a coaxial line with capacitance loading, and therefore belonging to this group of resonators. In the present paper it is shown that by introducing curvilinear co-ordinates, a rigorous solution of the problem is possible in principle.

621.396.611.4 2386
Design of the II-Type Cavity Resonator—V. L. Patrushev and O. V. Romanova. (*Zh. Tekh. Fiz.*, vol. 20, pp. 798–801; July, 1950.) A formula is derived for determining the length of the plunger used for tuning the resonator. The discussion is illustrated by two numerical examples which have been verified experimentally.

621.396.615.17 2387
Theory of the Symmetrical Multivibrator—N. A. Zheleztsov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 788–797; July, 1950.) The tube characteristic is assumed to consist of a number of linear sections. Analysis of the movement of the operating point along these sections makes it possible to determine the buildup of the discontinuous oscillations, and to prove the singularity and stability of the discontinuous periodic solution.

621.396.645 2388
Application of the Properties of Newtonian Potentials to the Design of Frequency-Modulation Amplifiers—P. Belgodère and A. Fromageot. (*Onde élect.*, vol. 31, pp. 18–32; January, 1951.) Use of a constant-gain amplifier to provide constant group-transmission time, (856 of 1950) requires that the width of the pass band be unnecessarily large. From a theoretical treatment, an alternative method of design is derived. This has yet to be checked experimentally, in particular as regards tolerances on tuning frequencies and circuit parameters.

621.396.645:535.247.4 2389
A Balance Indicator with High Input Impedance using a Cathode Follower—Dighton. (*See* 2513.)

621.396.645:621.317.083.4 2390
Sensitive Null Detector—M. G. Scroggie. (*Wireless World*, vol. 57, pp. 175–178; May, 1951.) Description of a selective af bridge amplifier for use at frequencies between 50 and 1,500 cps, with a "magic eye," milliammeter, or telephones as the output indicator. An agc circuit permits a range of signal input of 10 μv to 10v, and the time constant is such that a transient indication is given of a change in input at any signal level.

621.396.645:621.317.6 2391
The Determination of Amplifier Sensitivity with the Aid of the Noise Diode—Squires. (*See* 2478.)

621.396.645.012.8 2392
Network Representation of Input and Output Admittances of Amplifiers—F. W. Smith. (Proc. I.R.E., vol. 39, p. 439; April, 1951.) Comment on 2194 of 1949 (Vallese).

621.396.645.211 2393
Maximum Output from a Resistance-Coupled Triode Voltage Amplifier—J. M. Diamond. (Proc. I.R.E., vol. 39, pp. 433–434; April, 1951.) Simple formulas are derived for optimum load resistance with respect to output voltage, and for maximum voltage swing obtainable.

621.396.645.36.029.4 2394
Bridge-Compensated Differential Amplifiers—J. Labus. (*Arch. elekt. Übertragung*, vol. 4, pp. 437–440; October, 1950.) Sensitive push-pull amplifiers used for special purposes (e.g., electrocardiography), are liable to interference from ac fields at the input terminals. Several previously proposed circuits for eliminating this interference are briefly reviewed, and a method using a resistance-bridge network connected across the input, is described. This suppresses the interfering voltages before they reach the grids of the first-stage tubes, and hence prevents the production of harmonics.

621.396.645.37.029.4 2395
Independent Control of Selectivity and Bandwidth—O. G. Villard, Jr. (*Electronics*, vol. 24, pp. 121–123; April, 1951.) The feedback circuit in an RC af amplifier is designed to have constant amplitude/frequency but variable phase/frequency characteristics. The feedback is positive at the resonance frequency, negative at frequencies far from it. The complete circuit is shown for an amplifier with a constant percentage bandwidth/frequency variation, and a choice of three bandwidths at any desired selectivity.

621.396.822 2396
Thermal Fluctuation of Charge in Linear Circuits—E. A. N. Whitehead. (*Elliott Jour.*, vol. 1, pp. 32–34; March, 1951.) Derives the usual expressions for the noise power developed across an impedance; the noise generators are considered as being in parallel with the various circuit components.

621.392 2397
Transmission Lines and Networks [Book Review]—Johnson. (*See* 2355.)

621.392.025 2398
Alternating Current Circuits [Book Review]—R. M. Kerchner and G. F. Corcoran. Publishers: J. Wiley & Sons, New York, N. Y., 3rd edn. 1950, 586 pp., $5.50. (Proc. I.R.E., vol. 39, p. 448; April, 1951.) "... One of the very best summaries of the elementary background of ac circuit analysis . . ."

GENERAL PHYSICS

534.01+538.56 2399
Theory of Waves and Oscillations in Nonhomogenous Discrete Structures—P. E. Krasnushkin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1065–1083; September, 1950.) A general discussion applicable to various discrete oscillating systems, such as molecular chains of certain organic compounds, nonuniform waveguides and strings, the ionosphere, etc. The following two types of oscillations are met with in such systems: (a) "collective" oscillations of sinusoidal form spread more or less uniformly over all elements of the system, and (b) "local" oscillations of exponential form in parts of the space occupied by the system. In order to investigate the nature of these two types of oscillations and the conditions of their appearance, consideration is given to the general case of a chain structure consisting of cells, each of which represents an oscillating system with one degree of freedom. The local oscillations approximate in frequency and shape the natural oscillations of isolated cells, and the collective oscillations appear as a result of the interaction between the local oscillations, depending on the degree of resonance coupling, a conception introduced by Mandelstam. The loosening of the coupling may result in the appearance of collective oscillations only within parts of the system, and their penetration into other parts in the form of exponential "tails." Since the collective oscillations are essentially standing waves, the regions limiting them are called wave barriers. Three different types of these barriers are specified, and their effect on the operation of the system is discussed.

535.12 2400
Wave Propagation in an Anisotropic Medium and the Corresponding Principal Directions—M. Pastori. (*Nuovo Cim.*, vol. 6, pp. 187–193; May 7, 1949.) At any point within the medium, at least three principal directions exist such that the sum of the squares of the velocities of the three corresponding wavefronts is constant.

535.215:538.221 2401
The Surface Photoelectric Effect in Ferromagnetics—S. V. Vonsovski and A. V. Sokolov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 197–200; January 11, 1951. In Russian.) The anomalies in the photoelectric effect in ferromagnetics, observed by Cardwell (*Phys. Rev.*, vol. 76, p. 125; 1949) are discussed from the standpoint of the interaction between the outer *s*- and inner *d*-electrons, a concept developed by Vonsovski (2074 of 1947). The photoelectric current and the effective work function in ferromagnetics depend on the value of their spontaneous magnetization.

537.529 2402
A Review of Spark Discharge Phenomena—F. M. Bruce. (*Jour. Brit. IRE*, vol. 11, pp. 121–135; April, 1951.) The Townsend and streamer theories are discussed; the latter requires further evidence for its full substantiation, while the range of application of the former is likely to be greatly increased. Results of investigations in progress will be of importance, not only in the field of measurement over a wide variety of waveforms, but also in the extended use of gaseous insulation. Consideration is given to the uniform-field gap for standardizing methods of measurement. HF breakdown is not discussed here; this was dealt with in a paper noted in 613 of April.

537.533.8 2403
Some Peculiarities of the Secondary-Electron Emission from Thin Films of Calcium Chloride—V. N. Favorin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 916–922; August, 1950.)

537.562:537.311 2404
Convergence of the Chapman-Enskog Method for a Completely Ionized Gas—R. Landshoff. (*Phys. Rev.*, vol. 82, p. 442; May 1, 1951.) A note relevant to 335 of March (Cohen, Spitzer, and Routly).

537.71 2405
Generalized Electrical Formulas—V. P. Hessler and D. D. Robb. (*Elec. Eng.*, vol. 70, pp. 332–336; April, 1951.) "A set of generalized electrical formulas is developed to which units of any absolute system may be applied. The generalization is accomplished with the aid of two additional constants, n and u. The general formulas may be reduced to the usual rationalized or unrationalized forms, or to the Gaussian or Heaviside forms, by the substitution of tabulated numerical values of the constants n and u in the general form."

538.221 2406
Single-Domain Structure in Ferromagnetics, and the Magnetic Properties of Finely Dispersed Substances—E. Kondorski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 213–216; September 11, 1950. In Russian.)

538.221 2407
The Dependence of Magnetization Curves on Temperature and the Hysteresis Loop of High-Coercivity Alloys—Ya. S. Shur and N. A. Baranova. (*Compt. Rend. Acad. Sci.* (URSS), vol 74, pp. 225–228; September 11, 1950. In Russian.)

538.24 2408
The Effect of Directed Stresses on the Shape of the Magnetization Curve in Strong Fields—L. V. Kirenski and L. I. Slobodskoi. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 457–459; September 21, 1950. In Russian.) A formula (1) expressing the intensity of magnetization is quoted, and relations between the various constants are discussed, particularly for the case when the elastic stresses in the sample are directed along the magnetizing field.

538.249 2409
On Certain Laws Governing Magnetic Viscosity—R. V. Telesnin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 659–660; December 11, 1950. In Russian.)

538.249 2410
The Dependence of Magnetic Viscosity on Temperature—R. V. Telesnin and E. F. Kuritsyna. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 797–798; December 21, 1950. In Russian.)

538.56 2411
Applications of the Radiation from Fast Electron Beams—H. Motz. (*Jour. Appl. Phys.*, vol. 22, pp. 527–535; May, 1951.) The radiation from beams of fast electrons, passing through a succession of transverse electric or magnetic fields of alternating polarity, is examined. The frequencies and the angular distribution of radiated energy are calculated; the coherence of the radiation is discussed. Several applications appear possible, one of which is

the production of millimeter waves of considerable power. Another is the monitoring of the speed of electrons having energies up to 10^9 ev.

538.56:535.42 2412

Diffraction of Centimetre Electromagnetic Waves by Metal Disks—H. Severin. (*Z. angew. Phys.*, vol. 2, pp. 499–505; December, 1950.) The diffraction phenomena observed along the axis of, and close up to, a conducting disk, for normal incidence of a plane wave, are compared with three approximate theoretical solutions. A wavelength of 10 cm and disks of thickness 2 mm and radius 0.5λ, 1λ, 1.5λ, and 2λ, were used. Best agreement with measurements is provided by a theory which assumes the disk to be covered with a layer of magnetic dipoles.

538.56:535.42 2413

On the Diffraction of a Plane Electromagnetic Wave by a Paraboloid of Revolution—C. W. Horton and F. C. Karal, Jr. (*Jour. Appl. Phys.*, vol. 22, pp. 575–581; May, 1951.) A theoretical investigation of the diffraction by the convex surface of the paraboloid. Expressions are derived for the components of the incident, scattered and refracted waves. The case of a perfectly conducting paraboloid and a wave front perpendicular to the axis of rotation, is considered. The variation of amplitude of the scattered wave with distance along axis of rotation is compared with the corresponding curve for a sphere of radius equal to the radius of curvature of the paraboloid at its nose.

538.56:535.42 2414

On the Diffraction of Electromagnetic Waves by Two Conducting Parallel Half-Planes—M. G. Cheney, Jr., and R. B. Watson. (*Jour. Appl. Phys.*, vol. 22, pp. 675–679; May, 1951.) The diffraction produced by the edges of two half-planes, arranged one behind the other relative to the signal source, is investigated experimentally and theoretically. Agreement between the two sets of results is not very close.

538.561 2415

The Problem of the Excitation of Electromagnetic Oscillations—B. Ya. Moyzhes. (*Zh. Tekh. Fiz.*, vol. 20, pp. 698–715; June, 1950.) A general method has been recently proposed by G. A. Grinberg ("Selected Problems of the Mathematical Theory of Electric and Magnetic Phenomena", published by the Academy of Sciences of U.S.S.R., 1948) for solving a large group of problems in connection with the excitation of waveguides and other systems by a given distribution of currents, or by slots for which the tangential component of the electric field is known. In this method one or, more generally, two independent equations, are derived from Maxwell's equation. Each of these equations includes a scalar function (field component or a component of auxiliary function potentials) for which the boundary conditions have to be established separately. In the present paper this method is discussed in detail and applied to the cases of a cylindrical waveguide and a sectoral horn.

