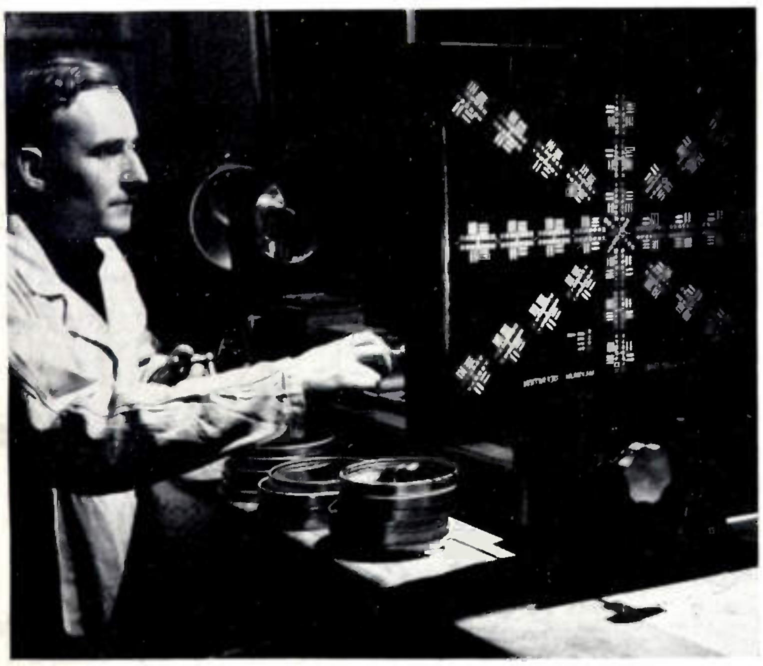
# Proceedings of the I·R·E

## A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

**June, 1949**

Volume 37 Number 6

North American Philips Co., Inc.

TELEVISION OPTICS

Optical test charts, hitherto mainly used for photographic-quality controls, now check performance of television picture-projection systems.

PROCEEDINGS OF THE I.R.E.



Waves and Electrons Section




Table of Contents Follows Page 32A



The Institute of Radio Engineers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COMPONENTS FOR EVERY APPLICATION
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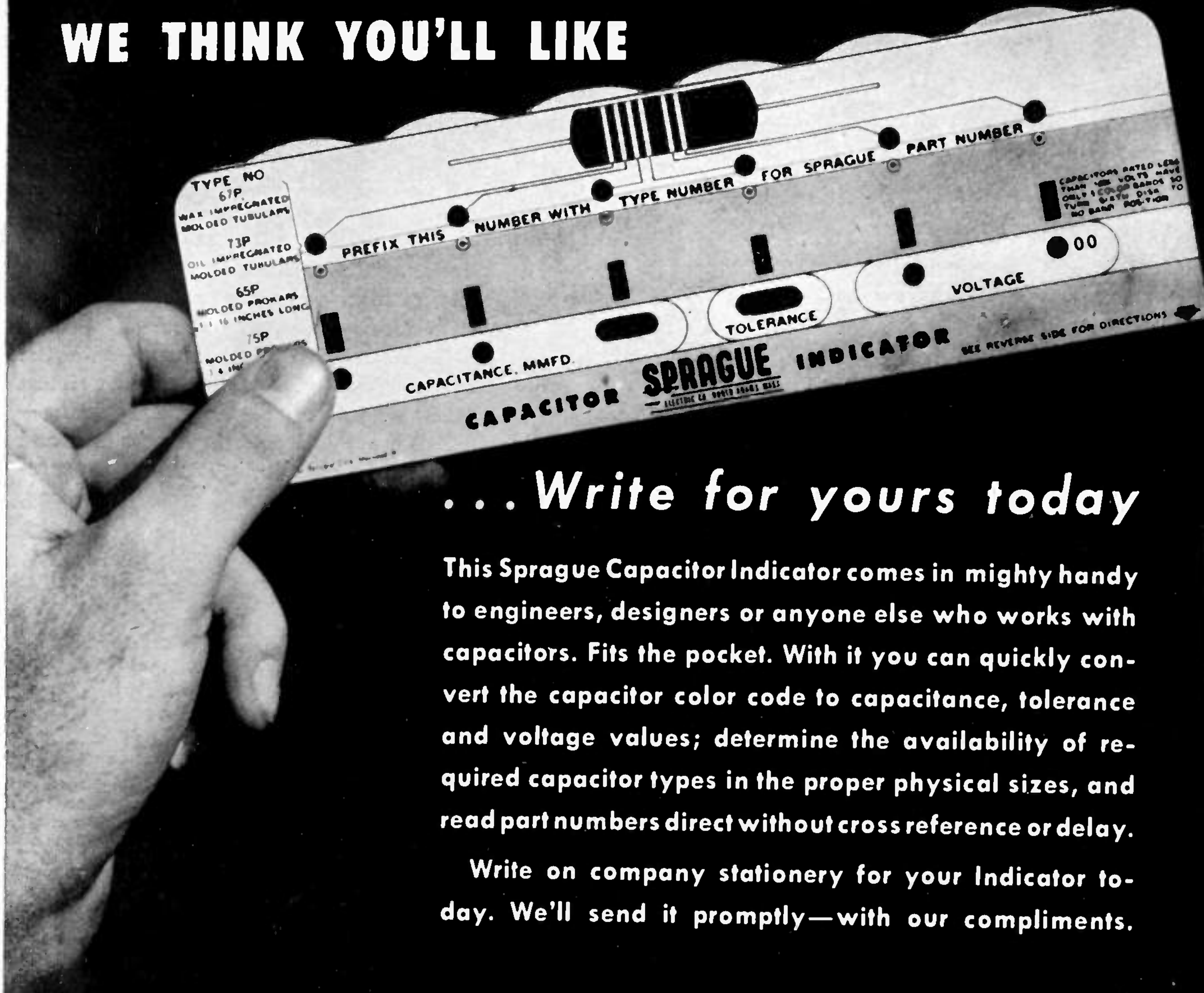
PROCEEDINGS OF THE I.R.E. June, 1949, Vol. 37, No. 6. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of contents will be found following page 32A

# A HANDY LITTLE GADGET WE THINK YOU'LL LIKE

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**The most important paper dielectric capacitor development since the war.**

Smaller than conventional paper tubulars—highly heat- and moisture-resistant—non-inflammable—rugged—vibration-resistant—conservatively rated for −40°C. to +85°C. operation.

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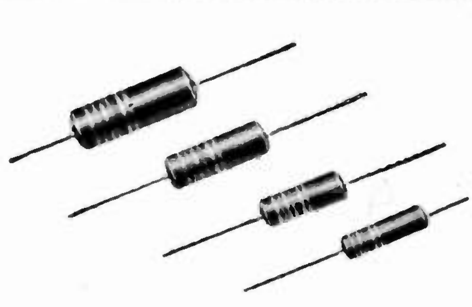

**SPRAGUE ELECTRIC COMPANY, NORTH ADAMS, MASSACHUSETTS**

# -hp- OSCILLATORS GIVE YOU PRECISE TEST VOLTAGES from ½ cps to 10 mc!

**-hp- MODEL 200C**

**From ½ cps to 10 mc,** there's an *-hp-* resistance tuned oscillator engineered to your exact need. Ten precision oscillators in all, including a portable unit that operates from batteries. Each has the familiar *-hp-* advantages of high stability, constant output, wide frequency range, low distortion and no zero set during operation. *-hp-* precision oscillators are used by radio stations, manufacturers, research laboratories and scientists throughout the world.

**SPECIFICATIONS OF -hp- OSCILLATORS**

| INSTRUMENT | FREQ. RANGE | OUTPUT | DISTORTION | FREQ. RESPONSE | PRICE |
|---|---|---|---|---|---|
| -hp- 200A | 35 cps to 35 kc | 1 watt/22.5v | Less than 1% | ± 1 db to 15 kc | $120.00 |
| -hp- 200B | 20 cps to 20 kc | 1 watt/22.5v | Less than 1% | ± 1 db to 15 kc | 120.00 |
| -hp- 200C | 20 cps to 200 kc | 100 mw/10v | Less than 1% to 20 kc | ± 1 db to 150 kc | 150.00 |
| -hp- 200D | 7 cps to 70 kc | 100 mw/10v | Less than 1% 10 cps to 70 kc | ± 1 db throughout range | 175.00 |
| -hp- 200H | 60 cps to 600 kc | 10 mw/1v | Less than 3% | ± 1 db, 60 cps to 600 kc | 350.00 |
| -hp- 200I | 6 cps to 6 kc | 100 mw/10v | Less than 1% above 10 cps | ± 1 db, 6 to 6000 cps | 225.00 |
| -hp- 201B | 20 cps to 20 kc | 3 w/42.5v | Less than ½% (1 watt output) | ± 1 db throughout range | 250.00 |
| -hp- 202B | ½ cps to 50 kc | 100 mw/10v | Less than 1% 1 to 1000 cps | ± 1 db, 10 to 50,000 cps | 350.00 |
| -hp- 202D | 2 cps to 70 kc | 100 mw/10v | Less than 2% 10 cps to 70 kc | ± 1 db, 7 cps to 70 kc | 275.00 |
| -hp- 204A (Battery Op'd.) | 2 cps to 20 kc | 2.5 mw/5v | Less than 1% | ± 1 db throughout range | 175.00 |
| -hp- 650A | 10 cps to 10 mc | 15 mw/3v | Less than 1% 100 cps to 100 kc | ± 1 db throughout range | 475.00 |

***For complete details on any -hp- instrument, write direct to factory or contact the -hp- technical representative nearest you.***

**HEWLETT-PACKARD CO.**
1855-D Page Mill Road, Palo Alto, California
Export Agents: Frazar & Hansen, Ltd.
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## ANOTHER -hp- SERVICE

### Person-to-Person Help With Your Measuring Problems

Almost anywhere in America, *-hp-* field representatives can give you personal help with your measuring problems. They have complete data on *-hp-* instruments, their performance, servicing and adaptability. Call the nearest *-hp-* field representative whenever, wherever you need help with a measuring problem.

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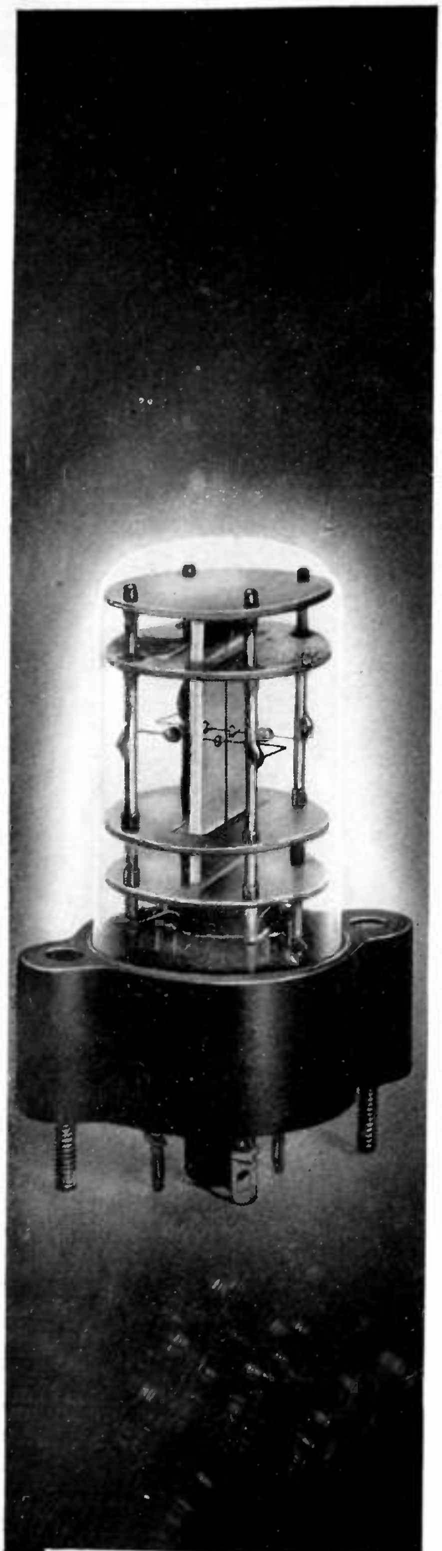
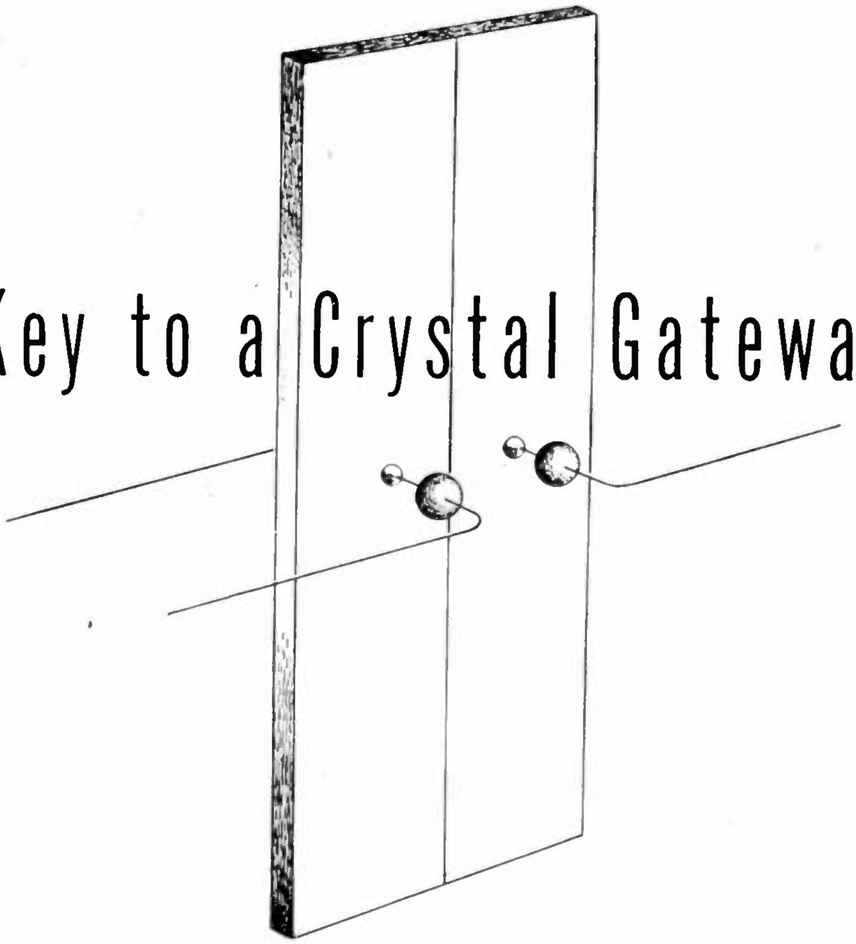
2017 S. Street N. W.
Decatur 8000

**-hp- laboratory instruments**
FOR SPEED AND ACCURACY



Above is a coaxial circuit crystal in its glass enclosure. At right the crystal is shown, 3½ times actual size, with connecting wires soldered in position. Weights on wires reflect energy back into crystal, so cut losses.

# Key to a Crystal Gateway



How would you solder a wire to a crystal? This must be done for most of those wafer-thin plates of quartz used in electrical circuits. They play a big part in the myriad-channel telephone system that utilizes coaxial cables.

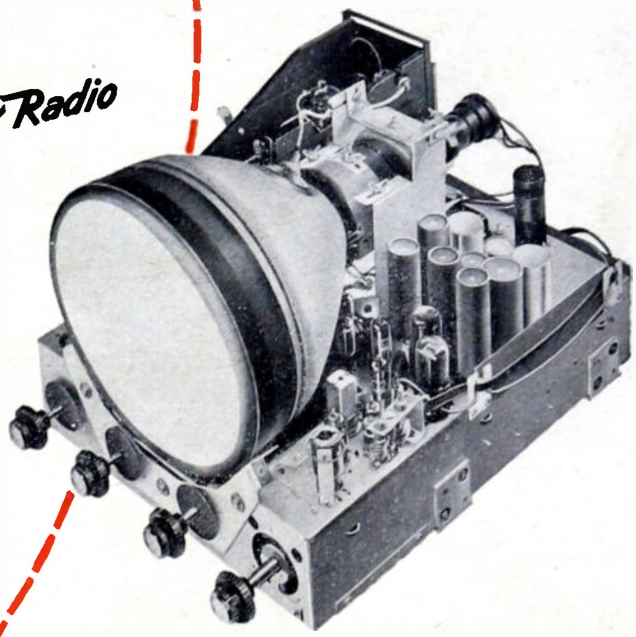
This is how Bell Laboratories scientists solved the problem: A spot of paste containing silver is deposited on the crystal and bonded to it by oven heat. The crystal is then vapor-plated with a thin layer of silver. Then a fine wire is soldered to the spot by a concentrated blast of hot air. The result is a rugged electrical connection to the surface of the crystal which does not interfere with its vibrations.

Sealed in glass tubes, the crystals are precise and reliable performers in the telephone system. Each is a crystal gate to a voiceway, separating *your* conversation from the hundreds of others which may be using a pair of coaxial conductors, at the same time.

This spot of paste, this tiny wire, this puff of air are among the tremendous trifles which concern Bell Telephone Laboratories in finding new ways to improve your telephone service.

**BELL TELEPHONE LABORATORIES**

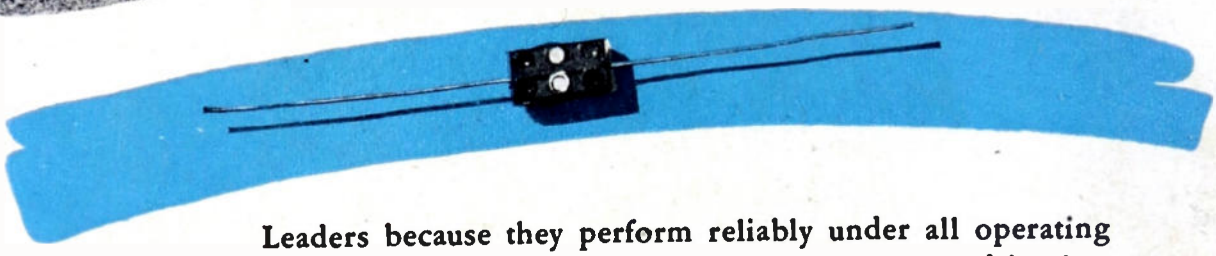
*EXPLORING AND INVENTING, DEVISING AND PERFECTING, FOR CONTINUED IMPROVEMENTS AND ECONOMIES IN TELEPHONE SERVICE*



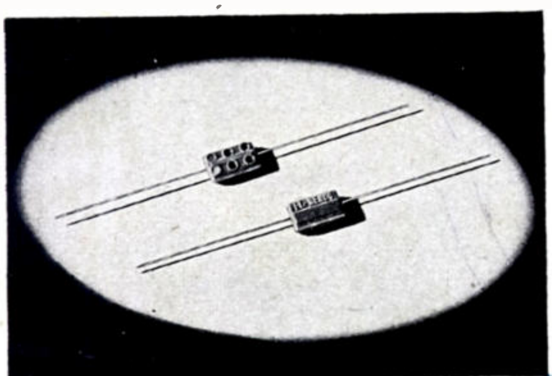
Chassis of the popular EMERSON Model 611 Television set. **HI-Q** components contribute their part to dependable operation.

EL-MENCO CAPACITORS

# Leaders



Leaders because they perform reliably under all operating conditions, these fixed mica dielectric capacitors are used in electronic applications wherever long life and successful performance are demanded.



**CM 15 MINIATURE CAPACITOR**

Actual Size $\frac{9}{32}$" x $\frac{1}{2}$" x $\frac{3}{16}$"
For Radio, Television and Other Electronic Applications

2 to 420 mmf. capacity at 500vDCw

2 to 524 mmf. capacity at 300vDCw

Temp. Co-efficient $\pm 50$ parts per million per degree C for most capacity values

6-dot standard color coded

Each tiny El-Menco Capacitor must pass life and humidity tests; meet standards set by the United States Army and Navy; pass tests at double their working voltages; prove their dielectric strength, temperature co-efficient and capacitance drift, and have their insulation resistance double-checked. These little leaders are molded in low-loss bakelite and wax-dipped for salt water immersion seal. They're available in a wide range, all impregnated, all precision-made, all JAN, RMA and RCM color-coded.

*Why not protect your product's performance with capacitors made under these rigid conditions?*

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Many years of reliable service in many types of application have established the Eimac 450T as the standout triode in its power class.

Recent technical achievements make the 450T a still better tube. Adoption of the Pyrovac plate and a non-emitting grid have amplified this already rugged tube's ability to "take it." Life expectancy and overload handling qualities have been increased multifold.

Comprehensive technical data on the Eimac 450T are immediately available . . . write direct.

EIMAC TYPE 450TH
ELECTRICAL CHARACTERISTICS

| | |
|---|---|
| Filament: Thoriated tungsten | |
| Voltage | 7.5 volts |
| Current | 12.0 amperes |
| Amplification Factor (Average) | 38 |
| Direct Interelectrode Capacitances (Average) | |
| Grid-plate | 5.0 μufd. |
| Grid-filament | 8.8 μufd. |
| Plate-filament | 0.8 μufd. |
| Transconductance ($I_b$ = 500 ma., $E_b$ = 4000v.) | 6650 μmhos |

MAXIMUM RATINGS
Radio Frequency Power Amplifier and Oscillator
Class-C Telegraphy (Key-down conditions, 1 tube)
Frequencies below 40 Mc.

| | |
|---|---|
| D-C Plate Voltage | 6000 Max. Volts |
| D-C Plate Current | 600 Max. Ma. |
| Plate Dissipation | 450 Max. Watts |
| Grid Dissipation | 65 Max. Watts |

# EITEL-McCULLOUGH, INC.

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21T

OHMITE THIN RESISTORS
SINGLE-UNIT BRACKET FIXED TYPE
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Distribute heat more evenly throughout resistor and conduct heat away quickly.
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## the possibilities of this combination of
# DU MONT CATHODE-RAY EQUIPMENT

**TYPE 263-B**
**HIGH-VOLTAGE POWER SUPPLY**

The Type 263-B High-Voltage Power Supply is designed to complement the Type 250-H Cathode-ray Oscillograph, a slightly modified version of the Type 250. This combination operates the Type 5RP-A Cathode-ray Tube in the Type 250-H instrument at accelerating potentials as high as 13,700 volts, permitting the photographic recording of writing rates as high as 40 inches per microsecond. The light output of the Type 250-H with the Type 263-B is 12 times greater than that of the Type 250 alone.

**TYPE 250**
**CATHODE-RAY OSCILLOGRAPH**

An excellent general-purpose cathode-ray oscillograph designed for observing and recording recurrent and transient phenomena. Contains both a-c and d-c amplifiers. Linear sweeps, driven and recurrent, from 1 cycle to 150 kc with automatic beam blanking on driven sweep. Deflection factor through a-c amplifier, 0.015 rms volt/in. max.; through d-c amplifier, 2 d-c volts/in. Built-in calibrator permits quantitative measurements. The high-current Type 5CP-A Cathode-ray Tube is operated at 3000 accelerating volts in this instrument to provide brilliance adequate for nearly all laboratory work.

The Types 5CP-A and 5RP-A Cathode-ray Tubes used in these instruments are manufactured by Du Mont, as are the cathode-ray tubes used in all Du Mont instruments.

**TYPE 2542**
**PROJECTION LENS**

With the addition of Type 2542 Projection Lens, the Type 250-H becomes a projection oscillograph capable of projecting its trace up to 30 feet for a picture size as large as 12 feet square. Advantages of such projections for lecture or demonstration work are readily apparent.

Lastly, the Du Mont Types 271-A and 314-A Oscillograph-record Cameras may be readily mounted on the Types 250 and 250-H for permanent recording.

**DETAILED LITERATURE ON REQUEST**
•
**EQUIPMENT DEMONSTRATED**
**No obligation**



# DUMONT *for Oscillography*

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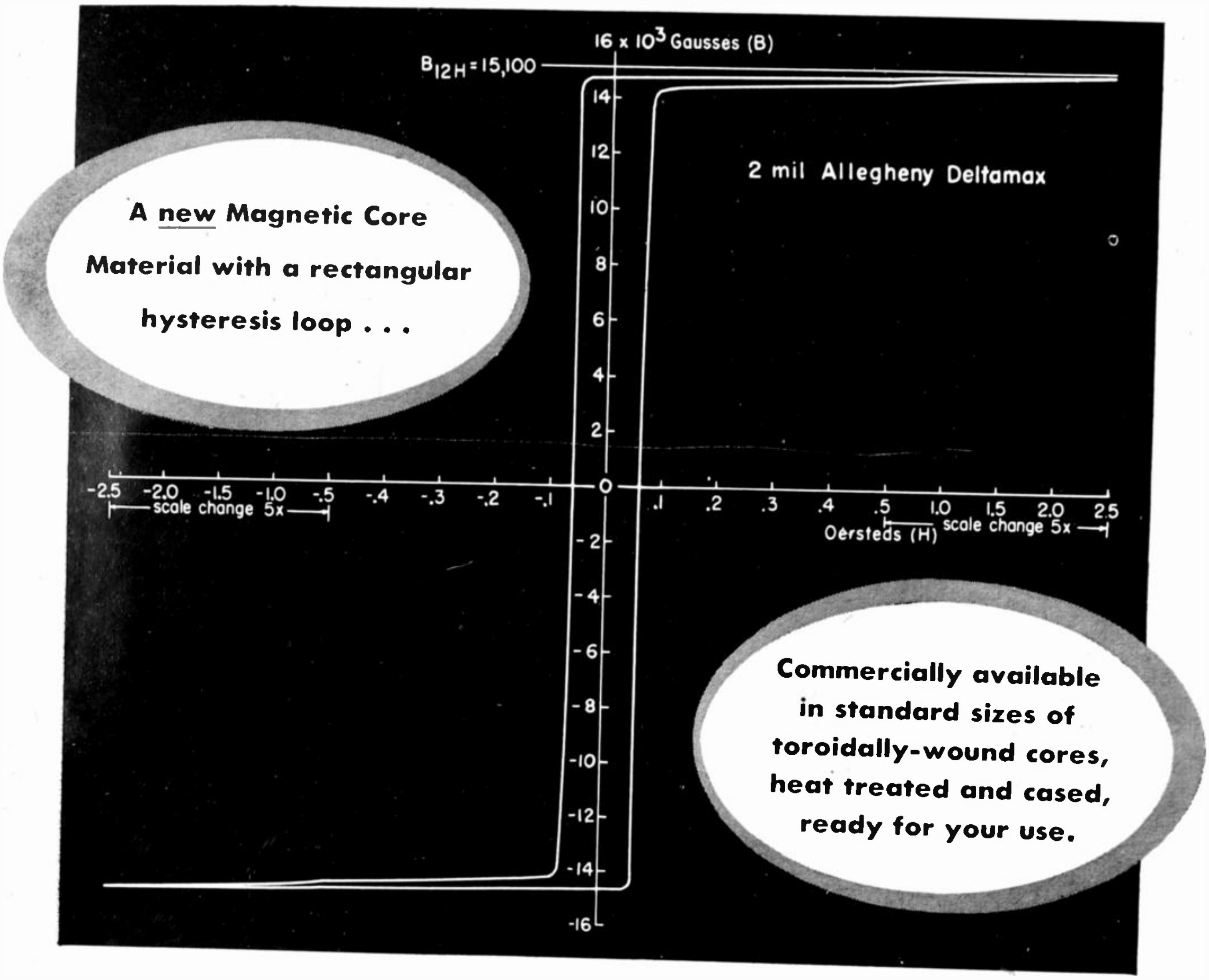
# DELTAMAX -now available!




## Where can YOU use a Magnetic Material with these specialized, dependable characteristics?

The properties of Deltamax are invaluable for many electronic applications, such as new and improved types of mechanical rectifiers, magnetic amplifiers, saturable reactors, peaking transformers, etc. This new magnetic material is available now as "packaged" units (cased cores ready for winding and final assembly) distributed by the Arnold organization. Every step in manufacture has been fully developed; designers can rely on complete consistency in each standard size of core.

Deltamax is the most recent extension of the family of special, high-quality electrical materials produced by Allegheny Ludlum, steel-makers to the electrical industry. It is an orientated 50% nickel-iron alloy, characterized by a rectangular hysteresis loop with sharply defined knees, combining high saturation with low coercivity.

● *Call on us for technical data.*

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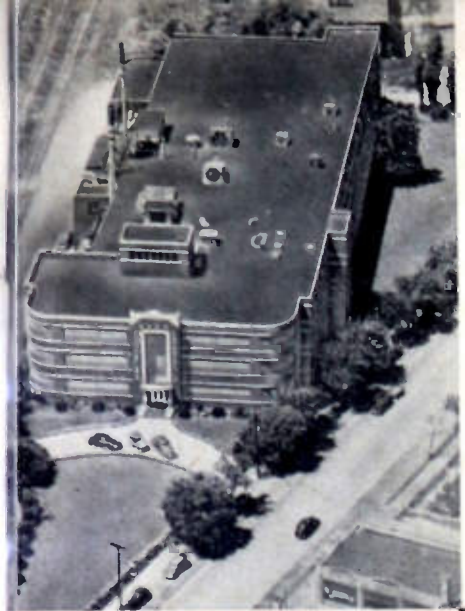
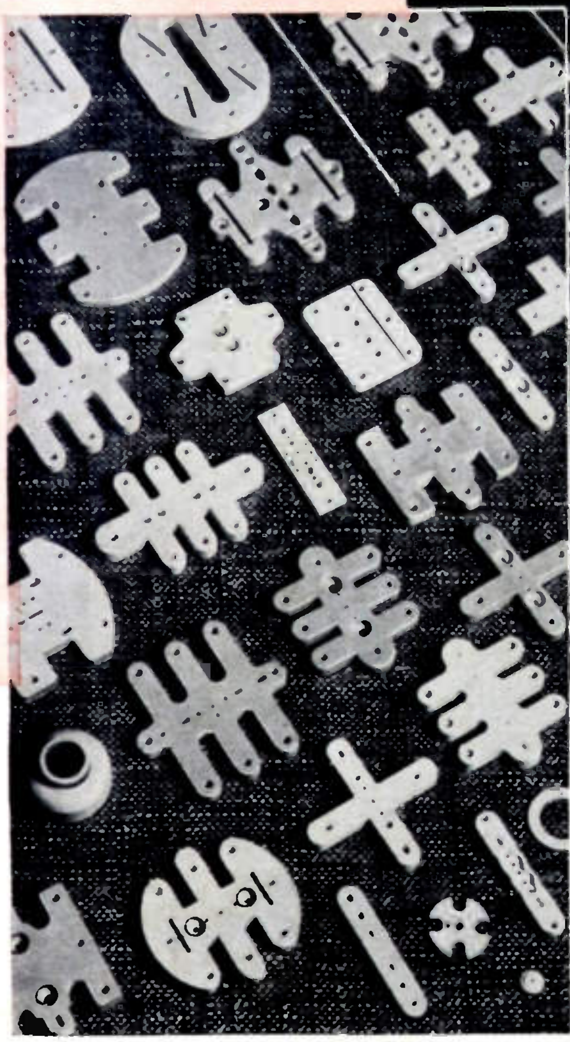
W&D 2379



# die pressed

***Die pressing enables AlSiMag to produce many shapes and sizes in large quantity . . . speedily, and at low cost***



One of several batteries of AlSiMag presses. These presses range from fully automatic high speed presses to large presses handling pieces up to 14" square.

For better, speedier service, AlSiMag makes its dies in its own modern die shop. This die shop serves only our customers, operates on a non-profit basis. Superior die pressing equipment and technique is another reason why American Lava Corporation is known as Headquarters for Custom Made Technical Ceramics.

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# NEWS and NEW PRODUCTS

JUNE, 1949

## Subminiature Filamentary Pentode

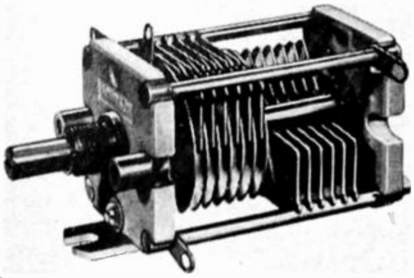
The **Raytheon Mfg. Co.,** 55 Chapel St., Newton 55, Mass., claims that, for the first time, a filamentary pentode is available with performance equivalent to that of many heater-cathode types. The type


1AD4 is a new subminiature sharp-cutoff pentode, shielded for rf applications, with a nominal mutual conductance rating of 2000 micromhos and an average plate current of 3 milliamperes, with 45 volts plate and screen supply. The filament rating is 1.25 volts, 100 ma.

This new type, which was originally developed for the Thermionics Branch of the Evans Signal Laboratory, is now available commercially from Raytheon.

## Non Fluctuating Variable Capacitor

A new line, 167, of variable capacitors, which, the manufacturer claims, are free of fluctuation in capacity, even under the most severe operating conditions, has been engineered by **E. F. Johnson Co.,** Waseca, Minn.



This new line of variables possesses new and perfected ceramic soldering. The ceramic band is stronger than the steatile end plates themselves. There are no eyelets, nuts or screws to work loose, causing stator wobble and capacity fluctuations. Stator terminals, mounting posts and rotor bearings are all ceramic soldered. Efficient operation is obtainable on frequencies as high as 500 megacycles.

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

## New Type Wow Meter

A flutter, wow, and drift percentage analyzer, model 491, type A—the first such instrument, the manufacturer claims, with a built-in high-gain amplifier and limiter—is in production by **Amplifier Corp. of America,** 398-1 Broadway, New York, N. Y.

Three distinct and simultaneous readings may be made of flutter, wow, and drift. Hum, noise, switching surges, and other extraneous transients have no effect on the reading or stability of the instrument.

Three scales: 0.3%, 1.0%, and 3.0%, are calibrated for flutter, wow, and drift readings. Variable capacitive and resistive elements enable accurate calibration and flutter measurements to be made on all recording and playback equipment whose normal linear rotational speed falls within ±5% of standard.

## Transformer for Reluctance Cartridges

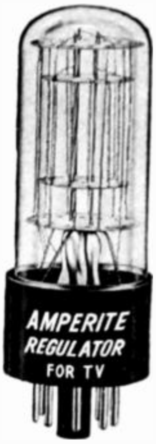
The Acrosound T1-100, a transformer to preamplify and equalize the output of reluctance-type pickup cartridges, is the first of a line of components to be put on the market by **Acro Products Co.,** 5328 Baltimore Ave., Philadelphia 43, Pa.

The unit is used to replace the usual tube preamplifier and equalizer, and, the manufacturer claims, does so with the elimination of hum, distortion and instability usually associated with preamplifiers.

New design principles that have been applied in the design of the T1-100 provide full undistorted high-frequency response, plus a rising bass characteristic for low-frequency equalization. The output of the transformer provides sufficient voltage to energize the tuner, phonograph, or other medium-level high-impedance input channels.

## Television Ballast Tube

A new television ballast tube which is hermetically sealed and filled with helium, producing rapid cooling, has been placed on the market by **Amperite Co., Inc.,** 561 Broadway, New York 12, N. Y.




The tube is produced with as many as five separate controlling elements. Furthermore, to withstand the possible overload in television receivers, some of the 2.5-watt elements are designed to withstand 40 watts, an overload of 2,000%. Voltage breakdown between elements is 13,000 volts dc.

Diameter of the tube is 1¼", seated height is 3"

## Recent Catalogs

• • • A new bulletin, F-2, entitled "How to Select Flexible Waveguide Assemblies," shows three distinct types of flexible waveguide constructions and their uses, applications, and characteristics. Of special interest is a discussion of rigid-flexible combination assemblies showing the attachment of flexible to rigid waveguides with the use of flanges, by **Technicraft Laboratories, Inc.,** Thomaston & Waterbury Rds., Thomaston, Conn.

• • • A new, detailed information catalog concerning the greatly expanded line of capacitors developed to serve the entire field can be secured from the **Illinois Condenser Co.,** 1616 N. Throop St., Chicago 22, Ill.

• • • A bulletin describing the type 110 Slide-Wire Resistance Boxes, designed as a laboratory substitute for the more expensive decade resistance boxes, is available to those interested from **Technology Instrument Corp.,** 1058 Main St., Waltham 54, Mass.

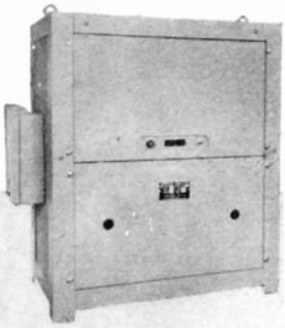
• • • A booklet stating reasons for galvanic cell corrosion, as well as methods for helping to overcome it (complete in all engineering details including drawings and other essential illustrative material), may be obtained from **International Nickel Co., Inc.,** New York 5, N. Y.

*(Continued on page 26A)*

## General Application

| Model | Load Range Volt-Amperes | *Regulation Accuracy |
|---|---|---|
| 150 | 25-150 | 0.5% |
| 250 | 25-250 | 0.2% |
| 500 | 50-500 | 0.5% |
| 1000 | 100-1000 | 0.2% |
| 2000 | 200-2000 | 0.2% |

*Models available with increased regulation accuracy.