538.566 2416

Synthesis and Analysis of Elliptic Polarization Loci in Terms of Space-Quadrature Sinusoidal Components—M. G. Morgan and W. R. Evans, Jr. (PROC. I.R.E., vol. 39, pp. 552–556; May, 1951.) A mathematical analysis of elliptically polarized waves, by means of which the elliptic locus, resulting from three mutually orthogonal component field vectors, may be specified in terms of those vectors, or the vectors specified in terms of the locus. The simpler case of two-component vectors is considered first.

538.566:535.43 2417

Electromagnetic Scattering from Spheres with Sizes Comparable to the Wavelength—A. L. Aden. (*Jour. Appl. Phys.*, vol. 22, pp. 601–605; May, 1951.) The general formula for the back-scattering cross section of a sphere is difficult to evaluate for a complex refractive index, owing to the lack of the necessary tables of Bessel functions. The evaluation may be carried out by transforming the formula by means of logarithmic derivative functions. Back-scattering cross sections were measured for water spheres with sizes comparable to the wavelength (16.23 cm) using a standing-wave method. The water was contained in a thin hemispherical shell of dielectric, mounted on an aluminium disk. Very good agreement with theory was obtained.

538.566:621.396.67 2418

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2—Geometrical Representation of the Polarization of a Plane Electromagnetic Wave—G. A. Deschamps. (PROC. I.R.E., vol. 39, pp. 540–544; May, 1951.) The polarization and amplitude of an elliptically polarized plane wave may be specified by three quantities which define the ellipse traced out by the field vector, and which may be represented, according to a method used by Poincaré in optics, by the co-ordinates of a point on a sphere. Methods of solving problems using this concept are discussed.

538.566:621.396.67 2419

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas. Part 3—Elliptically Polarized Waves and Antennas—M. L. Kales. (PROC. I.R.E., vol. 39, pp. 544–549; May, 1951.) A complex vector algebra is presented by means of which an elliptically polarized wave may be completely specified. The field at a point may be resolved into two space components, in general not in the same direction, having time variations in phase quadrature. Thus if the two space vectors are U_r and U_i, the complex vector given by U_r+jU_i completely defines the field at the point. The algebraic properties of such vectors are discussed, and resolution of the field into orthogonal elliptically polarized components, and the concept of phase, are studied. Relations useful in measurements and antenna problems are given.

537.311.33 2420

Semi-Conductors. [Book Review]—D. A. Wright. Publishers: Methuen and Co., London, Eng., 130 pp., 7s. 6d. (*Wireless Eng.*, vol. 28, p. 164; May, 1951.) "Specially concerned with the theory of electron flow in semi-conductors, and across the boundary between them and either a metal or a vacuum. . . . This monograph will undoubtedly be of great use to students of the electron physics of solids."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523+551+621.396.93 2421

Recent Work of the Radiophysics Division C.S.I.R.O.—E. G. Bowen. (PROC. I.R.E. (Australia), vol. 12, pp. 99–108; April, 1951.) Developments in radio techniques for meteorology, astronomy, and navigation, are described. A digital computer has been designed and built by the Division to deal with its internal computing requirements.

523.72:621.396.822 2422

Radio Helioscopy—M. Waldmeier. (*Naturwiss.*, vol. 38, pp. 1–4; January, 1951.) A survey based on papers presented at the General Assembly of the International Scientific Radio Union. Results of solar-radiation measurements within the wavelength range 1 cm to 10 m, are discussed, and the adequacy of theories so far advanced, is examined. Most recent observations tend to support theories according to which the radiation disturbances are caused by coronal plasma oscillations excited by corpuscular rays or protuberances in motion.

523.746:538.12 2423

The Propagation of the Electromagnetic Field of Sunspots in the Sun's Atmosphere—P. E. Kolpakov and Ya. P. Terletski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 185–188; January 11, 1951. In Russian.) Because of its electrical conductivity, the highly ionized atmosphere of the sun might be expected to act as a screen for the electromagnetic field of the sunspots. That indications of this field are nevertheless observed in the middle and upper layers of the sun's atmosphere, is due to motion of the latter. A mathematical analysis of the propagation process, leading to the derivation of two differential equations (bottom of p. 187) is presented.

537.591:[523.854:621.396.822 2424

Cosmic Rays as a Source of Galactic R.F. Radiation—V. L. Ginzburg. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 377–380; January 21, 1951. In Russian.) Galactic rf radiation cannot be explained by the thermal radiation of interstellar electrons. It is suggested that it may be produced by the radiation from the electronic component of cosmic rays traveling at relativistic velocities in the magnetic fields near and between the stars. On this assumption, and knowing the intensity of the magnetic field, it is possible to determine the concentration of the corresponding cosmic particles. Results of a detailed analysis are given separately for the cases of general galactic emission, and emission from discrete sources. The possibility of cosmic-ray radiation in the magnetic field of the earth is also discussed. The results obtained are not conclusive.

550.38 2425

Geomagnetic Indices for the Period from 22nd Dec. 1950 to 31st March 1951—(*Z. Met.*, vol. 5, p. 123; April, 1951.) Observations made at Niemegk are presented in chart form.

551.510.535 2426

Sporadic E Movements on 21 June 1949—N. C. Gerson. (*Tellus*, vol. 3, pp. 56–59; February, 1951.) An analysis of the reports from about 300 American radio amateurs, of radio contacts made by reflection from sporadic-*E* regions. Between midnight and 0400 hours, two clouds of E_s ionization were reported which drifted westward across the United States with average velocities of about 250 km/hr.

551.510.535 2427

Magneto-Ionic Triple Splitting of Ionospheric Waves—O. E. H. Rydbeck. (*Onde élect.*, vol. 31, pp. 70–81 and 153–156; February and March, 1951.) French version of paper noted in 1147 of June.

551.510.535:621.396.11 2428

The Effect of the Lorentz Polarization Term on the Vertical Incidence Absorption in a Deviating Ionosphere Layer—J. M. Kelso. (PROC. I.R.E., vol. 39, pp. 412–419; April, 1951.) Using the two parabola approximations, and neglecting the effect of the earth's magnetic field, the influence of the Lorentz term on the apparent scale height and total absorption of a Chapman layer is calculated for a wave reflected in the layer. The absorption is increased by 4 per cent or more when the Lorentz term is included. See also 638 of 1950.

551.594.25 2429

The Origin of the Electric Charge on Rain—J. A. Chalmers. (*Quart. Jour. R. Met. Soc.*, vol. 77, pp. 249–259; April, 1951.) Previously measured values of charge on rain during periods when point discharge occurs, can be accounted for quantitatively on reasonable assumptions in terms of the process of ion capture. The results show that separation of charge must operate at levels down to about 800 m, and the consequences of this are discussed in relation to theories of the process of separation of charge.

621.317.79:621.396.822 2430

High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in KT_0 Units—Röschlau. (*See* 2487.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.319.53 2431

High-Voltage Pulse Modulators for Radar Pulse Transmitters—H. Tigler. (*Arch. elekt. Übertragung*, vol. 5, pp. 47–51 and 91–98; January and February, 1951.) Descriptive review of 9 different circuits for pulse generation using vacuum tubes, thyratrons, and spark discharge systems. Control of the spark gap, and voltage multiplication by means of Marx circuits, are discussed. This circuit and the spark discharge are especially suitable for short pulses at high power and high voltage.

621.396.93+523+551 2432

Recent Work of the Radiophysics Division C.S.I.R.O.—Bowen. (*See* 2421.)

621.396.933 2433

A Source of Error in Radio Phase Measuring Systems—R. Bateman, E. F. Florman, and A. Tait. (PROC. I.R.E., vol. 39, pp. 436–438; April, 1951.) Discussion on 2515 of 1950.

621.396.933 2434

A General Survey of Electronics in Air Transport—C. H. Jackson. (*Jour. Brit. IRE*, vol. 11, pp. 139–155; Discussion, pp. 156–159; April, 1951.) Communications, navigation aids, and aids to approach and landing, are reviewed and related to standards of safety. Details of technique and function are not considered.

621.396.933 2435

Radio on the Airways—(*Wireless World*, vol 57, pp. 199–202; May, 1951.) A general description of mf omnidirectional beacons and radio ranges, and also vhf marker beacons, as used on the main air routes in Great Britain. The beacons and radio ranges, operating in the 200 to 400-kc band, provide airway entrance markers and 4-direction course indication, respectively. Position information is given by the vertically radiating vhf marker system.

621.396.9 2436

Radar Systems and Components [Book Review]—Bell Telephone Laboratories. Publishers: B. Van Nostrand Co., New York, N. Y., and Macmillan and Co., London, Eng., $7.50 or 56s. (*Engineering* (London), vol. 171, p. 392; April 6, 1951.) A collection of papers covering the magnetron, the klystron, the resonant cavity, radar antennas, etc. Valuable for the specialist and for the general student.

MATERIALS AND SUBSIDIARY TECHNIQUES

338.987.4:621.396/.397].6 2437

Conservation of Critical Materials—W. W. MacDonald. (*Electronics*, vol. 24, pp. 84–87; April, 1951.) Discussion of design modifications to economize in the use of scarce metals while maintaining receiver performance. Television receivers are particularly considered.

537.311.33 2438

Electrical Properties of Grey Tin—G. Busch, J. Wieland, and H. Zoller. (*Helv. Phys. Acta*, vol. 24, pp. 49–62; February 15, 1951. In German.) Grey tin of high purity was prepared by prolonged cooling of spectroscopically pure metallic tin, and numerous alloys were made by adding small amounts of Al. Conductivity was determined by measuring the Q factor of a coil with a core of grey tin powder, at frequencies up to 30 mc; at 0° C the value found was $5\times10^3\Omega^{-1}$ cm^{-1}. Hall effect and variation of resistivity with applied magnetic field were measured by conventional dc methods. Remarkably large variations of resistivity were observed. The experiments show that grey tin is a semiconductor of high electrical conductivity, with properties very similar to those of Si and Ge.

537.311.33 2439

The Effect of Pressure on the Electrical Resistance of certain Semi-Conductors—P. W. Bridgman. (*Proc. Amer. Acad. Arts and Sci.*, vol. 79, pp. 127–148; April, 1951.) Measurements made on Ge, Si, and several oxides, are described. The resistances of all the oxides decrease with rising temperature up to 200°C, but there is no common type of variation with change of pressure up to 50,000 kg/cm². The Ge and Si were investigated under hydrostatic conditions to 30,000 kg/cm² at room temperature only. Differences of behavior as between n- and p-types are indicated and discussed.

537.311.33 2440

The Diffusion of the Current Carriers in Semiconductors with Mixed Conductivity—V. A. Lashkarev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 929–932; August 11, 1950. In Russian.) It is shown that "bi-polar" diffusion is due to a thermodynamically unbalanced state. The conditions are derived under which such a state is established, all excitation except thermal being excluded.

538.221 2441

Interaction between the *d*-Shells in the Transition Metals: Part 2—Ferromagnetic Compounds of Manganese with Perovskite Structure—C. Zener. (*Phys. Rev.*, vol. 82, pp. 403–405; May 1, 1951.) A discussion of the correlation between conductivity and ferromagnetism found by van Santen and Jonker (656 of April).

538.221 2442

Ferromagnetism in the Manganese-Indium System—W. V. Goeddel and D. M. Yost. (*Phys. Rev.*, vol. 82, p. 555; May 15, 1951.) About 25 alloys were prepared, with Mn contents ranging from 3 to 91 per cent by weight, in steps of about 4 per cent. Over half (3 to 50 per cent Mn) were found to show ferromagnetism believed to be due to a single phase (Mn_2In). Alloys containing up to 49 per cent Mn appear to be composed of In+Mn_2In, no eutectic being formed.

538.221 2443

An Investigation of the Magnetic Properties of Alloys of Manganese with Nickel and Cobalt—F. Gal'perin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 515–518; December 1, 1950. In Russian.)

538.221 2444

An Investigation of the Magnetic Properties of Well Ordered Alloys—F. Gal'perin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 647–650; December 11, 1950. In Russian.)

538.221:621.395.625.3 2445

Mixed Ferrites for Recording Heads—R. Herr. (*Electronics*, vol. 24, pp. 124–125; April, 1951.) Short discussion of the advantages of using ferrite materials instead of laminations in magnetic recording heads.

538.249 2446

Variation with Frequency of the Magnetic After-Effect in Powder Cores—R. Feldtkeller and H. Hettich. (*Z. angew. Phys.*, vol. 2, pp. 494–499; December, 1950.)

546.431.22 2447

Jumps in the Conductivity of Barium Titanate—N. A. Tolstoi. (*Zh. Tekh. Fiz.*, vol. 20, pp. 970–974; August, 1950.)

546.431.82 2448

X-Ray Investigations of the Ferroelectricity of Barium Titanate—W. Känzig. (*Helv. Phys. Acta*, vol. 24, pp. 175–216; April 10, 1951. In German.)

546.431.82 2449

The Nature of Electromechanical Oscillations in $BaTiO_3$ Ceramics—N. A. Roi. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 937–940; August 11, 1950. In Russian.) Discussion, with experimental curves, is presented for the following cases: (1) alternating field applied to nonpolarized sample (quasi-electrostriction); (2) weak alternating field applied to weakly polarized sample (linearized quasi-electrostriction); (3) weak alternating field applied to strongly polarized sample (linear piezoelectric effect).

546.431.82:548.55 2450

Elastic and Electromechanical Coupling Coefficients of Single-Crystal Barium Titanate—W. L. Bond, W. P. Mason, and H. J. McSkimin. (*Phys. Rev.*, vol. 82, pp. 442–443; May 1, 1951.) A report of measurements made on large multidomain single crystals.

548.0:537 2451

Ferroelectricity—B. T. Matthias. (*Science*, vol. 113, pp. 591–596; May 25, 1951.) A general discussion of known ferroelectric materials, and an examination of explanatory theories that have been advanced.

549.514.51 2452

Zero-Temperature-Coefficient Quartz Crystals for Very High Temperatures—W. P. Mason. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 366–380; April, 1951.) Crystals with zero temperature coefficient of frequency were obtained by making measurements of a series of rotated Y-cuts in the thickness shear mode, and a series of rotated X-cuts in the longitudinal length mode, and hence determining the orientation for AT-, BT-, CT- and DT-type crystals with low temperature coefficients, passing through zero value at a prescribed temperature. Calculations are given for crystals operating at 200°C. An AT-type crystal was investigated experimentally, and the calculated results agreed reasonably well with measured values. The maximum temperature at which the temperature coefficient of an AT-type crystal can have zero value is 190°C.

620.193.21:679.5 2453

Outdoor Weather Aging of Plastics under Various Climatological Conditions—S. E. Yustein, R. R. Winans, and H. J. Stark, (*ASTM Bull.*, no. 173, pp. 31–43. Discussion, p. 43; April, 1951.) A report on electrical and mechanical tests carried out on five types of clear transparent sheet plastics, six types of laminated material, and five types of moulded terminal bars, after prolonged exposure to tropical, dry desert, temperate, subarctic and arctic conditions.

621.3.015.5:621.315.61 2454

Electrical Breakdown over Insulators in High Vacuum—P. H. Gleichauf. (*Jour. Appl. Phys.*, vol. 22, pp. 535–541; May, 1951.) Experimental investigations in the pressure range $5\times10^{-3}-10^{-7}$ mm Hg are described.

621.315.61:539.23 2455

The Electric Tunnel Effect across Thin Insulator Films in Contacts—R. Holm. (*Jour. Appl. Phys.*, vol. 22, pp. 569–574; May, 1951.) Previous calculations apply to either very weak or very strong electric fields. The important practical case of intermediate-strength fields is here considered. The image force is neglected, but it is shown how this can be allowed for, approximately. Calculated values of tunnel resistivity are plotted against applied voltage for metallic and for semiconducting contact members. Some inadequacies of the theory are discussed.

621.315.61.011.5 2456

"Heat Developed" and "Powder" Lichtenberg Figures and the Ionization of Dielectric Surfaces Produced by Electrical Impulses.—A. M. Thomas. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 98–109; April, 1951.) Some experiments on both types of figures are reported, and their characteristics outlined. An explanation is suggested of the mode of formation of "heat developed" figures which are associated with the state of the surface of certain kinds of solid dielectrics. "Powder" figures are used to investigate the effect of repeated impulses of alternating polarity, and they show that the effect of a discharge of given polarity is not cancelled by a succeeding discharge of opposite polarity. The phenomena are discussed in relation to theories of surface breakdown and spark discharge.

621.315.612 2457
Contribution to the Study of Physico-Chemical Phenomena in the Ceramics Industry—R. Lecuir. (*Ann. Radioélect.*, vol. 6, pp. 20–50; January, 1951.) A discussion of the forming and sintering of oxides not possessing the plasticity characteristic of clays, which therefore require organic additions to the mix varying according to the forming technique used. The effect of the state of aggregation of the initial powder and of the application of high pressures on the compactness of the product, are examined together with other factors affecting the amount of shrinkage on sintering and the sintering temperature required.

621.315.612.011.5 2458
Dielectric Losses in Ceramic Dielectrics and in Barium Titanate at High Frequencies—A. L. Khodakov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 529–533; May, 1950.) Measurements were made of tan δ at frequencies from 10 to 200 mc and at temperatures from 15° to 180°C. In the case of $BaTiO_3$, tan δ decreases with temperature but remains practically constant within the frequency range specified.

621.318.2 2459
The Determination of the Optimum Parameters of Magnetic Systems with Permanent Magnets—A. Ya. Sochnev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 65–68; January 1, 1951. In Russian.)

621.775 2460
Special Nickel-Iron Alloys Prepared by Powder Metallurgy—Thien-Chi N'Guyen and B. Michel. (*Ann. Radioelect.*, vol. 6, pp. 3–19; January, 1951.) The first of a series of articles on the manufacture of magnetic materials, particularly Ni-Fe alloys. A brief general discussion of ferromagnetism is presented. The superiority of powder-metallurgy techniques for producing these alloys is shown by a study of the surface texture and crystal structure of a sintered 50 per cent Ni alloy.

661.1.037.5 2461
The Physical Aspect of Glass-Metal Sealability in the Electronic Tube Industry—G. Trébuchon and J. Kieffer. (*Glass Ind.*, vol. 32, pp. 165–174, 202, 240–247, 255 and 290–295; April to June, 1951.) English translations of paper noted in 2253 of 1950, 135 of February, and 929 of May.

666.2:549.623.5 2462
Glass/Mica Vacuum-tight Seals—J. Labeyrie and P. Léger. (*Le Vide*, vol. 6, pp. 936–940; January, 1951.) See 1437 of 1950 (Labeyrie).

537.311.33 2463
Electrons and Holes in Semi-Conductors [Book Review]—W. Shockley. Publishers: D. Van Nostrand Co., New York, N. Y., 1950, 543 pp., $9.75. (Proc. I.R.E., vol. 39, pp. 449–450; April, 1951.) "This excellent book might almost be said to consist of a set of three monographs of increasingly rigorous treatment. Part I, 'Introduction to Transistor Electronics', is entirely descriptive . . . Part II . . . is 'Descriptive Theory of Semiconductors'. . . . In Part III, 'Quantum Mechanical Foundations', many of the concepts presented in earlier portions of the book are subjected to rigorous examination."

549.514.51:621.396.611.21 2464
Quartz Vibrators and their Applications [Book Review]—P. Vigoureux and C. F. Booth. Publishers: H. M. Stationery Office, London, Eng., 30/-. (*Jour. Brit. IRE*, vol. 11, p. viii; April, 1951.) The properties of quartz, manufacture of crystals, and applications in telecommunications, etc., are considered.

MATHEMATICS

512.831 2465
The Principle of Minimized Iterations in the Solution of the Matrix Eigenvalue Problem—W. E. Arnoldi. (*Quart. Appl. Math.*, vol. 9, pp. 17–29; April, 1951.)

517.534:538.566 2466
On the Method of Saddle Points—B. L. van der Waerden. (*Appl. Sci. Res.*, vol. B2, no. 1, pp. 33–45; 1951.) In attempting to solve the problem of radio propagation over a plane earth, earlier investigators encountered a difficulty in evaluating the integral $\int_c we^{-\lambda u}dv$ by the method of steepest descents, because of the proximity of a pole to a saddle point. The problem is here transformed to one of integration in the u-plane. A solution is obtained in which the part of the integral corresponding to the pole can be readily separated out.

681.142 2467
Mechanized Reasoning. Logical Computors and their Design—D. M. McCallum and J. B. Smith. (*Electronic Eng.*, vol. 23, pp. 126–133; April, 1951.)

681.142 2468
Visual Presentation of Binary Numbers—E. H. Lenaerts. (*Electronic Eng.*, vol. 23, pp. 140–141; April, 1951.) By using a raster timebase, the pulses representing a number can be displayed as vertical deflections of the trace, or by a pattern of bright dots on a background of fainter dots representing the zeros. In a better method described, the pulse train is superimposed upon the frame-deflection timebase, while the brightness of the spot is modulated by clock pulses timed to occur in all positions where a spot is possible. The pulses thus appear as short vertical lines with dots interposed which represent the zeros.

681.142:517.9 2469
Solution of a System of Linear Equations with a Slightly Unsymmetrical Matrix by Using a Network Analyzer—H. L. Knudsen. (*Trans. Dan. Acad. Tech. Sci.*, no. 2, 16 pp.; 1950. In English.) The method is based on iteration, only a few steps being necessary when the asymmetry is only slight. No equipment other than the network analyzer is required.

681.142:621.316.726 2470
Automatic Frequency Control—Pinkerton. (*See* 2364.)

MEASUREMENTS AND TEST GEAR

531.765:529.786 2471
Comparing Outputs from Precision Time Standards—J. M. Shaull and C. M. Kortman. (*Electronics*, vol. 24, pp. 102–107; April, 1951.) Description of equipment developed at the National Bureau of Standards for monitoring the time-keeping of a group of standard quartz clocks. The chronograph records time differences of two clocks to within 1 ms, using spark-generating equipment whose rate is controlled by one clock, while the drum speed is governed by the other. A motor-driven switch connects each of several clocks in turn to the spark generator every 15 minutes, thus providing intercomparison data. The chronoscope uses a 3 inch cr tube, one clock and frequency divider being applied to produce a circular sweep with small fixed marker dots at 0.1-ms intervals, while a pulse from the circuit of the second clock produces a larger bright spot on the sweep. The chronograph and chronoscope are locked in time phase, so that observation of the position of the bright spot enables the time difference between the two clocks to be determined to within 20 μs. See also *Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 14–16; January, 1951.

535.322.4:546.217 2472
A Phase-Shift Refractometer—C. W. Tolbert and A. W. Straiton. (*Rev. Sci. Instr.*, vol. 22, pp. 162–165; March, 1951.) A description of apparatus for measuring small changes in the dielectric constant or refractive index of air, by determining the phase change of a 9.375-kmc wave over a 3 foot path. The test path may be confined to a waveguide through which air is drawn, or it may be the space between two antenna systems. The waveguide system can be used to measure rapid changes in the refractive index with an error <1 part in 10^6; the antenna method has larger errors because of flexibility of supports and external reflections.

621.317.083.4:621.396.645 2473
Sensitive Null Detector—Scroggie. (*See* 2390.)

621.317.335.3†+621.317.374].029.64 2474
Measurement of Dielectric Constant and Losses of Solid Dielectrics by means of Waveguides—G. D. Burdun. (*Zh. Tekh. Fiz.*, vol. 20, pp. 813–821; July, 1950.) Waves of the H_{01} mode are excited in a rectangular waveguide, the end of which is closed by a piece of the dielectric under investigation. The distribution of the dielectric field intensity inside the waveguide is measured by means of a probe and indicator, and from these measurements the properties of the dielectric are determined. The theory of the method is discussed, and the results of measurements with various dielectrics on wavelengths between 1.6 and 3.2 cm are presented. The accuracy of the method is to within about 1 or 2 per cent.

621.317.35:621.397.5 2475
Notes on TV Waveform Monitor Frequency Response—W. L. Hurford. (Proc. I.R.E., vol. 39, pp. 562–563; May, 1951.) A comparison of monitors having (a) very wide response band, (b) the response specified in the IRE Standards on Television (see 2035 of 1950), and (c) a sharp cutoff at twice the bandwidth of the IRE curve. From the response of these devices to clean, sharp pulses, and to pulses with spikes, it is concluded that "the IRE response is an excellent choice."