## Extra Heavy Loads

| Model | Load Range Volt-Amperes | *Regulation Accuracy |
|---|---|---|
| 3,000 | 300-3000 | 0.2% |
| 5,000 | 500-5000 | 0.5% |
| 10,000 | 1000-10,000 | 0.5% |
| 15,000 | 1500-15,000 | 0.5% |

*Models available with increased regulation accuracy.

# the first line of STANDARD electronic AC voltage regulators and nobatrons

### GENERAL SPECIFICATIONS

- Harmonic distortion : max. 5% basic or 2% "S" models
- Input voltage range: either 95-125 or 190-250 volts
- Output: adjustable between either 110-120 or 220-240 volts
- Input frequency range: 50-60 cycles
- Power factor range: down to 0.7 P. F.

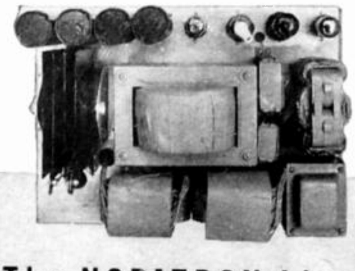
**All AC Regulators and Nobatrons may be used at no load.**

Special Models designed to meet your unusual applications.

**Write for the new Sorensen catalogue. It contains complete specifications on standard Voltage Regulators and Nobatrons.**

Special Transformers, D. C. Power Supplies, Saturable Core Reactors and Meter Calibrators made to order; please request information.

SORENSEN & Company, Inc.
Stamford, Connecticut
*Represented in all principal cities.*



## The NOBATRON Line

| Output Voltage DC | Load Range Amps. |
|---|---|
| 6 | 5-15-40-100 |
| 12 | 5-15-50 |
| 28 | 10-30 |
| 48 | 15 |
| 125 | 5-10 |

Regulation Accuracy—.25% from ¼ to full load.

## 400 Cycle Line

Inverter and Generator Regulators for Aircraft

Single Phase and Three Phase

| Model | Load Range Volt-Amps. | Reg. Accuracy |
|---|---|---|
| D 100 | 10-100 | 0.5% |
| D 500 | 50-500 | 0.5% |
| D 1200 | 120-1200 | 0.5% |
| D 2000 | 200-2000 | 0.5% |

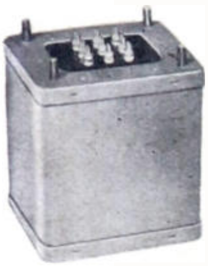
## 3-Phase Regulation

Star-connected three-phase systems can be handled effectively. Other three-phase systems must be reviewed by our Engineering Dept.

VA Capacities up to 45 KVA.

# FREED "PRODUCTS of EXTENSIVE RESEARCH" INSTRUMENTS & COMPONENTS!

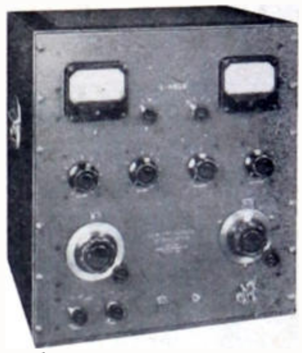
## HIGH FIDELITY OUTPUT TRANSFORMERS



| Type No. | Primary matches following typical tubes | Primary Impedance | Secondary Impedance | ± ½db from | Maximum level |
|---|---|---|---|---|---|
| F1950 | Push pull 2A3's, 6A5G's, 300A's, 275A's, 6A3's, 6L6's | 5000 ohms | 500, 333, 250, 200, 125, 50 | 20-30000 cycles | 15 watts |
| F1951 | Push pull 2A3's, 6A5G's, 300A's, 275A's, 6A3's, 6L6's | 5000 ohms | 30, 20, 15, 10, 7.5, 5, 2.5, 1.2 | 20-30000 cycles | 15 watts |
| F1954 | Push pull 245, 250, 6V6, 42 or 2A5 A prime | 8000 ohms | 500, 333, 250, 200, 125, 50 | 20-30000 cycles | 15 watts |
| F1956 | Push pull 245, 250, 6V6, 42 or 2A5 A prime | 8000 ohms | 30, 20, 15, 10, 7.5, 5, 2.5, 1.2 | 20-30000 cycles | 15 watts |
| F1958 | Push pull 6B5, 6A6, 53, 6F6, 59, 79, 89, 6V6, Class B 46, 59 | 10,000 ohms | 500, 333, 250, 200, 125, 50 | 20-30000 cycles | 15 watts |
| F1959 | Push pull 6B5, 6A6, 53, 6F6, 59, 79, 89, 6V6, Class B 46, 59 | 10,000 ohms | 30, 20, 15, 10, 7.5, 5, 2.5, 1.2 | 20-30000 cycles | 15 watts |
| F1962 | Push pull parallel 2A3's, 6A5G's, 300A's, 6A3's, 6L6 | 2500 ohms | 500, 333, 250, 200, 125, 50 | 20-30000 cycles | 30 watts |
| F1963 | Push pull parallel 2A3's, 6A5G's, 300A's, 6A3's, 6L6 | 2500 ohms | 30, 20, 15, 10, 7.5, 5, 2.5, 1.2 | 20-30000 cycles | 30 watts |
| F1966 | Push pull 6L6 or Push pull parallel 6L6 | 3800 ohms | 500, 333, 250, 200, 125, 50 | 20-30000 cycles | 50 watts |
| F1967 | Push pull 6L6 or Push pull parallel 6L6 | 3800 ohms | 30, 20, 15, 10, 7.5, 5, 2.5, 1.2 | 20-30000 cycles | 50 watts |

## "Q" INDICATOR

No. 1030 Frequency range from 20 cycles to 50 kilocycles. "Q" range from .5 to 500. "Q" of inductors can be measured with up to 50 volts across the coil. Indispensable instrument for measurement of "Q" and inductance of coils, "Q" and capacitance of capacitors, dialectric losses, and power factor of insulating materials.



## INCREMENTAL INDUCTANCE BRIDGE

**IMPEDANCE RANGE:** One millihenry to 1000 henries in five ranges. Inductance values are read directly from a four dial decade and multiplier switch. This range can be extended to 10,000 henries by the use of an external resistance.

**INDUCTANCE ACCURACY:** Within plus or minus 1% through the frequency range from 60 to 1000 cycles.

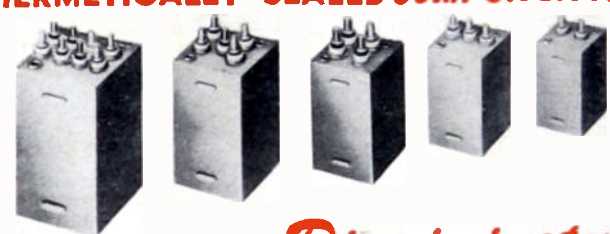
## COMPARISON BRIDGE

No. 1010 An invaluable instrument for precision laboratory adjustment and incoming inspection of resistors, capacitators and inductors . . . Entirely self-contained, A.C. operated and includes a three frequency oscillator, an A.C. bridge and a null detector.

## NULL DETECTOR

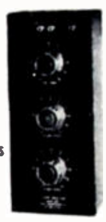
No. 1140 For bridge measurements, providing visual null indications or aural indications when used in conjunction with headphones. The unit may also be used as a high gain amplifier for general laboratory work. Functionally, the instrument consists of a high gain linear amplifier with a 30 db input attenuator in addition to the variable gain control. Output voltage is 40 volts undistorted into 1 megohm load, and 10 volts into 20,000 ohms.

## HERMETICALLY SEALED *COMPONENTS*



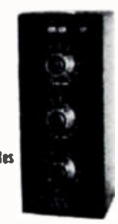
## *Decade Inductors*

No. 1160
10 x 1 HY steps
10 x .1 HY steps
10 x .01 HY steps
500-15,000 cycles

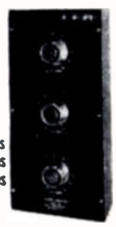


No. 1161
10 x .1 HY steps
10 x .01 HY steps
10 x .001 HY steps
2000-50,000 cycles

No. 1162
10 x .01 HY steps
10 x .001 HY steps
10 x .0001 HY steps
10,000-300,000 cycles



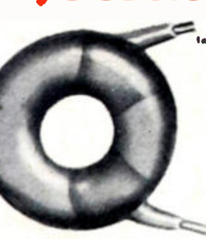
No. 1164
10 x 10 HY steps
10 x 1 HY steps
10 x .1 HY steps
50-1000 cycles



## *Discriminators*

For telemetering and remote control applications using audio and supersonic frequency subcarriers.

## *Toroidal Inductors*



| Type T1 1000-15,000 cycles Inductance Value | Type # | Type T1-2 2000-30,000 cycles Inductance Value | Type # | Type T1-3 10,000-300,000 cycles Inductance Value | Type # |
|---|---|---|---|---|---|
| 5 MHY | F-800T | 1 MH | F-1800 | .5 MH | F-1850 |
| 10 MHY | F-801T | 2 MH | F-1801 | 1 MH | F-1851 |
| 15 MHY | F-802T | 3 MH | F-1802 | 2 MH | F-1852 |
| 30 MHY | F-803T | 4 MH | F-1803 | 3 MH | F-1853 |
| 50 MHY | F-804T | 5 MH | F-1804 | 4 MH | F-1854 |
| 75 MHY | F-805T | 10 MH | F-1805 | 5 MH | F-1855 |
| 100 MHY | F-806T | 15 MH | F-1806 | 10 MH | F-1856 |
| 200 MHY | F-808T | 30 MH | F-1807 | 15 MH | F-1857 |
| 500 MHY | F-809T | 50 MH | F-1808 | 20 MH | F-1858 |
| 750 MHY | F-810T | 75 MH | F-1809 | 30 MH | F-1859 |
| 1 HY | F-811T | 100 MH | F-1810 | 40 MH | F-1860 |
| 1.5 HY | F-813T | 150 MH | F-1811 | 50 MH | F-1861 |
| 2 HY | F-815T | 200 MH | F-1812 | 75 MH | F-1862 |
| 3 HY | F-819T | 300 MH | F-1813 | 100 MH | F-1863 |
| 4 HY | F-821T | 400 MH | F-1814 | | |
| 5 HY | F-823T | 500 MH | F-1815 | | |

High quality toroidal coils wound on molybdenum permalloy dust cores. All those listed above can be supplied in hermetically sealed cans, commercial type construction or open units. Other types can be supplied out of stock on special orders.

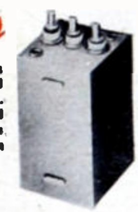
## LOW FREQUENCY HI "Q" COILS

| | |
|---|---|
| #1900 | 100 HY |
| #1901 | 75 HY |
| #1902 | 50 HY |
| #1903 | 25 HY |
| #1904 | 10 HY |
| #1905 | 5 HY |
| #1906 | 1 HY |

Available from stock in the indicated inductance values.

## *Filters*

Narrow band pass filters for remote control and telemetering applications. High pass, low pass, band pass and band elimination filters for communication and carrier systems.



**SEND FOR COMPLETE DESCRIPTIVE LITERATURE**

# FREED TRANSFORMER CO., Inc.

DEPT. PJ 1718-36 WEIRFIELD ST., (RIDGEWOOD) BROOKLYN 27, NEW YORK

# *Everything* IN CARBON *but Diamonds!*

## PLUS GRAPHITE, MOLDED METALS SINTERED ALNICO II

**HUNDREDS of STANDARD ITEMS . . . *thousands of "Specials"***

Write for details on any type

Anodes • Battery Carbons
Bearing Materials
Brazing Furnace Boats
Brushes of all types for rotating electrical equipment
Carbon and Graphite Contacts
Chemical Carbons • Clutch Rings
Dash Pot Plungers
Electric Furnace Heating Elements
Electrolytic Anodes • Friction Segments
Glass Molds • Ground Rods (carbon)
Mercury Arc Rectifier Anodes
Metal-Graphite Contacts
Power Tube Anodes
Rail Bonding Molds
Rare Metal Contacts
Resistance Welding and Brazing Tips
Seal Rings (for gas or liquid)
Special Molds and Dies
Spectrographite No. 1
Trolley and Pantograph Shoes
Voltage Regulator Discs
Water Heater and Pasteurization Electrodes
Welding Carbons
Welding Plates and Paste

## *. . a dependable source of supply*

The unique electrical, mechanical, physical and chemical properties of Stackpole carbon, graphite and carbon-graphite products solve countless problems of friction, temperature, arcing, corrosion, shaft sealing, voltage regulating and others. So broad is the line of standard Stackpole products, so extensive the facilities for "specials" that it is practical to list only a few of them here. Let Stackpole engineers recommend and quote on your next requirements.

**STACKPOLE**

STACKPOLE CARBON CO., ST MARYS, PA.

HARDEST WORKERS ON YOUR COMMUNICATIONS TEAM

# G-E BEAM POWER TRANSMITTING TUBES

GL-2E26

GL-807

GL-829-B

GL-813

**—always ready for dependable service—minutes or hours of it!**

**—need low drive, so ask less of your power supply.**

**—replacements are convenient to obtain...you can secure new tubes fast from your local G-E tube distributor!**

DESIGNERS of equipment give first place to General Electric beam power *economy* tubes. Their low drive requirements — a characteristic of this type — pay off in less space needed for the driving stages of a transmitter. That's Saving No. 1! And drain on the battery or other source of power supply is reduced . . . Saving No. 2.

If you operate police, taxicab, or ambulance radio equipment — if you maintain an airport, ship-to-shore, or other communications system — the benefits of a more compact transmitter and lower power consumption are matched by G-E tube dependability. These beam power types are amply proved in tough service!

A complete line of General Electric tubes is available, spanning the range of outputs and frequencies in communications work. Designers and builders of equipment, through their nearby G-E electronics office, may call on experienced G-E tube engineers to help select the right types for new circuits.

Transmitter owners will find that same-day, often same-hour service is given by their local G-E distributor on tube replacements. From coast to coast, stocks are in readiness for *your* emergency call! Get to know your G-E tube distributor; he's equipped to serve you fast . . . and well! *General Electric Company, Electronics Department, Schenectady 5, New York.*

Ratings (ICAS) for typical operation, Class C plate-modulated

| Type | Plate voltage | Plate current | Driving power (approx) | Power output (approx) | Freq. at max ratings |
|---|---|---|---|---|---|
| GL-2E26 | 500 v | 54 ma | 0.15 w | 18 w | 125 mc |
| GL-807 | 600 v | 100 ma | 0.4 w | 42.5 w | 60 mc |
| GL-829-B | 600 v | 150 ma | 0.9 w | 70 w | 200 mc |
| GL-813 | 2,000 v | 200 ma | 4.3 w | 300 w | 30 mc |

*Photo by courtesy N. Y. State Police*

**GENERAL ELECTRIC**

180-H4

FIRST AND GREATEST NAME IN ELECTRONICS

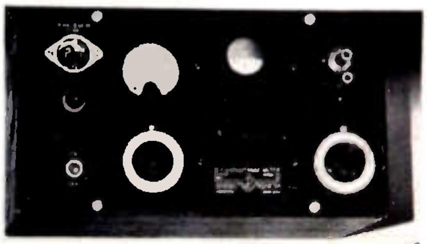
# NEWS—NEW PRODUCTS

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

*(Continued from page 18A)*

## RF Capacitance Meter

A new type 1612-A rf capacitance meter designed to rapidly measure and test small capacitors (up to 1,200 $\mu\mu$f.) is now in production at **General Radio Co.**, 275 Massachusetts Ave., Cambridge 39, Mass.



Measurement is made by a substitution method in which the capacitance of a calibrated air capacitor is reduced to reestablish resonance after an unknown capacitor is connected. Resonance is indicated by maximum deflection of a meter. Two ranges are provided, 0 to 80 $\mu\mu$f and 0 to 1,200 $\mu\mu$f, and range switching is accomplished automatically as the dial is rotated. Measurements are made at a frequency of 1 Mc.

Maximum meter deflection can also be used as an indication of loss, and hence as a basis for rejection of high-loss capacitors when compared to a standard sample.

The meter operates from a 115-volt source, ac or dc. Dimensions are: 12″×6⅝″×7½″, weight 11 lbs.

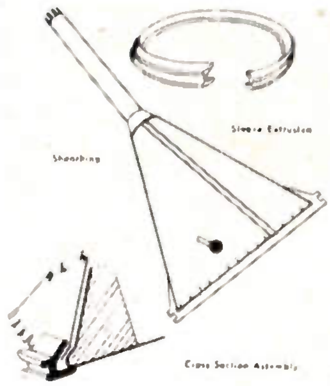
## Three Speed, all Size Record Changer

As a solution to the recent addition of two new-type records to be played at different speeds, **General Instrument Corp.**, 829 Newark Ave., Elizabeth, N. J., offers a changer designed to automatically play 10 or 12″ 78 rpm, 7, 10, and 12″ 33⅓ rpm, and the new 45 rpm records with the same pickup arm.

The new changer has a conventional spindle diameter, and the problem of the large hole diameter of the 45-rpm record is overcome by the use of plastic adaptor buttons. These buttons are inserted in the center of the 45 rpm records, reducing the hole diameter to the size of the standard spindle.

## Plastic Insulation for Metal TV Tubes

A new mounting and insulating sleeve constructed of a polyethylene compound, "Aeroflex," which also guards the tube against shock and vibration in shipping is being marketed by **Anchor Plastics Co.**, 533 Canal St., New York 13, N. Y.




This sleeve fits over the tube cone and is held in place by an extruded mounting ring; thus all metal portions of the tube are covered, eliminating accidental contact.

"Aeroflex" has been adapted to this new use through the development of a process in the extrusion of thermoplastics. The compound is extruded in circular shapes, eliminating the difficulty in bending.

## Recent Catalogs

• • • A new bimonthly exterior house organ, devoted to the latest developments in the connector and electrical specialty fields, will be sent on request to those writing to the Editor, *Cannonade*, **Cannon Electric Dev. Co.**, 3209 Humboldt St., Los Angeles 31, Calif.

• • • A data folder on capacitors with necessary information concerning design, capacitance, containers, chassis layouts, and mountings and sockets by **P. R. Mallory & Co., Inc.**, 3029 E. Washington St., Indianapolis 6, Ind., is available on request.

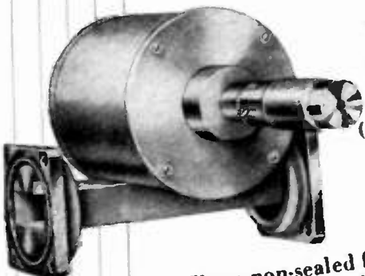
• • • A 28 page illustrated catalog, CDM-2A, describing stocked permanent magnets has been issued by **General Electric Co., Chemical Dept.**, Pittsfield, Mass. A wide variety of cast and sintered alinco magnets, as well as special magnetic alloys, are listed.

• • • The March issue of *CEC Recordings* has an interesting article concerning a new sampling probe that raises leak detector sensitivity 1,000 times. This may be secured by writing to **Consolidated Engineering Corp.**, 620 N. Lake Ave., Pasadena 4, Calif.

*(Continued on page 45A)*

PRD

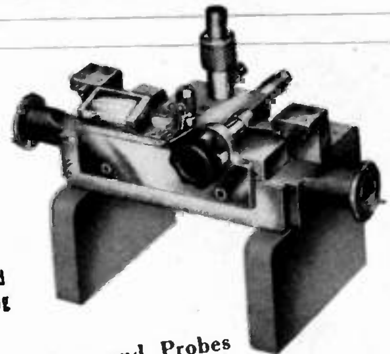
# MICROWAVE MEASUREMENT COMPONENTS Now Available..



TYPE 551-A — REACTION TYPE FREQUENCY METER
(1" x ½" waveguide)

High Q cavity; Precise and permanent calibration; Extraneous mode suppression

• These non-sealed frequency meters will soon be augmented by a new line of hermetically sealed, temperature compensated units covering the frequency range from 500 to 40,000 megacycles per second. Also available: frequency standarized signal sources.

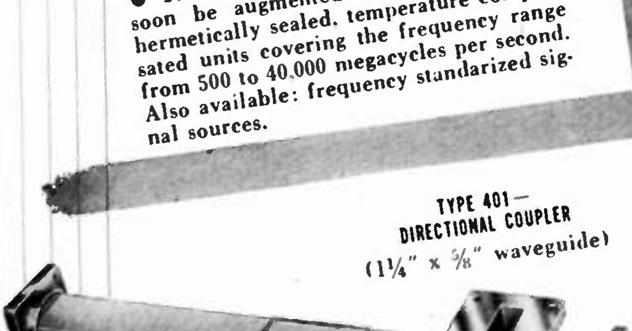


TYPE 211 — PRECISION WAVEGUIDE SLOTTED SECTION
(0.420" x 0.170" I.D.)

Broadband operation; Crystal and bolometer detection; Ball bearing carriage support

• Similar Slotted Sections and Probes in standard rectangular waveguide and coaxial line sizes make possible precise impedance measurements over the microwave spectrum from 1000 to 40,000 megacycles per second.

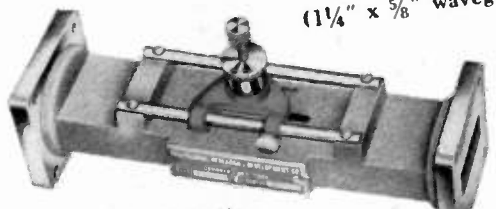
TYPE 401 — DIRECTIONAL COUPLER
(1¼" x ⅝" waveguide)



High directivity; Minimum frequency sensitivity; Broadband operation

• This unit is representative of a group of mono-directional broadband couplers covering in four waveguide sizes the frequency range from 4000 to 10,000 megacycles per second.

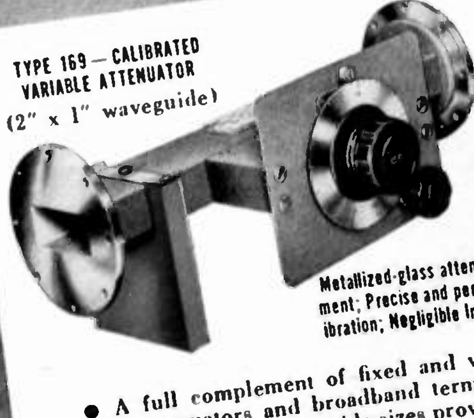
TYPE 302 — SLIDE SCREW TUNER
(1¼" x ⅝" waveguide)



Wide range impedance matching; Simplified rapid adjustment; Broadband operation

• Also available: similar units in standard waveguide sizes, fixed and tunable crystal and bolometer mounts, dielectric tuning devices for coaxial lines.

TYPE 169 — CALIBRATED VARIABLE ATTENUATOR
(2" x 1" waveguide)



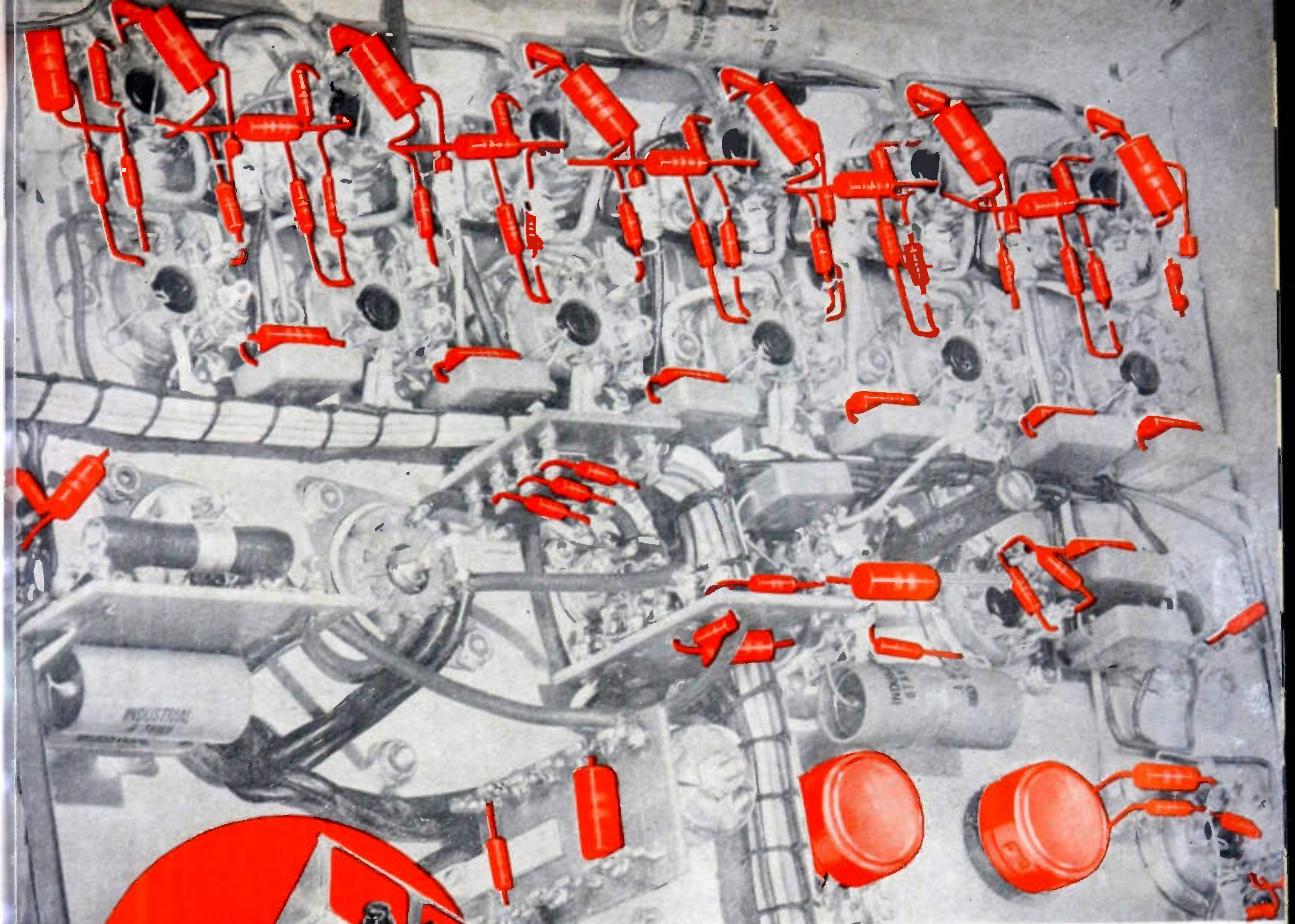
Metallized-glass attenuating element; Precise and permanent calibration; Negligible Insertion loss

• A full complement of fixed and variable attenuators and broadband terminations in standard waveguide sizes provides coverage for the frequency range from 2600 to 40,000 megacycles per second. Fixed pads and terminations are available for standard coaxial transmission lines.

**The items presented above are representative of the complete PRD line of precision microwave measurement and test equipment. These units embody basically new design principles calculated to provide the microwave research engineer with the ultimate in accuracy and reliability. A skilled staff of engineers and physicists is constantly pioneering the advance to the higher frequency regions of the microwave spectrum and stands ready to assist in the solution of your microwave problems. An illustrated catalog may be obtained by writing on company letterhead to Dept. R-3.**

*Polytechnic* **RESEARCH & DEVELOPMENT COMPANY, Inc.**

202 TILLARY ST., BROOKLYN 1, N. Y.




# 105 ALLEN-BRADLEY FIXED AND ADJUSTABLE RESISTORS

## in this automatic scaling unit

"Our Model 163 Automatic Scaling Unit," says John L. Kuranz, of Nuclear Instrument & Chemical Corporation, Chicago, "contains 103 Allen-Bradley fixed resistors and two Bradleyometers. To achieve the dependability so necessary in this type of research tool, our tests show that only A-B resistors can be used. In addition to their functional superiority, their small size makes our portable instruments really portable."

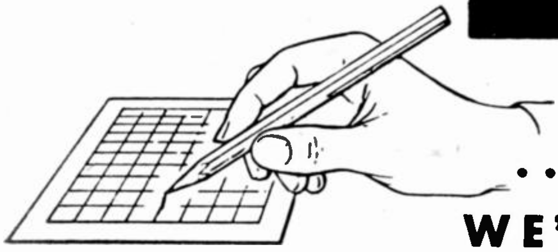
Where superlative performance is a *must* . . . where unfailing dependability is a basic requirement . . . as in nuclear research . . . you will find Allen-Bradley quality resistors. If you *must* meet high-quality standards . . . specify Allen-Bradley resistors.

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

**ALLEN-BRADLEY RADIO RESISTORS**

QUALITY

## FONO-FITTED TORQUE E-V DRIVE CARTRIDGES



# ... you draw the curve
# WE'LL BUILD THE CARTRIDGE

*Your specifications . . . your special requirements in phono pickup cartridges . . . are ideally "custom-solved" through E-V creative engineering . . . unusual manufacturing facilities . . . and inherent advantages of exclusive TORQUE DRIVE.**

**CUSTOM RESPONSE**

Smooth upper response with roll off frequency to your specifications or wide range, peak-free response to 10 kc. You draw the curve, we'll build the cartridge.

**VOLTAGE**

E-V TORQUE DRIVE cartridges provide the highest compliance per volt output. For example, the E-V 14 cartridge tracks at 5 grams with excellent wave form down through 50 c.p.s. on the RCA 12-5-31V record at 1 volt at 1,000 c.p.s.

**TRACKING FORCE**

With the high compliance and low mass of the driving system, needle forces at 5 grams for both one and three mil records are used in everyday production by leading manufacturers. Cartridges with even lower needle force with slight reduction in voltage are thoroughly practical. 3 gram tracking pressures are definitely in sight.

**COMBINATION**
***One and Three Mil***

E-V TORQUE DRIVE again leads in twin needle cartridge design. Tracking force of 5 grams on both one and three mil records precludes weight changing. Straight line needle position assures accurate set down when used with changers. Approximately the same output is obtained on both stylii. The E-V Twin-Tilt cartridge mounts in any arm with ½" mounting holes with no modification except adjustment for correct needle force.

**MOISTURE PROOFING**

The cartridge is *entirely filled* with DC4 Silicone jelly—the material that is used for inhibiting moisture on aircraft wiring. Tests indicate that it increases the life of an ordinary crystal *some 20 times*. This is a *plus* feature, found in all E-V crystal cartridges.

*Our engineering staff and full facilities are at your service. Contact us today.*

**Modern One-Acre Plant with Complete Internal Facilities for Quality-Controlled Volume Production**

From original conception to final product, you get full benefit of the unusual E-V processes. Here are E-V laboratories, where constant research keeps making new contributions to the Art. Here we make tools and dies . . . die cast, plate, screw machine, stamp, mold plastics, and assemble. Here we use specially designed test equipment for quality control. With all these facilities, we produce high standard acoustical products in quantity, with utmost economy. Come—see this plant in action.

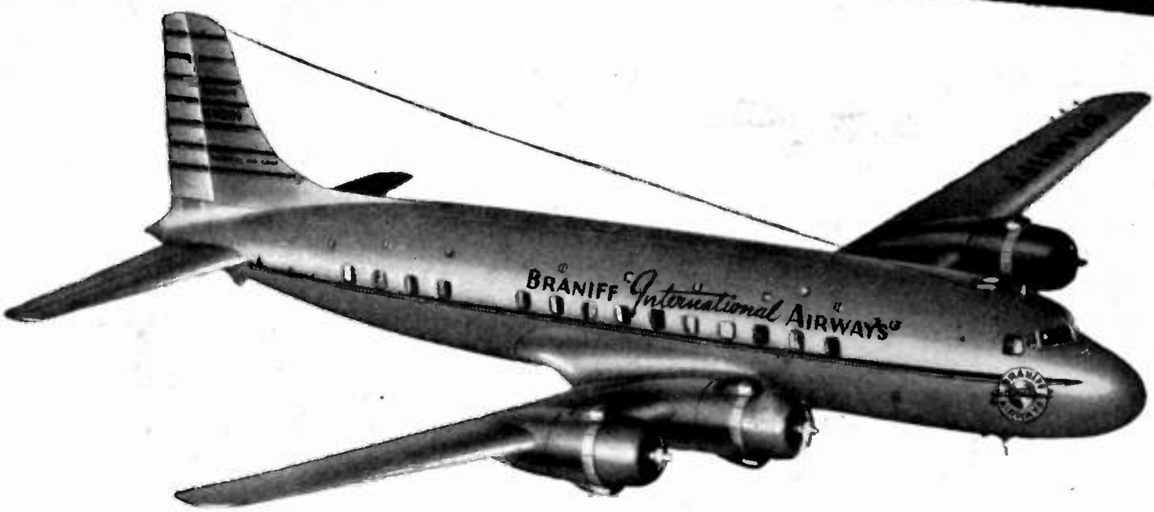
**NO FINER CHOICE THAN**

**Electro-Voice**

*E-V Pat. Pending
Licensed under Brush patents

*Export: 13 East 40th St., New York 16, U.S.A.
Cables: Arlab*

**ELECTRO-VOICE, INC. • BUCHANAN, MICHIGAN**




# WILCOX

## FIRST CHOICE OF

## BRANIFF *International* AIRWAYS

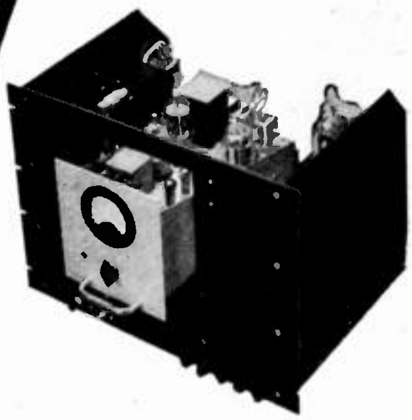
### BRANIFF EQUIPS GROUND STATIONS WITH WILCOX TYPE 364A TRANSMITTER

**DESIGN SIMPLIFIES SERVICE**
Conventional circuit design, fewer numbers and types of tubes, plus open mechanical construction simplify tube stocking problems and speed maintenance. The entire transmitter portion of the Type 364A is built on a drawer-type chassis, instantly withdrawable from the front of the panel.

**RELAY RACK MOUNTING SAVES SPACE**
Compact design requires only 15 inches of rack space for installation, frequently utilizing space already available.

**.005% FREQUENCY STABILITY WITHOUT TEMPERATURE CONTROL**
Through the use of a newly developed crystal, troublesome thermostatic temperature controls and crystal ovens are no longer necessary to provide adequate frequency stability.

**SIMPLIFIED CONTROL FOR REMOTE LOCATION**
Modulation over a single telephone pair and carrier control by means of a simplex circuit allow the transmitter to be readily located at a remote point.



WILCOX
Type 364A Transmitter
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# PROCEEDINGS OF THE I.R.E.

## (Including the WAVES AND ELECTRONS Section)

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## PROCEEDINGS OF THE I.R.E.



## INSTITUTE NEWS AND RADIO NOTES SECTION



## WAVES AND ELECTRONS SECTION







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† Deceased.

# Elmer W. Engstrom

DIRECTOR, 1949

Elmer W. Engstrom, vice-president in charge of research at the RCA Laboratories in Princeton, N. J., was born in Minneapolis, Minn., on August 25, 1901. After receiving the B.S. degree in electrical engineering from the University of Minnesota in 1923, he entered the employment of the General Electric Co., where he worked on transmitters, receivers, and sound motion pictures. He became an Associate Member of the IRE in 1925.

In 1930 Mr. Engstrom joined RCA, where he was successively employed in research positions of increasing responsibility, culminating in his being appointed to a vice-presidency in 1945. Meanwhile, he transferred to Member grade in the IRE in 1938, and was awarded the Fellow grade in 1940.

Mr. Engstrom has been a member of a number of Institute Committees, including the Board of Editors, Awards, and Nominations. A past president of the Princeton Chapter of Sigma Xi and current president of the Industrial Research Institute, he is also chairman of the Visiting Committee of the Navy Industrial Association to the Naval Research Laboratory.

# CLINTON B. DeSOTO

## 1912–1949

On April 1, 1946, Clinton B. DeSoto (M'46–SM'49) assumed the duties of technical editor of the PROCEEDINGS OF THE I.R.E. Born in Ogilvie, Minnesota, in 1912, he attended the University of Wisconsin's School of Journalism.

Mr. DeSoto became a licensed radio amateur in 1926, and in 1930 he joined the American Radio Relay League headquarters staff as assistant to the secretary of the League, transferring to the editorial staff as assistant editor of *QST* in 1942. He became executive editor of *QST* in 1943 and editor in 1944.

The author of a number of books and magazine articles dealing with radio topics, Mr. DeSoto handled the revision and production of the 1943, 1944, and 1945 editions of "The Radio Amateur's Handbook." He was associated with the development of radio remote-control systems for military and amateur applications. From 1933 to 1936 he served as secretary of the Connecticut Valley Section of The Institute of Radio Engineers. He also did the publicity work for the 1948 and 1949 IRE National Conventions.

After his acceptance of the position of Technical Editor, Mr. DeSoto capably fulfilled the duties of his post, and added substantially to the clarity, arrangement, and promptness of issuance of the PROCEEDINGS OF THE I.R.E.

The following expressions concerning him were signed by his associates:

The death of Clinton B. DeSoto, a Senior Member of The Institute of Radio Engineers and Technical Editor of the PROCEEDINGS OF THE I.R.E., is a shock to me personally and to the entire Institute. During his three years as Technical Editor, his keen insight into the publication business made him a tower of strength on the headquarters' editorial staff; and his handling of publicity for our annual conventions was truly inspired. His passing leaves a vacancy in our staff which will be extremely difficult to fill, and a gap in our hearts that can never be filled.

STUART L. BAILEY
President

I have known Clint twenty years. He brought to the IRE a wonderful combination of professional editorial ability, knowledge of the printing arts, a keen sense of publicity values, and personal charm.

GEORGE W. BAILEY
Executive Secretary

Clinton DeSoto, scholar, gentleman, and capable administrator, leaves behind him a host of grieving friends. He combined literary skill, unusual ability to bring out the best in his associates, and unfailing willingness successfully to handle any task, no matter how troublesome or time-consuming. The IRE, whose Technical Editor he has been, has lost one of its finest representatives, and deeply regrets his untimely passing.

A. N. GOLDSMITH
Editor

# Tropospheric Effects in Ionosphere-Supported Radio Transmission*

GREENLEAF W. PICKARD†, FELLOW, IRE AND HARLAN T. STETSON†, ASSOCIATE, IRE

*Summary*—The importance of meteorological effects in tropospheric transmission has been presented by the authors in previous papers. More extended observations in the 40- to 50-Mc band, recently analyzed, substantiate high correlations with temperature, atmospheric refraction, and wind direction.

Since ionospheric-supported transmissions must pass at least twice through the troposphere, investigations have been made for possible meteorological effects at frequencies from the broadcast band to 10 Mc and comparisons made with known tropospheric reception at frequencies in the 40 to 50 Mc band. Correlation coefficients as high as $0.8 \pm 0.1$ are shown.

Reception from WBBM, Chicago, 850 miles distant, based on 3 years data, show a direct correlation with surface refraction and with atmospheric temperature changes. Night fields rise from a minimum one day before a temperature rise equal to 10°F to a maximum on the day such temperature change occurs. Similarly, a drop in field occurs one day prior to an equal temperature fall.

Studies of reception at Needham of WWV, 5 Mc (Beltsville, Md.) 373 miles distant, show similar correlations as those determined for the broadcast band. Furthermore, night fields of WWV, 5 Mc tend to a maximum one day preceding a prevailing SE wind directed broadside to the transmission path. Little correlation was found with NE winds directed along the path. Variations in field intensities, range from 15 to 30 per cent of mean fields dependent upon meteorological conditions at the receiving end. Such variations may be attributed to atmospheric bending, reflection transmissions due to changing lapse rate at the lower atmospheric levels, and possibly to as yet undetermined troposphere-ionosphere relations.

FOR THE PAST DECADE, interest in tropospheric effects on radio transmission has been almost entirely centered on the very high frequencies, largely because of the wartime importance of radar. But inasmuch as all ionosphere-supported transmission must pass at least twice through the troposphere, all portions of the radio spectrum must, to some extent, at least, be affected by happenings in the lower atmosphere.

In one respect only is there a fundamental difference with respect to frequency; the duct transmission so often observed at vhf is absent at the lower radio frequencies, since these are beyond cutoff for any duct dimension possible in the troposphere. In all other respects, tropospheric effects are not dependent on frequency, save only that reflection from discontinuities, such as temperature or humidity inversions, and at interfaces of air masses of different refractive index is greater at the lower the frequency. And the changes in wave bending due to varying refractive lapse rate may be expected to be the same at all presently used frequencies.

It might be expected that correlations of radio transmission with tropospheric changes at any one point on the transmission path would be most prominent on relatively short distance transmission, for the reason that only on such paths would the atmosphere involved be in any sense homogeneous. This has been confirmed by an examination of 3 years continuous field strength recording at the MIT Cosmic Terrestrial Research Laboratory at Needham, Mass., from station XEWW at Mexico City, Mexico, operating at 9.5 Mc and 2300 miles distant. This is a two-hop path, and transmission must, therefore, pass four times through the troposphere, in three widely separated areas; obviously tropospheric conditions would, in general, be different in these areas, so it was not unexpected that analysis of the field data showed no definite correlation with Boston weather elements.

In 1928 one of the present authors[1] examined the relation between night fields at Newton Center, Mass., from broadcast station WBBM at Chicago, then operating at 1330 kc and 850 miles distant, and also the night fields at Pasadena, Calif., from broadcast station KPO at San Francisco, Calif., operating at 680 kc and 343 miles distant, with surface temperature and barometric pressure at different points along the transmission paths. The highest correlation found was with surface temperature at the receiving point, and the relation found was direct, and not inverse as Austin[2,3] had previously found for his day fields.

In 1933 Colwell[4] found the field at Morgantown, W. Va., from broadcast station KDKA at Pittsburgh, Pa., 60 miles distant, showed marked correlation with surface pressure changes, and particularly with the relative location of high- and low-pressure areas to the transmission path.

In this early work there was little attempt to suggest a mechanism by which tropospheric changes might affect radio transmission. But in 1937 Watt, Wilkins, and Bowen in England[5] and Friend and Colwell in this country[6] independently observed and reported reflections

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† Cosmic Terrestrial Research Laboratory, Massachusetts Institute of Technology, Needham, Mass.

[1] G. W. Pickard, "Some correlations of radio reception with atmospheric temperature and pressure," Proc. I.R.E., vol. 16, pp. 765–773; June, 1928.

[2] L. W. Austin, "Field intensity measurements in Washington on the Radio Corporation stations at New Brunswick and Tuckerton, New Jersey," Proc. I.R.E., vol. 12, pp. 681–693; December, 1924.

[3] L. W. Austin and I. J. Wymore, "Radio signal strength and temperature," Proc. I.R.E., vol. 14, pp. 781–785; December, 1926.

[4] R. C. Colwell, "Cyclones, anticyclones, and the Kennelly-Heaviside layer," Proc. I.R.E., vol. 21, pp. 721–725; May, 1933.

[5] R. A. W. Watt, A. F. Wilkins, and E. G. Bowen, "The return of radio waves from the middle atmosphere," *Proc. Roy. Soc.*, ser. A, vol. 161, pp. 181–196; July, 1937.

[6] A. W. Friend and R. C. Colwell, "Measuring the reflecting regions in the troposphere," Proc. I.R.E., vol. 25, pp. 1531–1541; December, 1937.

from layers between 1 and 12 km in height, with occasional echoes from over 15 km. Friend and Colwell suggested that these tropospheric reflections might explain Colwell's earlier observations.

In this present paper, the authors have re-examined WBBM Chicago to Boston transmission with respect to Boston weather elements, and in addition have analyzed over three years each of Needham recorded fields from WWV 5 Mc, XEWW 9.5 Mc, WWV 10 Mc, and W2XMN at 44.1 Mc for correlation with tropospheric changes. They had previously[7] established certain tropospheric correlations with the W2XMN field at Needham with two years of recording: here another year of data has been added. For the reason that the transmissions other than W2XMN generally contained day-to-day and even hour-to-hour ionosphere-controlled changes of much greater magnitude than the tropospheric amplitudes, the analysis used throughout was purely statistical. In this form of analysis, some sharply defined tropospheric event, such, for example, as a large temperature change, is taken as an epoch, and received fields are repeatedly summed up around a series of such epochs. Field changes which are not related to the epoch distribute randomly around the epoch and tend to cancel out, such cancellations being the more complete the longer the series, while related field changes continually add up to maxima and minima.

A statistical examination of 1945, 1946, and 1947 values of critical frequencies has been made with $f^\circ E$ and $f^\circ F_2$ distributions around tropospheric events, and no definite correlation became apparent. However, relation of troposphere and ionosphere cannot be ruled out; Mihran[8] has recently shown an apparent correlation.

Fig. 1 shows the relation of a purely tropospheric radio transmission, W2XMN at Alpine, N. J., operating at 44.1 Mc as received at Needham, Mass., 167 miles away, to Boston; mean temperature changes taken as epochs. It will be noted that this figure contains an element of prediction as field maxima and minima precede by 1 day the Boston temperature changes taken as epochs. The correlation between W2XMN fields and Boston temperatures is best seen by the dotted line showing Boston temperature distribution around Boston temperature rise (moved 1 day to the left) against the full-line curve for W2XMN.

In the absence of tropospheric soundings, surface refraction is our only available measure of refractive lapse rate, which determines the bending of radio wave fronts. In order to compare radio transmission with tropospheric reflections from inversions and interfaces, it is, at present, necessary to do this indirectly; that is, to compare some surface event which is related to inversions. One such surface event is a prevailing wind from the ocean, particularly during the summer months. As the transmission path from Alpine to Needham runs NE-SW, a SE wind will reach the path most directly.

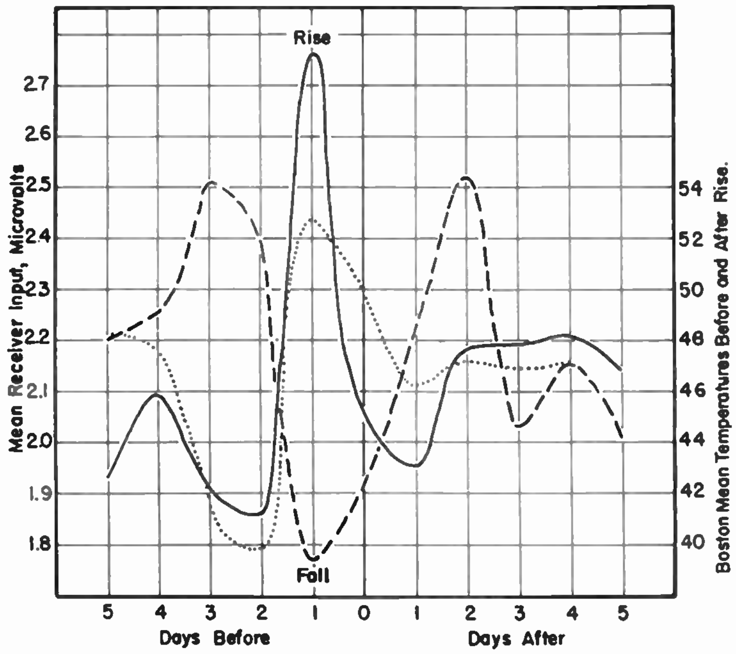


Fig. 1—Distribution of W2XMN fields around Boston mean temperature rise and fall of ≥10°F. 80 cases, 1945 to 1948.

In Fig. 2 are shown in full-line W2XMN fields against SE prevailing winds at Boston as epoch. Despite the imperfection of SE winds as an index of inversions, a clear relation with an amplitude of 15 per cent is shown.

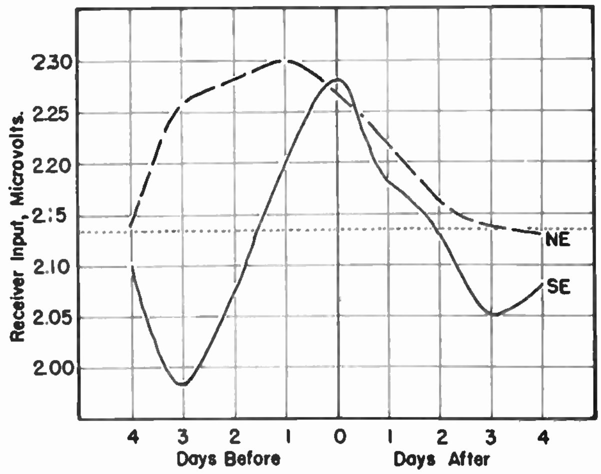


Fig. 2—Distribution of W2XMN fields around Boston SE and NE winds. Winter and summer, 1945 to 1948.

A less definite correlation with NE winds is shown in broken line. This figure is based upon all SE and NE winds, both summer and winter, with 61 cases of SE wind and 83 cases of NE wind.

[7] G. W. Pickard and H. T. Stetson, "A study of tropospheric reception at 42.8 Mc and meteorological conditions," Proc. I.R.E., vol. 35, pp. 1445–1450; December, 1947.

[8] T. G. Mihran, "A note on a new ionospheric-meteorological correlation," Proc. I.R.E., vol. 36, pp. 1093–1096; September, 1948.

In Fig. 3, based upon 23 cases, is shown the distribution of W2XMN fields around SE winds at Boston, during June, July, August, and September. Here the maximum is reached 1 day after the SE wind's advent, and the amplitude has increased to 34 per cent, from the 15 per cent of the preceding figure.

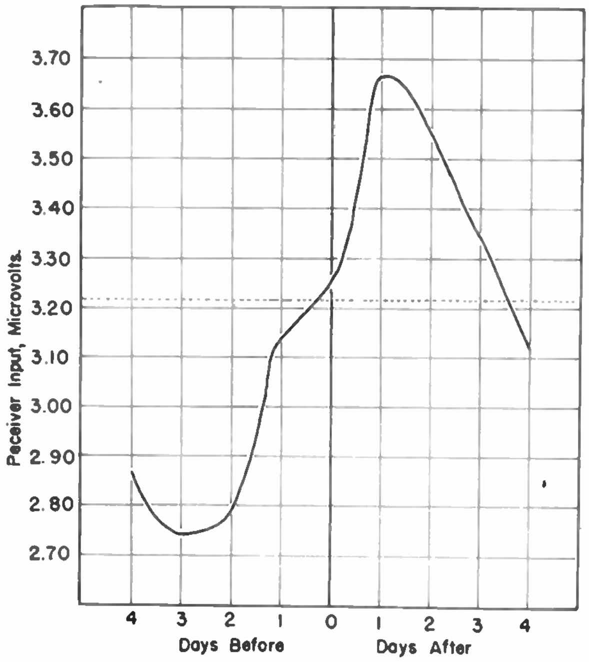


Fig. 3—Distribution of W2XMN fields around Boston SE wind, June, July, August, and September only.

Having shown that a purely tropospheric transmission, at a frequency too low for duct transmission, is highly related to tropospheric changes in temperature and wind direction favorable for inversions, we next examine similar relations at lower frequencies. Fig. 4 shows the field distribution of an ionosphere-supported transmission, WBBM at Chicago, operating at 780 kc, and received at Needham, 850 miles distant, around epochs of Boston mean temperature rise and fall. In the 1928 investigation of the same station and path, then at 1330 kc, it was found that the night fields from Chicago rose and fell with Boston temperatures; this figure shows the same thing. Around a temperature rise as epoch, the night field rose from a minimum one day before to a maximum on the epoch and one day after. The relation to a temperature fall is seen to be rather closely inverse, consisting of a field fall from a maximum one day before to a minimum one day after the epoch. And from four days before to five days after the epoch, the curves run inversely. A portion of the temperature distribution around the temperature rise is shown in dotted line; the correlation here is $r = 0.8 \pm 0.07$ or a ratio of correlation coefficient to probable error of over 11.

We shall next examine transmission from WWV's transmitter at Beltsville, Md., 373 miles distant from Needham. Owing to the shortness of the path from Beltsville to Needham, the transmission passes through the troposphere at a fairly high angle, which would not favor reflections; nevertheless, a clear correlation is evident for night fields.

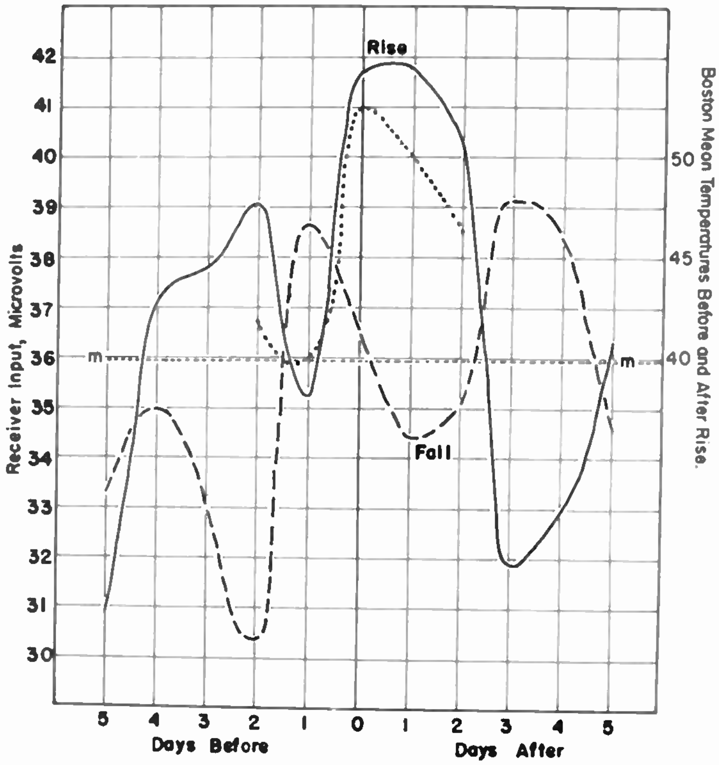


Fig. 4—Distribution of WBBM fields around Boston mean temperature rise and fall of $\geq 10°$F. 80 and 72 cases, respectively, 1945 to 1948.

In Fig. 5 is shown the distribution of WWV 5-Mc night fields at Needham with respect to Boston temperature rise and fall. It will be seen that the curves run inversely throughout, and also that the relation to Boston

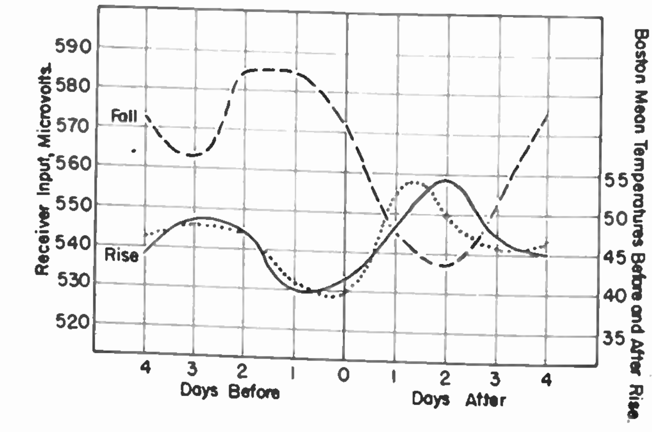


Fig. 5—Boston mean temperature rise and fall of $\geq 10°$F as epoch: WWV-5 21–03 E.S.T., 90 and 80 cases respectively, 1945 to 1948.

temperature is inverse, as has already been shown for WBBM night fields at broadcast frequencies. Again for comparison is shown in dotted line the distribution of Boston temperatures around rises; this has been moved one day to the right, as WWV 5-Mc fields lag temperature by one day. The correlation between the dotted and full line curves is $r = 0.8 \pm 0.1$, or a ratio of correlation coefficient to probable error of 8.

In Fig. 6 is shown the distribution of WWV 5-Mc

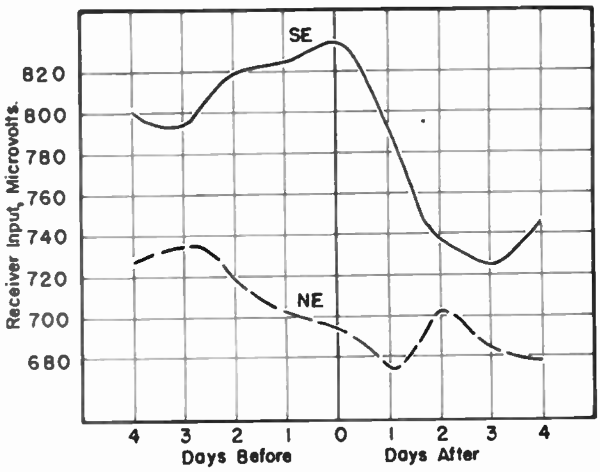


Fig. 6—Distribution of WWV-5 night fields around Boston SE and NE winds.

night fields with respect to SE and NE winds. The full-line curve is for SE winds, and reaches a maximum on the epoch day, whereas the NE wind distribution shows little correlation. Again, in view of the unfavorable angle which this transmission makes with the nearly horizontal reflecting layers, it is perhaps surprising that any correlation should be shown.

In conclusion, it may be stated that this investigation of the effects of meteorological changes upon the propagation of radio waves from the broadcast band to FM frequencies has entailed an extended analysis of measurements made for basic studies at the Cosmic Terrestrial Research Laboratory at Needham between 1945 and 1948 of over 30,000 hourly values. Both day and night transmissions have been analyzed with respect to temperature, atmospheric refraction, including pressure and humidity factors, and with wind direction.

Examination in detail of this large amount of computational work has revealed that there are variations of from 15 to 30 per cent in mean-field values dependent upon meteorological conditions. There have been individual instances where field variations as great as 2 to 1 in ionospheric-propagated transmission have resulted from meteorological factors affecting the waves necessarily traversing the troposphere. Investigations of conditions at various points along the path indicate that it is the atmospheric state at the receiving end, rather than at the transmitting end or midpoint, that shows the greatest correspondence between transmission conditions and the meteorological elements at the earth's surface. It is apparent that lack of uniformity in the lower atmosphere introduces variable refraction and reflection from inversions and air-mass interfaces. This causes significant effects in communication at relatively low frequencies.

# The Reliability of Ionospheric Height Determinations*

LAURENCE A. MANNING†, ASSOCIATE MEMBER, IRE

*Summary*—**A study is made of the uncertainties existing in determinations of electron distribution based on virtual height data. The effect of an assumed distribution in the valley between two layers upon the necessary upper-layer form is investigated, and a method for evaluating the maximum error in height determination is given.**

## Introduction

In a previous paper[1] the writer has shown that analysis of ionospheric virtual height versus frequency records, by a process of integral transformation, will yield the true distribution of electrons in the upper atmosphere. In the course of that analysis it was found necessary to make a number of idealizing approximations. It is the purpose of the present paper to investigate the significance of the most restrictive of these assumptions.

It was assumed that:

(a) The distribution function $Z(M)$, specifying the height at which an electron density $M$ is found, is single-valued. This condition is violated by an ionosphere of more than one layer.

(b) The ionosphere is lossless, or that the electron collision frequency is zero. In the $E$ layer the assumption is not upheld.

(c) Geometrical optics is valid. At the top of the path the assumption is invalid.

* Decimal classification: R113.602.21. Original manuscript received by the Institute, August 11, 1948; revised manuscript received, December 20, 1948.

The work described in this paper has been supported in part by the Office of Naval Research and the Army Signal Corps.

† Stanford University, Stanford, Calif.

[1] L. A. Manning, "The determination of ionospheric electron distribution," Proc. I.R.E., vol. 35, pp. 1203–1208; November, 1947.

(d) The earth's magnetic field does not affect propagation.

Assumption (b), (c), and (d) have been treated by Rydbeck.[2,3] Assumption (a) is of greater practical importance, and will be considered forthwith.

## Basic Distribution Analysis

As in the previous paper, the abbreviation $M=KN$ will be used for electron density. It is proportional to the true electron density $N$, but has units of squared frequency. The function $M(z)$ specifies the density at hieght $z$, while the inverse function $Z(M)=M^{-1}(z)$ gives the heights at which the density $M$ is to be found. In terms of the distribution function $M(z)$ it is possible to write the virtual height versus frequency integral expressing the experimentally measured time delays as

$$z_v(f) = \int_0^{z_{\max}=M^{-1}(f^2)} \frac{dz}{(1-M/f^2)^{1/2}} \tag{1}$$

where $f$ is frequency. When only one layer maximum is present, $M(z)$ is monotonically increasing, and the inverse function $Z(M)$ is single-valued. Then it is possible to write (1) in the equivalent form:

$$z_v(f) = \int_0^{f^2} \frac{Z'(M)dM}{(1-M/f^2)^{1/2}}. \tag{2}$$

It was shown that, in cases in which the inversion of $M(z)$ is valid, the integral may also be inverted; the distribution function is then given in terms of the experimentally determined virtual height versus frequency function by

$$Z_T(f_v) = \frac{2}{\pi}\int_0^{f_v} \frac{z_v(f)df}{(f_v^2-f^2)^{1/2}} \tag{3}$$

in which $Z_T(f_v)$ is the true height of penetration of a wave of vertical-incidence frequency $f_v$.

## Correction for Second Layer

Virtual-height data is not sufficient to determine the electron distribution at heights beyond the first layer maximum. Nevertheless, the integral of (3) can still be evaluated at frequencies higher than the first-layer critical frequency. From physical considerations it was indicated in the first paper that, when the frequency $f_v$ is much greater than a lower critical frequency, the error of assuming the representation of (2) to be possible must approach zero. The error which will actually result from applying (3) when several layers exist will now be investigated.

In Fig. 1 is shown an electron-distribution curve representing a lower layer of critical frequency $f_0$; it shields a valley at heights between $z_0$ and $z_2$. At heights greater than $z_2$ the electron density rises again to a higher maximum. Note that $Z(M)$ is triple-valued when $f_1^2<M<f_0^2$. In order to apply (2) when $f>f_0$, it is

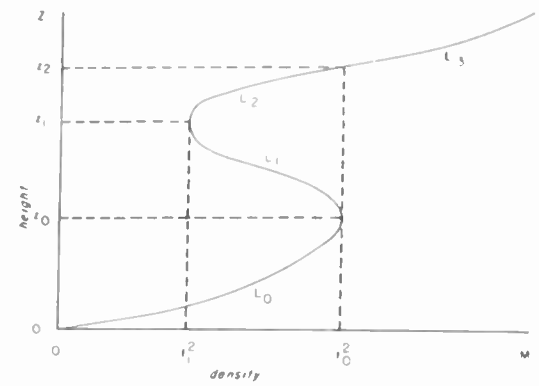


Fig. 1—A multiple-valued ionospheric height distribution.

necessary to break the curve up into single-valued components $L_0$, $L_1$, $L_2$, and $L_3$, and then integrate over each part of the path separately. If $f>f_0$,

$$z_v(f) = \int_0^{f_0^2} \frac{Z_0'(M)dM}{(1-M/f^2)^{1/2}} + \int_{f_1^2}^{f_0^2} \frac{Z_1'(M)dM}{(1-M/f^2)^{1/2}} + \int_{f_1^2}^{f_0^2} \frac{Z_2'(M)dM}{(1-M/f^2)^{1/2}} + \int_{f_0^2}^{f^2} \frac{Z_3'(M)dM}{(1-M/f^2)^{1/2}} \tag{4}$$

where $Z_1(M)$ is the distribution over the part of the curve $L_1$, and is measured from the height $z_1$ downwards. Distribution $Z_2(M)$ is also measured from the valley minimum. The two inner integrals represent the retardation in the valley. The form of (4) shows that, for $f>f_0$, the retardation is exactly the same as would be obtained from a new distribution $Z_b(M)$ defined by

$$Z_b(M) = Z_0(M) + Z_1(M) + Z_2(M) + Z_3(M). \tag{5}$$

inasmuch as application of (2) to (5) yields just (4). When $f_1<f<f_0$, the virtual height resulting from $Z_b(M)$ does not correspond to the experimental retardation, which is dependent on $Z_0(M)$ alone. Since $Z_0$, $Z_1$, $Z_2$, and $Z_3$ are all defined in such a way as to be monotonically increasing and single-valued, $Z_b(M)$ is also. Fig. 2 shows the form of the new distribution.

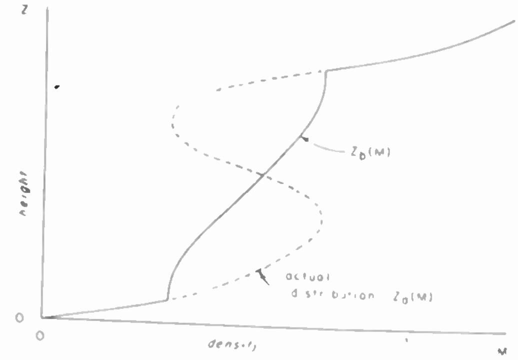


Fig. 2—A single-valued ionization distribution giving identical retardation at high frequencies as the actual distribution.

[2] O. E. H. Rydbeck, "On the propagation of radio waves," *Chalmers Tek*, Hogskol, Handl. No. 34, 158 pp.; 1944.

[3] O. E. H. Rydbeck, "Theoretical survey of the possibility of determining the distribution of free electrons in the upper atmosphere," *Chalmers Tek*, Hogskol, Handl. No. 3, 74 pp.; 1942.

The new distribution gives the same retardation as the old distribution when $f>f_0$, but since it is single-valued the representation of (2), and the resultant inversion of (3), are possible. The true height corresponding to $z_v(Z_b)$ is uniquely $Z_b$. Since $Z_b$ is the same as the original distribution for $z>z_2$, analysis of $z_v(Z_b)$ leads to the correct true height $Z_a$ when $z>z_2$. To find the true distribution $Z_a$ for $0<z<z_0$, the experimental virtual height $z_v=z_v(Z_0)$ is correct.

When $f>f_0$, the experimental $z_v(Z_a)$ is equal to $z_v(Z_b)$. When $f<f_0$, the experimental $z_v$ is $z_v(Z_0)$ while the virtual height $z_v(Z_b)$ which leads to the correct *upper-layer* distribution is $z_v(Z_0+Z_1+Z_2)$. Their difference, $z_v(Z_1+Z_2)=\Delta z_v$, is the addition which must be applied to the experimental virtual-height curve before it will give the true distribution for $z>z_2$.

$$\Delta z_v = \int_{f_1^2}^{f^2} \frac{Z_1'(M)+Z_2'(M)}{(1-M/f^2)^{1/2}}\,dM$$
$$\equiv \int_{z_1}^{M_1^{-1}(f^2)} \frac{dz}{(1-M_1/f^2)^{1/2}} + \int_{z_1}^{M_2^{-1}(f^2)} \frac{dz}{(1-M_2/f^2)^{1/2}}. \tag{6}$$

Error arises in computing the true height of the upper layer if the experimental data are used, because the virtual height in the region $f_1<f<f_0$ is too small by the amount $\Delta z_v$. The magnitude of the height error can be ascertained by substituting $\Delta z_v$ of (6) in (3).

$$\Delta Z_T(f_v) = \frac{2}{\pi}\int_{f_1}^{f_0} \frac{\Delta z_v(f)df}{(f_v^2-f^2)^{1/2}}. \tag{7}$$

The true height of the layer is always greater than that computed directly from the experimental data by the amount $\Delta Z_T$.

### Nature of the Single-Valued Distribution

When the experimental data are analyzed by applying (3) directly, a single-valued distribution is obtained which is lower than the true distribution by the error height $\Delta Z_T$. Fig. 5 shows such a distribution. Now, we wish to show that the single-valued distribution thus obtained is consistent with the experimental virtual-height curve, and that it is *mathematically* a correct solution to the true-height problem.

If $v(f)$ is the component of virtual height resulting from passage of the wave through the valley of the multiple-valued distribution, and $\delta Z(M)$ is the difference between the heights of the single- and multiple-valued curves for $M>f_0^2$, the condition that both curves give the same total virtual height is

$$\int_{f_0^2}^{f^2} \frac{-\delta Z'(M)dM}{(1-M/f^2)^{1/2}} = v(f). \tag{8}$$

Reference to (2) shows that the integral on the left is the increased virtual height the single-valued distribution must give, for heights above $M=f_0^2$, in order to compensate for its lack of valley retardation. The valley retardation $v(f)$ is given in terms of the valley-height distribution $V(M)=Z_1(M)+Z_2(M)$ by

$$v(f) = \int_{f_1^2}^{f_0^2} \frac{V'(M)dM}{(1-M/f^2)^{1/2}}. \tag{9}$$

By taking $\delta Z(M)$ and $V'(f)$ to be zero for $M$ less than $f_0^2$ and $f_1^2$, respectively, the lower limits in the integrals of (8) and (9) may be made zero. Then, application of the inversion of (3) demonstrates that

$$\delta Z(f_v) = h - \frac{2}{\pi}\int_0^{f_v} \frac{v(f)df}{(f_v^2-f^2)^{1/2}} \tag{10}$$

where $h$ is the valley width. The above equation gives the height difference between a multiple-valued distribution and a single-valued distribution having the same virtual height. Equation (7) give the relation between a multiple-valued distribution and the single-valued distribution obtained by analyzing directly the virtual height curve corresponding to the multiple-valued distribution. If we can show that $\delta Z(f_v)=\Delta Z_T(f_v)$, we will have shown that the true height obtained by applying (3) to the experimental data is always consistent with the experimental virtual height. We will also have shown that (6) and (7) can be used to find new true-height curves containing arbitrarily assumed valley distributions and giving the correct virtual height, because the height correction $\Delta Z_T(f_v)$ being added to the experimentally determined true height will exactly compensate for the assumed valley retardation by virtue of (8) through (10).

To show that (7) and (10) are identical, it is first necessary to put (6) in (7), and (9) in (10). Remember that $V(M)=Z_1(M)+Z_2(M)$ is the valley distribution. Thus, letting $f^2=g$,

$$\Delta Z_T(g_v) = \frac{1}{\pi}\int_0^{g_0}\int_0^{g} \frac{V'(M)dMdg}{(g-M)^{1/2}(g_v-g)^{1/2}} \tag{11}$$

and

$$\delta Z(g_v) = h - \frac{1}{\pi}\int_{g_0}^{g_v}\int_0^{g_0} \frac{V'(M)dMdg}{(g-M)^{1/2}(g_v-g)^{1/2}}. \tag{12}$$

These integrals can be shown to be equal by changing the order of integration, and noting that the valley width $h=\int_0^{g_0}V'(M)dM$. Both integrals then become

$$\Delta Z_T(g_v) = \delta Z(g_v)$$
$$= \frac{2}{\pi}\int_0^{g_0} V'(M)\sin^{-1}\left(\frac{g_0-M}{g_v-M}\right)^{1/2} dM, \tag{13}$$

and the mathematical correctness of the directly obtained solution of the true height problem is demonstrated despite the presence of more than one layer maximum.

### Effect of Assumed Valley Distributions

From ionospheric-sounding data it is physically impossible to determine the distribution of electrons in the valley of the true-height curve. Since the valley distribution is needed to evaluate the error heights of (6) and (7), or their equivalent, (13), it is not possible to correct the upper-layer distribution exactly. What can be done is to make a number of assumptions as to the valley distribution, and then consider the resulting corrections as first approximations to the true corrections, or as limits to the error of the analysis.

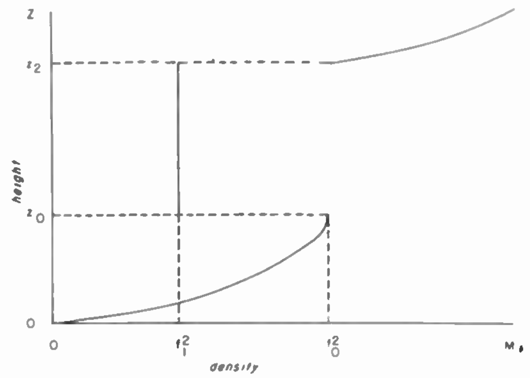


Fig. 3—An assumed constant-density valley distribution.

The simplest assumption about the valley is that it is of constant electron density $M=f_1^2$, as in Fig. 3. Then the virtual-height correction of (6) becomes

$$\Delta z_v = \int_{z_0}^{z_2} \frac{dz}{(1-f_1^2/f^2)^{1/2}} = \frac{h}{(1-f_1^2/f^2)^{1/2}}$$

where $z_2 - z_0 = h$ is the valley width. The error height $\Delta Z_T$ of (7) is now

$$\begin{aligned}\Delta Z_T &= \frac{2}{\pi}\int_{f_1}^{f_0} \frac{h}{(1-f_1^2/f_0^2)^{1/2}} \cdot \frac{df}{(f_v^2-f^2)^{1/2}} \\ &= \frac{2}{\pi} h \sin^{-1}\left(\frac{f_0^2-f_1^2}{f_v^2-f_1^2}\right)^{1/2}\end{aligned} \qquad (14)$$

when $f_v > f_0$. If $f_v < f_0$, $\Delta Z_T = 0$. Note that when at the critical frequency, with $f_v = f_0$, the true-height correction is equal to the width of the valley. For higher frequencies the error decreases, and approaches zero when $f_v/f_0$ approaches infinity.

The true-height curve obtained by direct application of (3) to the experimental records does not have discontinuous jumps in height at a critical frequency. The discontinuity in the true-height curve which actually exists across the valley must be obtained by adding the discontinuous correction factor; since the correction factor is zero for $f < f_0$, and invariably equal to $h$ at $f = f_0$, application of (7) corrects for the inability of the single-valued analysis to cope with a multiple-valued situation.