621.317.39:[621.318.4+621.319.4 +621.396.611.1 2476
Devices for the Measurement of the Temperature Coefficients of Coils, Capacitors and Oscillatory Circuits—C. Schreck. (*Fernmeldetech. Z.*, vol. 4, pp. 30–36; January, 1951.) Review of the development of methods and apparatus necessitated by the continual demand for higher frequency constancy. "Static" temperature coefficients (i.e., those related to external heating effects) have received more attention than "dynamic" coefficients (related to internal heating effects).

621.317.4:621.317.755 2477
The Electron-Beam Ferroscope—P. E. Klein. (*Arch. tech. Messen.*, no. 181, pp. T34–T-24; February, 1951.) Description and circuit details of a cro unit for displaying the magnetization curve of high-permeability iron-alloy samples. The voltage drop across a resistor in the test-circuit primary is applied to the X-plate amplifier; the output from the secondary is fed to the Y-plate amplifier, either direct or through an integrating circuit. By means of a 3-position switch the waveform of the primary current, secondary voltage, or B-H characteristic of the magnetic circuit, can be displayed. Typical traces are shown.

621.317.6:621.396.645 2478
The Determination of Amplifier Sensitivity with the Aid of the Noise Diode—W. K. Squires. (*Sylvania Technologist*, vol. 4, pp. 35–

37; April, 1951.) A measure of amplifier performance designated "sensitivity factor" is introduced. It is expressed quantitatively as the ratio of standard noise output to actual noise output at maximum gain, and its use enables the gain and noise factor to be correlated. A method of measuring the sensitivity factor is described.

621.317.7+621.396.611].029.63 2479
Circuits and Measurement Apparatus for the 30-cm Band—E. Safa. (*Onde élect.*, vol. 31, pp. 33–43; January, 1951.) Illustrated review outlining the characteristics of tubes and associated coupling circuits, and giving details of generators, wavemeter, wattmeter, and curve tracer, designed for the uhf band.

621.317.715:621.396.611.33/.34 2480
Coupling of A.C. Galvanometer to A.C. Amplifier—C. T. J. Alkemade and P. M. Endt. (*Appl. Sci. Res.*, vol. B2, no. 1, pp. 46–52; 1951.) A galvanometer with a 50-cps magnetic field is considered. Untuned transformer coupling provides higher gain than capacitor coupling, but magnetic relaxation phenomena in the transformer cause slow changes in sensitivity, which make capacitor coupling preferable.

621.317.725 2481
An Instantaneous Peak Voltmeter—M. W. Tobin, H. Grundfest, and R. L. Schoenfeld. (*Rev. Sci. Instr.*, vol. 22, pp. 189–190; March, 1951.) The operation of the diode capacitor peak-voltmeter circuit is discussed, and a circuit is described which provides a measurement of pulse amplitude unaffected by the amplitude of the previous pulse. This is used for investigating pulses of duration 0.1–20 ms at frequencies as low as 0.2 cps.

621.317.725 2482
Audio-Frequency Valve Voltmeter—S. Kelly. (*Wireless World*, vol. 57, pp. 215–218; June, 1951.) Details are given of a self-calibrating, portable instrument designed for a voltage range of 1 mv to 10 v in four decades, with an input impedance >10 MΩ across 10 pF and an output impedance <500Ω.

621.317.726:621.3.011.6 2483
Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances—Elger. (*See* 2356.)

621.317.76:621.396.615 2484
An Instrument for Recording the Frequency Drift of an Oscillator—W. W. Boelens. (*Philips Tech. Rev.*, vol. 12, pp. 193–199; January, 1951.) The meter was designed for measuring the frequency variation of the local oscillator of FM receivers in the 88 to 108-mc band. The reference frequencies, of which there are ten in the band 80.4 to 118.5 mc, are obtained from a 4.232-mc crystal oscillator by a multiplication and mixing process. The frequency drift is indicated by the variation of a direct current, which can be read on a dc meter or applied to a recording instrument.

621.317.79:621.3.018.78†:621.396.61 2485
Measurement of Distortion in Broadcast Transmitters—Müller. (*See* 2566.)

621.317.79:621.396.67 2486
A Phase Front Plotter for Testing Microwave Aerials—C. A. Cochrane. (*Elliott Jour.*, vol. 1, pp. 29–30; March, 1951.) A search antenna, servo-controlled via an rf phase discriminator, is used to find lines of constant phase, accurate to within about $\pi/16$ near the antenna. The search antenna is an open-circuited circular waveguide, suitable for wavelengths near 3.2 cm.

621.317.79:621.396.822 2487
High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in kT_0 Units—H. Röschlau. (*Arch. elekt. Übertragung*, vol. 4, pp. 427–434; October, 1950.) Requirements for a receiver to have high sensitivity, appropriate for the investigation of cosmic noise sources on a wavelength of 1.5 m, are examined. The choice of input tube and input circuit are discussed in detail, a cavity-resonator tank circuit being used on account of its high resonance resistance, together with a pentode with regeneratively-coupled screen grid. Three alternative methods of performing the absolute calibration are described; a method using a noise diode Type SA102 was most exact and gave a value better than 0.05 kT_0, for the sensitivity. Details are given of the method used for coupling the receiver to the antenna array.

621.317.799†:621.385.012 2488
Tube Characteristic Tracer using Pulse Techniques—H. M. Wagner. (*Electronics*, vol. 24, pp. 110–114; April, 1951.) A full description of an instrument designed primarily to obtain characteristic curves in the positive-grid region for small tubes used at high pulse power levels. The curves are displayed on a cro screen and can be recorded photographically. See also 2576 of 1950 (Leferson) and 692 of April (Graffunder and Schultes).

621.396.615.015.7.001.4† 2489
Radar Test Generator—K. S. Stull. (*Electronics*, vol. 24, pp. 93–95; April, 1951.) Circuit details and description of equipment providing triggered or free-running pulses of duration 0.25, 0.5, or 1.0 μs, and also cw signals in the range 47–76 mc, for testing wide-band circuits. The output voltage is variable from 0.1 μv to 0.1 v.

537.7 2490
Electrical Measurements and the Calculation of the Errors Involved: Part 1 [Book Review]—D. Karo. Publishers: Macdonald & Co., London, Eng., 1950, 191 pp., 18s. (*Nature*, (London), vol. 167, p. 745; May 12, 1951.) "... an extremely useful book, particularly for finals students, research workers, and engineers."

621.317.755 2491
Encyclopedia on Cathode-Ray Oscilloscopes and their Uses [Book Review]—J. F. Rider and S. D. Uslan. Publishers: J. F. Rider, New York, N. Y., 1950, 992 pp., $9.00. (*Electronics*, vol. 24, pp. 282, 284; April, 1951.) "Although the authors explain in the foreword to this book that some readers having special interest may find that it has limited coverage, the book quite adequately backs up its title for the average reader. . . . A novel aid . . . is a collection of synthesized waveform patterns, a total of 1580 extending over 79 pages. These are provided for those readers who do not have an harmonic wave analyzer available."

621.396.615.17 2492
Time Bases (Scanning Generators) [Book Review]—O. S. Puckle. Publishers: Chapman and Hall, London, Eng., 2nd edn, 387 pp., 30s. (*Electrician*, vol. 146, p. 1131; April 6, 1951.) This edition includes a new chapter on Miller capacitance timebases, as well as many other modifications and additions to the original 1943 edition.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

620.179.16 2493
Nondestructive Testing of Large Forgings by an Ultrasonic Method—W. Felix. (*Schweiz. Arch. angew. wiss. Tech.*, vol. 17, pp. 107–113; April, 1951.) An account of experience gained over a period of several years in the operation of actual tests using pulse-type equipment.

621.316.7.076.7+621-526 2494
Control Systems and their Application to Industry—W. R. Blunden. (*Jour. Inst. Eng.* (Australia), vol. 23, pp. 89–94; April/May, 1951.) A general survey, in which the different types of control system are classified both according to operation and application, and some of the more important types are described briefly.

621.317.083.7:551.508.1 2495
Cosmic-Ray Radiosonde and Telemetering System—M. A. Pomerantz. (*Electronics*, vol. 24, pp. 88–92; April, 1951.) Details of equipment comprising four GM counters which trigger a multivibrator controlling the keying of an uhf transmitter. Altitude and temperature are indicated by modulation intervals and modulation frequency.

621.365.5 2496
High-Frequency Generators and their Applications—W. Burkhardtsmaier. (*Telefunken Ztg.*, vol. 23, pp. 73–82; September, 1950.) Commercially available generators, with outputs ranging from 1.5 to 70 kw, are described and illustrated. Industrial heating applications are considered. A frequency of 400 kc is used for inductive heating, and 20 mc for dielectric heating.

621.365.54†:621.793 2497
Metal Evaporator uses High-Frequency Heating—R. G. Picard and J. E. Joy. (*Electronics*, vol. 24, pp. 126–128; April, 1951.) Difficulties due to nonwetting of filament, and its reaction with the material to be evaporated, are avoided by the use of a water-cooled hf heating coil. The method enables metals, that would react with the usual tungsten or tantalum filament, to be evaporated.

621.383:551.576 2498
Electronic Theory for Design of a Radar System using Light Waves (in particular for a Cloud Height Indicator)—A. Baude. (*Onde élect.*, vol. 31, pp. 44–48 and 90–101; January and February, 1951.) The principle and theory of distance measurement by pulsed light waves is discussed. An illustrated description is given of two sets of apparatus developed for measurement of cloud height and estimation of layer thickness. At 1,500m, error may be about 10m. Clear echoes from above 10km have been obtained despite intervening cloud layers.

621.384.611.2† 2499
The 300 MeV Synchrotron at the Massachusetts Institute of Technology—(*Engineer* (London), vol. 191, pp. 440–442; April 6, 1951.) A general account.

621.385.833 2500
Theory of the Three-Electrode Electrostatic Lens—É. Regenstreif. (*Ann. Radioélect.*, vol. 6, pp. 51–83 and 114–155; January and April, 1951.) An expanded account of work noted in 1213, 1743 and 2314 of 1950, 1455 of July, and 1742 of August.

621.385.833 2501
The Visualization of Atomic Distances by means of the Electron Microscope—L. Wegmann. (*Helv. Phys. Acta*, vol. 24, pp. 63–71; February 15, 1951. In German.) By stopping out certain concentric lens zones, it is theoretically possible to effect an improvement of the resolving power of uncorrected electron lenses sufficient to render visible the atomic lattices of crystals. The practical difficulties are explained briefly.

621.385.833 2502
Scattering Phenomena in Electron Microscope Image Formation—C. E. Hall. (*Jour. Appl. Phys.*, vol. 22, pp. 655–662; May, 1951.)

621.385.833 2503
Electronoptical Theory of the Deflection of an Extended Electronoptical Image by means of Crossed Electrical Deflection Systems—J. Himpan. (*Ann. Phys.*, (Lpz.), vol. 8, pp. 405–422; February 15, 1951.) The discussion presented is valid for mutually perpendicular deflection systems with plates of any type and arrangement, and it takes third-order aberrations into account. The conditions under

which deflection in two dimensions can be performed without introducing distortion are established. For deflections greater than those permitted in the ideal case, six different types of defect are recognized, viz., three types of over-all image distortion, and three of image-point deformation. Simple formulas are derived expressing these defects. Their magnitudes, in particular cases, are calculated by use of constants, which can be determined from a few measurements.

621.385.833 2504

Calculation of the Optical Constants of Powerful Magnetic Electron Lenses—W. Glaser. (*Ann. Phys.*, (Lpz.), vol. 8, p. 423; February 15, 1951.) Corrections to paper noted in 703 of April.

621.385.833:537.533.72 2505

The Significance of the Concepts 'Focus' and 'Focal Length' in Electron Optics and Strong Electron Lenses with Newtonian Image-Formation Equation—W. Glaser and O. Bergmann. (*Z angew. Math. Phys.*, vol. 1, pp. 363–379; November 15, 1950.) The functions determining the relation between the position of the object and that of the image, for linear magnification, are more complex in the case of electron optics than in the case of light. These functions are approached, in the neighborhood of two conjugate points, by the Newtonian osculating equation for the image function, including terms up to the fourth order. Each pair of conjugate points thus possesses corresponding foci, principal points, and focal lengths resulting from the Newtonian equation. If the osculating cardinal elements are independent of the pair of conjugate points chosen, they characterize by themselves the formation of the image, and are identical to the magnitudes defined in the usual way for light. Such fields, for which the image-formation equation of ordinary optics is strictly valid, are termed "fields with Newtonian representation."