The result of (8) is useful mainly in establishing a limit to the error in neglecting the valley density. If it is assumed that there are no electrons in the valley, $f_1 = 0$, and $\Delta Z_T = (2/\pi)h \arcsin(f_0/f_v)$. If $h$ is assumed to be larger than the actual valley width, the error height of this assumption will be larger than the true correction, and a bound is placed on the inaccuracy of the upper-layer height determination. In curve (a) of Fig. 4 is plotted the error height in percentage of the valley

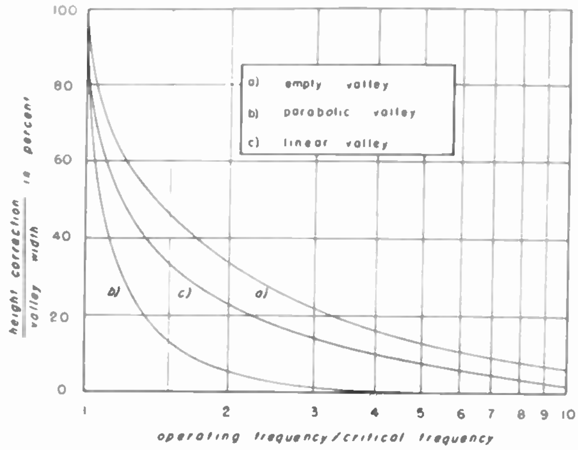


Fig. 4—Height corrections computed for three assumed valley distributions.

width, as a function of the extent by which the frequency exceeds the lower critical frequency. In Fig. 5 are shown the upper-layer electron distributions which would result in a typical case from assuming an unionized valley of various widths. The record used is that of Fig. 9 of the previous paper.

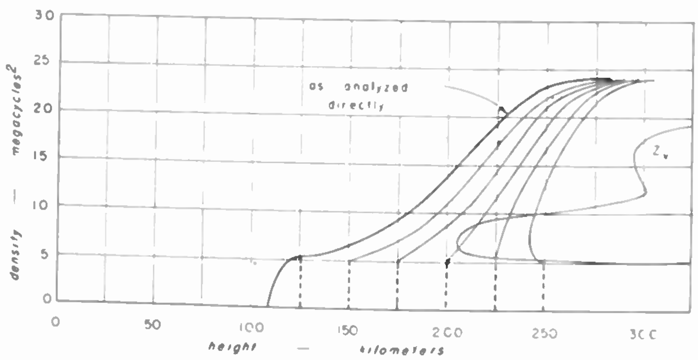


Fig. 5—An example of an electron distribution corrected for assumed empty valleys of various widths.

Note that a valley width of greater than about 110 km is impossible physically in Fig. 5, since the $M(z)$ curve would become multiple-valued. A further restriction on the width of the valley may be obtained by noting that the computed true height must be less than the virtual height, as witness (1). The true height can only exceed the virtual height in the physically impossible case in which $M(z)$ is multiple-valued, and more than one electron density is ascribed to a given height. As applied to Fig. 5, this condition demands that the valley width be less than about 77 km. The first condition should serve as well, if it were possible to get the form of the curves with sufficient accuracy when $M$ is but slightly greater than $f_0^2$. If we now assume the largest possible valley width of 77 km, and then assume the least possible number of electrons in the valley, the error in the computed true height of the $F$-layer maximum is only 25 km. On the basis of other assumptions as to valley distribution, the true height error at the $F$-layer maximum would be less.

It is rather likely that the lower-layer ionization suddenly drops to zero beyond the height $z_0$. A more reasonable assumption is that the ionization drops off as the tail of a Chapman region; as an approximation valid near $z_0$, we shall assume the density to drop off parabolically. Let the parabola be $M=f_0^2[1-(z/h)^2]$, so that in a width $h$ of valley the ionization density decreases from $f_0^2$ to zero. Then,

$$\Delta z = \int_h^{h(1-f^2/f_0^2)^{1/2}} \frac{dz}{\{1-f_0^2[1-(z/h)^2]/f^2\}^{1/2}}$$

$$= -\frac{hf}{2f_0}\ln\frac{f_0-f}{f_0+f},$$

and the error height is

$$\Delta Z = -\frac{2}{\pi}\int_0^{f_0}\frac{hf}{2f_0}\,\frac{\ln\dfrac{f_0-f}{f_0+f}}{(f_v^2-f^2)^{1/2}}\,df.$$

Integrated by parts, the error height becomes

$$\Delta Z = \frac{2}{\pi}h\sin^{-1}\frac{1}{r} - \frac{h}{\pi}(r^2-1)^{1/2}\ln\frac{r^2+1}{r^2-1} \qquad (15)$$

where $r=f_v/f_0$. Equation (15) is plotted as curve (b) in Fig. 4.

One other easily computed valley distribution is that for which the density varies linearly from zero to $f_0^2$ in a width $h$. Then $\Delta Z_v=2h\,f^2/f_0^2$, and

$$\Delta Z = \frac{2}{\pi}h\left\{r^2\sin^{-1}\frac{1}{r} - (r^2-1)^{1/2}\right\} \qquad (16)$$

where as above $r=f_v/f_0$. Equation (16) is also in Fig. 4 as curve (c).

If it is desired, the valley may be divided into a number of regions. The first region might be parabolic and of width $h_p$, the second of constant density zero and width $h_0$, while the last section might be linear of width $h_l$. The total correction $\Delta Z_T$ is then the sum of those found separately for all the sections of the valley.

The true-height corrections which have been computed provide a means of establishing limits to the height errors. Whenever a valley distribution is assumed, and a corrected true-height curve obtained, analysis of this distribution function by the application of (1) will give the experimental virtual-height curve. Thus, merely on analytic grounds, the uncorrected true-height curve is as good a solution to the problem presented by the experimental virtual-height curve as is a corrected curve embodying a valley. The argument in favor of assuming a valley distribution must be based either on physical reasoning as to probable layer shapes, or upon considerations of continuity between records obtained at different times of day.

It seems worth noting that the present method of analysis obtains the maximum possible amount of information from the experimental curves. All uncertainties in the distribution are effects of the physical situation and so penetrate such approximate methods of true-height analysis as Booker and Seaton's parabolic layer-matching process.[4]

## Conclusion

An analysis of virtual-height records possessing critical frequencies invariably involves ambiguity with respect to that portion of the electron distribution beyond the first layer maximum. Direct integral analysis of the experimental virtual-height curves determines a distribution having an inflection in height versus density at the density corresponding to the critical frequency. The distribution thus found shows no valley between separate layers, and is consistent with the experimental virtual-height curves. For physical reasons, one may believe that several separate layers actually exist. It is then possible to find a true-height correction which makes the computed true height include the postulated valley, and yet be consistent with the virtual-height data. A limit on the width of an assumed valley is imposed when it is found that the corrected true-height curve is multiple-valued in electron density. By assuming the largest valley which still permits interpretation with a physically realizable distribution, a limit is found to the error in determination of upper-layer true height. The uncertainty found applies to any method of true-height analysis from virtual-height data.

[4] H. G. Booker and S. L. Seaton, "Relation between actual and virtual ionospheric height," *Phys. Rev.*, vol. 57, pp. 87–94; January, 1940.

# Antenna Impedance Measurement by Reflection Method*

EDWIN ISTVÁNFFY†

**Summary—Two methods of measuring the impedance of radiators are described. They are based on measurement of the power reflected by the radiators. These methods are particularly useful for finding the radiation resistance of thick half-wave dipoles.**

Numerous papers published during the last few years relating to antenna-impedance calculations and measurements are concerned with determining the radiation resistance of antennas in which the radius of the antenna conductor cannot be neglected. Although recently published new methods of calculation give better approximations, they are not adequate for a dipole in which $\Omega = 2 \log_n (2l/a) \leq 10$, where $2l$ and $a$ are the length and the radius of the dipole, respectively, and $\Omega$ is an often-used parameter determined by the length-to-thickness ratio of the conductor.

Impedance measurements commonly introduce discontinuities through coupling the dipole to the measuring equipment. These difficulties may be avoided by placing the antenna in the beam of a directional transmitter-receiver—like a radar target—and measuring the power reflected to the receiver.

The amplitude of the received signal depends on the *reflected power*, and the target distance for maximum reflected power depends on the *phase of the antenna impedance.*

By varying the length of a dipole, while retaining a constant thickness, the resonant length for a given wavelength may be found from the changes in reflected power.

Another check is to compare dipoles of different thicknesses, whose resonant lengths have been determined. If their resonant lengths correspond exactly, maximum reflection must occur at exactly the same target distances.

When the resonant lengths of dipoles having different length-to-thickness ratios are determined, a comparison of their reflective properties from the same target distance permits the ratio of their radiation resistances to be calculated.

If the field strength produced by the transmitter at the dipole is $E_t$ the induced voltage in the dipole is $V = E_t h_{eff}$ where $h_{eff}$ is the effective length of the dipole. The impedance of a dipole adjusted to its resonant length is equal to its radiation resistance ($R_r$). Thus, the reflected power is

$$P_r = \frac{E_t^2 h_{eff}^2}{R_r} \quad (1)$$

and the ratio of the reflected powers for two dipoles is

$$\frac{P_r''}{P_r'} = \left\{\frac{h_{eff}''}{h_{eff}'}\right\}^2 \frac{R_r'}{R_r''}. \quad (2)$$

The directional properties of resonant-length dipoles of different thicknesses show only small variations and, for small differences in length,

$$\frac{h_{eff}''}{h_{eff}'} = \frac{l''}{l'} \quad (3)$$

where $l''/l'$ is the ratio of the resonant lengths of the two dipoles. The ratio of resistances, as derived from (2) and (3), is

$$\frac{R_r''}{R_r'} = \frac{P_r'}{P_r''}\left(\frac{l''}{l'}\right)^2. \quad (4)$$

Although this is a comparative method, the absolute value of the radiation resistances may also be found, provided the value of one of the tested dipoles is known.

Another method of determining radiation resistance is based on measured phase-angle data as a function of dipole length in electrical degrees. This method is not comparative. Using an impedance circle diagram, it is possible to determine the radiation resistance by plotting one part of the impedance locus curve.

## 1. Setup and Technique

The measurements were carried out on an athletic field. Horizontally polarized waves were used. Fig. 1 shows the equipment as set up for making measurements. On the table are the transmitting and receiving parabolic reflectors, which are of similar design. In

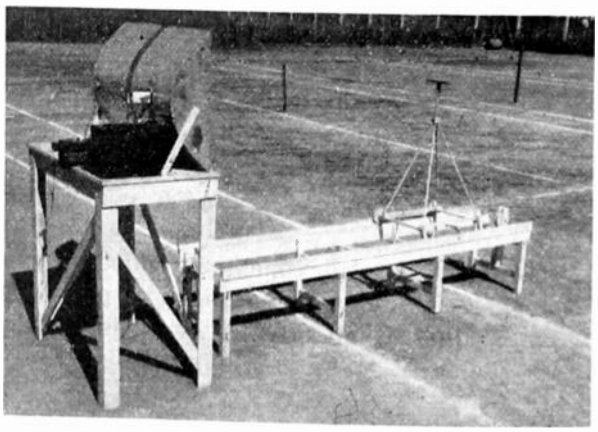

Fig. 1—The setup of equipment for making reflection measurements.

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front of the reflectors is a wooden track 2.5 meters (8.2 feet) long and 0.5 meter (1.6 feet) high, on which a little wooden carriage, which carries the dipole under test, can be moved by strings from behind the reflectors. The focal line of the reflectors was 1.5 meters (4.9 feet) above the ground and the axis of the dipole under test was set in each case to this height. An indicator fixed to the side of the carriage and a scale on the track, marked at every half centimeter, enabled the target distance to be determined within an accuracy of about 1 millimeter (0.04 inch).

A battery-operated 955 acorn triode produced an output power of about 0.1 watt at a wavelength of 55.2 cm (540 Mc). The transmitting dipole and its oscillator were placed in the focus of one of the parabolic reflectors, the apertures of which measured 69 by 50 cm (27 by 20 inches). For reducing the direct pickup by the receiver, the focus was 7 cm (2.75 inches) inside the aperture. The power gain of the reflector was 9.

The receiving dipoles and crystal rectifier were mounted in the other reflector, from which two wires were brought out through an rf filter to a meter. The detector was a silicon crystal with gold electrode and operated as a square-law rectifier.

Initially, a heavier carriage was used, but, even without a dipole, it caused significant changes in the meter deflection as it was moved. The structure of the carriage was gradually reduced to that shown in Fig. 1. With an initial deflection of 40 divisions on the receiving meter caused by direct pickup from the transmitter, a change of about 0.5 division or less resulted from moving the carriage along the track.

Tests made with dipole-supporting structures showed that, for thin antennas, the use of dielectric materials near the dipoles, especially at the ends, tended to decrease the resonant length and impair the response at resonance. To avoid these effects, the holder, shown in Fig. 2, was placed at the center of the dipole. The dipole is lying on two 1-mm (0.04-inch) phenol fiber plates 10 mm (0.39-inch) apart, and is held in position by a small rubber strip.

Some of the transmitted power picked up directly by the receiver and a relatively small amount reflected by the ground and distant objects produce an initial reading on the meter. Some efforts were made at first to compensate for these effects, but later it was obvious that this was unnecessary. The direct pickup causes a steady field at the receiver, to which the reflected field from the target dipole is added and when the target is moved, the phase of the reflected wave rotates, causing phase additions and oppositions with the steady field and alternate maximum and minimum deflections of the meter.

If the distance to the dipole target is large compared to that between the transmitting and receiving dipoles, the distance between adjacent maxima or minima is half the free-space wavelength. It may be shown that, with square-law detection, if the reflecting powers of two dipoles being tested are unequal, the difference in the meter deflections increases as the steady field becomes stronger. Further, there is an improvement in the accuracy of positioning the dipole for maximum deflection of the meter with increasing steady field, and the target-distance measurements make possible the calculation of the phase of the impedance.

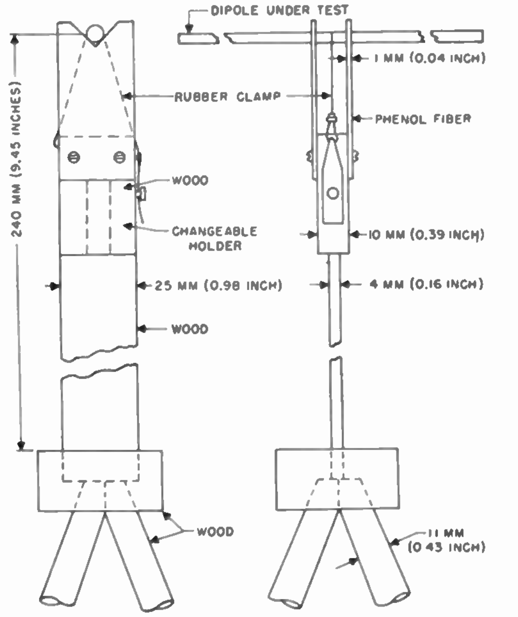


Fig. 2—The dipole support is changeable and is mounted at the top of the carriage.

If a dipole is made longer than its resonant length so that the current lags 45° behind the voltage, this will cause a time lag at the receiver that may be compensated for by moving the target 1/16 wavelength nearer to the transmitter-receiver. In this way, if the resonant length is known, the phase of the impedance can be determined.

## 2. Measurements

Five series of dipoles were tested; their diameters being 0.14, 0.48, 3, 13.6, and 40 mm (0.006, 0.019, 0.118, 0.535, and 1.575 inches). The first two sizes were of bare copper wire, and a series of different lengths ranging from 230 to 280 mm (9 to 11 inches) were prepared. The 3-mm dipoles were brass rods and a similar series were made. The dipoles of 13.6- and 40-mm diameters were brass tubes with 1-mm (0.04 inch) wall thickness and were adjustable in length. Both could be used with closed and open ends.

First, the field strengths were measured along the track of the target. For this purpose, a tuned dipole fitted with a thermocouple was placed on the holder and the readings on the linear scale of the meter were recorded for different distances. The plotted field power curve, which is proportional to the square of the field strength, is shown in Fig. 3. At a distance of 2.2 meters, there is a flat minimum caused by reflections from the earth. From about 2 to 2.4 meters, the field power changes very little and, therefore, this range was considered to be most suitable for measurements.

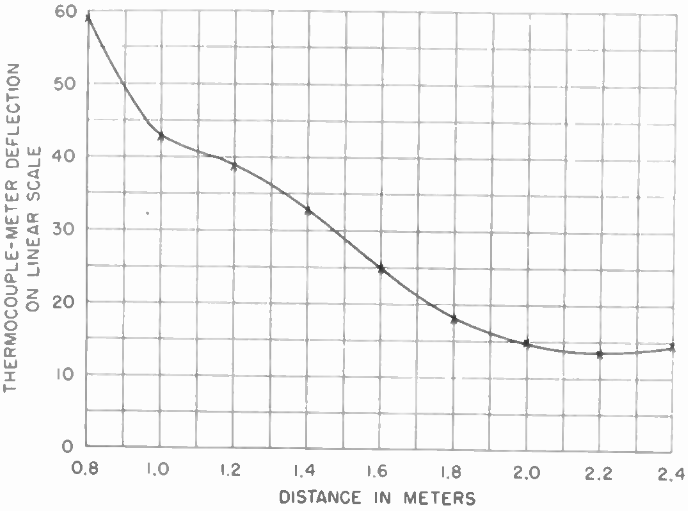


Fig. 3—The measured field-power curve along the track of the target. From about 2.0 to 2.4 meters, the field power is approximately constant and this range is most suitable for the measurements.

Fig. 4 shows the maxima and minima of the receiver meter deflections when moving a 3-mm dipole of resonant length along the track. A deflection of 41 divisions was obtained in the absence of a dipole. The square root values of the readings are proportional to the field strengths.

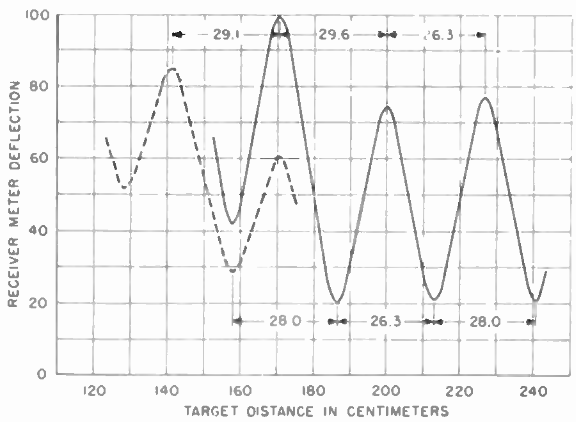


Fig. 4—Maximum and minimum deflections when moving a dipole along the track. The dotted curve is for 120 volts and the solid curve for 150 volts on the plate of the oscillating tube.

From Fig. 4, it is evident that the distances between adjacent maxima and minima show some irregularities. The average value, however, calculated from the sum of the last two maxima and minima, gives accurately the free-space half-wavelength used (276 mm).

Fig. 5 shows curves of the measured maximum deflections plotted against dipole lengths for different dipole thicknesses. The thinnest dipole gave the sharpest resonance curve, while the curve of the 40-mm dipole is so flat that the resonant length can hardly be determined from these measurements.

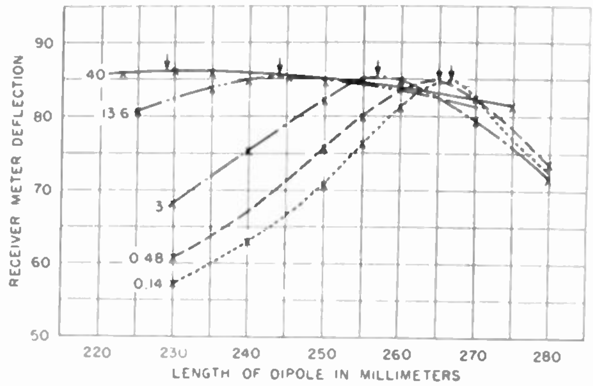


Fig. 5—Measured maximum deflections for dipoles of different lengths and diameters. The figures identifying the curves are the diameters of the dipoles in millimeters.

The target distances are plotted against dipole lengths in Fig. 6 for the five different diameters. On the co-ordinate scale, only target distances for the 0.14-mm dipoles are shown. For the other curves, only relative changes in distance are needed. The resonant lengths are marked with arrows. The inflection tangents are also drawn, giving a rough check of the correctness of the resonant

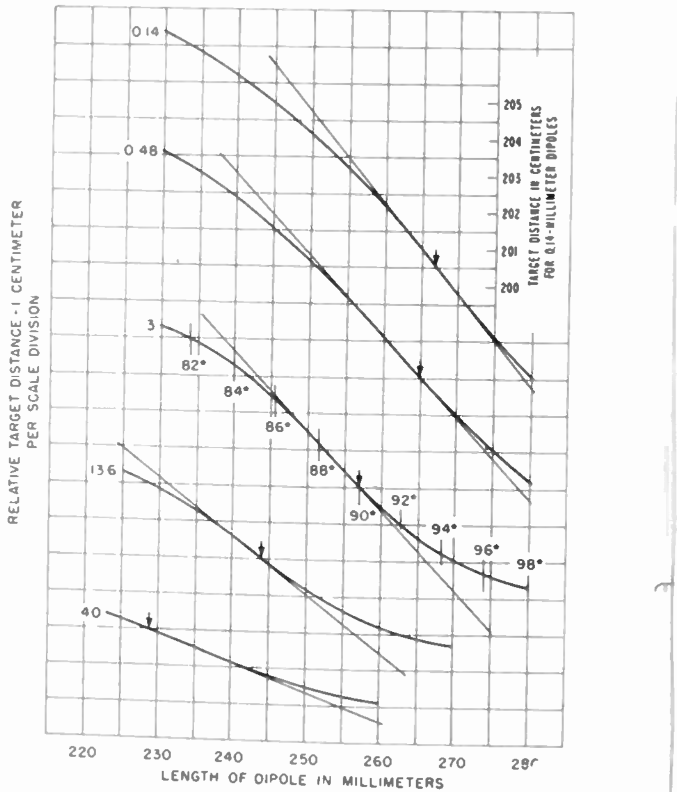


Fig. 6—The measured changes of target distances for dipoles of different lengths. The resonant lengths are marked with arrows. Actual distances are given for only one dipole thickness, 0.14 millimeter, the remaining curves being plotted against relative distances.

lengths. As may be seen, the angle of the tangents to the horizontal will be smaller with increasing dipole thickness.

## 3. Comparison of Reflected Power and Calculation of Radiation Resistance

To produce the highest useful deflection of the receiver meter, measurements were made at a target distance of 1.7 meters (67 inches) on the previously checked five resonant-length dipoles of different thicknesses.

The results are given in Table I. For the 0.48-, 3-, and 13.6-mm dipoles, the measured target distances are equal, which is a statistical proof of the correctness of the lengths. For the 40-mm dipole, the small difference in target distance was probably caused by the nonuniform current distribution around the surface of the dipole and for that reason no correction was made.

For the 0.14-mm dipole, the somewhat uncertain results of comparison were probably caused by the fact that a straight position of this very thin wire could not be secured. For this reason, it is not included in Table I.

For calculating the ratio of the reflected powers, the effect of the steady field must be eliminated. The corrected values were calculated by

$$\alpha_{cor} = (\sqrt{\alpha} - \sqrt{\alpha_0})^2, \tag{5}$$

in which $\alpha_0$ was 41 divisions, the value for the steady field. These corrected values are also given in Table I.

The loss resistance of the 0.48-mm dipole must not be neglected in these comparisons. Calculations showed that the reflected power decreases by about 2 per cent as a result of the loss resistance. Therefore, the ratios of radiation resistance calculated with (4) must be multiplied by 1.02. These ratios are included in Table I.

TABLE I

| | | | | |
|---|---|---|---|---|
| Diameter of dipole in millimeters | 0.48 | 3 | 13.6* | 40* |
| Diameter of dipole in inches | 0.019 | 0.118 | 0.535 | 1.575 |
| Resonant length in millimeters | 265 | 257 | 244 | 229 |
| Target distance in centimeters | 171.4 | 171.4 | 171.4 | 171.6 |
| Dipole length / Free-space half wavelength | 0.96 | 0.931 | 0.885 | 0.83 |
| Maximum receiver meter deflection | 96.5 | 97.5 | 98 | 98.5 |
| Corrected meter deflection, without steady field | 11.7 | 12.0 | 12.25 | 12.4 |
| Ratio of resistances | 1 | 0.935 | 0.828 | 0.72 |
| Radiation resistance based on 63 ohms for 3-millimeter dipole | 67.5 | 63 | 55.8 | 48.5 |

* With closed ends.

To obtain numerical values of radiation resistance, use was made of one of the values of radiation resistances, determined by the phase method, dealt with in Section 4. The 3-mm dipole was chosen as its radiation resistance of 63 ohms was one of the most reliable values found, being only slightly influenced by the limitations of the measurement methods. The radiation resistances for the other dipoles were calculated from this value and the ratios, and are given in the last line of the table.

## 4. Calculation of Antenna Impedance from Phase Measurements

To calculate antenna impedance from phase measurements, an impedance circle diagram was prepared in chart units[1] as shown in Fig. 7. This represents an en-

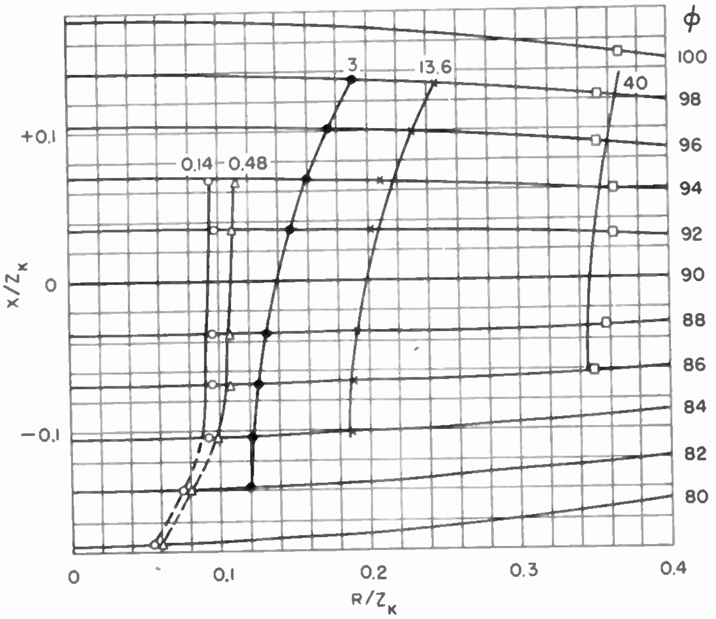


Fig. 7—Impedance locus curves plotted on an impedance circle diagram in chart units.

largement of only a small part of the usual diagram. The horizontal distances from the origin are equal to the resistive components, and the vertical distances to the reactive components of the impedances in chart units, i.e., the unit of the chart is the characteristic impedance of the line. The constant $\phi$ circles correspond to the lengths of the line in electrical degrees; $\phi = 90°$ represents the resonant length of a dipole.

As an example, let us take the diagram of the 3-mm dipole. In Fig. 6, the resonant length was 257 mm, which is equal to 90 electrical degrees. Thus 1° corresponds to 2.8 mm change in length. Fig. 6 gives values from 82 to 98° for this dipole. The vertical distances, measured from the target distance of the resonant-length dipole, are proportional to the phase angles of the impedances. Half the free-space wavelength, i.e., 276 mm, corresponds to 360°, thus 1 mm on the vertical scale is equal to 1.3° of phase angle. For $\phi = 96°$, $\Delta D = -23$ mm, which corresponds to $23 \times 1.3 = +30°$ of phase angle. If we draw a line with a slope of 30° from the origin, then the intersection with the $\phi = 96°$ circle will give one point of the impedance spiral diagram.

A number of points are marked on Fig. 7 for dipoles of different thicknesses. The irregularities in some points indicate the presence of some disturbing effects, which seem to be greatest for the thinnest dipoles and when they are tuned off resonance.

[1] Massachusetts Institute of Technology, Radar School Staff, "Principles of Radar," McGraw-Hill Book Co., New York, N. Y., 1946; chap. 8, p. 58.

The radiation resistance at resonant length is obtained at the intersection of the impedance locus curve and the $X=0$ abscissa. It is customary to call this reading the "standing-wave ratio." Multiplying this value by the calculated characteristic impedance of the resonant-length dipole, we get the radiation resistance at the resonant length. The results are given in Table II, where the calculated values of the characteristic impedance and the parameter $\Omega$ are also tabulated. The radiation resistances of the first two dipoles could not be determined with certainty.

TABLE II

| | | | | | |
|---|---|---|---|---|---|
| Diameter of dipole in millimeters | 0.14 | 0.48 | 3 | 13.6 | 40 |
| Diameter of dipole in inches | 0.006 | 0.019 | 0.118 | 0.535 | 1.575 |
| $\Omega = 2 \log_n (2l/a)$ | 16.2 | 13.8 | 10.2 | 7.15 | 4.87 |
| Characteristic impedance of resonant length dipole in $\Omega$, $Z_k$ | 830 | 678 | 456 | 269 | 130 |
| Standing-wave ratio from Fig. 7 | 0.092 | 0.107 | 0.138 | 0.2 | 0.35 |
| Radiation resistances at resonant lengths | 76* | 72.5* | 63 | 53.9 | 45.5 |
| Radiation resistances at resonant lengths without loss | 73.5* | 71.8* | 63 | 53.9 | 45.5 |

* The position of these curves was somewhat uncertain.

Comparing these values with those of Table I, it may be seen that the agreement is fair enough, although two quite different methods have been used. The somewhat greater values for the thickest dipoles, as measured by the amplitude method, could probably be explained by the small change in directional characteristics of the dipoles of shorter length.

Some data for dipoles of varying-length-to-thickness ratio, calculated by different theories and also measured by the described methods, are given in Table III.

TABLE III[2,3]

| $\Omega = 2 \log_n (2l/a)$ | 16.2 | 13.8 | 10.2 | 7.15 | 4.87 |
|---|---|---|---|---|---|
| Hallen-Bowkamp, 1st order | 63.1 | 61.54 | 58 | — | — |
| Hallen-Bowkamp, 2nd order | 68.7 | 66.5 | 60.4 | — | — |
| Gray, modified, 1st, 2nd order | 69.7 | 69.2 | 67.3 | — | — |
| King-Middleton, 1st order | 69.2 | 68.1 | 65 | — | — |
| King-Middleton, 2nd order | 70.85 | 70.82 | 70.77 | — | — |
| Schelkunoff | 66.7 | 65.2 | 61.4 | — | — |
| Measurements | | | | | |
| Amplitude | — | 67.5 | 63 | 55.8 | 48.5 |
| Phase | — | — | 63 | 53.9 | 45.5 |

[2] D. Middleton and R. King, "The thin cylindrical antenna: a comparison of theories," *Jour. Appl. Phys.*, vol. 17, pp. 273–284; April, 1946.

[3] S. A. Schelkunoff, "Electromagnetic Waves," fourth edition, D. Van Nostrand Company, New York, N. Y., October, 1945; p. 464.

For $\Omega = 10.2$, best agreement with measured values is obtained with calculations based on the Schelkunoff and King-Middleton first-order formulas.

For the same dipole, some other measured and calculated values may be compared. For the free-space half-wave dipole, the measured impedance values according to Fig. 7 are: $R=82$ ohms, $X=53$ ohms, and phase angle $=32.5°$. According to Schelkunoff, these values are: $R=75$ ohms, $X=45$ ohms, and phase angle $=31°$. The King-Middleton second order gives: $R=88$ ohms, $X=42.5$ ohms, and phase angle $=25.8°$. The measured resonant length of this dipole was less by 0.095 radian than $\pi/2$, if $\pi/2$ corresponds to the free-space half-wavelength. According to the Schelkunoff and King-Middleton second-order calculations, this value was 0.094 radian.

For $\Omega < 10$, unfortunately, no data were available for comparison.

Measurements made with open-ended tubes for the 13.6- and the 40-mm dipoles showed no practical differences in resonant lengths and in reflected powers.

### 4.1 Limitations

The slight effect of the carriage and probable small changes in earth reflections when moving the target may cause some irregularities in the steady field and in the meter deflections, especially for thin and detuned dipoles. This was substantiated by irregularities observed between adjacent maxima and minima when moving the target. Deviations in the impedance locus curves may result. Further, difficulties in obtaining accurately straight positions of thin wires made the equipment most suitable for determining the impedances of relatively thick dipoles. For the thickest dipoles, changes in target distance versus dipole length were small, and the accuracy of setting limited the results. With the amplitude method of comparing reflected powers, the effect of these factors was reduced.

During measurements, the anode voltage was maintained at $160 \pm 0.2$ volts, and the crystal detector proved to be very stable. The accuracy of the wavelength measurements was about $\pm 1$ mm.

## CORRECTION

The following errors have been brought to the attention of the editors by a communication from Leon Riebman, author of the paper, "Theory of the superregenerative amplifier," which appeared on pages 29–34 of the January, 1949, issue of the PROCEEDINGS OF THE I.R.E.

In equations (2) and (4) the limits on the integral $\epsilon^{1/2 \int_0^\tau P dt}$ should be $\tau$, not $r$.

On page 30, under definitions of symbols,

$t_Y$ (not $t_r$) = duration of decay phase

$t_Y + t_S = t_D$ (not $t_0$)

$G^R(t) = G(t_i, t_i + t_D)$, not $G(t_i t_i + t_D)$.

# Measured Impedance of Vertical Antennas Over Finite Ground Planes*

A. S. MEIER† AND W. P. SUMMERS‡, ASSOCIATE, IRE

*Summary*—**An investigation was made to obtain some fundamental information concerning the relation of the impedance of a vertical antenna over a finite ground plane as a function of the size and shape of the ground plane when dimensions of the ground plane are relatively small in terms of wavelength. It was found that the input impedance is a damped oscillating function of wavelength and ground-plane dimensions, the impedance of a circular ground plane varying from ±5 to ±20 per cent. Similar variations were observed on a square ground plane, which were approximately 50 per cent of those of the circular ground plane except when the dimensions of the ground plane were small. In general, it was found that the impedance is quite critical with respect to the size and shape of the ground plane, and relatively independent of the thickness of the antenna.**

**Measurements were made at microwaves by a modified Chipman method capable of measuring small differences in antenna impedance. Recent emphasis on improved microwave measurement techniques led to the investigation of the merits of the Chipman method as compared to those of the more conventional slotted-line standing-wave method. Although the two methods are shown to be closely related from a theoretical viewpoint, practical mechanical and electrical limitations are encountered which dictate preference of method entirely dependent on the application.**

## I. Introduction

In measuring experimentally the impedance of unbalanced antennas, it is the general practice, in many cases, to select arbitrarily a ground plane of convenient dimensions, the size and shape being dictated solely by practical considerations. A great many impedance problems involve unbalanced antenna systems for which practical ground planes cannot be constructed conveniently in excess of five or six wavelengths, and there is little information in the literature to guide the investigator in determining the relation between the antenna input impedance and ground-plane configuration. The investigation covered in Section II of this paper was undertaken with a view toward providing some much needed basic information on simple ground-plane structures to determine the antenna impedance as a function of the size and shape of the ground plane when the dimensions of the ground plane are relatively small in terms of wavelength.

The method of measurement is discussed in Section III. The equipment and techniques of measurement were developed for scale-model antenna measurements at microwaves, with particular emphasis on a high order of accuracy and a wide range of impedance.

* Decimal classification: R221×R244.21. Original manuscript received by the Institute, June 1, 1948; revised manuscript received, September 29, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 22, 1948. The basic development of microwave measuring equipment and techniques was undertaken at the direction of the Army Air Forces, Wright Field, Dayton, Ohio, under War Department Contract 33-038ac13493.

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## II. Antenna Impedance over Finite Ground Planes

Antenna impedance measurements were made on circular and square ground planes varying in dimensions from 1.7 to 6.1 wavelengths. Two antennas, a "thick" antenna 0.0258λ in diameter and a "thin" antenna 0.0060λ in diameter, were measured at a fixed frequency and were varied in length from 0.163λ to 0.529λ. The actual wavelength used was 12.28 cm. Details of the antenna and ground-plane mountings are shown in Fig. 1.

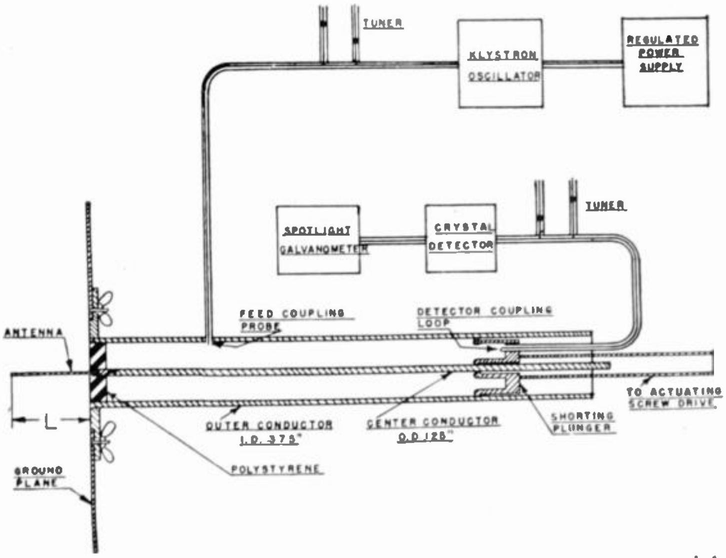


Fig. 1—Antenna and ground-plane mounting, Chipman coaxial measuring line and associated components.