A study is made of these strong fields, and examples of them are given which approximate the fields actually existing in electron lenses. In order to keep a physical significance for focal length, whatever the magnification, as close an approximation as possible to the empirical field must be obtained by one of the Newtonian type. A method for doing this is indicated, and experimental methods of determining the focal points and focal lengths of such approximate fields are examined.

621.385.833:621.311.1 2506

Modern Low-Power High-Voltage Generators—J. Vastel. (*Ann. Radioélect.*, vol. 6, pp. 84–94; January, 1951.) Various forms of generator suitable for supplying voltages in the range 30–80 kv, constant to within about 1 part in 100,000, are discussed, and the causes of ripple in the output are analyzed. Particular equipments described for supplying electron microscopes, etc., use low- and high-frequency oscillators (600 cps–46 kc), in association with voltage multipliers, giving output currents of about 100 μa and voltages of 60–80 kv.

621.386 2507

The Intensification of X-Ray Fluorescent Images—W. S. Lusby. (*Elec. Eng.*, vol. 70, pp. 292–296; April, 1951.) Paper given at AIEE Winter General Meeting, New York, January, 1951. The intensifier has a Cs-Sb photocathode arranged close to the ZnS input screen, an accelerating voltage of 30 kv causing the emitted photoelectrons to impinge on an Al-backed Zn-CdS output screen of reduced size at the far end of the 17-inch tube. Requirements for medical and industrial applications are discussed. Experimental installations, giving a brightness amplification of slightly over 100 times, have been put into operation.

621.386.1:621.385 2508

Radiographic Examination of Electronic Valves—H. B. van Wijlen. (*Philips Tech. Rev.*, vol. 12, pp. 207–209; January, 1951.) The examination of the electrode structure of tubes by means of X rays is described. A resolution of 6 μ is obtained with an image of approximately the same size as the object.

621.387.4† 2509

A Secondary-Electron Photon Counter—S. F. Rodionov and A. L. Osherovich. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 461–463; September 21, 1950. In Russian.) A description is given of a photomultiplier device for counting "visible" photons (λ=3600–6500 Å). A Sb-Cs photocathode developed by Kubetski is used, and light fluxes of the order of $10^{-14}-10^{-15}$ lumens can be measured.

621.387.4† 2510

The Discharge Mechanism for Oversize Pulses in Counters with Vapour Filling—H. Neuert. (*Ann. Phys.* (Lpz.), vol. 8, pp. 341–349; February 15, 1951.)

621.387.462† 2511

Silver Bromide Crystal Counters—K. A. Yamakawa. (*Phys. Rev.*, vol. 82, pp. 522–526; May 15, 1951.)

621.387.464† 2512

The Scintillation Counter—W. Hanle. (*Naturwiss.*, vol. 38, pp. 176–185; April, 1951.) A survey of the development and applications of photomultiplier-type counters, with a list of 156 references.

621.396.645:535.247.4 2513

A Balance Indicator with High Input Impedance using a Cathode Follower—D. T. R. Dighton. (*Jour. Sci. Instr.*, vol. 28, pp. 101–102; April, 1951.) "The use of the cathode follower circuit with high value grid resistances is discussed, and it is shown that a high-slope pentode can give an impedance conversion of 10^7. A cathode follower circuit, suitable for a null-balance indicator for photometric work, is described. The input impedance is 500 MΩ, the grid current about 10^{-10}a, and the detection limit 1 to 2 mv change of grid potential. A simple method of compensating for slow variations of heater voltage is employed."

621.38 2514

Survey of Modern Electronics [Book Review]—P. G. Andres. Publishers: J. Wiley and Sons, New York, N. Y., 1950, 522 pp., $5.75. (*Jour. Appl. Phys.*, vol. 22, pp. 685–686; May, 1951.) "Written as a text for a short survey course for students in electrical engineering . . . ."

621.365.54† 2515

Induction Heating [Book Review]—N. R. Stansel. Publishers: McGraw-Hill Publishing Co., London, Eng., 1949, 212 pp., 34s. (*Nature* (London), vol. 167, p. 700; May 5, 1951.) Of the nature of a handbook giving formulas and data relating to the electrical and thermal quantities involved.

PROPAGATION OF WAVES

538.566 2516

Is there a Zenneck Wave in the Field of a Radiator?—H. Ott. (*Arch. elekt. Ubertragung*, vol. 5, pp. 15–24; January, 1951.) Application of the modified saddle-point integration method (1024 of 1946) confirms the existence of a surface wave, but with coefficient half that of Sommerfeld's residuum, and only in the region of the boundary (earth) surface and for large values of refractive index. It is a component of a more general "surface effect" of fundamental significance. The validity of other theoretical solutions is discussed.

538.566 2517

The Nonexistence of the Surface Wave in the Radiation from a Dipole over a Plane Earth—T. Kahan and G. Eckart. (*Arch. elekt. Übertragung*, vol. 5, pp. 25–32; January, 1951.) Discussion of Sommerfeld's treatment of the problem, drawing attention to the mathematical error involved (2892 of 1949). Objections to Ott's theory (2516 above) are pointed out.

538.566 2518

The Radiation Principle—A. G. Sveshnikov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 917–920; August 11, 1950. In Russian.) The radiation principle (equation 2), introduced by Sommerfeld to ensure the unique solution of the wave equation (1), varies in accordance with the region to which the latter is applied. Accordingly, either the principle of limiting absorption proposed by Ignatovski in 1905 or the principle of limiting amplitude proposed by Tikhonov and Samarski (*Zh. Eksp. Teor. Fiz.*, vol. 18, no. 2, p. 243; 1948) should be applied.

538.566:517.534 2519

On the Method of Saddle Points—van der Waerden. (*See* 2466.)

621.396.11 2520

Comparison of Ionospheric Radio Transmission Forecasts with Practical Results—A. F. Wilkins and C. M. Minnis. (*Proc. IEE*, vol. 98, pp. 209–220; May, 1951.) "The production of muf forecasts for oblique transmission involves numerous operations on basic information obtained at vertical incidence. At each stage, errors are introduced whose cumulative effect determines the difference between predicted and observed circuit performance. The sources of the errors are examined and tentative values assigned to them with special reference to F_2 region. The computed value of the total error is compared with results obtained on commercial and Service circuits, and with observations made by other means. It is concluded that, although on the average, agreement is good, discrepancies remain which need further examination after the elimination of known sources of error. In a few cases, comparisons of predicted and actual times of fades due to ionospheric absorption have been made. Although the agreement between these times is reasonably good, it is believed that predictions of the actual field strength may be in error by large amounts."

621.396.11 2521

Evaluation of Ionosphere Observations—W. Becker. (*Arch. elekt. Übertragung*, vol. 4, pp. 391–400; October, 1950.) Development of an earlier theoretical paper (2844 of 1944). The magnitudes of inaccuracies due to using ray theory instead of rigorous wave theory are investigated, neglecting the earth's curvature and magnetic field, and assuming a vertical distribution of ionization decreasing gradually to zero both above and below a layer of maximum density. Application of the calculated results to the evaluation of fixed-frequency and swept-frequency records shows that the ray theory is reliable down to low values of relative thickness of ionosphere layers for all frequencies except those within a small range around the critical frequency and, for very oblique incidence, those below 0.2 times the critical frequency.

621.396.11.029.45 2522

The Ionospheric Propagation of Low- and Very-Low-Frequency Radio Waves over Distances less than 1,000 km—R. N. Bracewell, K. G. Budden, J. A. Ratcliffe, T. W. Straker, and K. Weekes. (*Proc. IEE*, vol. 98, pp. 221–236; May, 1951.) Results are summarized of experimental work performed at the Cavendish Laboratory over a period of years. Waves of frequency 16 to 30 kc are reflected as if from a sharp horizontal boundary at a height of 72±3 km (with the sun overhead), and waves of frequency 30 to 150 kc at about 75 km at oblique incidence, and perhaps 10 km higher at vertical incidence. The polarization is approximately circular at steep incidence, and linear on 16 kc at oblique incidence (65°). Absorption increases rapidly with frequency; differences are observed in behavior around sunrise at steep and oblique incidence; sudden ionospheric disturbances are associated with decreases in the apparent height of reflection. Present theories of reflection of very long waves are outlined.

621.396.11.029.64:621.396.621.087.4 2523

A Receiver for Measuring Angle-of-Arrival in a Complex Wave—F. E. Brooks, Jr. (Proc. I.R.E., vol. 39, pp. 407–411; April, 1951.) Descriptions of the design, construction, and calibration of a field-strength and wave-direction recorder developed at the University of Texas for operation on a wavelength of 3.2 cm. Two plane wave fronts can be measured simultaneously. The absolute amplitude of the dominant wave can be determined to within ± 1 db over a range of 70 db, the relative amplitude of the weaker to within ± 0.5 db. The application of phase-interferometry technique enables the angles of incidence to be measured to within $\pm 0.01°$.

621.396.812.029.62/.63 2524

Investigations of the Influence of the Troposphere on the Propagation of Ultra-Short Waves—R. Schachenmeier. (*Arch. elekt. Übertragung*, vol. 5, pp. 1–9; January, 1951.) Description of a method developed in 1941 to enable reliable calculations of field strength to be made from a knowledge of tropospheric conditions. The theory is based on the combined effects of atmospheric refraction of the ray, and the diffraction effect due to the earth's curvature. Values calculated from meteorological observations are in satisfactory agreement with measured field strengths at meter and decimeter wavelengths, for different land/sea paths. The cases of propagation beyond and within the optical range are treated separately.

621.396.812.029.64 2525

Attenuation of Radio Signals caused by Scattering—A. H. LaGrone, W. H. Benson, Jr., and A. W. Straiton. (*Jour. Appl. Phys.*, vol. 22, pp. 672–674; May, 1951.) An equation is developed for determining the total energy scattered during passage of a beam through unit volume of a scattering medium. Curves are plotted from which the attenuation due to scattering can be found.

621.396.812.3:551.510.535 2526

Multiple Reflections and Undulations in the F_2-Region of the Ionosphere—S. S. Banerjee and R. R. Mehrotra. (*Science and Culture* (Calcutta), vol. 16, pp. 72–73; August, 1950.) Anomalous variations of the amplitude of multiple reflections are observed even when transmitter and receiver are close together. In some cases, at frequencies between 6 and 11 mc (low compared with f_cF_2 at the time), the amplitude of any higher-order echo was greater than that of any lower-order echo. The effect may be caused by undulations in the lower structure of the F_2 layer as suggested by Ratcliffe (193 of 1949).

621.396.812.3:551.510.535 2527

Anomalous Behaviour of Multiply Reflected Echoes from the Ionosphere—S. N. Mitra. (*Science and Culture* (Calcutta), vol. 16, pp. 425–426; March, 1951.) An alternative explanation of the phenomenon noted by Banerjee and Mehrotra (2526 above). It is suggested that on normal quiet days the effect may be due to interference beween the two magneto-ionic components of the downcoming wave. Experiments have shown that the use of polarized waves removed the anomaly, the first reflection becoming always the strongest.

621.397.8.08 2528

U.H.F. TV Propagation Measurements—Cook and Artman. (*See* 2563.)

RECEPTION

621.396.621 2529

The Development of Commercial Receivers [by Telefunken]—H. Hart, G. Schaffstein, and G. Vogt. (*Telefunken Ztg.*, vol. 23, pp. 83–92; September, 1950.) Descriptions are given of the first four communications receivers developed by Telefunken after the war, for press and government use, viz., the EPK/1, for telegraphy and telephony; the E11-1/48 and EPH/L/2 "Hell" system teleprinter receivers, and the "Ball E1" relay and monitor receiver.

621.396.621 2530

The Development of the Telefunken Broadcast Receiver since 1945—W. F. Ewald. (*Telefunken Ztg.*, vol. 23, pp. 97–105; September, 1950.) Account of the difficulties surmounted in restoring production in the absence of nearly all normal facilities, and descriptions, with constructional and circuit details, of a range of portable and table models marketed up to 1950.

621.396.621.087.4:621.396.11.029.64 2531

A Receiver for Measuring Angle-of-Arrival in a Complex Wave—(*See* 2523.)

621.396.822 2532

Receiver Noise—W. Kleen. (*Fernmeldetech. Z.*, vol. 4, pp. 19–25, 56–63 and 182; January, February, and April, 1951.) Review of present-day theory and technique for determining optimum noise factor. Relations connecting antenna and circuit noise with transmission range are derived, and the physical basis of tube noise is described. Cosmic noise is discussed. A quantitative study is made of input-circuit noise for disk-seal tubes. Noise figures for the klystron, traveling-wave tube, etc., are discussed. For wavelengths below about 2m, the inherent noise of the receiver determines the signal/noise ratio. Over 50 bibliographical references are given.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2533

Symposium on Information Theory [London, 1950]—F. A. Fischer. (*Fernmeldetech. Z.*, vol. 4, pp. 79–85; February, 1951.) German report and comment. See also 984 of May (Jackson).

621.39.001.11 2534

Geometrical Interpretation in Hilbert Space of the Properties of Periodic or Pulsed Systems—R. Vallée. (*Ann. Télécommun.*, vol. 6, pp. 61–66; March, 1951.) The simpler properties of Hilbert space are summarized; a method is given for representing voltage and current as vectors in this space. Ohm's law is interpreted as a matrix operation: $\vec{V}=(Z)\vec{I}$, and Joule's law as a scalar product: $W=\vec{V}.\vec{I}$. Power factor is taken as the real part of $\vec{V}.\vec{I}/|\vec{V}|.|\vec{I}|$. A geometrical interpretation of the quantity of information contained in a periodic signal is deduced with a generalization of the above concepts for the case of a closed linear network. Similar relations are deduced for pulsed systems, the analogy resting on a time-frequency relation.

621.39.001.11:[621.317.083.7+621.398 2535

Application of Shannon's Communication Theory to Telecontrol, Telemetry, or Measurement—J. Loeb. (*Ann. Télécommun.*, vol. 6, pp. 67–76; March, 1951.) That part of Shannon's theory relating to sources capable of producing a finite number of different symbols is summarized; the properties of transmission channels in the absence and in the presence of noise, and the necessity for matching source to channel by suitable coding, are discussed. Loss of information due to noise is evaluated in the light of the concept of conditional entropy. The theory of a transmission channel in the presence of noise is treated as a generalization of which the Baudot-Verdan system of radiotelegraphic communication is a particular anticipatory application.

621.39.001.11:621.396.619.14 2536

Correlation Functions and Spectra of Phase- and Delay-Modulated Signals—L. A. Zadeh. (Proc. I.R.E., vol. 39, pp. 425–428; April, 1951.) "A delay-modulated signal may be regarded as the response of a delay modulator to the carrier of the signal. By using this point of view, a general expression for the correlation function of a delay-modulated signal is obtained. This expression is given in an operational form in which the operand is the correlation function of the carrier, and the operator is the correlation function of the delay modulator. The general result is applied to the determination of the correlation function of a delay-modulated signal having a periodic carrier, and, more particularly, to the determination of the correlation function of a phase-modulated signal."

621.395.44 2537

A Twelve-Channel Carrier Telephone System for Use on Open Wire Lines—T. B. D. Terroni. (*Strowger Jour.*, vol. 7, pp. 180–193; April, 1951.) A development of the cable system noted in 2891 of 1950 is described.

621.396.5 2538

U.S.W. Radiotelephony Technique—W. Runge. (*Telefunken Ztg*, vol. 23, pp. 67–72; September, 1950.) Meter-wave apparatus developed by Telefunken since the end of the war includes: (a) an FM mobile telephone installation especially for the police; (b) an easily portable FM transmitter for reporting and emergency purposes; (c) a fixed transmitter-receiver link of very high quality for connecting studio to broadcast transmitter. Descriptions are given of (a) and (b), (c) being reserved for a later paper.

621.396.5:621.395.722 2539

The New London Radio-Telephony Terminal—C. W. Sowton and D. B. Balchin. (*P.O. Elec. Eng. Jour.*, vol. 44, pp. 25–30; April, 1951.) The technical problems encountered in the construction of the new 48/80-circuit international radio telephone terminal in London are discussed, and the special control and supervision equipment installed is described with an outline of circuit operation. The use of automatic control circuits and standardized equipment results in considerable economies in staff and apparatus.

621.396.61/.62:623.6 2540

Progress in Military (Land Forces) Radio-communications—Morand. (*Onde élect.*, vol. 31, pp. 3–17; January, 1951.) Description of equipment of American pattern selected for production in France from 1945 onward. Concise details and illustrations are given of (a) three short-range telephony intercommunication sets, two using FM, the third PHM for use in vehicles; (b) two medium-range portable W/T-R/T sets using AM; (c) long-range equipment comprising a mobile station with trailer power unit; power is 250 w for R/T, 400 w for W/T, frequency range 2 to 18 mc; a 4-channel multiplex mobile station with transmitter power 50 w, designed to replace land-line systems. The trend of technical development in miniaturization, tropicalization, etc., is outlined.

621.396.65 2541

Short-Range Communication by V.H.F. Radio—(*GEC Telecommun.*, vol. 1, no. 2, pp. 61–79; 1946.) A summary of the factors governing the choice of systems for point-to-point and mobile services. Advantages of the vhf band mentioned are the reduced noise level, the convenient size of antenna systems, and the restriction of transmissions to the service area. Simplex, duplex, and relay systems using both AM and FM, are discussed, and a complete range of equipment for both fixed and mobile stations is described.

621.396.712 2542

150-kW Medium-Wave Broadcast Transmitter at Daventry—(*Engineering* (London), vol. 171, pp. 506–507; April 27, 1951.) A description of the new British Third Program transmitter and antenna system. The transmitter, which uses air-cooled tubes, consists of two identical 100-kw units which can be paralleled. The fading-free area is increased by connecting the coaxial feeder across an insulator at a point 460 feet (about two-thirds of the height) up the mast radiator.

621.396.933 2543
A General Survey of Electronics in Air Transport—Jackson. (*See* 2434.)

621.396.933 2544
The M.C.A. [Ministry of Civil Aviation] V.H.F. Area Coverage Network: Audio Frequency Distribution—J. L. French. (*Electronic Eng.* (London), vol. 23, pp. 146–148; April, 1951.) General description, with block diagram, of the af equipment and its operation. See also 2545 below.

621.396.933 2545
The M.C.A. [Ministry of Civil Aviation] V.H.F. Area Coverage Network: Provision of Transmitting Station Equipment—D. H. C. Scholes. (*Electronic Eng.* (London), vol. 23, pp. 148–150; April, 1951.) General description of the modified Type-T.1131 transmitter and its temperature-controlled crystal unit. See also 2544 above.

SUBSIDIARY APPARATUS

621-526+621.316.7.076.7 2546
Control Systems and their Applications to Industry—Blunden. (*See* 2494.)

621.526 2547
Servomechanisms with Linearly Varying Elements—M. J. Kirby. (*Elec. Eng.*, vol. 70, p. 343; April, 1951.) Digest of paper presented at the AIEE Fall General Meeting, Oklahoma, 1950. An analytical method is presented for determining the stability of a servomechanism in which one or more elements vary linearly with time.

621.316.722 2548
High-Voltage Stabilization by means of the Corona Discharge between Coaxial Cylinders—S. W. Lichtman. (Proc. I.R.E., vol. 39, pp. 419–424; April, 1951.) 1950 IRE National Convention paper. The design and performance of corona-discharge voltage-regulator tubes for operation with currents of 10 to 200 μa at voltages between 700 v and 40 kv, are described. The dependence of the mode of operation and efficiency on circuit parameters is discussed.

621.396.6.017.71.012.3 2549
Estimating Temperature Rise in Electronic Equipment Cases—R. J. Bibbero. (Proc. I.R.E., vol. 39, pp. 504–508; May, 1951.) The discussion is concerned with airborne equipment. Charts are presented as an aid in calculating temperature rises. Corrections are given for pressure variations, and the effects of case color, high aircraft speed, etc., are considered.

621.396.78† 2550
Power Supplies for Large Transmitters—H. Kropp. (*Fernmeldetech. Z.*, vol. 4, pp. 25–30; January, 1951.) Review of different types of rectifiers for low- and high-voltage supplies.

TELEVISION AND PHOTOTELEGRAPHY

621.397 2551
Cathode-Ray Picture Telegraphy—F. Schröter. (*Telefunken Ztg.*, vol. 23, pp. 111–118; September, 1950.) Inherent tube and circuit factors tending to reduce the resolution attainable in practice in a cathode-ray tube are discussed. For picture telegraphy, "flying-spot"-type scanning systems appear to be most suitable, used in conjunction with a cathode-ray tube at the receiver. The electronoptical development is described of systems of this type having the following properties: ability to transmit directly, by reflection scanning, unprepared material such as manuscripts, drawings and photographs; elimination of mechanical aspects capable of affecting quality of transmission; the possibility of varying the aspect ratio and of emphasizing particular parts of the material; immediate readability at the receiver on a long-lag screen producing the same sharpness and co-ordinate fidelity as at the transmitter.

621.397.5 2552
B.B.C. Television—T. H. Bridgewater. (*Electronic Eng.*, vol. 23, pp. 120–125; April, 1951.) A brief history of the outside-broadcasts section of the B.B.C. television service, together with a comparison of the characteristics and performance of the cable and radio links used to convey pictures from the pickup point to the main transmitter at Alexandra Palace. Future developments, which will greatly extend the scope of such broadcasts, are outlined. See also 752 of April.

621.397.5:535.62 2553
Quality of Color Reproduction—D. L. MacAdam. (Proc. I.R.E., vol. 39, pp. 468–485; May, 1951; *Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 487–512; May, 1951.) A discussion of methods of evaluating the quality of color reproduction in television in which, as in methods already used in color photography, subjective judgments are compared with color measurements made; e.g., those made on the I.C.I. system.

621.397.5:621.317.35 2554
Notes on TV Waveform Monitor Frequency Response—Hurford. (*See* 2475.)

621.397.5:778.5 2555
American Television Film Recording Equipment—R. B. Hickman. (*Jour. Telev. Soc.*, vol. 6, pp. 167–169; October–December, 1950.) The method used with the R.C.A. Kinephoto equipment to perform the conversion from the 30 frames per second of U.S. television to the 24 frames per second of standard motion-picture projection, is described. Either an electronic or a mechanical shutter can be used to blank out the image during the pull-down interval. Details of exposure and processing of the film are given.

621.397.611.2 2556
Television Camera Tubes—E. L. C. White and J. D. McGee. (*Wireless Eng.*, vol. 28, pp. 163–164; May, 1951.) Comment on 1264 of June (Bedford).

621.397.62 2557
Some Aspects of Single Side-Band Receiver Design—W. M. Lloyd. (*Jour. Telev. Soc.*, vol. 6, pp. 135–149; October–December, 1950.) The discrepancies which appear in the response of the receiver to a unit step are discussed theoretically in relation to those features of the frequency characteristics which give rise to them. The experimentally obtained step-responses of two typical receivers are shown.

621.397.62:621.385.2:546.289 2558
An Analysis of the Germanium Diode as Video Detector—Whalley, Masucci and, Salz. (*See* 2589.)

621.397.62:621.396.67 2559
Indoor Television Aerial—H. Page. (*Wireless World*, vol. 57, pp. 168–170; May, 1951.) The antenna consists of a horizontal slot, about $\lambda/2$ long, in a vertical conducting sheet with a gain of 4 db over a vertical $\lambda/2$ dipole. Details of construction and performance are given, showing negligible change of gain, impedance, and radiation patterns for a 10 per cent frequency change.

621.397.621.2 2560
Material-Saving Picture Tube—L. E. Swedlund and R. Saunders, Jr., (*Electronics*, vol. 24, pp. 118–120; April, 1951.) The use of an es focusing system instead of the usual magnetic type, economizes in alnico-5 and copper. Performance of the new electron gun is at least equal to that of the magnetic type, and may even be the better.