A polystyrene bead was used to support the antenna, as shown in Fig. 1. In calculating the antenna impedance at the base of the antenna, the measured impedance was transformed through the polystyrene section which was treated as a uniform length of transmission line having a characteristic impedance of 41.5 ohms and an electrical length of 22.13°. The characteristic impedance of the measuring line was 66.0 ohms. Any error introduced by the polystyrene section does not necessarily invalidate results, since the basic ground-plane investigation involves only variations of impedance.

Measurements were made with the antenna and ground plane projecting through a second-floor window, with the ground plane located approximately two wavelengths beyond the outside wall of a brick building.

### *Circular Ground Planes*

Initial measurements were made with the 0.0060λ-diameter antenna on three circular ground planes of widely varying size in order to determine the approximate magnitude of impedance variation. The measured impedances are shown in Fig. 2. In Fig. 2 it will be

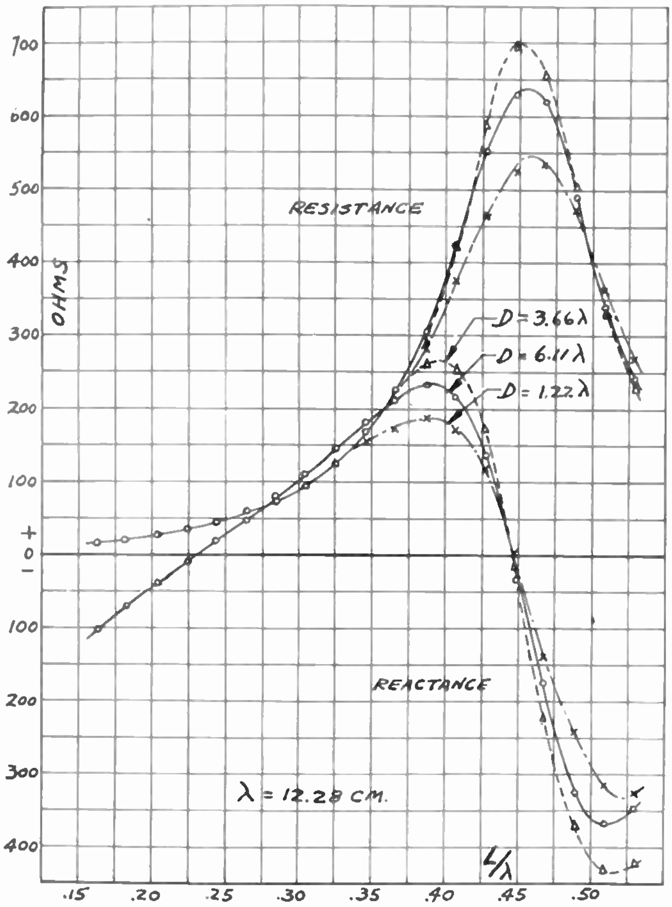


Fig. 2—Antenna impedance as a function of antenna length $L$ for three circular ground planes of varying diameter $D$.

noted that the impedance of the antenna of the 6.11λ-diameter ground plane falls within the values of the impedance of the 1.22λ- and 3.66λ-diameter ground planes which suggests, as would be expected, that the variation in impedance is an oscillating function of wavelength and ground-plane diameter. The magnitude of the impedance variation for the three arbitrarily chosen ground planes is somewhat larger than was first anticipated. It was observed that a considerable variation exists; the resonant resistance varies between 33 and 43 ohms, the antiresonant resistance varies between 545 and 700 ohms, and the maximum reactance values vary between +185 to +265 ohms and −335 to −435 ohms.

Figs. 3 and 4 show the results of measurements of the two antennas over ground planes varying from 1.7λ to 3.5λ in diameter. The results indicate comparable amplitudes of oscillation for both "thick" and "thin" antennas and a noticeable damping of oscillation as the ground plane increases in size. Rather large magnitudes of variation are evident which indicate that the antenna impedance is very critical with respect to the ground-plane dimensions.

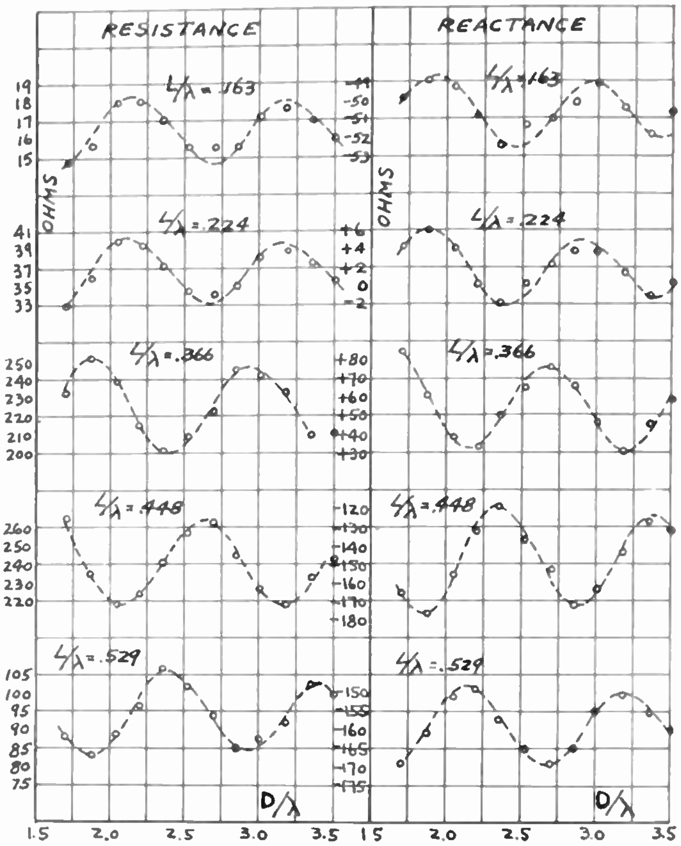


Fig. 3—Circular ground plane—Impedance of 0.0258λ-diameter antennas for ground-plane diameters $D$ varying from 1.7 to 3.5λ.

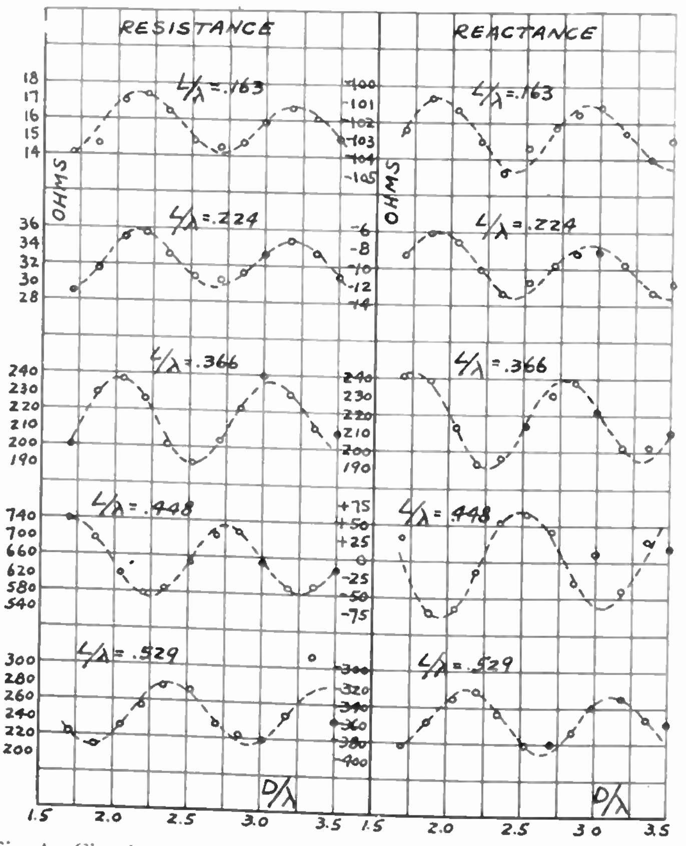


Fig. 4—Circular ground plane—Impedance of 0.0060λ-diameter antennas for ground-plane diameters $D$ varying from 1.7 to 3.5λ.

Measurements were next extended to a ground-plane diameter of six wavelengths in increments of 2 cm for both "thick" and "thin" antennas. Two antenna lengths were chosen which approximated resonant and anti-resonant antenna lengths. The results of these measurements are shown in Figs. 5 and 6. The variations in resistance amount to approximately ±5 to ±15 per cent and comparable variations in reactance exist except when the reactance approaches zero and the percentage variation thus becomes meaningless.

*Square Ground Planes*

Measurements of impedance were made on square ground planes, the same two antenna lengths of both "thick" and "thin" antennas being used so that a direct comparison could be made with the circular ground planes shown in Figs. 5 and 6. The results of the square

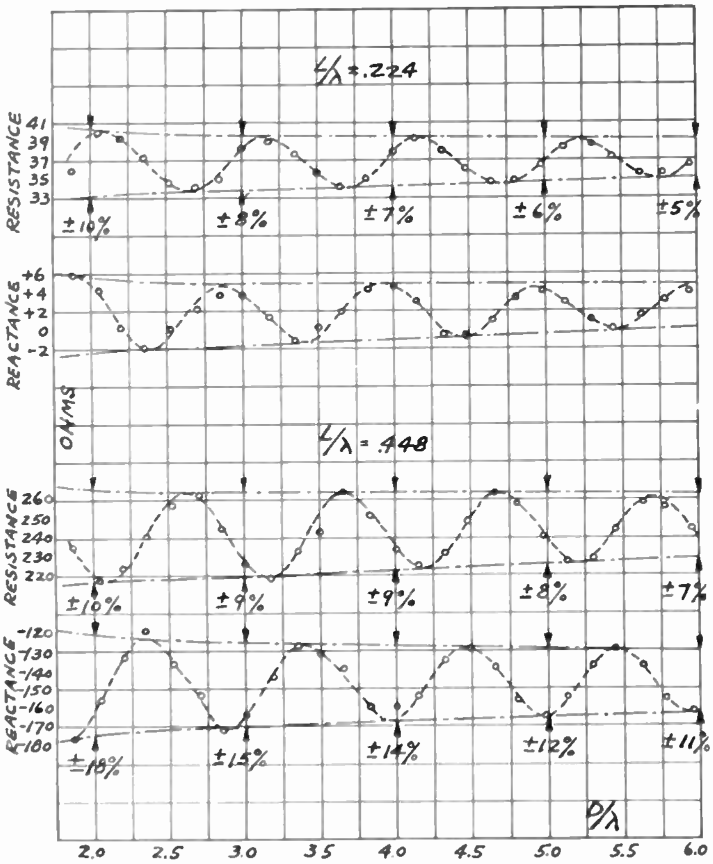


Fig. 5—Circular ground plane—Impedance of 0.0258λ-diameter antennas extended to 6λ-diameter ground plane.

ground-plane measurements are shown in Figs. 7 and 8. The results indicate comparable amplitudes of oscillation for both "thick" and "thin" antennas and a gradual damping of oscillation as the ground-plane increases in size. The larger variations in impedance occur for the smaller ground-plane sizes, increasing rapidly when the lateral dimensions of the ground planes are less than 2.5λ.

The amplitudes of oscillation of the square ground planes are considerably less than those of the circular

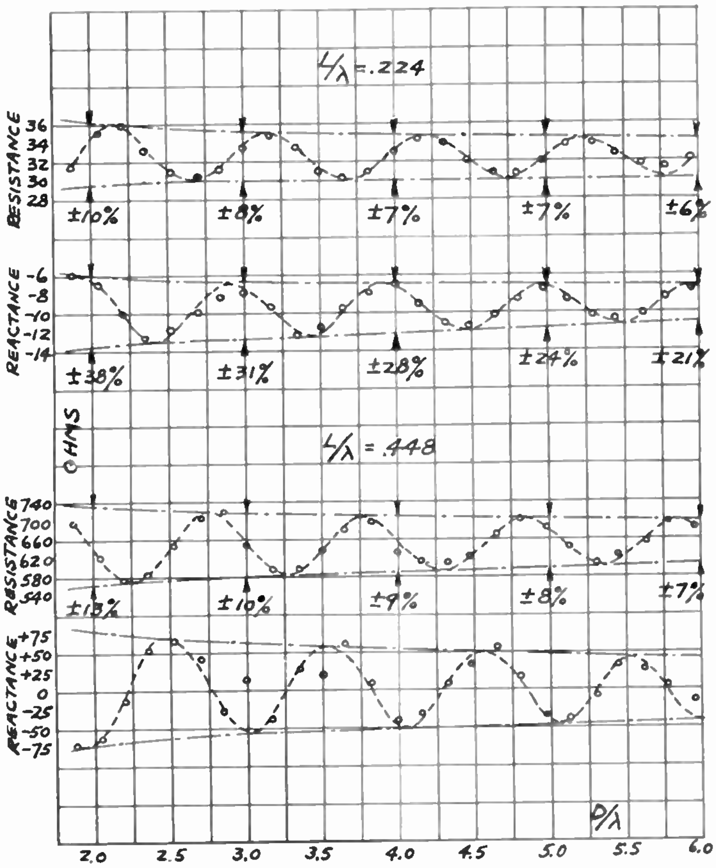


Fig. 6—Circular ground plane—Impedance of 0.0060λ-diameter antennas extended to 6λ-diameter ground plane.

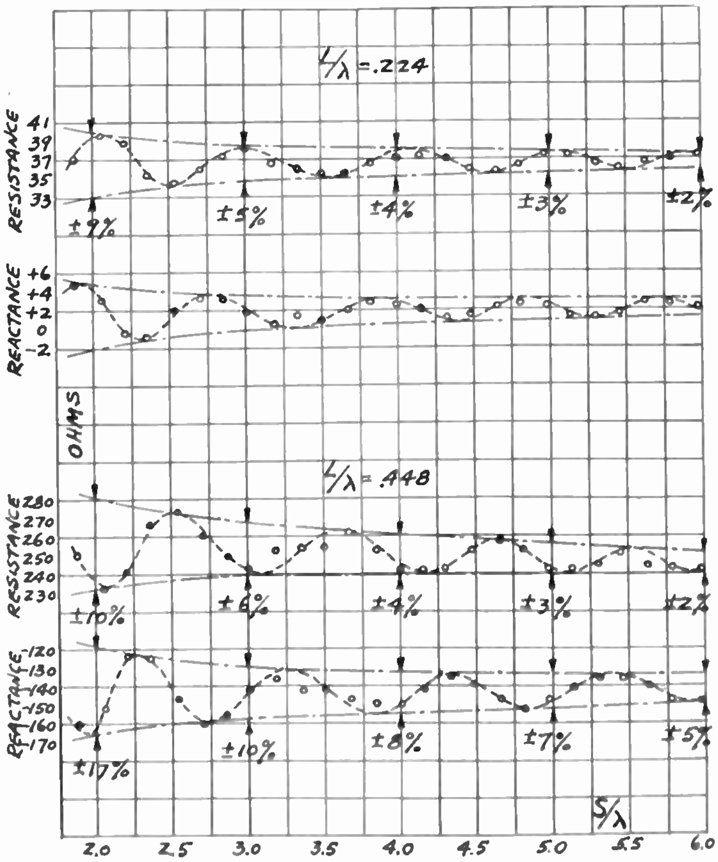


Fig. 7—Square ground plane—Impedance of 0.0258λ-diameter antennas as a function of ground-plane side dimension *S*.

ground planes, and the damping of oscillation is more rapid. In general, the variations in impedance are approximately 50 per cent of the variations of the circular ground planes except when the dimensions of the ground planes are small.

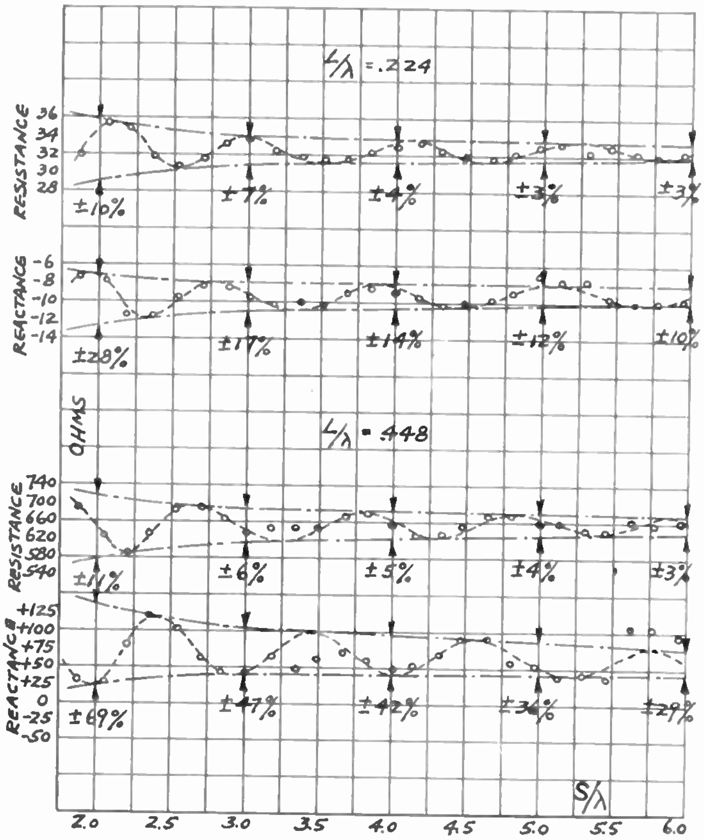


Fig. 8—Square ground plane—Impedance of 0.0060λ-diameter antennas as a function of ground-plane side dimension $S$.

Difficulty was experienced with the measuring equipment when the square ground-plane measurements were taken. Deterioration of contact occurred due to excessive wear of the fingers of the movable shorting plunger. Measurements were further complicated at this point by the oscillator tube giving out, causing fluctuations in the output voltage. Measurements of the square ground planes were repeated using a new oscillator tube which was adjusted as close to the original frequency as the cavity adjustment permitted, and temporary repairs were made to the contact fingers in order to complete the investigation. After repeating data, it was found that the contact difficulty had not been entirely corrected, so that the curves of Figs. 7 and 8 were finally plotted from composite data of the two sets of measurements. This accounts for some of the irregularities in the data.

## III. Method of Measurement

A modified Chipman[1] method was used to measure the impedance of the antennas discussed in Section II. The Chipman methods has not been used extensively at microwaves, so that a discussion of the application is given in the following analysis, which differs from the original Chipman paper in that the derivation is made to afford a direct comparison with the more conventional standing-wave slotted-line method.

Fig. 1 shows the arrangement of the measuring system. The measuring line consists of a coaxial line fitted with an adjustable shorting plunger containing a very small detecting loop. The line is excited near the load end of the line by means of a voltage probe coupled to the generator. The actuating screw which drives the plunger is connected to a vernier system graduated to 1/10,000 cm. As in the standing-wave slotted-line method, two quantities determine the unknown load impedance: (1) the magnitude of the "standing-wave ratio" of current through the shorting plunger as the electrical length of the measuring line is changed by movement of the plunger, and (2) the position of this simulated "standing-wave" on the transmission line.

### Chipman Method

Consideration is given to the ideal case, in which it is assumed that the measuring line is lossless and the generator impedance is zero.

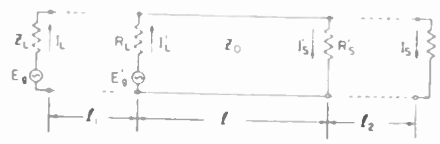


Fig. 9—Generalized transmission line for computing the equivalent standing wave due to the load and plunger impedance, Chipman method.

Referring to Fig. 9, the complex load impedance $Z_L$ is transformed to a real impedance $R_L'$ through an appropriate length of transmission line $l_1$, and, similarly, the shorting plunger and receiver impedance $Z_s$ is transformed to a real impedance $R_s'$ through a corresponding length of line $l_2$.

Considering the voltage relation in terms of real impedances,

$$E_g' = I_L'R_L' + I_s'[R_s'\cos\beta l + jZ_0\sin\beta l] \quad (1)$$

$$E_g' = \frac{I_s'}{Z_0}[Z_0(R_L' + R_s')\cos\beta l + j(Z_0^2 + R_L'R_s')\sin\beta l]. \quad (2)$$

Let

$$\frac{R_L'}{Z_0} = \frac{1}{\rho_L} \quad \text{and} \quad \frac{R_s'}{Z_0} = \frac{1}{\rho_s},$$

$\rho_L$ representing the mismatch between the load impedance and the transmission line alone and $\rho_s$ that due to the mismatch of the plunger impedance and line.

[1] R. A. Chipman, "A resonance curve method for the absolute measurement of impedance at frequencies of the order of 300 Mc," *Jour. Appl. Phys.*, vol. 10, pp. 27–38; January, 1939.

Then

$$|I_s'|^2 = \frac{|E_g'|^2}{Z_0^2\left[\left(\frac{1}{\rho_L}+\frac{1}{\rho_s}\right)^2+\left(1-\frac{1}{\rho_L^2}\right)\left(1-\frac{1}{\rho_s^2}\right)\sin^2\beta l\right]}. \tag{3}$$

From (3), when $\rho_L > 1$ and $\rho_s > 1$, $|I_s'|^2$ is maximum when $\sin^2\beta l = 0$ (i.e., when $l = n\lambda/2$), and $|I_s'|^2$ is minimum when $\sin^2\beta l = 1$ (i.e., when $l = (2n-1)\lambda/4$).

$$\frac{|I_s'|_{\max}}{|I_s'|_{\min}} = \rho_0 = \frac{\rho_L\rho_s + 1}{\rho_L + \rho_s} \tag{4}$$

where $\rho_0$ is the apparent "standing-wave ratio" of current measured directly on the line. From (4),

$$\rho_L = \frac{\rho_0\rho_s - 1}{\rho_s - \rho_0} = \frac{Z_0}{R_L'}. \tag{5}$$

The terms $\rho_L$ and $\rho_s$ represent standing-wave ratios associated with the load- and plunger-impedance mismatches to the transmission line, respectively.

It will be noted that (4), expressed in terms of reflection coefficients, reduces to

$$|K_0| = |K_L| \cdot |K_s|$$

where

$$|K| = \frac{\rho - 1}{\rho + 1}. \tag{6}$$

Equation (4), in terms of reflection coefficients, corresponds to Chipman's original derivation:

$$\frac{|I_s'|_{\max}}{|I_s'|_{\min}} = \rho_0 = \frac{1 + |K_L| \cdot |K_s|}{1 - |K_L| \cdot |K_s|}. \tag{7}$$

Determination of high standing-wave ratios by direct measurement of current maxima and minima as in (4) and (7) is usually limited by the range of the galvanometer, especially when a square-law detector is used. High standing-wave ratios, however, may be measured more accurately by determining the current distribution near a maximum-current position.

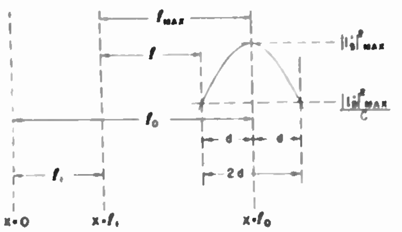


Fig. 10—Configuration for determination of the equivalent standing-wave ratio by determining the current distribution near a current maximum position, Chipman method.

Referring to Fig. 10, $|I_s'|_{\max}^2$ will be reduced to $|I_s'|_{\max}^2/C$ by moving the shorting plunger a distance $d$ from the current maximum position. Then, from (3),

$$\frac{|I_s'|_{\max}^2}{|I_s'|^2} = C$$

$$= \frac{\left(\frac{1}{\rho_L}+\frac{1}{\rho_s}\right)^2+\left(1-\frac{1}{\rho_L^2}\right)\left(1-\frac{1}{\rho_s^2}\right)\sin^2\beta l}{\left(\frac{1}{\rho_L}+\frac{1}{\rho_s}\right)^2}. \tag{8}$$

Then,

$$(C-1)(\rho_L+\rho_s)^2 = (\rho_L^2-1)(\rho_s^2-1)\sin^2\beta l. \tag{9}$$

Solving for $\rho_L$ in (9),

$$\rho_L = \frac{\rho_0\rho_s - 1}{\rho_s - \rho_0}$$

where

$$\rho_0 = \sqrt{\frac{C-1}{\sin^2\beta l}+1}. \tag{10}$$

Since

$$l = l_{\max} - d = n\lambda/2 - d,$$

$$\rho_0 = \sqrt{\frac{2(C-1)}{1-\cos 2\beta d}+1}. \tag{11}$$

Using the transmission-line equation to find the unknown load impedance $Z_L$,

$$\frac{1}{\rho_L} = \frac{R_L'}{Z_0} = \frac{Z_L + jZ_0\tan\beta l_1}{Z_0 + jZ_L\tan\beta l_1}. \tag{12}$$

From Fig. 10.

$$\tan\beta l_1 = \tan\beta(l_0 - l_{\max}) = \tan\beta l_0. \tag{13}$$

Solving for $Z_L$ from (12) and (13),

$$Z_L = R_L + jX_L = \frac{\rho_L Z_0(1+\tan^2\beta l_0)}{\rho_L^2+\tan^2\beta l_0} + j\frac{(1-\rho_L^2)Z_0\tan\beta l_0}{\rho_L^2+\tan^2\beta l_0}. \tag{14}$$

The quantity $\rho_s$ maybe determined from short-circuit measurements by use of (4), and once $\rho_s$ and $\rho_0$ are known, (14) may be directly applied to determine the unknown load impedance $Z_L$.

### *Standing-Wave Method*

The standing-wave slotted-line method of impedance measurement consists of a fixed length of transmission line terminated in a load impedance $Z_L$, the generator being tightly coupled to the line. A traveling detector measures the voltage or current distribution along the line from which the unknown impedance $Z_L$ is calculated.[2-4]

The complex load impedance $Z_L$ of Fig. 11 is transformed to a real impedance $R_L'$ through an appropriate length of transmission line $l_1$. Considering, then, the voltage relationship in terms of real impedances,

$$E = I_L'[R_L' \cos \beta l + jZ_0 \sin \beta l]. \qquad (15)$$

Then,

$$|E|^2 = |I_L'|^2[R_L'^2 \cos^2 \beta l + Z_0^2 \sin^2 \beta l]. \qquad (16)$$

Let

$$\frac{R_L'}{Z_0} = \frac{1}{\rho_L}.$$

Then

$$|E|^2 = |I_L'|^2 Z_0^2\left[\frac{1}{\rho_L^2} + \left(1 - \frac{1}{\rho_L^2}\right)\sin^2 \beta l\right]. \qquad (17)$$

Thus, when $\rho_L > 1$, $|E|^2$ is maximum when $\sin^2 \beta l = 1$, (i.e., when $l = l_{min} \pm \lambda/4 = (2n-1)\lambda/4$), and $|E|^2$ is minimum when $\sin^2 \beta l = 0$ (i.e., when $l = l_{min} = n\lambda/2$).

| | Lossless Case | Attenuated Case |
|---|---|---|
| 1. Measured $SWR$ | $\rho_0 = \sqrt{\frac{2(C-1)}{1-\cos 2\beta d} + 1}$ | $\rho_a \cong \sqrt{\frac{2(C-1)}{1-\cos 2\beta d} + 1}$ |
| 2. $SWR$ corrected for Attenuation | . . . . . . . . . . . . | $\rho_0 \cong \frac{\rho_a - \alpha l_0}{1 - \rho_a \alpha l_0}$ |

Then

$$\frac{|E|_{max}}{|E|_{min}} = \rho_L = \frac{Z_0}{R_L'}. \qquad (18)$$

As before, high standing-wave ratios may be readily measured by determining the voltage distribution near a voltage minimum instead of direct measurement of voltage maxima and minima as in (18). Referring to Fig. 11, $|E|_{min}^2$ will be increased to $C|E|_{min}^2$ by moving the traveling detector a distance $d$ from a voltage minimum position. Then

$$\frac{|E|^2}{|E|_{min}^2} = C = (\rho_L^2 - 1)\sin^2 \beta l + 1. \qquad (19)$$

Since $l = l_{min} - d = n\lambda/2 - d$, (19) reduces to

$$\rho_L = \sqrt{\frac{2(C-1)}{1 - \cos 2\beta d} + 1}. \qquad (20)$$

Equations (12) through (14) are directly applicable for calculating the unknown impedance $Z_L$, once $\rho_L$ and $l_0$ are known.

It will be noted that the Chipman and standing-wave methods are theoretically similar except for the introduction of the plunger impedance in the circuit. For low standing-wave ratios (when $\rho_s \gg \rho_L$) the plunger impedance has negligible effect as shown in Fig. 12 and the determination of impedance of the unknown, when assuming $\rho_s = \infty$, is exactly that of the standing-wave method.

### *Effect of Attenuation*

In the previous analysis, it is assumed that the measuring line is lossless. In a very few instances, it may be correctly assumed that the errors due to attenuation are negligible and in most cases, especially at microwaves and even at very low frequencies where high standing-wave ratios are encountered, attenuation corrections are of prime importance for accurate measurements. Simple corrections for attenuation may be made by correcting the directly measured values by the following relations:

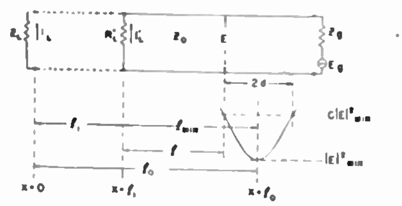

Fig. 11—Generalized transmission line for computing the standing-wave ratio by measurements about a voltage minimum, standing-wave method.

The relative magnitudes of the attenuation correction are shown in Fig. 12. The above formulas may be rigorously developed by assuming $\tanh \alpha l \cong \alpha l$ and $\alpha d \cong 0$, which are well within the experimental error of a well-designed measuring system. The term $\alpha$ is defined by the usual relation $\gamma = \alpha + j\beta$ where $\gamma$ is the propagation constant and $\alpha$ and $\beta$ the attenuation and phase constants of the measuring line.

In the Chipman method, the measured plunger-impedance mismatch must be corrected for attenuation. Although $\alpha$ may be determined within an experimental error of 5 per cent, applying this correction to the

[2] 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., 1945; pp. 36–41.
[3] F. J. Gaffney, "Microwave measurements and test equipments," Proc. I.R.E., vol. 34, pp. 775–780; October, 1946.
[4] R. F. Lewis, "Measurements in the uhf spectrum," *Electronics*, vol. 15, pp. 63–68; April, 1942.

plunger-impedance mismatch $\rho_s$ results in a much larger error in $\rho_s$. For the measuring line used in Section II, the value of $\rho_s$ corrected for attenuation was found to be of the order of 600 to 1 within an error of approximately 20 per cent. However, when $\rho_s \gg \rho_L$ as in the case of the antennas measured in Section II, very accurate measurements are obtained in spite of the apparently large error in $\rho_s$.

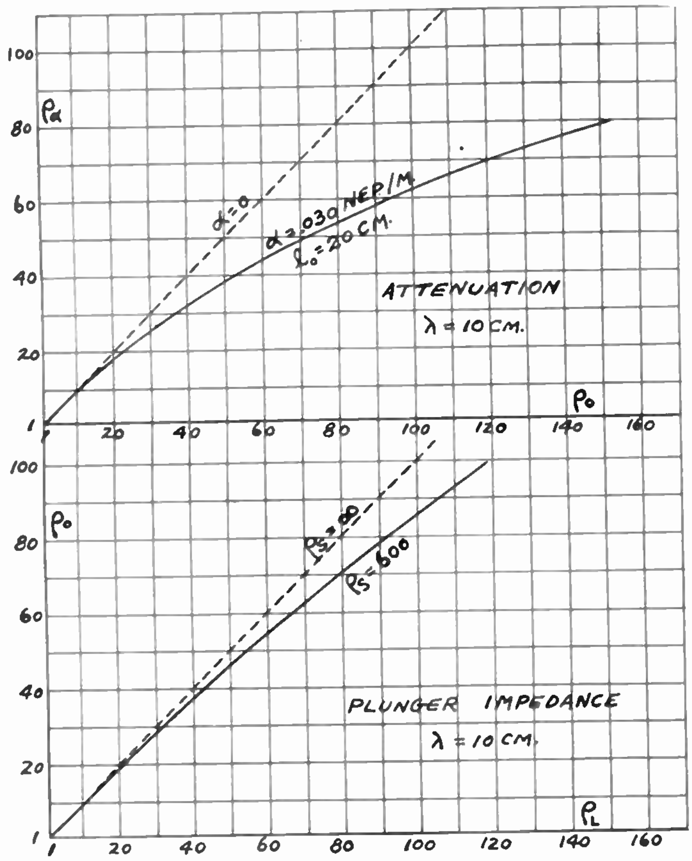


Fig. 12—Effects of attenuation and plunger impedance, Chipman method.

Careful consideration of the attenuation effect is essential in the calibration of the detecting system. Usually, a direct calibration is made on short circuit which is accurate only within certain limitations. In the Chipman method, a direct calibration may be made providing $\sin \beta l \gg 1/\rho_s + \alpha l$. In the measuring line used in Section II, a direct calibration can be made only in the range $14° < \beta l < 166°$ in order to satisfy this relation without exceeding an error of 1/10 per cent in power. Thus, it is essential to make a direct calibration more than 14° away from a current maximum position. The same condition applies to the standing-wave method for the condition $\sin \beta l \gg \alpha l$.

### *Effect of Coupling Impedances*

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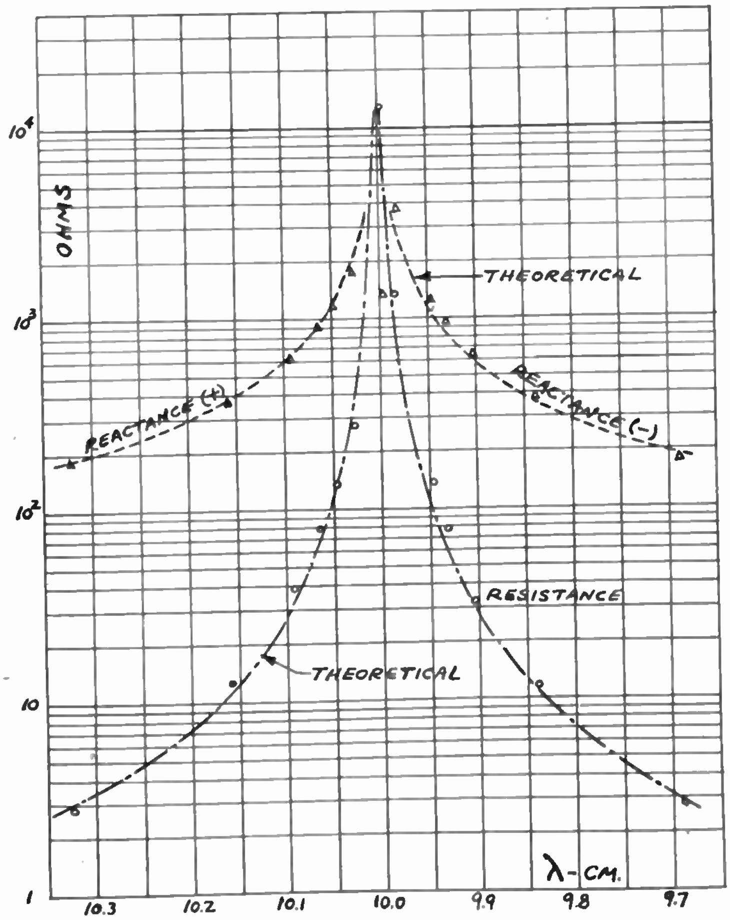


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A check of the accuracy of the measuring system used in Section II was made by measuring the apparent input impedance of a $1\frac{3}{4}$-wavelength short-circuited transmission line. The input impedance is of such magnitude that standing-wave ratios of 180 to 1 are involved. The probe was located at a distance two wavelengths from the short circuit, at a *voltage minimum* point, and adequate excitation was obtained with a small penetration when using a klystron oscillator with an output power of the order of two watts. The results of these measurements shown in Fig. 13 indicate the validity of neglecting the probe impedance for standing-wave ratios of practical significance. These measurements also show that the measuring system is capable of a relatively high order of accuracy in terms of absolute measurement of impedance.

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The previous analysis indicated that the Chipman and standing-wave methods of impedance measurement are comparable from a theoretical standpoint. Physical realization of the measuring systems bring about practical electrical and mechanical limitations which dictate preference of method entirely dependent upon the particular application.

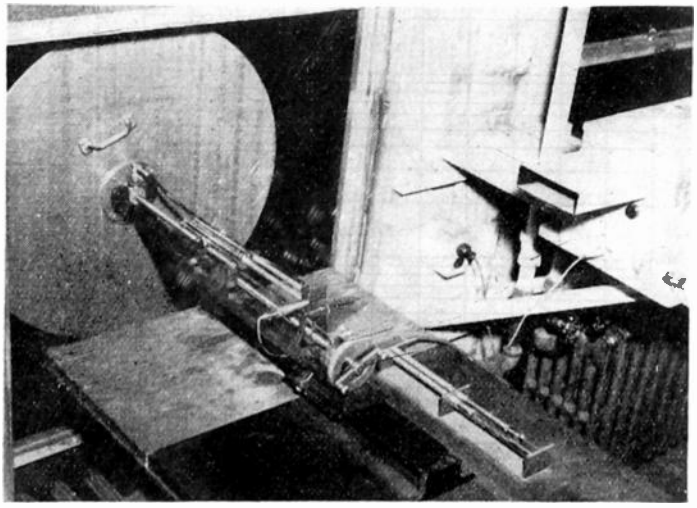

Fig. 14—Photograph of Chipman measuring line and associated components.