621.397.645 2561
New Video Circuits in Modern TV Sets—E. M. Noll. (*Radio-Electronics*, vol. 22, pp. 26–27; April, 1951.) Video amplifier circuits, used by a number of United States manufacturers, are illustrated and briefly described.

621.397.645 2562
Shunt-Regulated Amplifiers—V. J. Cooper. (*Wireless Eng.*, vol. 28, pp. 132–145; May, 1951.) Describes circuits used for modulating television transmitters, and designed to produce across a substantially constant load, large voltage swings regulated to ensure faithful reproduction. Numerous variants of the circuit are classified and analyzed. Practical applications and experimental results are given.

621.397.8.08 2563
U.H.F. TV Propagation Measurements—K. H. Cook and R. G. Artman. (*Tele-Tech*, vol. 10, pp. 50–51, 93, and 52–54, 82; March and April, 1951.) Measurements of peak field intensity of the vision signal, and observations of relative picture quality were made at 130 locations within 25 miles of the experimental transmitter at Kansas City, under typical broadcasting conditions. Vision frequency was 507.25 mc, radiated power, 3,450 kw. Equipment is described and results are reported and discussed.

621.396.615.17 2564
Time Bases (Scanning Generators) [Book Review]—Puckle. (*See* 2492.)

TRANSMISSION

621.396.61 2565
The First High-Power Transmitter Built Since 1945 [in Germany]—K. Müller. (*Telefunken Ztg.*, vol. 23, pp. 31–38; September, 1950.) Descriptions, with block diagrams and tube details, are given for the following: (a) 100-kw broadcast transmitter, 150 to 300 kc, at Königs Wusterhausen; (b) 5-/2-kw broadcast transmitters, 545 to 1,500 kc, for northwest Germany; (c) 20-kw broadcast transmitter, 545 to 1,500 kc, at Potsdam and Hanover; (d) 100-kw broadcast transmitter, 525 to 1,610 kc, at Berlin-Britz; (e) 30-kw telegraphy transmitter, 100 to 150 kc, at Bad Vilbel, Frankfurt am Main; (f) 60-kw telegraphy transmitter, 75 to 150 kc, also at Bad Vilbel. Innovations as compared with pre-1945 practice include: thoriated instead of plain tungsten cathodes in the directly heated tubes; single-circuit cooling; ignitron protecting devices for transmitters of power >20 kw; a simple thermostat control for the quartz crystals, giving frequency constancy to within 10^{-7} over periods of 24 hours.

621.396.61:621.317.79:621.3.018.78† 2566
Measurement of Distortion in Broadcast Transmitters—H. Müller. (*Telefunken Ztg.*, vol. 23, pp. 53–66, September, 1950.) Apparatus for the measurement of nonlinear distortion, developed by Telefunken from 1946 onwards, is described. The filter method of measuring harmonic distortion is adequate for the general monitoring of transmission quality. For more stringent requirements, particularly when investigating the nature of the distortion and its frequency dependence towards the upper transmission-frequency limit, a two-tone method such as that of von Braunmühl (456 of 1935) is used, enabling symmetrical and asymmetrical distortion to be separated. For the range 30 to 150 cps, a search-tone method is used, enabling the individual harmonics to be separated.

621.396.61:621.385.4 2567
A 30-Watt Transmitter for 430 Mc/s Employing the Transmitting Valve QQE 06/40

(AX 9903)—(*Phillips Tech. Commun.* (Australia), no. 2, pp. 14–17; 1951.) See also 2062 of September (Dorgelo and Zijlstra).

621.396.78† 2568
Power Supplies for Large Transmitters—H. Kropp. (*Fernmeldetech. Z.*, vol. 4, pp. 25–30; January, 1951.) Review of different types of rectifiers for low- and high-voltage supplies.

TUBES AND THERMIONICS

537.533.8 2569
Secondary Electron Emission from Aluminum Oxide—A. R. Shul'man and I. Yu. Rozentsveig. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 497–500; September 21, 1950.) A report on an experimental investigation of the effect of temperature on the secondary-emission coefficient of Al_2O_3. No variation with temperature was observed.

537.534.8 2570
Positive Emission from Thermionic Cathodes—K. H. Steigerwald. (*Z. angew. Phys.*, vol. 2, pp. 491–493; December, 1950.) Excitation of the fluorescent screen of an electron microscope was observed even when the negative voltage applied to the control electrode was sufficient to cut off electron emission from the cathode. The effect was traced to the emission of positive ions which release secondary electrons at the control electrode, the ion current being of the order of 10^{-9}–10^{-8} A for the tungsten hairpin cathode used, and occurring only in the range of cathode temperatures 1,300°k to 1,800°k. The emission is thought to depend on a vaporization process.

537.58 2571
Elements of Thermionics—W. E. Danforth. (Proc. I.R.E., vol. 39, pp. 485–499; May, 1951.) A survey, intended primarily for workers in other fields, of the principal experimental and theoretical developments in thermionics. From a simple basis of statistical mechanics, relations are derived including the Richardson equation, the Schottky field-effect equation, and the Fowler equation for emission from a normal impurity semiconductor.

621.314.632+621.315.592† 2572
The Characteristics and Some Applications of Varistors—Stansel. (*See* 2363.)

621.385 2573
Reliability in Miniature and Subminiature Tubes—P. T. Weeks. (Proc. I.R.E., vol. 39, pp. 499–503; May, 1951.) The meaning of the term "reliability" as applied to tubes is discussed. Reliability is found to be a function not only of tube design and quality, but also of the relation between tube ratings and the operating conditions and requirements. Specific features discussed include ruggedness, operating temperature, emission stability, and life, and consideration is given to the general effect of reducing tube size.

621.385 2574
Valve Development of Telefunken since the Cessation of Hostilities (1945)—H. Rothe. (*Telefunken Ztg.*, vol. 23, pp. 93–96; September, 1950.) On account of difficulties due to the condition of the various plants, no considerable development was possible till the second half of 1948. Since then the broadcast-receiver 11 series has been completed in glass-envelope tubes and, for usw FM receivers, in metal-envelope tubes. A new series of miniature tubes ("pico" series) was marketed in mid-1949. Some half-dozen power tubes intended for communications, broadcast transmitters, and industrial generators, are also very briefly described.

621.385:621.386.1 2575
Radiographic Examination of Electronic Valves—van Wijlen. (*See* 2508.)

621.385:621.396.619.16 2576
Signal-Retardation Electron Tubes with Delay Modulation—É. Labin. (*Onde élect.*, vol. 31, pp. 82–89; February, 1951.) An electron beam, modulated in density by the signal to be retarded, is passed between a pair of deflecting plates before injection into a retarding chamber within which delay is effected by means of a magnetic field, which causes the beam to follow a helical trajectory. The amount of the delay is dependent on the angle of entry into the chamber, and this is controlled by the deflector plates, to which the required "delay" modulation voltage is applied. In the nonmodulated condition, delay may be of the order of 100 μs. Modulated, the maximum obtainable delay is an inverse function of the modulation frequency; at the limiting frequency, dispersion due to space charge may limit the output current to <1 μa. This and other limitations of the method are discussed, and illustrations are given of tubes constructed to test the validity of the principle.

621.385.012:621.317.799† 2577
Tube Characteristic Tracer Using Pulse Techniques—Wagner. (*See* 2488.)

621.385.029.6 2578
Pulse Technique in High-Power Valve Development—A. M. Hardie. (*Metrop. Vick. Gaz.*, vol. 23, pp. 350–360; April, 1951.) The trend of high-power tube development is discussed, with particular reference to the future use of demountable tubes on sw service. Design problems are enumerated. A recording technique is described for presenting positive-grid characteristics in a form suitable for engineering applications. Either point-by-point or photographic recording may be used. Some examples of the latter are presented.

621.385.029.63/.64 2579
Amplification of the Traveling Wave Tube—B. Friedman. (*Jour. Appl. Phys.*, vol. 22, pp. 443–447; April, 1951.) A simpler and more exact method is presented for solving the transcendental equation given by Chu and Jackson (3549 of 1948) for wave propagation in the helix of the traveling-wave tube. The propagation is considered as a perturbed form of that in the cold helix. The dependence of amplification factor on geometrical parameters and operating conditions is determined explicitly. The tube will not amplify if the dc beam current is too high.

621.385.029.63/.64 2580
Effect of Hydrostatic Pressure in an Electron Beam on the Operation of Traveling-Wave Devices—P. Parzen and L. Goldstein. (*Jour. Appl. Phys.*, vol. 22, pp. 398–401; April, 1951.) Small velocity spreads in the electron beam appear to cause a decrease in gain and noise figure of a traveling-wave tube. The physical explanation is that the velocity spread introduces interactions between the electrons in which the external circuit takes no part, this interaction being of the nature of a hydrostatic pressure.

621.385.029.63/.64 2581
Travelling-Wave Tubes with Dispersive Helices—F. N. H. Robinson. (*Wireless Eng.*, vol. 28, pp. 110–113; April, 1951.) Oscillation occurs in traveling-wave-tube amplifiers when reflection takes place at mismatches between the ends of the helix and the external circuit. The difficulty of obtaining good matching over the wide frequency band of normal tubes has led to the development of a dispersive helix in which the phase velocity varies rapidly with frequency. This is achieved by making the diameter of the helix very small. Amplification then occurs for only a limited range of frequencies over which correct termination of the helix is possible. By this means the beam current required to produce a given gain is much reduced. Noise factor is also comparatively low.

621.385.032.213:537.533.8 2582
Secondary-Emission Cathodes of High Stability—B. D. Tazulakhov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 773–787; July, 1950.) The preparation of cathodes possessing a high stability under high temperatures and heavy current loads, was investigated experimentally. The requirements which the active and intermediate layers of the cathodes should satisfy are defined, and tables showing the properties of various suitable materials are given. The performance of complex BaO emitters deposited on Ag, Cu, Ni, nichrome, Mo, and Ta, is discussed in detail, and experimental curves showing the secondary emission from these cathodes are plotted. It is claimed that in the production technique proposed, the thickness control of the emissive layer is much simpler than in the usual methods, where it is more of the nature of an art than of a technological process.

621.385.15:621.385.831 2583
Voltage-Controlled Secondary-Emission Multipliers—A. J. W. M. van Overbeek. (*Wireless Eng.*, vol. 28, pp. 114–125; April, 1951.) Secondary-emission tubes have, in some cases, a much shorter life than normal tubes. This objectionable feature has been overcome by using a coating of Cs_2O on the dynodes and keeping their temperature below 180°C. The constructions of various experimental tubes are shown, and their characteristics described. A variable-μ tube and a very-high-slope tube with four stages of multiplication are shown. The use of grid dynodes is discussed. Some circuits in which secondary-emission tubes offer specific advantages are described, including generators of sinusoidal and nonsinusoidal oscillations, and trigger circuits.

621.385.16:537.312.5 2584
Magnetic Electron Multipliers for Detection of Positive Ions—L. G. Smith. (*Rev. Sci. Instr.*, vol. 22, pp. 166–170; March, 1951.) Two designs of 15-stage multipliers with crossed electric and magnetic fields are described. BeCu dynodes are used, of width $\frac{3}{8}$ inch for fields of 250 to 460 oersted, and $\frac{1}{8}$ inch for fields of 300 to 1,100 oersted. From their performance it is concluded that a multiplier of this type could be designed to have a rise time between 10^{-10} and 10^{-11} seconds.

621.385.2 2585
Effect of Variable Mass of the Electron on the Space-Charge Limited Current in a Diode—S. Visvanathan. (*Canad. Jour. Phys.*, vol. 29, pp. 159–162; March, 1951.) "The change in the current-potential distribution, due to the relativistic variation of the mass of the electron, has been calculated by suitable series expansions in the case of a plane parallel diode, and has been shown to be considerable in the case of large power tubes."

621.385.2 2586
The Transformation of Heat into Electrical Energy in Thermionic Phenomena—R. Champeix. (*Le Vide*, vol. 6, pp. 936–940; January, 1951.) An experiment is described and theory is adduced showing that, in a thermionic diode, the standing current vanishes when the two electrodes are at the same temperature, independently of the composition of the electrodes. Practical suggestions are made for the design of a diode without standing current, and for the determination of the actual source of emission of electrons from oxide-coated cathodes.