In the standing-wave method, very close machining tolerances must be maintained in the cross-sectional dimensions in both conductor and probe spacings. If variations exist in the spacing of the traveling probe with respect to either conductor of the measuring line, random variations in measured voltage will occur which cannot be compensated for by any simple procedure. This necessitates extremely accurate machining, which is difficult and expensive. (Fig. 14.)

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As previously pointed out, accurate measurements by the Chipman method may be made when $\rho_s \gg \rho_L$, $\rho_s$ and $\rho_L$ being the standing-wave ratios associated with the plunger- and load-impedance mismatches, respectively. With the present measuring line operating at a wavelength of 10 cm, $\rho_s \cong 600$, which indicates that very good accuracy can be obtained when $\rho_L < 50$, i.e., when $\rho_L$ is relatively small compared to $\rho_s$. Fairly good measurements may be made to values of $\rho_L = 200$, as shown by the results of Fig. 13, but poor results on an absolute basis are obtained for higher values when $\rho_L$ and $\rho_s$ are comparable in magnitude. This indicates that the Chipman method is not particularly adaptable to the measurement of extremely high standing-wave ratios.

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# Some Properties of Radiation from Rectangular Waveguides*

J. T. BOLLJAHN†, ASSOCIATE, IRE

*Summary*—**Certain exact relationships between the radiation pattern and impedance characteristics of a radiating rectangular waveguide with vanishingly thin walls are developed. In particular, it is shown that the ratio of radiation intensities in certain preferred directions and the power gain in these directions are related in a simple manner to the reflection coefficient inside the waveguide. Although the information obtained is restricted in the sense that it applies to only a few discrete points on the radiation pattern, it is quite general as regards the manner in which the waveguide is broken or perforated to allow radiation. The results are shown to apply equally well if conducting sheets, having arbitrary shapes but lying on specified planes, are present in the vicinity of the radiating waveguide.**

## I. INTRODUCTION

IT IS POSSIBLE to write various free-space solutions to Maxwell's equations in which there appear "null planes" or planes having no tangential component of electric field. It is readily shown that thin conducting sheets may be inserted into the null planes of such a field without affecting the validity of the solution. The conducting sheets may, of course, coincide with all or any parts of the null planes. It follows that many well-known solutions to simple radiation problems are, at the same time, solutions to a variety of diffraction problems involving conducting sheets whose dimensions are completely arbitrary.

The problems which may be solved by simply inserting conducting sheets into the null planes of a known field are usually of little importance in themselves. In some cases, such as those discussed below, however, solutions obtained in this fashion may be used in conjunction with the principles of superposition and reciprocity to obtain more information than can be extracted from these general principles alone.

## II. CONSTRUCTION OF THE FIELDS FOR THE WAVEGUIDE PROBLEM

Consider two dipole sources $a$ and $b$ (Fig. 1) which are polarized in the $z$ direction and which have their centers in the plane $z=0$ at points symmetrical about the $y$ axis. If the distance $r$ is large compared with the wavelength, the fields in the vicinity of the origin due to $a$ and $b$ may be approximated by plane waves with directions of propagation making angles of $\pi+\theta_0$ and $\pi-\theta_0$ with the $y$ axis, respectively. If the two sources are fed with signals of equal amplitude and phase, the resultant electric field in the vicinity of the origin calculated on the basis of the plane-wave approximation is readily shown to be

$$E_z = 2E_0 \cos(kx \sin\theta_0) e^{jky\cos\theta_0} \tag{1}$$

where the time factor $e^{j\omega t}$ has been omitted, $k=2\pi/\lambda$, and $E_0$ is the rms amplitude of the field at the origin due to either source (rationalized mks units are employed throughout the discussion). We will be concerned with the case $r=\infty$, so that the plane-wave approximation becomes exact, and (1) is valid for all finite values of $x$, $y$, and $z$.

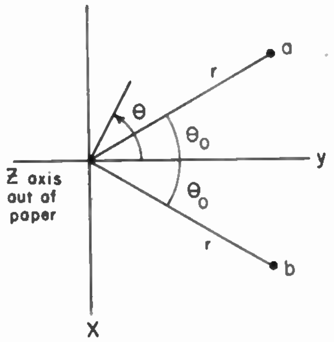


Fig. 1

It is seen that the electric field vanishes along planes which are parallel to the plane $x=0$ and which occur at values of $x=\pm n\lambda/4 \sin\theta_0$, where $n$ may have odd integral values. In addition, it is noted that any plane parallel to the plane $z=0$ has no tangential component of electric field. Thin conducting sheets may now be located so that they coincide with parts of these null planes without changing the solution given by (1).

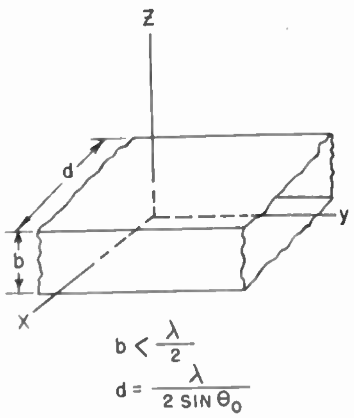


Fig. 2

We will consider in detail a configuration of conducting sheets which consists in part of a rectangular waveguide whose walls fit into the field as specified (Fig. 2). Only the case in which the waveguide is cut off for all modes except the $TE_{01}$ will be considered. This restric-

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being tightly coupled to the line. A traveling detector measures the voltage or current distribution along the line from which the unknown impedance $Z_L$ is calculated.[2-4]

The complex load impedance $Z_L$ of Fig. 11 is transformed to a real impedance $R_L'$ through an appropriate length of transmission line $l_1$. Considering, then, the voltage relationship in terms of real impedances,

$$E = I_L'[R_L' \cos \beta l + jZ_0 \sin \beta l]. \qquad (15)$$

Then,

$$|E|^2 = |I_L'|^2[R_L^{2\prime} \cos^2 \beta l + Z_0^2 \sin^2 \beta l]. \qquad (16)$$

Let

$$\frac{R_L'}{Z_0} = \frac{1}{\rho_L}.$$

Then

$$|E|^2 = |I_L'|^2 Z_0^2\left[\frac{1}{\rho_L^2} + \left(1 - \frac{1}{\rho_L^2}\right)\sin^2 \beta l\right]. \qquad (17)$$

Thus, when $\rho_L > 1$, $|E|^2$ is maximum when $\sin^2 \beta l = 1$, (i.e., when $l = l_{\min} \pm \lambda/4 = (2n-1)\lambda/4$), and $|E|^2$ is minimum when $\sin^2 \beta l = 0$ (i.e., when $l = l_{\min} = n\lambda/2$).

| | Lossless Case | Attenuated Case |
|---|---|---|
| 1. Measured *SWR* | $\rho_0 = \sqrt{\dfrac{2(C-1)}{1-\cos 2\beta d} + 1}$ | $\rho_a \cong \sqrt{\dfrac{2(C-1)}{1-\cos 2\beta d} + 1}$ |
| 2. *SWR* corrected for Attenuation | . . . . . . . . . . . . . | $\rho_0 \cong \dfrac{\rho_a - \alpha l_0}{1 - \rho_a \alpha l_0}$ |

Then

$$\frac{|E|_{\max}}{|E|_{\min}} = \rho_L = \frac{Z_0}{R_L'}. \qquad (18)$$

As before, high standing-wave ratios may be readily measured by determining the voltage distribution near a voltage minimum instead of direct measurement of voltage maxima and minima as in (18). Referring to Fig. 11, $|E|_{\min}^2$ will be increased to $C|E|_{\min}^2$ by moving the traveling detector a distance $d$ from a voltage minimum position. Then

$$\frac{|E|^2}{|E|_{\min}^2} = C = (\rho_L^2 - 1)\sin^2 \beta l + 1. \qquad (19)$$

Since $l = l_{\min} - d = n\lambda/2 - d$, (19) reduces to

$$\rho_L = \sqrt{\frac{2(C-1)}{1-\cos 2\beta d} + 1}. \qquad (20)$$

Equations (12) through (14) are directly applicable for calculating the unknown impedance $Z_L$, once $\rho_L$ and $l_0$ are known.

It will be noted that the Chipman and standing-wave methods are theoretically similar except for the introduction of the plunger impedance in the circuit. For low standing-wave ratios (when $\rho_s \gg \rho_L$) the plunger impedance has negligible effect as shown in Fig. 12 and the determination of impedance of the unknown, when assuming $\rho_s = \infty$, is exactly that of the standing-wave method.

### *Effect of Attenuation*

In the previous analysis, it is assumed that the measuring line is lossless. In a very few instances, it may be correctly assumed that the errors due to attenuation are negligible and in most cases, especially at microwaves and even at very low frequencies where high standing-wave ratios are encountered, attenuation corrections are of prime importance for accurate measurements. Simple corrections for attenuation may be made by correcting the directly measured values by the following relations:

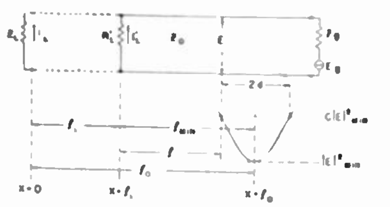

Fig. 11—Generalized transmission line for computing the standing-wave ratio by measurements about a voltage minimum, standing-wave method.

The relative magnitudes of the attenuation correction are shown in Fig. 12. The above formulas may be rigorously developed by assuming $\tanh \alpha l \cong \alpha l$ and $\alpha d \cong 0$, which are well within the experimental error of a well-designed measuring system. The term $\alpha$ is defined by the usual relation $\gamma = \alpha + j\beta$ where $\gamma$ is the propagation constant and $\alpha$ and $\beta$ the attenuation and phase constants of the measuring line.

In the Chipman method, the measured plunger-impedance mismatch must be corrected for attenuation. Although $\alpha$ may be determined within an experimental error of 5 per cent, applying this correction to the

[2] 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., 1945; pp. 36–41.
[3] F. J. Gaffney, "Microwave measurements and test equipments," Proc. I.R.E., vol. 34, pp. 775–780; October, 1946.
[4] R. F. Lewis, "Measurements in the uhf spectrum," *Electronics*, vol. 15, pp. 63–68; April, 1942.

plunger-impedance mismatch $\rho_s$ results in a much larger error in $\rho_s$. For the measuring line used in Section II, the value of $\rho_s$ corrected for attenuation was found to be of the order of 600 to 1 within an error of approximately 20 per cent. However, when $\rho_s \gg \rho_L$ as in the case of the antennas measured in Section II, very accurate measurements are obtained in spite of the apparently large error in $\rho_s$.

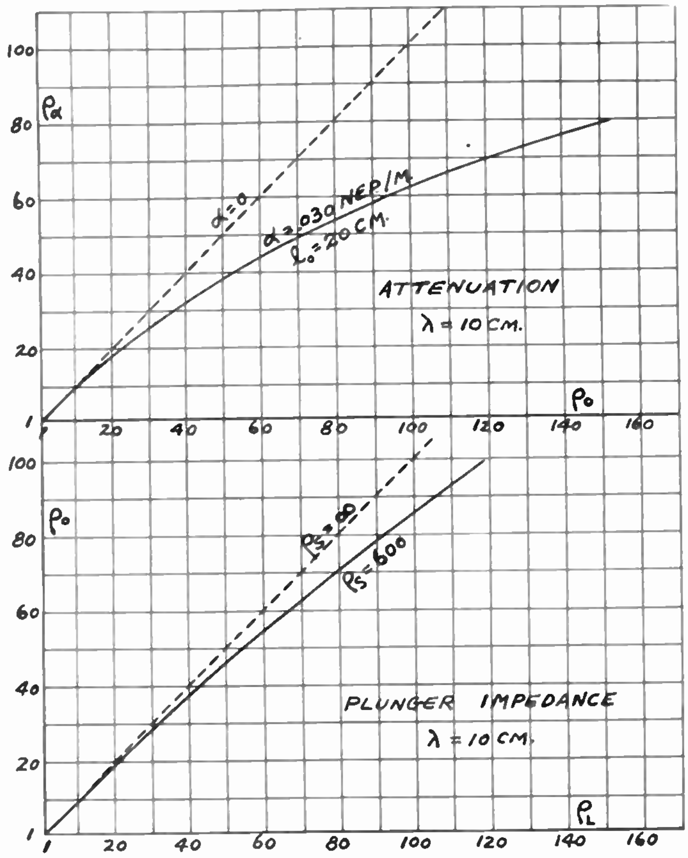


Fig. 12—Effects of attenuation and plunger impedance, Chipman method.

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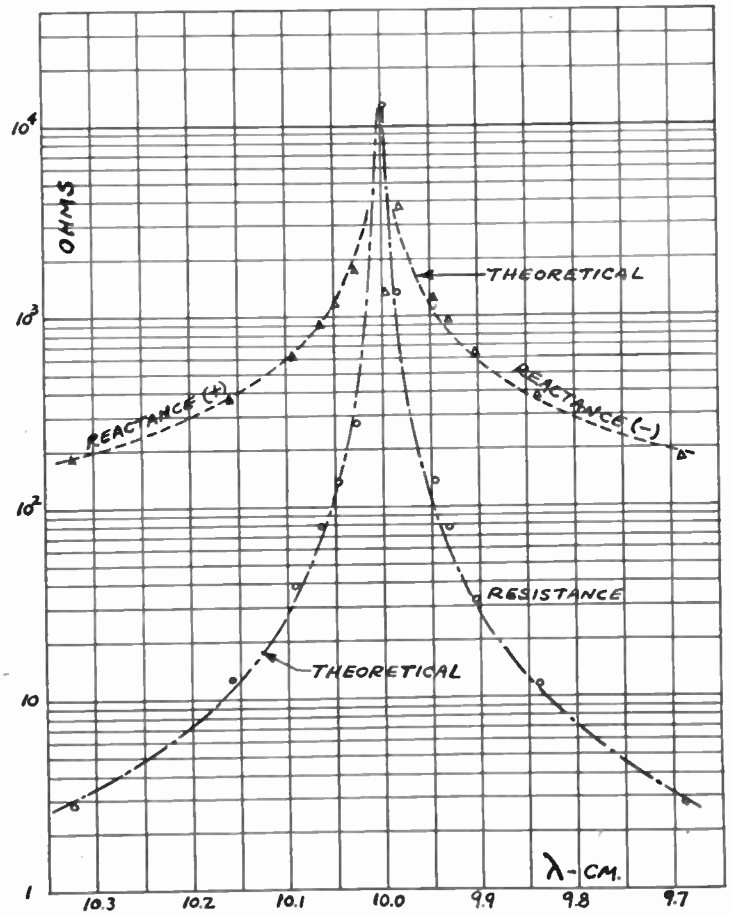


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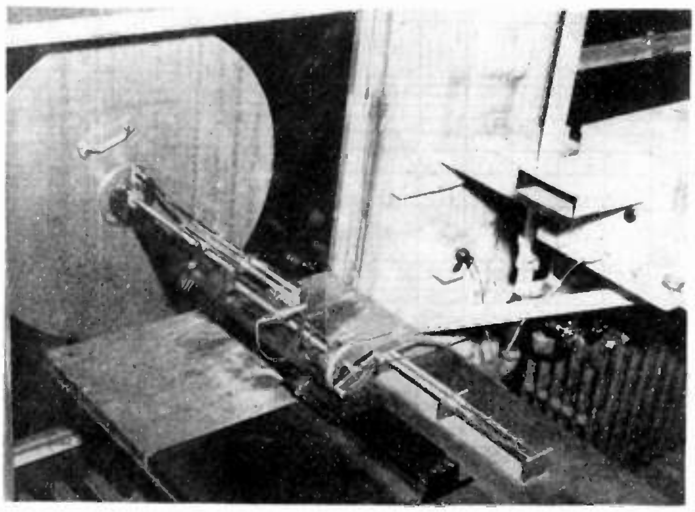

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## II. Construction of the Fields for the Waveguide Problem

Consider two dipole sources $a$ and $b$ (Fig. 1) which are polarized in the $z$ direction and which have their centers in the plane $z=0$ at points symmetrical about the $y$ axis. If the distance $r$ is large compared with the wavelength, the fields in the vicinity of the origin due to $a$ and $b$ may be approximated by plane waves with directions of propagation making angles of $\pi+\theta_0$ and $\pi-\theta_0$ with the $y$ axis, respectively. If the two sources are fed with signals of equal amplitude and phase, the resultant electric field in the vicinity of the origin calculated on the basis of the plane-wave approximation is readily shown to be

$$E_z = 2E_0 \cos (kx \sin \theta_0) e^{jky\cos\theta_0} \qquad (1)$$

where the time factor $e^{j\omega t}$ has been omitted, $k=2\pi/\lambda$, and $E_0$ is the rms amplitude of the field at the origin due to either source (rationalized mks units are employed throughout the discussion). We will be concerned with the case $r=\infty$, so that the plane-wave approximation becomes exact, and (1) is valid for all finite values of $x$, $y$, and $z$.

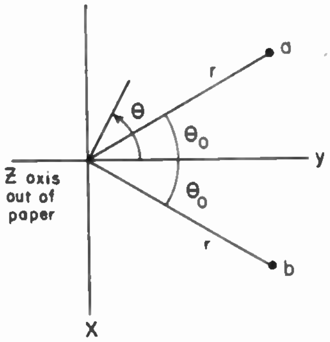


Fig. 1

It is seen that the electric field vanishes along planes which are parallel to the plane $x=0$ and which occur at values of $x=\pm n\lambda/4 \sin \theta_0$, where $n$ may have odd integral values. In addition, it is noted that any plane parallel to the plane $z=0$ has no tangential component of electric field. Thin conducting sheets may now be located so that they coincide with parts of these null planes without changing the solution given by (1).

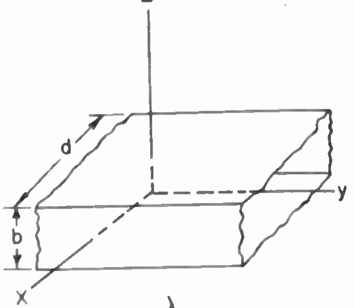


Fig. 2

We will consider in detail a configuration of conducting sheets which consists in part of a rectangular waveguide whose walls fit into the field as specified (Fig. 2). Only the case in which the waveguide is cut off for all modes except the $TE_{01}$ will be considered. This restric-

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tion specifies that the "narrow" faces of the waveguide lie along two adjacent null planes and that $\theta_0 > 30°$. This value of $\theta_0$ corresponds to cutoff conditions for the $TE_{02}$ mode. We will specify, further, the form of the waveguide to be considered by stating that all four of its boundary walls are continuous for all points to the left of the origin (i.e. for $y<0$). For $y>0$ on the other hand, the waveguide walls may extend any desired distance and may be arbitrarily separated or perforated as long as the conducting portions remain in the appropriate null planes. There may be present, in addition, any number of conducting sheets occupying any portions of the other null planes. Fig. 3 shows some examples of allowable systems.

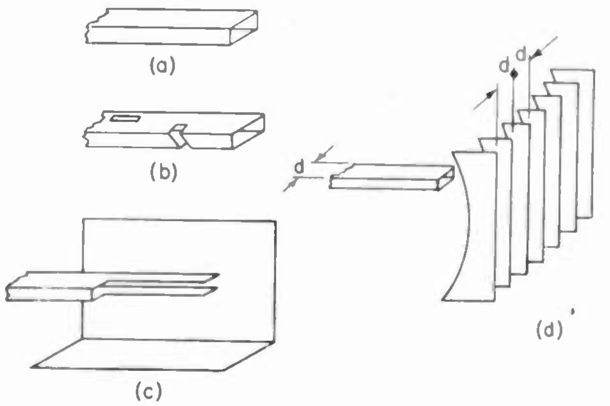


Fig. 3

The waveguide and its array of appropriately located conducting sheets may now be regarded as a receiving antenna receiving signals simultaneously from the two sources $a$ and $b$ and feeding them into the infinitely long waveguide to the left of the origin. Although the procedure followed in the foregoing steps fixed the angle $\theta_0$ simply by the choice of location of the two sources, we are ultimately interested in the radiation properties of a waveguide with given dimensions at a specified frequency. For this reason, $\theta_0$ should be considered as a dependent rather than an independent variable (i.e. $\theta_0 = \sin^{-1}(\lambda/2d)$).

## III. Gain Calculations

If one of the two sources is removed, the problem is no longer a simple one for which the solution is the same with or without conductors. To find the fields in this case would require the solution of a boundary-value problem of considerable difficulty even for the simplest configurations included in the general description above. By virtue of the restriction that the waveguide will propagate only the $TE_{01}$ mode, however, we may at least write the form of the electric field for points inside the continuous portion of the waveguide which are sufficiently far removed from the origin that all the higher-order modes are essentially damped out. The electric field in a $TE_{01}$ wave traveling to the left is of exactly the same form as the field given by (1) so we may write for the field due to source $a$ alone

$$E_a = h_a E_0 \cos(kx \sin\theta_0) e^{iky\cos\theta_0}. \tag{2}$$

The complex factor of proportionality $h_a$ relates the amplitude and phase of the $TE_{01}$ wave inside the waveguide (referred to the origin) to the field $E_0$ which would be produced at the origin by source $a$ if all the conductors were absent. Similarly, for the field due to source $b$ alone,

$$E_b = h_b E_0 \cos(kx \sin\theta_0) e^{iky\cos\theta_0}. \tag{3}$$

Since, by the principle of superposition, the sum of these must equal the field $E_z$ given by (1) it follows that

$$h_a + h_b = 2. \tag{4}$$

In the case of the general configuration of conductors, subject only to the conditions outlined above, we can proceed on simple grounds no further than (4). There are, however, two special classes of antennas contained within the scope of this general configuration which are of considerable interest and which allow additional development. These are:

(i) Systems having physical symmetry about the plane $x=0$. For this class, it follows immediately that

$$h_a = h_b = 1. \tag{5}$$

(ii) Systems for which it is known from measurements or from other considerations that either $h_a$ or $h_b$ = zero. This would be the case if the $z$ component of the radiation pattern were zero in the direction of one of the two sources (such as would be expected, for example, with the antenna shown in Fig. 3(c)). For antennas of this type

$$\begin{aligned} h_a &= 2 \\ h_b &= 0. \end{aligned} \tag{6}$$

Proceeding with class (i) antennas, we may substitute from (5) into (2) and (3) to get

$$E_a = E_b = E_0 \cos(kx \sin\theta_0) e^{iky\cos\theta_0}. \tag{7}$$

Either of the sources thus produces a $TE_{01}$ wave traveling to the left of the form given by (7). The power abstracted from the incident field due to either source by the receiving antenna may now be calculated using the well-known expressions developed by integrating the Poynting vector over a cross section of the waveguide.[1,2] This calculation yields

$$P_{\theta_0} = \frac{E_0^2 b\, l \sqrt{1 - \left(\frac{\lambda}{2d}\right)^2}}{2\eta} \tag{8}$$

where $\eta = \sqrt{\mu_0/\epsilon_0}$, $b$ and $d$ are the transverse dimensions of the waveguide and the subscript $\theta_0$ indicates that a

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

[2] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, New York, N. Y., 1944.

line connecting the origin and the distant source makes this angle with the waveguide axis.

One of the fundamental parameters characterizing the operation of an antenna is the receiving cross section.[3] This quantity is defined for a receiving antenna upon which a plane wave is incident by the equation

$$A\frac{E_0^2}{\eta} = P' \tag{9}$$

where

$A$ = receiving cross section

$E_0$ = amplitude of electric field in incident wave (rms)

$P'$ = available power, or power delivered to a termination which provides a conjugate match to the antenna impedance.

The receiving antenna may be represented by an equivalent circuit (Fig. 4) in which the waveguide is replaced by a transmission line and the antenna impedance or equivalent generator impedance is specified in terms of the reflection coefficient $\Gamma_0$ (referred to plane $y=0$) which would be present on the waveguide if the

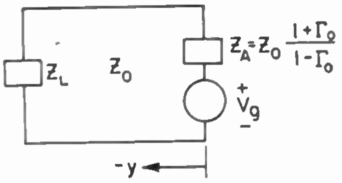


Fig. 4

antenna system were transmitting. The conditions under which $P_{\theta_0}$ was calculated correspond to the case $Z_L = Z_0$ in the equivalent circuit. A straightforward analysis of the circuit shows that this value of power is related to the available power by the expression

$$P_{\theta_0}' = \frac{P_{\theta_0}}{1-|\Gamma_0|^2}. \tag{10}$$

Combining (8), (9), and (10) yields for this antenna

$$A_{\theta_0} = \frac{bd\sqrt{1-\left(\frac{\lambda}{2d}\right)^2}}{2(1-|\Gamma_0|^2)}. \tag{11}$$

It is of interest to note that the numerator of this expression is the projected area of a cross section of the waveguide on a wave front of the incident plane wave.

The receiving cross section is related to the power gain through a simple factor of proportionality[3]

$$G = \frac{4\pi}{\lambda^2}A \tag{12}$$

where $G$ is the power gain over an isotropic radiator in the direction from which the incident plane wave arrives to give the corresponding value of $A$. Combining (11) and (12), we obtain the expression

$$G_{\theta_0} = \frac{2\pi bd\sqrt{1-\left(\frac{\lambda}{2d}\right)^2}}{\lambda^2(1-|\Gamma_0|^2)} \tag{13}$$

which gives the gain of the symmetrical antenna in the direction $\theta_0$ in terms of the physical dimensions of the waveguide and the reflection coefficient in the waveguide when the system is operating as a transmitting antenna.

For class (ii) antennas, a similar analysis using (6) instead of (5) yields

$$G_{\theta_0} = \frac{8\pi bd\sqrt{1-\left(\frac{\lambda}{2d}\right)^2}}{\lambda^2(1-|\Gamma_0|^2)}. \tag{14}$$

It should be remarked that (13) and (14) apply only to the $z$ components of electric field in the appropriate directions $\theta_0$. Symmetry arguments show that there will be only a $z$ component of electric field produced by antennas having physical symmetry about the plane $z=0$. but that, in general, there may be $x$ and $y$ components also.

These relationships have been developed for the idealized problem of antennas being fed through waveguides of infinite length. At this point, it is of interest to consider briefly the manner in which the results would be modified in the more practical case involving a waveguide of finite length. If, in the receiving case with both sources operating, the waveguide were closed with a conducting plate a finite distance to the left of the origin and a matched termination were placed inside the waveguide near the closed end, the $TE_{01}$ waves inside the remaining portion of the waveguide could still travel only to the left. The form of the fields on the outside of the waveguide, on the other hand, would have to be modified by a correction term to account for the diffraction of the incident waves about the closed end. The main part of this correction term would take the form of a perturbation field spreading out from the location of the closed end. The amount by which the amplitude of the $TE_{01}$ wave inside the waveguide would be changed from the value calculated above, would depend on the amount by which this perturbation field modified the fields in the vicinity of the open part of the waveguide. It is reasonable to assume that this modification could be made negligibly small by moving the closed end sufficiently far from the origin. It is thus concluded that the results calculated above should apply to practical systems with long waveguides, but no attempt is made to specify limits in numerical terms.

## IV. Some Additional Properties

Consider, now, a configuration of conductors meeting the requirements of the general case given above, but with the sources moved to positions $c$ and $d$ (Fig. 5).

[3] H. D. Friis and W. D. Lewis, "Radar antennas," *Bell Sys. Tech. Jour.*, vol. 26, pp. 219–318; April, 1947.

As before, the waveguide is continuous to the left of the origin, but in this case, it extends only a finite distance to the left, ending in an open termination. Proceeding as before, the total electric field in this problem is shown to be

$$E_z = 2E_0 \cos(kx \sin\theta_0) e^{-jky\cos\theta_0}. \quad (15)$$

Inside the waveguide, this field constitutes a $TE_{01}$ wave moving to the right.

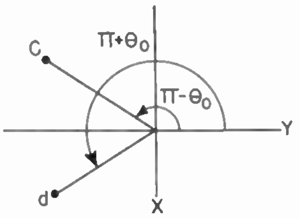

Fig. 5

We will now close the left end of the waveguide with a shorting plate and place a matched termination inside the waveguide near this end. In addition, we will install a source of energy near this terminated end and adjust its amplitude and phase so that its direct wave (traveling to the right in the waveguide) is of exactly the form given by (15). Since the incident waves from sources $c$ and $d$ and the incident $TE_{01}$ wave reach the antenna end of the waveguide with the same relative amplitudes and phase as they had when the remote end of the waveguide was open (neglecting the perturbation field as discussed in the preceding section) it follows that (15) still gives the electric field both inside and outside the waveguide. This field may be interpreted as the superposition of the individual fields due to the two external sources $c$ and $d$ and the source inside the waveguide.

The $TE_{01}$ waves excited in the waveguide by the external sources can travel only to the left since energy can enter the waveguide only at the antenna end, and the matched termination precludes the possibility of reflected waves being set up. The fields due to these sources, therefore, are of the form

$$E_c = h_c E_0 \cos(kx \sin\theta_0) e^{jky\cos\theta_0}, \quad (16)$$

$$E_d = h_d E_0 \cos(kx \sin\theta_0) e^{jky\cos\theta_0}. \quad (17)$$

The internal source, on the other hand, will produce waves in both directions, a direct wave traveling to the right and a reflected wave traveling to the left. Designating these fields by the subscript $e$ we have

$$E_e = 2E_0 \cos(kx \sin\theta_0)\left\{e^{-jky\cos\theta_0} + \Gamma_0 e^{+jky\cos\theta_0}\right\} \quad (18)$$

where the reflection coefficient $\Gamma_0$ is again referred to the plane $y=0$. Equation the sum $E_c+E_d+E_e$ to the field $E_z$ given by (15) leads to the condition

$$h_c + h_d + 2\Gamma_0 = 0. \quad (19)$$

Solving for $\Gamma_0$ and replacing the numerical factor 2 by its equivalent from (4) yields

$$\Gamma_0 = -\frac{h_c + h_d}{h_a + h_b}. \quad (20)$$

The $h$'s have been introduced as constants of the receiving antenna. A simple application of reciprocity allows (20) to be re-stated in terms of the distant fields produced by the waveguide antenna when transmitting. Thus

$$\Gamma_0 = -\frac{E_{zc} + E_{zd}}{E_{za} + E_{zb}} \quad (21)$$

where the fields $E_z$ are measured (in amplitude and phase) at the points previously occupied by the distant sources. For class (i) antennas, (21) reduces to

$$\Gamma_0 = -\frac{E_{zc}}{E_{za}}. \quad (22)$$

Thus a measurement of the reflection coefficient inside the waveguide specifies the ratio of the radiation fields in these directions.

Combining the information in (22) with that in (13) allows evaluation of the gain in the direction of point $c$.

$$G_{\pi-\theta_0} = \frac{2\pi b d \sqrt{1 - \left(\frac{\lambda}{2d}\right)^2}}{\lambda^2\left(\frac{1}{|\Gamma_0|^2} - 1\right)}. \quad (23)$$

## V. Application to Another Type of Waveguide Antenna

Returning now to free-space conditions, let us consider the fields produced when all four sources, $a$, $b$, $c$, and $d$ are operating simultaneously. If the common phase of $c$ and $d$ differs from that of $a$ and $b$ by $\pi$, the total electric field may be written as the field given by (1) minus that given by (15)

$$E_z = j4E_0 \cos(kx \sin\theta_0) \sin(ky \cos\theta_0). \quad (24)$$

This field has the same null planes as the fields considered previously plus another set occurring at the planes

$$y = \frac{m\lambda}{2\cos\theta_0}$$

where $m$ may have integral values (including zero).

We will again consider antennas fed by a single waveguide of finite length with continuous walls to the left of the origin. A few examples of configurations fitting into this system of null planes are shown in Fig. 6.

In this case, the left end of the waveguide may be closed with a shorting plate without the appearance of correction terms in the equations for the fields, provided the plate is located on one of the new set of null planes.

It is of interest to note, in passing, that the null planes now available allow the construction of completely closed systems such as that shown in Fig. 6(d).

For such systems, the outside fields must still be given by (24), but the inside region is completely isolated from the external sources, and hence the fields in this region are no longer related to the outside fields. If a hole of any size is cut anywhere in the conducting boundary, however, it may be argued that since (24) is a solution, it is the only solution and is valid in both regions.

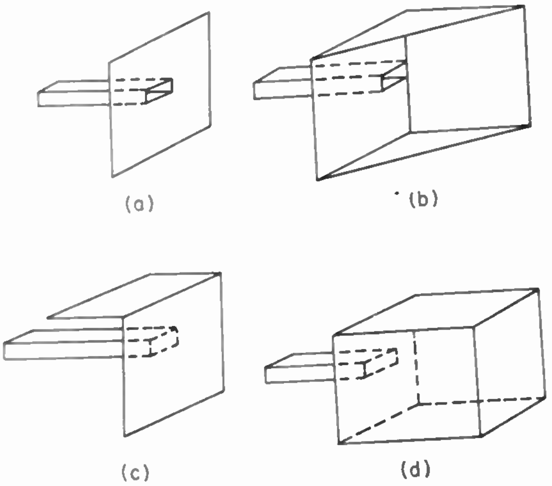


Fig. 6

Returning to the antenna problem, it is seen that the waves inside the waveguide now consist of one $TE_{01}$ wave traveling to the right and another of equal amplitude traveling to the left with relative phases such that cancellation occurs at $y=0$. This condition is not altered if we again place a matched termination and a source of energy in the waveguide near the left end with the amplitude and phase of the source adjusted so that its direct wave (traveling right) is of the proper form (i.e. the negative of (15)). The wave components present in this case are summarized in the diagram of Fig. 7. The

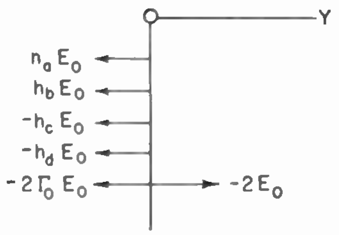


Fig. 7

only wave traveling to the right is the direct wave due to the internal source since all components entering at the antenna end are absorbed in the matched termination at the left end. Since the total field must vanish for $y=0$ in accordance with (24), we have the result

$$h_a + h_b - h_c - h_d = 2(1 + \Gamma_0). \tag{25}$$

For systems whose radiation patterns have zeros in the direction of sources $c$ and $d$ such as might be the case, for example, with the antenna of Fig. 6(a), (25) is simplified in that $h_c$ and $h_d$ are both zero. If, in addition, the antenna is of the symmetrical type (class (i)), (25) reduces to

$$h_a = h_b = 1 + \Gamma_0. \tag{26}$$

Proceeding as in the previous gain calculation we obtain

$$G_{\theta_0} = \frac{2\,|\,1+\Gamma_0\,|^2 \pi b d \sqrt{1 - \left(\dfrac{\lambda}{2d}\right)^2}}{\lambda^2(1 - |\,\Gamma_0\,|^2)}. \tag{27}$$

In this case, both the magnitude and phase of $\Gamma_0$ must be known in order to evaluate $G_{\theta_0}$. The reflection coefficient is again referred to the plane $y=0$ which is automatically fixed in a physical system as the location of one of the conducting sheets normal to the axis of the waveguide (or a parallel plane spaced $m\lambda/2 \cos\theta_0$ from such a conducting sheet). For antennas having radiation-pattern zeros in three of the four directions of interest so that $h_b = h_c = h_d = 0$ (Fig. 6(b)), the gain in the remaining direction becomes

$$G_{\theta_0} = \frac{8\,|\,1+\Gamma_0\,|^2 \pi b d \sqrt{1 - \left(\dfrac{\lambda}{2d}\right)^2}}{\lambda^2(1 - |\,\Gamma_0\,|^2)}. \tag{28}$$

## VI. Conclusions

Quantitative relationships between the pattern and impedance characteristics of various types of waveguide-excited radiators have been established. The results obtained are exact only for systems employing ideal conductors of the type assumed in the analysis. It appears justified, however, to conclude that these results will be good approximations for practical systems using good conductors whose thickness is small compared with the dimensions of the openings in the waveguide. Slotted-waveguide radiators (Fig. 3(b)) would not meet this requirement, in general, since the slot width in such antennas is ordinarily comparable with the wall thickness. The system shown in Fig. 6(a), on the other hand, meets the requirements quite well. In this case, the ends of the waveguide walls simply form part of the large sheet in which the waveguide terminates, so the "opening" in the system has the dimensions of this transverse sheet. It is believed that antennas of this type and of the type shown in Fig. 6(b) might find application as gain standards. Some development work on standard-gain antennas employing these principles will be undertaken at this laboratory in the near future.

Certain properties of the radiation patterns and the radiation resistance of loop and dipole elements near flat-plate and corner reflectors may be deduced by a procedure similar to that used above.