621.385.2:[546.27+546.289 2587
Crystal Diodes—R. W. Douglas and E. G. James. (*Proc. IEE*, vol. 98, pp. 157–168; Discussion, pp. 177–183; May, 1951.) The influence of small amounts of impurities on the electrical properties of semiconductors, and the mechanism of contact rectification, are discussed. The processing of Ge and Si for use in crystal diodes is considered in the light of the theory. The design and performance of

(a) a coaxial-type Si-crystal diode for use as a mixer at frequencies up to about 10 km, and (b) a wire-ended Ge-crystal diode are described. Particular attention is given to the frequency dependence of the rectification efficiency of the Ge diode, and to its application as a replacement for the thermionic diode.

621.385.2:546.289 2588
A New High-Conductance Crystal Diode—B. J. Rothlein. (*Sylvania Technologist*, vol. 4, p. 44; April, 1951.) The experimental Ge diode described is made by applying to the whisker contact an amount of metal paste so small that is does not add appreciably to the capacitance.

621.385.2:546.289:621.397.62 2589
An Analysis of the Germanium Diode as Video Detector—W. B. Whalley, C. Masucci, and N. P. Salz. (*Sylvania Technologist*, vol. 4, pp. 25–34 April, 1951.) Methods, including some rapid production-line tests, are discussed for the measurement of those characteristics of Ge diodes which are important in the detection of video signals. The forward and reverse conductances are assumed constant over the range of operation, and both loads with small and loads with large time constants are considered.

621.385.3+621.385.5 2590
Interelectrode Impedance in Triodes and Pentodes—E. E. Zepler and S. S. Srivastava. (*Wireless Eng.*, vol. 28, pp. 146–150; May, 1951.) Bridge measurements of capacitance and conductance were made at 1 mc and 32 mc, and the values were plotted against mutual conductance. Discrepancies between observed capacitance variations and the values indicated by North's theory (1450 of 1936) are discussed, and explanations are advanced for some of the effects.

621.385.3:546.289 2591
Effect of Auxiliary Current on Transistor Operation—H. J. Reich, P. M. Schulsheiss, J. G. Skalnik, T. Flynn, and J. E. Gibson. (*Jour. Appl. Phys.*, vol. 22, pp. 682–683; May, 1951.) Transistor gain characteristics may be improved by the flow of direct current between auxiliary electrodes, one of which is placed as close as possible to the collector. The best improvement in current gain is obtained with relatively large spacing between emitter and collector.

621.385.3:546.289 2592
Crystal Triodes—T. R. Scott. (*Proc. IEE*, vol. 98, pp. 169–177; Discussion, pp. 177–183; May, 1951.) The various forms of crystal triode developed up to date are reviewed. A brief résumé is given of the various materials for the manufacture of these triodes, and the types of control used to modify their characteristics. Testing procedure is discussed. Applications and circuit design are dealt with briefly.

621.385.3.029.64 2593
Passive Feedback Admittance of Disc-Seal Triodes—G. Diemer. (*Philips Res. Rep.*, vol. 5, pp. 423–434; December, 1950.) A discussion of the design of disk-seal triodes with a view to using the self-inductance of the grid wires to neutralize the feedback via the anode-cathode capacitance at microwave frequencies.

621.385.3.032.24 2594
Aspects in the Design and Manufacture of Planar Grids for Triodes at U.H.F.—W. J. Pohl. (*Electronic Eng.*, vol. 23, pp. 95–99; March, 1951.) Discussion, with calculations and application to practical manufacturing problems, of the relation between the grid dimensions and its ability to dissipate power. A recently developed method of producing planar ring-frame grids carrying highly tensioned wires, and the method of measurement of residual wire tension, are described. The most suitable material for constructing tensioned grids is tungsten wire with a copper coating of thickness about 15 per cent of the radius of the wire.

621.385.4:621.396.61 2595
A 30-Watt Transmitter for 430 Mc/s Employing the Transmitting Valve QQE 06/40 (AX 9903)—(*Philips Tech. Commun.* (Australia), no. 2, pp. 14–17; 1951.) See also 2062 of September (Dorgelo and Zijlstra.)

621.385.832.001.4 2596
A Note on Cathode-Resistance Stabilization of C.R.T. Gun Current—H. Moss. (*Electronic Eng.*, vol. 23, pp. 111–112; March, 1951.) An expression defining the increase in current stability produced by an autobias cathode resistor is deduced in terms of a stability factor S given by $2S=7E_c/E_d-5$, where E_c and E_d are the grid cutoff voltage and drive voltage, respectively. Graphs are drawn of S against cathode resistance for various cutoff voltages.

621.386.7 2597
Centering the Cathode in a Demountable X-Ray Tube—R. Fouret. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1651–1653; April 30, 1951.) Centering is performed while the tube is under vacuum, by an electrical method which consists of rendering minimum the capacitance between anode and cathode.

621.396.615.141.2 2598
On the Theory of the Anode Block of a Plane Magnetron—S. D. Gvozdover and V. M. Lopukhin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 955–960; August, 1950.) A mathematical discussion is presented of magnetrons with anode blocks of the hole-and-slot and slot types. To simplify the discussion, the space occupied by the anode block is divided into the interaction space, which does not include the anode resonators, and the space which includes these resonators. Equations for the electromagnetic fields in the interaction space are derived and solved, and the natural frequencies of oscillation are determined by an approximate method in which the complex impedances of the interaction space are matched to those of the resonators. The discussion is limited to the case of two-dimensional (plane) magnetrons, and the end effects, as well as the effects of couplings, are neglected.

621.396.615.141.2:537.525.92 2599
The Space-Charge in a Magnetron under Static Cut-Off Conditions: Planar or Quasi-planar Magnetron—G. A. Boutry and J. L. Delcroix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1413–1415; April 9, 1951.) The two static cutoff states [1828 of 1950 (Delcroix and Boutry)] are compared; the total space charge is the same for the two cases. The condition of lower energy level of the electron gas corresponds to the Brillouin state. The cutoff surface is the same for the two cases.

621.396.615.141.2:537.525.92 2600
The Space-Charge in a Magnetron under Static Cut-Off Conditions: Cylindrical Magnetron—J. L. Delcroix and G. A. Boutry. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1653–1655; April 30, 1951.) The two static cutoff states are compared (see 2599 above for corresponding consideration of planar magnetrons). The total space charge is different for the two cases; the condition of lower energy level of the electron gas corresponds to the "bidromic" state. The cutoff surface is not generally in the same position for the two states.

621.396.615.141.2:537.525.92 2601
Analysis of Synchronous Conditions in the Cylindrical Magnetron Space Charge—H. W. Welch, Jr., and W. G. Dow. (*Jour. Appl. Phys.*, vol. 22, pp. 433–438; April, 1951.) "In the multianode cylindrical magnetron there exist favored phase velocities of the electromagnetic wave around the interaction space between anode and cathode These velocities are characteristic of the resonant system attached to the anode segments. In the oscillating magnetron, the electronic space charge within the interaction space is presumed to maintain synchronism with one of these velocities. Certain of the conditions of synchronism, which can be discussed analytically, are treated in this paper The results, although based on restrictive assumptions, can be used in the interpretation of magnetron operation, and in predicting regions of efficient behavior. See also 1282 of June (Welch).

621.396.615.142 2602
The Limiting Efficiency of Oscillation Generation by Means of Velocity-Modulated Electron Beams in Drift-Space Valves with Fields of Finite Length—R. Gebauer and H. Kosmahl. (*Z. angew. Phys.*, vol. 2, pp. 478–486; December, 1950.) The concept of the ideal efficiency is introduced. This quantity is a measure of the greatest possible amount of hf energy which can be extracted from the tube, neglecting the velocity modulation, and is hence also an indication of the quality of the focusing. The optimum length of drift-space for a given modulation depth and control-gap length is calculated. The focusing properties of infinitely short and finite-length fields are compared and found to be equivalent only for vanishingly small modulation. The relation of the practically attainable limiting efficiency to the ideal efficiency, is defined. The practical limiting efficiency decreases with increase of modulation depth.

621.385.032.216 2603
Die Oxydkathode: 2. Teil—Technik und Physik [Book Review]—G. Herrmann and S. Wagener. Publishers: J. A. Barth, Leipzig, Germany, 2nd edn, 1950, 284 pp. (*Fernmeldetech. Z.*, vol. 4, p. 46; January, 1951.) A modern and exhaustive exposition of the subject. Vol. 1: 2985 of 1949.

MISCELLANEOUS

621.396 2604
Radio Progress during 1950—(Proc. I.R.E., vol. 39, pp. 359–396; April, 1951.) A survey based on material compiled by the 1950 Annual Review Committee of the IRE, and including 1,084 references. The material is grouped under the following headings: antennas and waveguides; audio techniques; electroacoustics; sound recording and reproducing; circuit theory, electron tubes and solid-state devices; electronic computers; facsimile; industrial electronics; measurements; mobile radio; modulation systems; navigation aids; piezoelectric crystals; radio transmitters; receivers; standards on symbols; television system; video techniques; wave propagation.

621.396(083.72) 2605
Standards on Abbreviations of Radio-Electronic Terms, 1951—(Proc. I.R.E., vol. 39, pp. 397–400; April, 1951. Reprints of this Standard, 51 IRE 21 S1, may be purchased while available, from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy.

621.396 Tesla 2606
The Life and Work of Nikola Tesla—A. Damianovitch. (*Bull. Soc. franç. Élect.*, vol. 1, pp. 85–99; February, 1951.) Lecture before the Société française des Électriciens, reviewing the pioneer work of Tesla in the field of ac and radio engineering.

621.39 2607
Electrical Engineers' Handbook—Electric Communication and Electronics [Book Review]—H. Pender and K. McIlwain. Publishers: Chapman and Hall, London, Eng., and J. Wiley and Sons, New York, N. Y., 4th edn, 1345 pp., 68s. (*Electrician*, vol. 146, p. 823; March 9, 1951.) An entirely rewritten edition, with contributions by 78 specialists. FM and pulse techniques in communications and radar are included for the first time. A bibliography is appended to each of the 23 sections.

FXC

OUTLINE OF ESTABLISHED AND POTENTIAL APPLICATIONS

WRITE FOR BULLETIN

OUTLINE OF APPLICATIONS	RECOMMENDED FERROXCUBE MATERIAL	SHAPE
1 TELEVISION FLYBACK CIRCUITRY		
a) Flyback transformers	3C	U-Core
b) Deflection yokes	3C	Ring segment
c) Correction coils—to improve sawtooth linearity	3C	Slug
2 RADIO RECEIVERS		
a) I F Transformers	Depends upon Frequency	Slug
b) R F Tuning Coils		Slug
i) fixed L		
ii) permeability tuning		
c) Antenna cores	4B	Rod
3 TELEPHONY (Voice Frequency and Carrier)		
a) Interstage transformers	3C	E-Core
b) Transformer for matching to co-axial cable	3C	E-Core
c) Loading coils	Special grade	Pot-Core
d) Filter circuits (not limited to telephony)	Special grade	Pot-Core
e) Delay lines (not limited to telephony)	Special grade	Pot-Core
4 PULSE NETWORKS AND TRANSFORMERS		
a) Signal-shaping	Depends upon Pulse width	Simple closed magnetic circuit
b) Power—to feed magnetron directly—built up from Ferroxcube rods		
c) Low-power—e.g., in computer applications	Special grade	
5 MODULATION APPLICATIONS		
a) Use of loss effects to achieve AM without FM in modulating Klystron output	4B	Rod
6 APPLICATION OF NON-LINEAR EFFECTS—e.g., in saturable core reactors		Toroid or rod with saturating circuit
a) Permeability tuning of diathermy apparatus	4B	
b) Pulse generation from sine waves		
c) Magnetic amplifiers and saturable core reactors		
7 RECORDING HEADS		
8 IGNITION COILS		
a) Automotive		
b) Aircraft		
9 MAGNETOSTRICTION APPLICATIONS		
a) Band-pass filters		
b) Transducers		

FERROXCUBE

THE MODERN CORE MATERIAL

FERROXCUBE CORPORATION OF AMERICA

A Joint Affiliate of Philips Industries and Sprague Electric Co., Managed by Sprague

50 East 41st Street • New York 17, New York • Factory: Saugerties, New York