## Acknowledgment

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# A Note on Filter-Type Traveling-Wave Amplifiers*

J. R. PIERCE†, FELLOW, IRE, AND NELSON WAX‡

*Summary*—**A small-signal analysis of systems in which an electron beam interacts with a circuit composed of discrete filter elements is given here. The effects of a line beam interacting with a series of gaps, which are capacitive elements of a filter structure, are calculated, and it is shown that an admittance can be introduced which arises from the presence of the electrons. This admittance is in parallel with the gap capacitance, and thus will alter the propagation factor of the filter circuit. It is shown that traveling-wave solutions exist for the combination of electron beam and filter circuit, and that there is a solution which has a positive real part, indicating that gain will be exhibited.**

TRAVELING-WAVE amplifiers with high efficiencies and large gains might be obtained if one could increase greatly the impedance of the helices used; unfortunately, the practicable range of helix impedance is relatively small, and one is thus led to a consideration of different kinds of circuits. This note presents the results of a small-signal analysis of iterated structures coupled to an electron beam; it is shown that traveling-wave solutions exist for such systems, and that the gain of the system will increase with increasing circuit impedance.[1] The main disadvantage in using filter-type circuits of high impedance, rather than helices, would be the narrower bandwidths of the former.

The system to be considered is shown in Fig. 1. A source of electromagnetic energy (not shown in the figure) is delivering power to a load, which terminates the lossless circuit in its characteristic impedance. The direction of propagation of energy is to the right, when the electrons are absent; i.e., the group velocity of the "undisturbed circuit wave" is positive. The average motion of the electrons is to the right; the electron stream is coupled capacitively to the circuit only at the gaps shown, which are similar capacitors in corresponding branches of every filter section.

The following assumptions and restrictions hold throughout this work:

1. Space-charge effects are neglected.
2. The differences of potential are never more than a few kilovolts, and, therefore, the magnetic effects of the moving electrons may be ignored.
3. The time taken for an electron to cross any of the gaps is small compared with the period of the ac oscillations.
4. The ac components of voltage, current, etc., are so small compared to their dc values that products and squares of ac quantities may be neglected.

*Notation:* Circuit quantities will be in upper-case letters, those for the electron stream in lower-case letters. dc quantities will have the subscript zero, and ac quantities (with the factor $e^{j\omega t}$ omitted) will have the subscripts appropriate to the gap considered.

*Units:* Mks units are used.

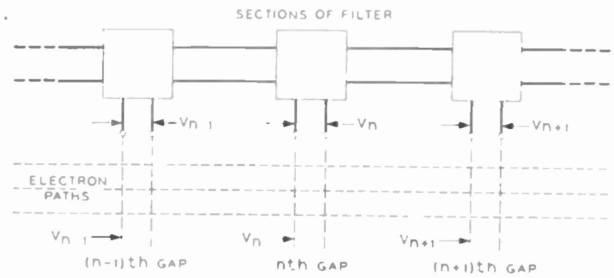


Fig. 1—A schematic representation of the system considered is shown here.

Let $v_n$ be the ac component of velocity of an electron just before it crosses the $n^{th}$ gap, and $V_n$ the ac voltage across the $n^{th}$ gap. Then $v_{n+1}$ will be related to $v_n$ and $V_n$ by[1]

$$v_{n+1} = \left[v_n + \eta \frac{M}{u_0} V_n\right] e^{-j\theta} \tag{1}$$

where

$\eta$ = the ratio of the charge of the electron to its mass
$u_0$ = the average electron velocity
$M$ = the modulation coefficient
$\theta$ = the transit angle,[1] $j = \sqrt{-1}$
$e = 2.718\ldots$

If a traveling-wave solution exists, then

$$v_{n+1} = v_n e^{\Gamma}, \tag{2}$$

with similar equations for the other ac quantities; $\Gamma$ is the propagation factor of the system.

The ac component of the convection current at the $(n+1)$st gap, $i_{n+1}$, is given, in terms of $i_n$, by[2]

$$i_{n+1} = i_n e^{\Gamma} = \left[i_n - j\theta\left(\frac{v_n}{u_0} + \eta \frac{M}{u_0^2} V_n\right) i_0\right] e^{-j\theta} \tag{3}$$

where $i_0$ is the dc convection current.

The ac component of velocity, $v_n$, may be eliminated from (1), (2), and (3); then one obtains

$$i_n = -\frac{j\theta M i_0 V_n}{2V_0} \frac{e^{\Gamma+j\theta}}{(e^{\Gamma+j\theta} - 1)^2} \tag{4}$$

with the dc voltage $V_0 = u_0^2/2\eta$.

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[1] Some similar work is to be found in Andre Blanc-Laprene and Pierre Sapostelle, "Contribution a L'Etude des Amplifications a Ondes Progressives," *Ann. de Telecommunications*, vol. 1, no. 12, pp. 283–302; December, 1946.

[2] J. R. Pierce and W. G. Shepherd, "Reflex oscillators," *Bell Sys. Tech. Jour.*, vol. 26, pp. 460–681; July, 1947.

The convection current $i_n$ will induce a current $-Mi_n$ in the circuit. One may thus introduce an admittance, $y_e$, which arises from the presence of the electrons,

$$y_e = -\frac{Mi_n}{V_n} = \frac{j\theta i_0 M^2}{2V_0}\frac{e^{\Gamma+j\theta}}{(e^{\Gamma+j\theta}-1)^2}, \quad (5)$$

is in parallel with the gap admittance, and is small compared to it.

Let the propagation factor for the filter, within the pass band, be $-j\beta$ when there are no electrons present. Then

$$\cosh(-j\beta) = 1 + \frac{Y_2}{2Y_1} \quad (6)$$

where $Y_1$ and $Y_2$ are the admittances of the series and shunt arms of the filter section, respectively.

One defines the function $F(y_e)$ by

$$F(y_e) = \frac{Y_2(y_e)}{2Y_1} \quad (7)$$

or

$$F(y_e) = \frac{Y_2}{2Y_1(y_e)}$$

where the admittance $y_e$ has been added to the gap admittance appropriately, in either the shunt or series arm of the filter, respectively. The parentheses indicate functional dependence.

One may expand $F(y_e)$ in a Maclaurin series and obtain

$$F(y_e) = F(0) + F'(0)y_e + \cdots \quad (8)$$

where $F(0) = Y_2/2Y_1$, $F'(0)$ etc., depend on the circuit used. Two simple examples of it are given below. The primes denote differentiation with respect to $y_e$.

The propagation factor of the system (circuit plus electron stream) is given by

$$\cosh\Gamma = 1 + F(y_e). \quad (9)$$

One obtains

$$\cosh\Gamma - \cos\beta = j\theta DF'(0)\frac{e^{\Gamma+j\theta}}{(e^{\Gamma+j\theta}-1)^2} \quad (10)$$

by combining (5), (6), (8), and (9), and using the first two terms in the expansion for $F(y_e)$.

$$D \equiv \frac{M^2 i_0}{2V_0}.$$

Let $\Gamma = -j(\beta+\gamma)$. One assumes here that $|\gamma| \ll |\beta|$; i.e., that the change in propagation factor arising from the presence of the electrons is small when the signal is small. This assumption is not necessary, though it is convenient, and the exact equations are given in the Appendix.

Equation (10) then becomes

$$\frac{\gamma^4}{4}[\cos(2\beta-\theta)+\cos\theta] + \gamma^3\sin(2\beta-\theta)$$

$$\frac{\gamma^2}{2}[\cos\theta + 2\cos\beta - 3\cos(2\beta-\theta)]$$

$$+ 2\gamma\sin\beta[1-\cos(\beta-\theta)] = j\theta DF'(0). \quad (11)$$

The approximations $\cosh\gamma \doteq 1+\gamma^2/2$, $\sinh\gamma \doteq \gamma$ have been used.

The nature of the roots of (11) is dependent on the function $F'(0)$. Two cases are considered: (a) the series arm, or (b) the shunt arm of the filter is the gap capacitance.

For case (a)

$$F(y_e) = \frac{Y_2}{2(Y_1+y_e)} \quad (12)$$

with

$$F'(0) = -\frac{Y_2}{2Y_1^2} = -\frac{Z_1^2}{2Z_2}. \quad (13)$$

For case (b)

$$F(y_e) = \frac{Y_2+y_e}{2Y_1} \quad (14)$$

with

$$F'(0) = \frac{1}{2Y_1} = \frac{Z_1}{2}. \quad (15)$$

In either case $F'(0) = jX$ where $X$ is a real quantity (positive for case (b) and negative for case (a)) when the filter is composed of pure reactances and the region of operation is within its pass band. It is particularly noteworthy that under these conditions, filters with inductive series arms have $0 \leqq \beta \leqq \pi$ (case b) and those with capacitive series arms have $-\pi \leqq \beta \leqq 0$ (case a). $\beta$ increases monotonically with frequency for both case (a) and case (b). Furthermore, the direction of propagation of the wave cannot be inferred from the phase velocity; one must resort to the group velocity.

Let $V_g$ be the group velocity, $v_p$ the phase velocity, $\omega = 2\pi f$, and $Im\Gamma$ the imaginary part of $\Gamma$.

The relationship between $V_g$, $v_p$, and $\omega$ is given by

$$\frac{1}{V_g} = \frac{1}{v_p} - \frac{\omega}{v_p^2}\frac{dv_p}{d\omega}; \quad (16)$$

since $v_p = -\omega/Im\Gamma$ one obtains

$$\frac{1}{V_g} = -\frac{dIm\Gamma}{d\omega} = -\frac{dIm\Gamma}{d\beta}\frac{d\beta}{d\omega}. \quad (17)$$

The wave will be propagated to the right, in the direction of the average motion of the electrons, if the imaginary part of $\Gamma$ is negative, and increasing with $\beta$ (or $\omega$), or positive, and decreasing with $\beta$.

Returning to (11) and using the expressions given above for $F'(0)$ one has

$$\frac{\gamma^4}{4}\left[\cos(2\beta-\theta)+\cos\theta\right]+\gamma^3\sin(2\beta-\theta)$$
$$+\frac{\gamma^2}{2}\left[\cos\theta+2\cos\beta-3\cos(2\beta-\theta)\right]$$
$$+2\gamma\sin\beta\left[1-\cos(\beta-\theta)\right]+\theta DX=0. \qquad (18)$$

Equation (18) may be simplified in an important special case, namely, when $\beta=\theta+2m\pi$ (case b), $m$ an integer. This corresponds to linear phase shift with frequency, since the transit angle $\theta$ is linear with frequency, for constant dc voltage. One obtains

$$\frac{\gamma^4}{2}\cos\theta+\gamma^3\sin\theta+\theta DX=0. \qquad (19)$$

$X$ is a positive quantity here.

Synchronism between the undisturbed circuit wave (i.e. when the electrons are absent) and the average motion of the electrons is given, for case (a), by $\theta=\beta=2m\pi$ ($\beta$ negative, $m$ an integer). Equation (19) still holds, with $X$ now negative. As $\beta$ varies from $-\pi$ to $0$ in the pass band, $\sin\theta$ will be negative in (19) as well.

The definition of $\gamma$, $\Gamma=-j(\beta+\gamma)$, indicates that gain will occur for the system if there is a root of (18) or (19) which has an imaginary part that is positive, and a real part that satisfies the condition that $V_g$ be positive. The first condition assures the existence of an increasing wave, and the second, that this wave shall be propagated to the right.

A circuit wave with a negative group velocity (propagation to the left, contrary to the electron motion) may appear to have a positive phase velocity equal to the electron velocity. It seems physically unreasonable that the over-all wave resulting when the electrons are present could have a negative group velocity and increase in traveling to the left; this would imply a bunching of the electrons to give up power in anticipation of the arrival of the wave.

The correspondence between discrete circuits, which have been treated here, and continuous circuits, which Pierce[3] has discussed in considerable detail, may be demonstrated by considering the limiting form of the filter circuit as the length of a section goes to zero, and the number of sections becomes infinite. It is found that the equation for the discrete case is indistinguishable from that for a corresponding continuous case if $|\delta/2|\cot\beta\ll 1$.

[3] J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.

To show the variation of gain and phase of the increasing wave with drift angle $\theta$, the roots of (19) have been obtained for the following sets of numbers: $M=1$, $V_0=1000$ volts, $i_0=0.001$ amperes, $F'(0)=\pm j\ 2{,}000$ ohms. The real and imaginary parts of the root satisfying the above conditions for system gain are shown in Figs. 2 and 3, as functions of $\theta$ and $\beta$. Two of the other

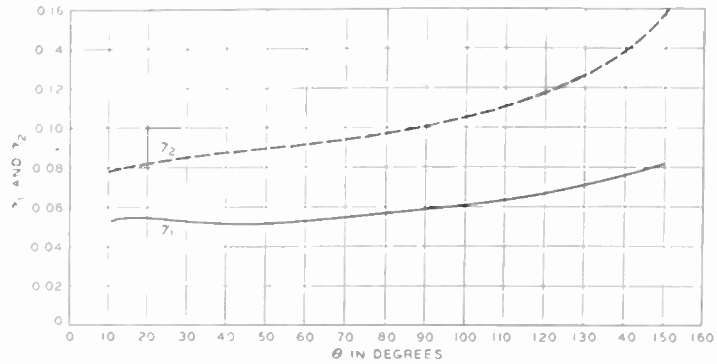


Fig. 2—The real part of the root $\gamma_1$ pertaining to phase and the imaginary part of the root $\gamma_2$ giving gain, shown as functions of $\theta$. The change per section of the ac quantities is $\exp[-j(\beta+\gamma_1)+\gamma_2]$. The equation which was solved was $\gamma^4/2\cos\theta+\gamma^3\sin\theta+0.001\theta=0$, $\beta=\theta$ (case b).

roots were real, in each case, indicating a shift in phase from the undisturbed circuit wave, and the remaining one was the complex conjugate of the one shown in the figures, introducing attenuation rather than gain. It should be noted that $F'(0)$ can be expressed in terms of the characteristic impedance and phase shift of the filter, if desired, and will not be, in general, a constant. In these calculations, $F'(0)$ was taken as a constant, for simplicity. The regions where this assumption is most likely to break down, namely, near the edges of the pass band, were omitted in the figures.

A check on the accuracy of the roots obtained is given in the Appendix.

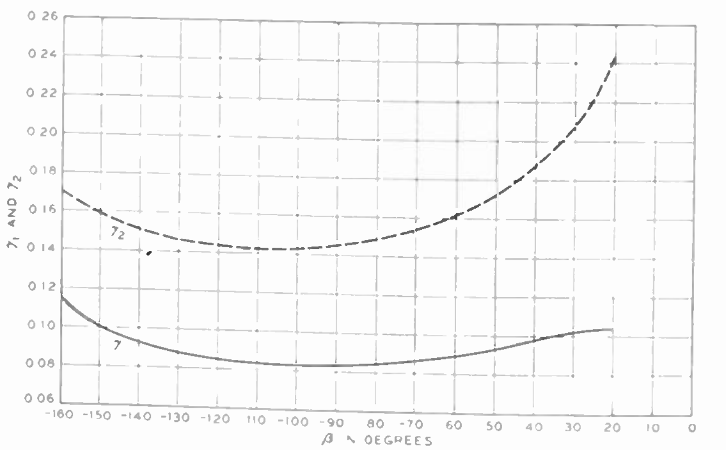


Fig. 3—The real part of the root $\gamma_1$ pertaining to phase and the imaginary part of the root $\gamma_2$ giving gain, shown as functions of $\beta$. The change per section of the ac quantities is given by $\exp[-j(\beta+\gamma_1)+\gamma_2]$. The equation which was solved was, $\gamma^4/2\cos\beta+\gamma^3\sin\beta-0.001\theta=0$, $\theta-\beta=2\pi$ (case a).

The authors thank their colleague, Miss Mary E. Moore, for her efficient performance of all the numerical calculations and her care in checking many of the algebraic details.

## Appendix

It was assumed, in deriving (11) from (10), that $|\gamma| \ll |\beta|$ where $\Gamma = -j(\beta+\gamma)$. This approximation is not invoked here, and the exact small-signal equations are obtained.

Let $e^{\Gamma} = \rho e^{-j\Phi}$. If this substitution is made in (10) then the pair of simultaneous quadratic equations with real coefficients

$$\begin{aligned} &\rho^4 \cos(4\Phi - \theta) - 2\rho^3[\cos 3\Phi + \cos(3\Phi - \theta)\cos\beta] \\ &\quad + \rho^2[\cos(2\Phi + \theta) + \cos(2\Phi - \theta) \\ &\quad + 4\cos 2\Phi \cos\beta + 2\theta DX \cos 2\Phi] \\ &\quad - 2\rho[\cos\Phi + \cos(\Phi + \theta)\cos\beta] + \cos\theta = 0 \end{aligned} \tag{20}$$

and

$$\begin{aligned} &\rho^4 \sin(4\Phi - \theta) - 2\rho^3[\sin 3\Phi + \sin(3\Phi - \theta)\cos\beta] \\ &\quad + \rho^2[\sin(2\Phi + \theta) + \sin(2\Phi - \theta) \\ &\quad + 4\sin 2\Phi \cos\beta + 2\theta DX \sin 2\Phi] \\ &\quad - 2\rho[\sin\Phi + \sin(\Phi + \theta)\cos\beta] + \sin\theta = 0 \end{aligned} \tag{21}$$

are obtained by expanding and separating real and imaginary quantities. The substitution in (10) shows directly that there are four waves.

Gain will occur for the system when $\rho > 1$ and when the direction of propagation is to the right (a positive group velocity). The use of (16) with $v_p = \omega/\Phi$ yields

$$\frac{1}{V_g} = \frac{d\Phi}{d\omega} = \frac{d\Phi}{d\beta}\,\frac{d\beta}{d\omega}. \tag{22}$$

As before, $V_g$ will be positive when $d\Phi/d\beta$ is positive.

In the body of this paper, the equation $\Gamma = -j(\beta+\gamma)$ was used; here the substitution $e^{\Gamma} = \rho e^{-j\Phi}$ is made. The connection between the two sets of quantities is, for $\gamma = \gamma_1 + j\gamma_2$,

$$\left.\begin{aligned} e^{\gamma_2} &= \rho \\ \beta + \gamma_1 &= \Phi \end{aligned}\right\}. \tag{23}$$

The roots of equation (18) may be used, with the aid of equations (23), as a first approximation in the solution of equations (20) and (21).

The root of equations (20) and (21) which corresponds to the values shown in Fig. 2 at $\beta = \theta = \pi/2$, $DX = 0.001$, is $\rho = 1.08$, $\Phi = 1.63$; this yields, using (23), $\gamma_1 = 0.06$, $\gamma_2 = 0.078$. The values obtained from (19) are $\gamma = 0.058$, $\gamma_2 = 0.101$. Similarly when $\beta = -\pi/2$, $\theta = 3\pi/2$, $DX = -0.001$, (20) and (21) yield $\rho = 1.16$, $\Phi = -1.49$, giving $\gamma_1 = 0.08$, $\gamma_2 = 0.149$. The approximate root had $\gamma_1 = 0.084$, and $\gamma_2 = 0.145$. Thus the approximation is fairly good even for the relatively large value of $X$ which was used.

# Correction

W. W. Mumford, whose paper, "Directional Couplers," appeared on pages 160–165 of the February, 1947, issue of the Proceedings of the I.R.E., has called to the attention of the editors an error in Fig. 6, which appeared on page 163. The correct figure is presented herewith.

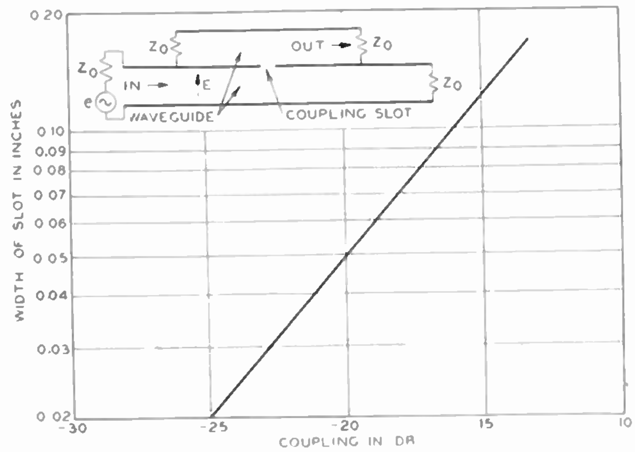


Fig. 6—Coupling ($10 \log_{10} P_{out}/P_{in}$) versus slot width for narrow slots across the wide face of 1-×2-inch rectangular waveguide. $\lambda_0 = 7.4$ centimeters. The wall thickness is 0.128 inch.

# Note on Space-Charge Considerations in Test-Diode Design*

EDWARD A. COOMES†, SENIOR MEMBER, IRE, AND JAMES G. BUCK†

*Summary*—**A computation has been made of the conditions for choice of anode and cathode diameters in test diodes that will give the minimum variation in the slope of the space-charge line with small variations in tube geometry.**

THE EXPERIMENTAL development of microwave tubes usually brings about attempted correlations between tube performance and cathode quality, particularly when oxide cathodes are used. Such experiments indicate that many instabilities, such as certain types of moding and sparking in multiple-cavity magnetrons and loss of efficiency in coplaner triodes, may to a large extent be decreased by increasing the electron emitting capacity of the cathode.[1,2] While this cathode property has been ascertained in various ways which are often dependent upon measuring techniques available, an accurate and conservative measure of the intrinsic electron emission may be taken as that value of the cathode current density $I^{2/3}$ versus $V$ plot at which the experimental points first deviate appreciably from the theoretical space charge line for the test tube which contains the cathode. If the interface plus coating resistance of the cathode is small, the emission $I_0$ taken at this point (see Fig. 1), which represents the maximum space-charge-limited current, corresponds to the primary emission with zero field at the cathode when held at its operating temperature. This point may be determined without recourse to considerations of the cathode characteristic in the Schottky region where phenomena may be varied and complicated. For example, if instead of $I_0$ in Fig. 1, $I_A$ measured at an arbitrary anode voltage $V_A$ were taken as the cathode emission, definite correlation between this quantity and microwave performance would be impossible.

The testing of experimental cathodes in diodes has been found useful both in the production of types needed for new tube developments and in deriving new fundamental information about the cathodes themselves.[1,3] This is particularly true when $I^{3/2}$ versus $V$ plots of emission data were made the basis of interpretation. The usefulness of this procedure is contingent on the choice of anode and cathode dimensions which will allow the expectancy of an accurate fit of experimental data to a computed space-charge line. A good choice can be made, even though small deviations from theoretical geometry must be tolerated for practical reasons.

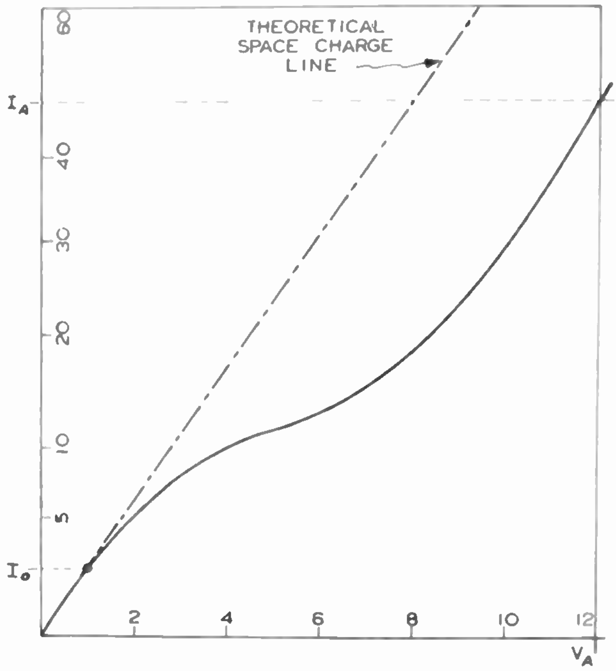


Fig. 1—Pulsed cathode emission characteristic of an oxide cathode in a cylindrical test diode. The current is plotted against anode voltage on $\frac{3}{2}$-power paper.

For test diodes of cylindrical symmetry the space-charge equations may be written

$$I = 2.33 \times 10^{-6} \frac{V^{3/2}}{r_c r_a \beta^2} \tag{1}$$

where

$I$ = cathode current density in amperes/cm²
$V$ = applied potential difference between anode and cathode in volts
$r_a$ = anode radius in centimeters
$r_c$ = cathode radius in centimeters.

The quantity $\beta$ is a function of ratio of anode to cathode radius, $\sigma = r_a/r_c$, and is given by the series

$$\beta = \sum_0^\infty A_n(\ln \sigma)^n. \tag{2}$$

The $A_n$'s have been given by Langmuir and Blodgett[4] as far as $n = 14$.

Equation (1) may be rewritten in the form

$$I^{2/3} = KV \tag{3}$$

* Decimal classification: R138. 1. Original manuscript received by the Institute, August 23, 1948; revised manuscript received, December 27, 1948. This work was done under U. S. Navy Bureau of ships Contract NObsr-30028.

† University of Notre Dame, Notre Dame, Ind.

[1] E. A. Coomes, J. G. Buck, A. S. Eisenstein, and A. Fineman, "Alkaline Earth Oxide Cathodes for Pulsed Tubes," NDRC Report No. 14-933, March 30, 1946.

[2] M. A. Pomerantz, "Magnetron cathodes," Proc. I.R.E., vol. 34, pp. 903–910; November, 1946.

[3] E. A. Coomes, "The pulsed properties of oxide cathodes," *Jour. Appl. Phys.*, vol. 17, p. 647; August, 1946.

[4] I. Langmuir and K. B. Blodgett, "Currents limited by space charge between coaxial cylinders," *Phys. Rev.*, vol. 22, pp. 347–356; October, 1923.

so that space-charge-limited values of cathode current density plotted on an $I^{2/3}$ versus $V$ plot will lie on a straight line of slope $K$, where

$$K = \text{const.}\left(\frac{1}{r_c r_a \beta^2}\right)^{2/3}. \tag{4}$$

The smallest change in $K$ with tube dimensions would be obtained when $K$ goes through a minimum; and a value of $\sigma$ may be chosen to bring about that condition in the following manner.

$$\text{For} \qquad K = K_{\min}; \qquad \frac{dK}{d\sigma} = 0 \tag{5}$$

from which $\sigma_{\min}$ is the value of $\sigma$ at which $K = K_{\min}$ may be obtained.

If values of $K$ are plotted as a function of cathode diameter, the family of curves in Fig. 2 is obtained which indicates that the $K_{\min}$ is reached for $\sigma \simeq 3$.

If (5) is solved with accuracy to two decimal places $\sigma_{\min}$ has a value of 3.16.

ACKNOWLEDGMENT

The authors appreciate the helpful criticism of Henry F. Ivey and L. R. Bittman of the Westinghouse Electric Corporation, Bloomfield, N. J., on this problem.

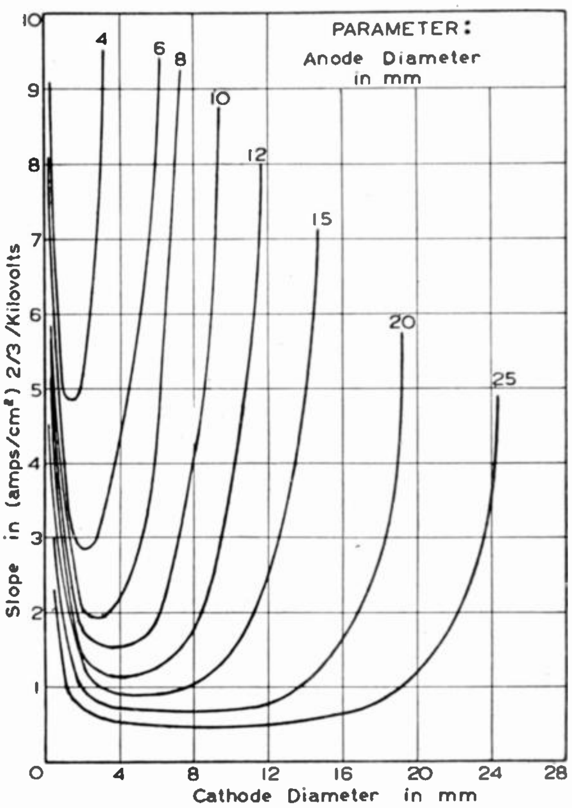


Fig. 2—The slope of the $I^{2/3}$ versus $V$ line for the space-charge law for cylinders plotted against cathode diameter. Note that the ratio of anode to cathode radius for minimum change in slope is approximately 3.

# The Synthesis of Electric Networks According to Prescribed Transient Conditions*

MORTON NADLER†, ASSOCIATE, IRE

*Summary*—**By use of the Laplace transform a network function may be obtained yielding a prescribed transient response. The procedure is based directly on the prescribed function of time, without consideration of amplitude or delay as a function of frequency. An example is given in which the Poisson-Stieltjes integral is employed for the physical realization of a network function which involves a transcendental term. In conclusion, certain questions are raised for further research in applied mathematics.**

## I. THE PROBLEM

MODERN ELECTRONICS, with pulse modulation, television and radar, servomechanisms, etc., is very concerned with transient response. The usual approach in designing for good transient response is through "wide-band" amplifiers. This is frequently interpreted to mean uniform absolute magnitude of response over a sufficiently great frequency band. The empiric rule relates the required bandwidth to the rise time of the pulse. For instance, at the 1947 IRE National Convention, Lee[1] stated that "the frequency at which the response is 90 per cent of maximum must be 1/4 to 1/2 the reciprocal of the time required for the front of the pulse to attain 90 per cent of maximum amplitude, the figure of 1/4 to 1/2 depending on the circuit." For insurance, some have even proposed a bandwidth fully as great as the reciprocal of the rise time. But even this extreme "safety factor" is not enough to guarantee a satisfactory response. Nicolas[2] uses the simple example of the reactive all-pass lattice

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† Tesla, Prague, Czechoslovakia.

[1] Reuben Lee, "Frequency response," (Abstract), *Tele-Tech*, vol. 6, p. 55; May, 1947.

[2] P. Nicolas, "Filtres electriques, largeurs de bande et tamps d'etablissement," *Rev. Gen. Elec.*, p. 27; July 8, 1939.

filter section. The reciprocal of the bandwidth is zero. Nevertheless the rise time is finite.

The requirement that the time required for the signal to traverse the network be uniform over the pass band is well known. At the same Convention of the IRE DiToro[3,4] discussed this question and proposed a definition of phase bandwidth complementing the amplitude bandwidth usually meant by the unmodified term "bandwidth." He advised making the phase bandwidth several times wider than the amplitude bandwidth.

While many workers have given formulas relating phase response to attenuation characteristic and vice versa,[5,6] the sphere of application of these formulas is restricted. Furthermore, while the frequency response (steady-state) and the time response (transient) are directly related in the solution of the differential equations involved, yet previous methods based on these relations have been largely "cut-and-try."

Whiteman[7] anticipates the present work to some extent. However, this author completely neglects the possibility that the network function obtained may be nonrational, and still be physically realizable.

## II. Fundamental Relations

A mathematical expression $\phi(p)$ describing an electrical network is called a network function. Formally, this function may be obtained by the substitution $p=j\omega$ in the impedance or admittance function of the network.[4] For the network function to correspond to a physically realizable, passive network, it is necessary and sufficient that (for two-terminal networks):

1. Zeros and poles either are real or occur in complex conjugate pairs.
2. The real and imaginary parts are, respectively, even and odd functions of $p$ on the imaginary $p$-axis.
3. None of the zeros and poles may be found in the right half-plane, and zeros and poles on the imaginary $p$-axis must be simple.
4. The real component of the function cannot be negative on the imaginary $p$-axis.

Further conditions for four terminal networks are given by Bode.[5]

As stated above, the network function may be obtained in a formal manner from the impedance or admittance function with the substitution $p=j\omega$. But the same function may be obtained in the analysis of networks by means of the operational calculus rigorously based on the Fourier-Mellin theorem:[8]

Let $F(t)$ be an arbitrary function of the real variable $t$, with only a finite number of maxima and minima and discontinuities, and $F(t)=0$ for $t<0$. If

$$g(p) = p\int_0^\infty e^{-pt}F(t)dt \quad [R(p) > c > 0], \tag{1}$$

then

$$F(t) = \frac{1}{2\pi j}\int_{c-j\infty}^{c+j\infty} e^{pt}\frac{g(p)}{p}dp, \tag{2}$$

provided

$$\int_0^\infty e^{-ct}F(t)dt$$

converges absolutely.

It should be noted that $g(p)/p$ is the Laplace transform of $F(t)$, written $L[F(t)]$.

In the analysis of the response $F_r(t)$ of a network $\phi(p)$ to an impressed driving function $F_d(t)$, one obtains the basic operational equation

$$L[F_r(t)] = \frac{L[F_d(t)]}{\phi(p)}. \tag{3}$$

This is an algebraic equation and *not* an integro-differential equation with $p$ and $1/p$ written in place of the differential and integral operators. Therefore, we need only solve (3) for $\phi(p)$ to obtain

$$\phi(p) = \frac{L[F_d(t)]}{L[F_r(t)]} = \frac{g_d(t)}{g_r(t)}. \tag{4}$$

When $F_d(t)=H(t)$, the unit step function $[H(t)=0$ for $t<0$, $H(t)=1$, $t\geqq 0]$, then $g_d(p)=1$, $L[H(t)]=1/p$. We then obtain

$$\Phi(p) = g_r(p) \qquad \left(\Phi = \frac{1}{\phi}\right). \tag{4a}$$

If the desired network performance can be specified in terms of response to the unit function $H(t)$, the network function is given directly, bearing in mind that admittance and impedance have become interchanged; if in (4) $\phi$ represents admittance (impedance), then in (4a) $\Phi$ represents impedance (admittance). In the balance of the discussion we shall employ $\phi$ indifferently to represent the network function of (4) or (4a).

It is now necessary to determine whether $\phi$ is physically realizable. For two-terminal networks, the conditions given above are necessary and sufficient.

When $\phi$ is a rational function, we may employ the standard methods of network synthesis.[4] But, in general, $\phi$ will not be a rational function. In this case it is necessary to use some convenient approximation method, transforming $\phi$ into a convergent or asymptotically convergent rational function.

[3] M. J. DiToro, "Phase and amplitude distortion in linear networks," Proc. I.R.E., vol. 36, pp. 24–36; January, 1948.
[4] M. J. DiToro, "Network distortion," (Abstract) *Tele-Tech*, vol. 6, p. 55; May, 1947.
[5] Hendrik W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand, Inc., New York, N. Y., 1945.
[6] Wilhelm Cauer, "Das Poissonsche Integral und seine Anwendungen auf die Theorie der linearin Wechselstromschaltungen (Netzwerke)," *Elek. Nach. Tech.*, p. 17; January, 1940.
[7] R. A. Whiteman, "A contribution to the theory of network synthesis," Proc. I.R.E., vol. 30, pp. 244–245; May, 1942.
[8] Louis A. Pipes, "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Co., Inc., New York, N. Y. 1946.

For instance, $\phi$ may be expanded as an infinite product, realizable by the Foster reactance theorem, or which may further be transformed into partial fractions[9] or into a continued fraction.[10] The network function may be expanded as a Taylor series, which may then be realized as a network of lattice structures after a method of Wiener and Lee,[11] or which, again, may be transformed further, into a continued fraction.[10]

In the illustrative example of Section IV a transcendental $\phi$ is expressed as a Poisson-Stieltjes integral, which is expanded as a continued fraction.

## III. The Poisson-Stieltjes Integral

Cauer gave a great deal of attention to the Poisson integral in the form of a Stieltjes integral. This integral has value in the physical realization of non-rational network functions.

Let $\phi(p)$ be a positive function, real for real $p$, regular and with positive real part in the right half-plane. Then[6]

$$\phi(p) = p\left[c + \int_0^\infty \frac{d\psi(x)}{p^2 + x}\right] \tag{5}$$

where

$$c = \lim_{p=\infty} \frac{\phi(p)}{p}; \quad d\psi(x) = \frac{R[\phi(j\sqrt{x})]}{\pi\sqrt{x}}\,dx. \tag{6}$$

Stieltjes has shown that if $\Psi(x)$ is an increasing function, with infinitely many points of growth, if the integrals

$$c_k = \int_0^\infty (-x)^{k-1} d\psi(x) \quad (k = 1, 2, 3, \cdots) \tag{7}$$

all exist, then the integral

$$\int_0^\infty \frac{d\psi(x)}{z + x} \tag{8}$$

may always be represented (except on the negative real axis) by a continued fraction of the form

$$\frac{1|}{|b_1 z} + \frac{1|}{|b_2} + \frac{1|}{|b_3 z} + \frac{1|}{|b_4} + \cdots \tag{9}$$

where the $b_i$ are all positive.[10]

[Note: (9) is a concise notation for the more familiar

$$\cfrac{1}{b_1 z + \cfrac{1}{b_2 + \cfrac{1}{b_3 z + \cfrac{1}{b_4 + \cdots}}}}$$

with an infinite number of members.]

[9] Murray F. Gardner and John L. Barner, "Transients in Linear Systems," John Wiley and Sons, New York, N. Y., 1945.

[10] Oskar Perron, "Die Lehre von den Kettenbruchen," Leipzig, 1913.

[11] Norbert Wiener and Y. W. Lee, U. S. Patent numbers 2,024,900, 2,124,599, and 2,128,257.

Comparing (8) with (5), when the conditions for expressing (8) in the form of (9) are satisfied, (5) may be also expressed as a continued fraction, except for the imaginary $p$-axis. Fortunately, in network synthesis, it is the real $p$-axis which is employed ($p = j\omega$). Hence we may write

$$\phi(p) = p\left[C + \int_0^\infty \frac{d\psi(x)}{p^2 + x}\right] \tag{5}$$

$$= pC + \frac{p|}{|b_1 p^2} + \frac{1|}{|b_2} + \frac{1|}{|b_3 p^2} + \frac{1|}{|b_4} + \cdots \tag{10}$$

$$= pC + \frac{1|}{|b_1 p} + \frac{1|}{|b_2 p} + \frac{1|}{|b_3 p} + \frac{1|}{|b_4 p} + \cdots \tag{10a}$$

which is immediately realizable as a ladder network of pure reactance.

Following Perron,[10] if

$$\left.\begin{aligned} B_0(p) &= 1, \quad B_1(p) = b_1 p^2, \quad B_2(p) = b_1 b_2 p^2 + 1, \\ B_{2n+1}(p) &= b_{(2n+1)} p^2 B_{2n}(p) + B_{(2n-1)}(p) \\ B_{2n}(p) &= b_{2n} B_{(2n-1)}(p) + B_{(2n-2)}(p) \end{aligned}\right\} \tag{11}$$

then the $b_i$ are obtained from

$$\frac{1}{b_{(2n+1)}} = \int_0^\infty [B_{2n}(j\sqrt{x})]^2 d\psi(x)$$

$$\frac{1}{b_{2n}} = \int_0^\infty \frac{1}{x} [B_{(2n-1)}(j\sqrt{x})]^2 d\psi(x).$$

Thus

$$\frac{1}{b_1} = \int_0^\infty d\psi(x) \qquad (n = 0)$$

$$B_1(p) = b_1 p^2$$

$$\frac{1}{b_2} = b_1^2 \int_0^\infty x\,d\psi(x) \qquad (n = 1)$$

$$B_2(p) = b_1 b_2 p^2 + 1$$

$$\frac{1}{b_3} = \int_0^\infty (1 - b_1 b_2 x)^2 d\psi(x) \qquad (n = 2)$$

etc.

## IV. Illustrative Example

We wish to construct a single-stage amplifier whose response to the unit function is approximately the curve $(2G/\pi)Si(2t/a)$. Here, $2G/\pi$ is a proportionality constant relating the response to the gain $G$ of the amplifier; $Si$ is the "sine integral" function

$$\int_0^t \frac{\sin t}{t}\,dt$$

and $t/a$ is the rise time in seconds. $Si$ is the response of an "ideal" low-pass filter with linear phase response and zero attenuation within the pass band and 100 per cent attenuation above the pass band. The function represents an oscillatory type of response (curve (a) in Fig. 1)

$$F_r(t) = \frac{2G}{\pi} Si \frac{2t}{a}; \qquad F_d(t) = g_m H(t). \tag{13}$$

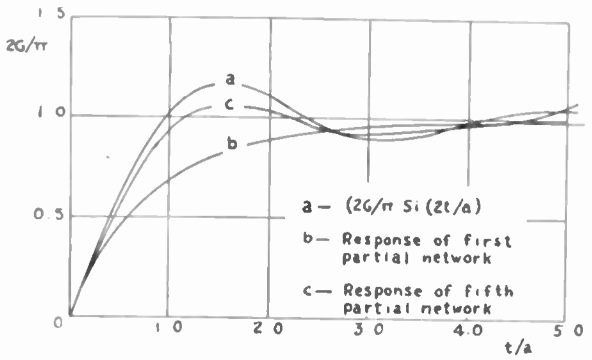



Fig. 1—Response of the first and fifth partial networks.

From a suitable table we obtain

$$g_r(t) = \frac{2G}{\pi} \cot^{-1} \frac{ap}{2}; \qquad g_d(t) = g_m. \tag{14}$$

This is real for real $p$, finite and with positive real part in the right half-plane, provided we take the principle value $0 \leqq \cot^{-1}(ap/2) \leqq \pi$. If we write $R_L = G/g_m$, we find the required *impedance*

$$\phi(p) = \frac{2R_L}{\pi} \cot^{-1} \frac{ap}{2}. \tag{15}$$

To represent (15) in the form of the Poisson-Stieltjes integral, we substitute in (5) and (6)

$$\left.\begin{aligned} C &= \lim_{p=\infty} \frac{2R_L}{\pi p} \cot^{-1} \frac{ap}{2} = 0 \\ d\psi(x) &= \frac{2R_L}{\pi^2 \sqrt{x}} R\left(\cot^{-1} \frac{aj\sqrt{x}}{2}\right) dx. \end{aligned}\right\} \tag{16}$$

But

$$\begin{aligned} R(\cot^{-1} jy) &= \frac{\pi}{2}; \qquad -1 < y \leqq 1 \\ &= 0; \qquad |y| > 1. \end{aligned}$$

Hence

$$\left.\begin{aligned} d\psi(x) &= \frac{R_L}{\pi\sqrt{x}} dx; \qquad x \leqq \frac{4}{a^2} \\ &= 0; \qquad x > \frac{4}{a^2}, \end{aligned}\right\} \tag{17}$$

and finally

$$\phi(p) = \frac{R_L p}{\pi} \int_0^{4/a^2} \frac{dx}{\sqrt{x}(p^2 + x)}. \tag{18}$$

From (11) and (12) in connection with (17) we can obtain the coefficients of the continued fraction representing (15). The first five coefficients are:

$$\left.\begin{aligned} b_1 &= \frac{\pi a}{4R_L}, \qquad b_2 = \frac{3aR_L}{\pi}, \qquad b_3 = \frac{5\pi a}{16R_L} \\ b_4 &= \frac{28aR_L}{9\pi}, \qquad b_5 = \frac{81\pi a}{256R_L}. \end{aligned}\right\} \tag{19}$$

If we denote the network represented by the partial fraction terminating with $b_n$ as the "$n$th partial network," the transient response of the first and fifth partial networks is given in Fig. 1. The convergence of the response toward the prescribed response is clearly shown.

The circuit diagram of the fifth partial network is given in Fig. 2. The first member is a shunt capacitance, with which is included the input capacitance of the electron tube.

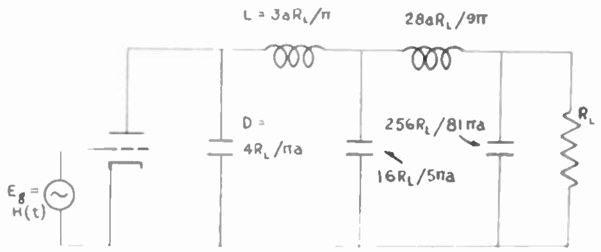



Fig. 2—Fifth partial network.

Since, at $p=0$, $\phi$ must be equal to $R_L$, the partial networks must terminate with a resistance of value $R_L$. In each case, $R_L$ represents the limit of the impedance of the infinite network following the partial network.

## V. Conclusions

Mathematical techniques exist that are adequate for the synthesis of finite networks whose transient response approximates a prescribed response as well as desired. The main obstacle to bringing this method of network synthesis from the realm of theory to the realm of practice is the difficulty of finding a suitable form for $F_r(t)$. Many plausible functions will be found to have Laplace transforms which are not positive functions. On the other hand, two functions equally good as approximations to the desired response, and each with satisfactory transforms, may lead to network functions which converge at widely different rates.

Another problem of importance will be the means of combining transient and steady-state prescribed conditions. For instance, in the design of television intermediate-frequency amplifiers, the steady-state conditions to be included are the rejection of one sideband, and of the sound channel.

## Acknowledgment

Acknowledgment is made to T. J. Higgins, professor of electrical engineering at the Illinois Institute of Technology, who first brought the Poisson-Stieltjes integral to the attention of the author.

# Transient Response of Cathode Followers in Video Circuits*

B. Y. MILLS†

*Summary*—**The behavior of cathode followers handling the irregular video signals of radar or television is discussed. It is shown that the signal-handling ability of a cathode follower may be reduced if it has a capacitive load. A general expression for the magnitude of this reduction is given, and its effect on design procedures is discussed.**

## I. Introduction

THE USE OF cathode followers for providing low-impedance outputs with a small current drain and low distortion is well known, but in general their behavior when operating into a capacitive load is the subject of some misconception. The object of the present paper is to show the limitations this behavior imposes on the design of cathode followers required for handling the irregular video signals of radar or television.

The exact shape of these signals depends on the nature of the circuits preceding the cathode follower. The signals may be represented approximately, however, as a series of positive and negative linear transitions of constant transition time $T$, as illustrated in Fig. 1. The transition time of the actual signals is conventionally

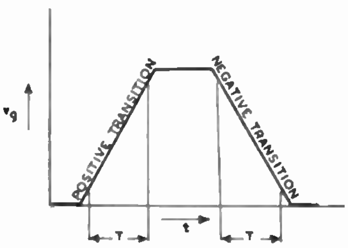


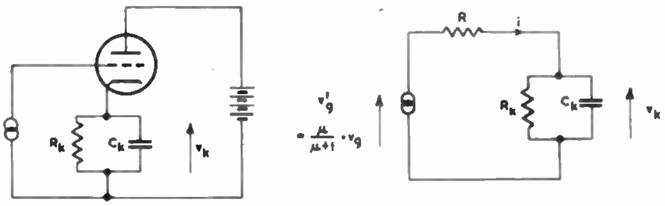

Fig. 1—A typical signal.

taken as the time required for the voltage to change from 10 to 90 per cent of its final value.[1] When applied to a linear transition, this value will be conservative, provided that there is little or no overswing in the original.

The requirement usually imposed on a cathode follower is that such transitions be reproduced with little distortion. It will be shown that the presence of capacitance in parallel with the cathode load resistor may then result in a reduction of the maximum signal amplitude which may be handled. Although this effect is qualitatively familiar to most circuit designers, no quantitative analysis immediately applicable to video signals has so far been published. It is common to see uneconomical designs based on an exaggerated respect for the limitation.

## II. The Equivalent Circuit

In the analysis it has been found convenient to use the equivalent circuit of Fig. 2(b). The cathode follower of Fig. 2(a) can be represented there by putting its internal resistance $R$ equal to $1/gm \cdot \mu/\mu+1$.




(a) (b)

Fig. 2—The cathode follower and equivalent circuit.

Nonlinearity of tube characteristics is neglected in the mathematical analysis, although its effect is demonstrated qualitatively in the final equations. As explained later, its presence does not greatly affect the usefulness of the results.

## III. Resistive Cathode Load

Before commencing a discussion of transient behavior in the equivalent circuit, it is desirable to write down some of the relations needed which apply to the simple resistive-loaded case. These relations hold when the transition time of the applied signal is long compared with the time constant of the cathode load $R_k C_k$, so that the capacitance $C_k$ may be ignored.

The gain of a cathode follower with cathode load $R_k$ is given by the ratio of the cathode voltage $v_k$ to the grid voltage $v_g$:

$$A = \frac{v_k}{v_g} = \frac{R_k}{R + R_k} \cdot \frac{\mu}{\mu + 1}. \tag{1}$$

A more useful concept turns out to be the gain of the equivalent circuit $A'$, given by the ratio of the cathode voltage $v_k$ to the input voltage of the equivalent circuit $v_g'$.

$$A' = \frac{v_k}{v_g'} = \frac{R_k}{R + R_k}. \tag{2}$$

It is also necessary to define the maximum output voltage swing $V_0$ as the maximum amount by which the output voltage can change in a given direction from any particular reference level. In a negative direction the limit is set by cathode current cutoff, or sometimes sim-

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[1] Transition times for common circuits are given by a number of authors. See, in particular, H. E. Kallmann, R. E. Spencer, and C. P. Singer, "Transient response," Proc. I.R.E., vol. 33, pp. 169–196; March, 1945.

ply an increase in distortion. In a positive direction the usual limitation is the drawing of grid current. If $I_0$ is the corresponding cathode-current change,

$$V_0 = R_k I_0 \tag{3}$$

## IV. Capacitive Cathode Load

Turning now to the case of an applied signal with a short transition time, we may divide the analysis into two parts. The first part deals only with the small-signal response. The second deals with the limitation on maximum output voltage caused by the short transition time of the applied signal.

As far as the small-signal response is concerned, it is sufficient to note that the transient response of a cathode follower is the same as that of a resistance-coupled pentode amplifier with plate load $A'R$ (consisting of $R$ and $R_k$ connected in parallel) shunted by a capacitance of $C_k$. The behavior of such a circuit to various signal forms is well known.[1] In particular its response to a linear transition is given by equation (9) of the Appendix. It is seen that, provided $\exp(-1.25T/A'RC_k)$ is small, the signal is reproduced with little distortion and delayed by a time $A'RC_k$.

A cathode follower which has an adequate small-signal transient response may, however, display severe transient distortion on a signal which would appear to have an amplitude well within the capabilities of the tube. The reason is that in reproducing accurately a linear transition, a cathode follower must supply a steady current of $dv_k/dt \cdot C_k$ to the capacitance $C_k$ during the rise (or fall) time of the signal. The available current swing through the load resistance $R_k$ is therefore reduced, and consequently the signal-handling capabilities of the tube may be severely restricted. The following relation between the maximum output-voltage swing possible with a signal transition time $T$ and the maximum available voltage swing (i.e., for $T=\infty$) is derived in the Appendix:

$$V_T = \frac{V_0}{1 + A'\dfrac{R_k C_k}{1.25T}} \tag{4}$$

where $V_T$ is the maximum output voltage swing with transition time of $T$, and $V_0$ is the maximum available output voltage swing.

It is assumed that $\exp(-1.25T/A'RC_k)$ is small or, in other words, that the transient distortion with a small input signal is not large. The equation applies to both positive and negative transitions.

In a cathode follower handling a positive signal and biased near anode-current cutoff, the limitation occurs usually on the negative-going trailing edge, and its effect is often apparent as a considerable lengthening of this portion of the signal. In such a case the distortion is the result of insufficient available voltage swing in a negative direction, which causes the cathode current to be cut off. The effective time constant of the cathode circuit is thereby changed from $A'RC_k$ to $R_kC_k$, with consequent distortion.

Equation (4) can be shown to apply also when $C_k$ is replaced by an open-circuited transmission line with propagation time considerably shorter than $T$, and with a total capacitance of $C_k$. This is analogous to the similar and more familiar representation of an open-circuited transmission line carrying a sine-wave signal of a frequency considerably less than the lowest resonant frequency of the line.

## V. Cathode Follower Design

For design purposes it is better to write (4) in a different form. In Fig. 3 are represented the cathode voltages in a cathode follower handling a typical positive signal. Consider the negative transition, as it is the one in which trouble is usually expreienced. It will be assumed that the signal amplitude is the maximum which can be handled without cathode-current cutoff.

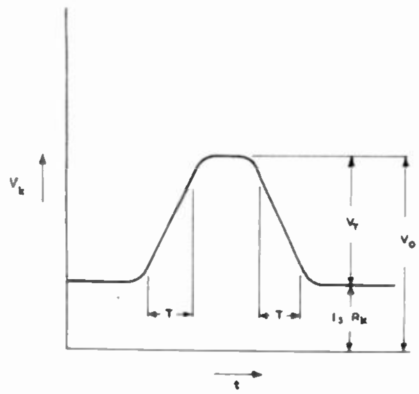


Fig. 3—Cathode voltages with a typical positive signal.

The cathode current necessary to maintain a steady voltage across $R_k$ equal to the minimum cathode voltage is taken as $I_s$. The available voltage swing $V_0$ is then equal to $V_T + I_sR_k$, whence (4) takes the form

$$V_T = \frac{V_T + I_s R_k}{1 + A'\dfrac{R_k C_k}{1.25T}}.$$

which, on rearranging, becomes

$$I_s = V_T \cdot \frac{A'C_k}{1.25T}. \tag{5}$$

In the design of a cathode follower in which the negative transition is the determining factor, it is therefore necessary to ensure that the minimum steady cathode current $I_s$ is greater than that obtained from (5). This equation is based on the ideal linear circuit of Fig. 2(b). Actually, of course, the internal resistance of a cathode follower increases steadily as the anode current is decreased. This will result in a higher possible value of $V_T$

before cutoff occurs because of the decrease in $A'$, but a lower value before transient distortion increases perceptibly. In practice, the equation as given will often represent a suitable compromise. If it is necessary to ensure that the internal resistance $R$ should never exceed a certain value, then $I_s$ must be increased by an amount equal to the cathode current corresponding to the maximum allowable internal resistance.

It will be observed that the signal-handling capacity apparently does not depend to any great extent on $R_k$, provided that the gain $A'$ is near unity. This is because the negative transition only has been considered. In a cathode follower designed to have the minimum value of $I_s$, the peak cathode current will be given by

$$I_{(\max)} = \frac{V_T}{R_k} + 2I_s. \tag{6}$$

The current $I_s$ is doubled because an extra current of this magnitude is required to reproduce the positive transition.

The peak current must, of course, be within the capabilities of the tube, and usually it is desirable that it should be as small as possible in order to keep the tube dissipation to a minimum.

Usually, therefore, the cathode resistor should be made as large as possible. In the most common practical circuit the signal is coupled to the cathode follower grid via a capacitor and grid-leak combination in which the cold end of the grid leak is at earth potential. The minimum grid voltage is then zero, or even negative, if the duty cycle of the signal is appreciable.

The maximum value of $R_k$ is then limited to that which produces a cathode current $I_s$ given by (5) when the grid has this minimum voltage. Often a more economical design is, therefore, one in which the cathode resistor is larger than this, and the grid leak is returned to a positive voltage sufficient to keep $I_s$ above the minimum value.

## VI. Acknowledgment

This work was carried out as part of the research program of the Division of Radiophysics, Council for Scientific and Industrial Research, Commonwealth of Australia.

## APPENDIX

### Maximum Voltage Output

A cathode follower with capacitive load may be represented by the simplified circuit of Fig. 4, on applying Thévénin's theorem to the euqivalent circuit of Fig. 2(b). The input signal to this circuit is then the output of the cathode follower in the absence of the capacitance $C_k$, while the voltage across the capacitance $C_k$ is the actual cathode voltage $v_k$. The input signal to be considered is a linear transition of magnitude $V_T$ and transition time $T$. It is required to find the relationship between this applied signal and the maximum change of cathode current $I_0$ at time $1.25T$.

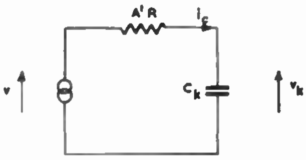


Fig. 4—Simplified equivalent circuit for calculation. of the instantaneous value of the cathode voltage $V_k$.

The slope of the applied signal is equal to $V_T/1.25T$; that is,

$$v = \frac{V_T}{1.25T} t. \tag{7}$$

It may be shown that, if a signal consisting of a linearly changing voltage with the above slope is applied to the circuit of Fig. 4, the magnitude of the current $i_c$ flowing into the capacitor during the transition time of the signal is given by

$$i_c = \frac{V_T}{1.25T} \cdot C_k \left\{ 1 - \exp\left( \frac{-t}{A'RC_k} \right) \right\}. \tag{8}$$

The voltage $v_k$ across the capacitor is then given by

$$\begin{aligned} v_k &= v - A'Ri_c \\ &= V_T \left[ \frac{t}{1.25T} - \frac{A'RC_k}{1.25T} \left\{ 1 - \exp\left( \frac{-t}{A'RC_k} \right) \right\} \right]. \end{aligned} \tag{9}$$

The total cathode current is the sum of current flowing into $C_k$ and $R_k$. The former is given by (8) above; the latter is equal to $v_k/R_k$. Thus, on adding these and rearranging, we have, for the cathode current,

$$i = \frac{V_T}{R_k} \left[ \frac{t}{1.25T} + \frac{A'R_kC_k}{1.25T} \left\{ 1 - \exp\left( \frac{-t}{A'RC_k} \right) \right\} \right] \tag{10}$$

The maximum value of $i$, equal to $I_0$, occurs at a time of $1.25T$. Inserting these values for $i$ and $t$ in (10) and rearranging, we have

$$V_T = \frac{I_0 R_k}{1 + A' \dfrac{R_kC_k}{1.25T} \left\{ 1 - \exp\left( \dfrac{-1.25T}{A'RC_k} \right) \right\}}. \tag{11}$$

Since the requirement of small transient distortion dictates a small value for $\exp(-1.25T/A'RC_k)$, this equation may normally be simplified to

$$V_T = \frac{V_0}{1 + A' \dfrac{R_kC_k}{1.25T}}. \tag{4}$$

# A Microwave Impedance Bridge*

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*Summary*—A six-arm waveguide structure is described which is a true microwave equivalent of a Wheatstone bridge. A theoretical analysis of the equivalent circuit of the device has been made, using the symmetry properties of the structure. The resulting relation among the admittances of the various arms is exactly that of a Wheatstone bridge with shunting susceptances across each pair of terminals. A device of this sort for use at 10 cm has been built and tested, and was found to behave as predicted.

With this bridge it is possible to measure any impedance to about the same accuracy as with a standing-wave detector. A valuable feature of this instrument is that the standard impedances required are three variable reactances (movable shorting plungers) and a $Z_0$ termination. The data are obtained in the form of three lengths, the positions of the movable shorts.

Since the device is the complete equivalent of a Wheatstone bridge, it can also be used as a four-terminal lattice section in filter design or in any other related application requiring the microwave equivalent of a lattice section. This allows all the greater flexibility which lattice sections have, as compared to tee or pi sections commonly used in microwave work.

## I. Introduction

Any COMPARISON method of measuring impedances at microwave frequencies requires two things: (1) some sort of device which can function as a bridge, and (2) an impedance standard. There have been a number of microwave devices invented which measure impedances by comparison. Few of them, however, are actually bridges in the sense that one uses the term at low frequencies. A true bridge of the Wheatstone type obeys the equation

$$AB = CD$$

or

$$B = \frac{C}{A} D. \tag{1}$$

Here, $A$, $B$, $C$, and $D$ are four impedances, three of of which are known and the other is unknown. By this criterion, for example, a magic-tee or hybrid junction is not a bridge, for it obeys the equation

$$A = B. \tag{2}$$

The difference between (1) and (2) is not academic. For (2) it is necessary to provide an adjustable impedance which will exactly equal the unknown. That is, the standard impedance will require both a variable resistance and a variable reactance. A variable resistance is difficult to provide at microwave frequencies, and this is a drawback to the use of a device based on (2). With a true bridge which obeys (1), however, it is possible to match any impedance by use of a fixed resistance only plus variable reactances, both of which, in practice, are quite simple to make at microwave frequencies. For example, let $B$ be the unknown. Then, if $A$ and $C$ are variable reactances consisting of shorted sections of transmission line of variable length, the ratio $C/A$ can be adjusted to have any real value desired. $D$ consists of a perfectly terminated line in series with a variable reactance, which is another shorted-section line. By adjusting the latter, $D$ can be made to have any arbitrary phase angle. Therefore $D$ can be adjusted to have the phase angle of the unknown, and the ratio $C/A$ can be adjusted to give the correct magnitude. The value of the unknown $B$, then, is determined from the three physical lengths of stubs $A$, $C$, and $D$.

A Wheatstone bridge satisfying (1) has six-terminal pairs, the four impedance arms, and an input and output arm, and therefore such a device at microwave frequencies requires connecting six waveguides or transmission lines in a suitable way so that the combination is a bridge. This is not topologically simple.

The device shown in Fig. 1 is such a combination, and is a true microwave equivalent of a Wheatstone

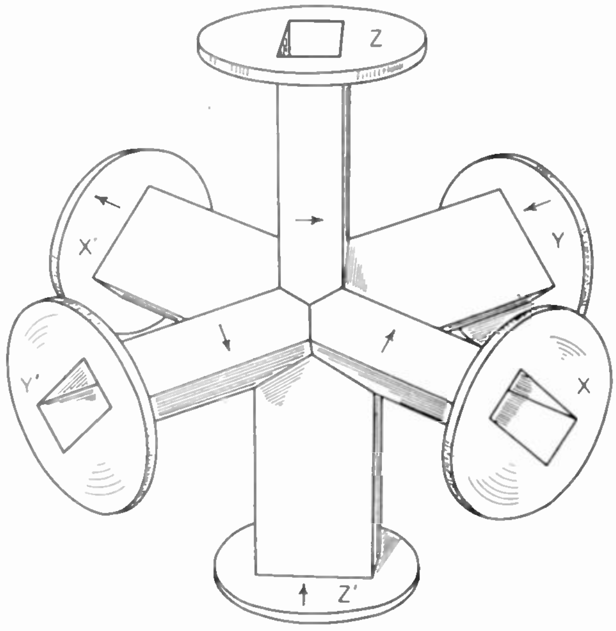


Fig. 1—A waveguide impedance bridge.

bridge. It was first suggested by R. Dicke, formerly of the MIT Radiation Laboratory. It is made by joining six pieces of rectangular waveguide at a symmetrical junction. Each waveguide then represents an arm of the bridge, or the input, or the detector. Opposite pairs of arms, such as $A$ and $C$ or $B$ and $D$, lie on common axes but are rotated 90° with respect to each other. A bridge of this sort has been constructed and tested. It is shown

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in Fig. 2. It was made of $3\times1\frac{1}{2}$ inch waveguide, which is the standard size used in the 10-cm region.

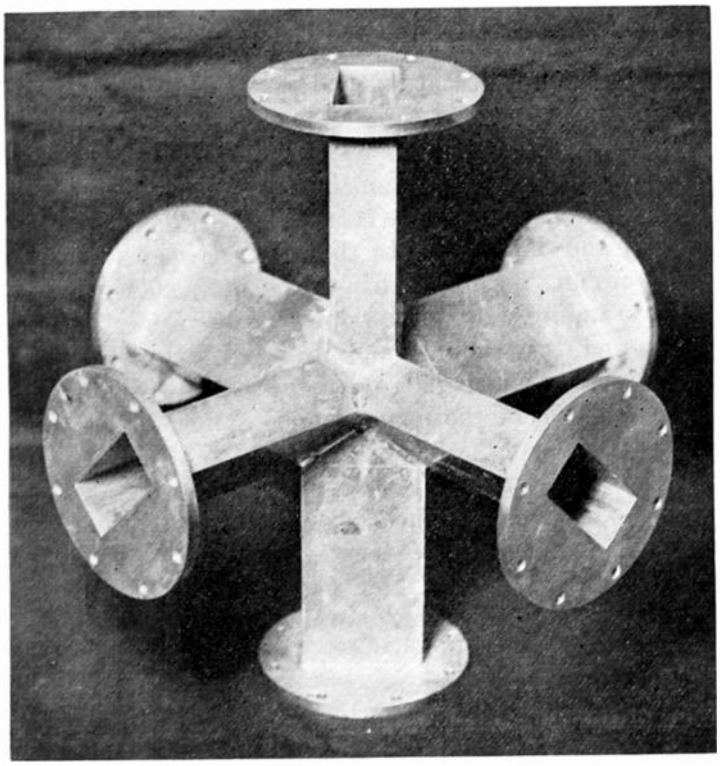

Fig. 2—A photograph of an early waveguide bridge.

That the device shown in Fig. 1 is a Wheatstone bridge appears plausible from consideration of the geometry and the way in which the various guides are connected. Although this intuitive picture indicates that the device is a true bridge, it was thought useful to show analytically that this is indeed the equivalent circuit.

The procedure was to select sets of waves in the structure, such that the field distribution due to any arrangement of terminating admittances and sources can be represented by linear combination of these waves. Then, upon consideration of these waves and the symmetry properties of the bridge, relations between the input admittance at one terminal pair and all the other five terminating admittances can be found. The waves that are set up are for the lowest mode in the waveguide only; that is, it is assumed that one is measuring the field sufficiently far from the junction of the guides so that all the higher modes which are present in the fringing fields at the junction are attenuated. The effect of the junction appears in the equation for the bridge, in the form of equivalent lumped susceptances at suitably chosen reference planes in the guides. The expression for the admittance relation among the six arms is

$$\begin{aligned} y_x y_{x'} y_z &+ y_x y_{x'} y_{z'} + y_x y_{x'} y_y + y_x y_{x'} y_{y'} + y_y y_{y'} y_x \\ &+ y_y y_{y'} y_{x'} + y_y y_{y'} y_z + y_y y_{y'} y_{z'} + y_z y_{z'} y_x \\ &+ y_z y_{z'} y_{x'} + y_z y_{z'} y_y + y_z y_{z'} y_{y'} + y_x y_y y_z \\ &+ y_{x'} y_{y'} y_z + y_x y_{y'} y_{z'} + y_{x'} y_y y_{z'} = 0, \end{aligned} \tag{3}$$

where

$$y_x = Y_x + jb \qquad y_{x'} = Y_{x'} + jb, \text{ etc.,}$$

and $Y_x$ is the terminating admittance on arm $x$, etc. This is the equation of a Wheatstone bridge with fringing susceptance of magnitude $jb$ across each terminal. The derivation of (3) will be given in a subsequent section.

In terms of this general admittance relation, the condition for a balanced bridge is that the input admittance is independent of the detector admittance, and a relation of the form (1) will fulfill this requirement. This balance equation, as modified by the fringing susceptances and using a slightly different notation, can be written as

$$(Y_A + jb)(Y_B + jb) = (Y_C + jb)(Y_D + jb). \tag{4}$$

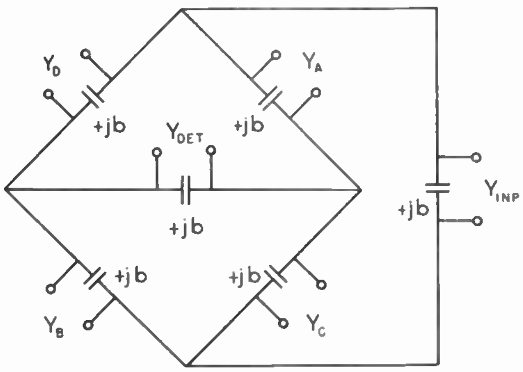


Fig. 3—The exact equivalent circuit of the waveguide bridge.

The equivalent circuit is shown in Fig. 3.

## II. Construction and Experimental Investigation

The bridge to be described here was made of $3\times1\frac{1}{2}$ inch waveguide and was operated in the $TE_{10}$ mode. Such a bridge can be made for rectangular waveguide of any dimensions. The ends of the waveguides which come together at the junction have triangular tongues cut in each side with angles as shown in Fig. 4. When the

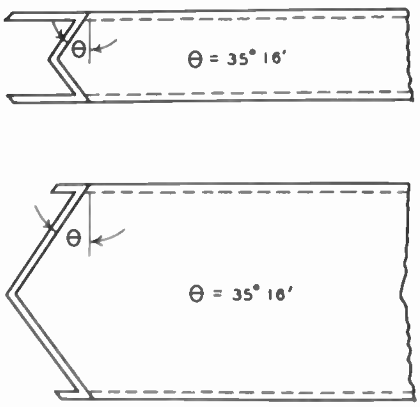


Fig. 4—One arm of the six-arm waveguide bridge.

six guides are placed together, these tongues mate to give the correct orientation for all the guides. The structure can then be clamped and soldered. Measurements were made on the bridge to test the correctness of the equivalent circuit shown in Fig. 3.

As is always the case, for such equivalent circuits to be applicable, the admittances or impedances must be referred to certain special planes in the various arms of the structure. In this case, the equivalent plane for any of the bridge arms is the position of a shorting plunger which will cause a short to appear in the corresponding branch of the equivalent circuit. It is clear that, if the admittances $Y_A$, etc., which are measured at such equivalent planes satisfy an equation of the form (4), they will, in general, no longer satisfy it after they have been transformed an arbitrary distance down the line

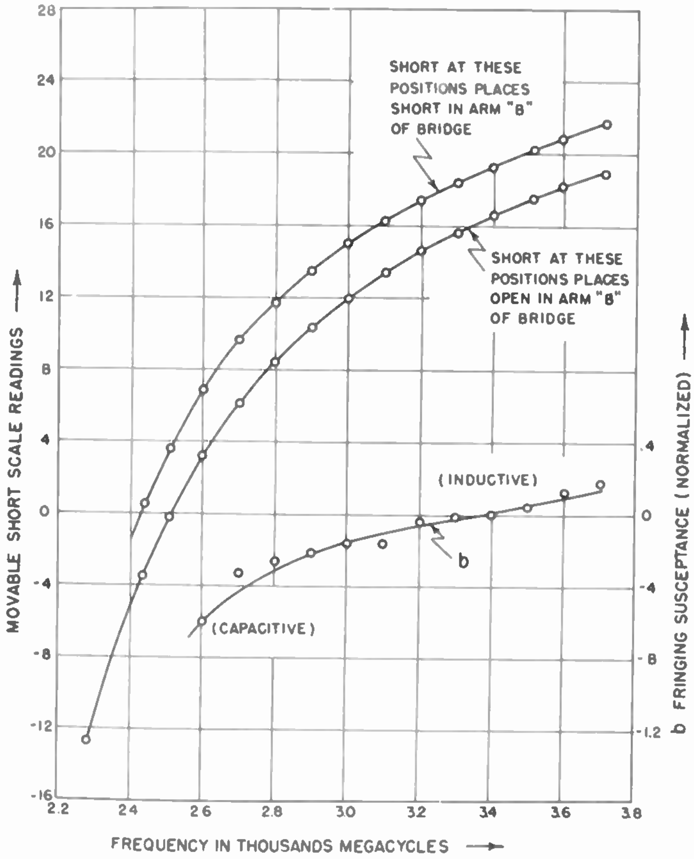


Fig. 5—Positions of critical planes versus frequency for one arm of six-arm bridge (without reactance tuning screw).

from the place at which they were originally defined. These equivalent planes can be located by moving shorts in two adjacent arms, such as $A$ and $C$, until the standing-wave ratio in the input circuit is infinite. In this case, the positions of the shorts determine the equivalent planes in $A$ and $C$. The position of the node in the standing wave in the input arm also defines the equivalent plane in that arm.

A more practical method for locating the equivalent plane was actually used. This method depends on the fact that when shorts are located at the equivalent planes of $A$ and $C$, corresponding to infinite admittances in the branch $A$–$C$, then the output of the detector is zero, independent of variations in $Y_B$ and $Y_D$. The admittances $Y_B$ and $Y_D$ were modulated by setting rotating electric fans near the open ends of arms $B$ and $D$, resulting in a modulated detector output. The short positions in $A$ and $C$ were varied until the detector output was reduced to zero, which determined the equivalent planes for these arms, and also in the input arm. Planes in the other arms are located in the same way, by choosing different input arms.

There is another position of the shorts which gives zero detector output. This corresponds to the case when $Y_A+jb$ and $Y_C+jb=0$. In this case, as seen from the input circuit, the bridge has a finite impedance and a finite standing wave ratio which distinguishes it from the former case when $Y_A$ and $Y_C$ were infinite. This second way of producing zero detector output permits the determination of the magnitude $jb$. Thus, if the position of the short, which makes the total admittance $Y_A+jb=0$, is a distance $l$ from the equivalent plane, $Y_A=-jY_0\cot 2\pi l/\lambda g$ and $b=Y_0\cot 2\pi l/\lambda g$. The value of this fringing susceptance was found to be about 0.2 $Z_0$ at 10 cm. Due to asymmetries in construction, this varied slightly from arm to arm by about 10 per cent. It was also frequency-dependent. Fig. 5 shows the value of this fringing susceptance and also the relative position of the equivalent plane as a function of frequency. The value of the fringing susceptance becomes zero and changes sign at 3.4 kMc. This fringing susceptance can be eliminated at any particular frequency by inserting a capacitative screw at the position of the equivalent plane or a quarter wavelength away, depending on the sign of the $jb$. No attempt was made to provide a broad-band susceptance which would eliminate the fringing susceptance over a wide range of frequency. In use as a bridge, a calibrated moveable short was placed on arm $D$ in such a position that normalized admittance $1+jY$ appeared at the equivalent plane. Moveable shorts were attached to arms $A$ and $C$ and the admittance to be measured was attached to arm $B$. If the fringing susceptance has been eliminated as described above, then $Y_B$ is given by the equation

$$Y_B=\frac{Y_C}{Y_A}Y_D.$$

Since $Y_C/Y_A$ is a dimensionless number $M$, $MY_D=M(1+jY)$ will cover the whole admittance plane. A Smith chart, Fig. 6, shows a comparison of some impedances measured first by the bridge described and then by a conventional standing-wave detector using a nodal-shift method.

Once properly adjusted, the bridge is simple to use, and measurements can be made rapidly. The unknown impedance is determined in terms of three physical lengths, the distances of the shorts from their correct reference planes, which is a very simple standard and is independent of detector calibration. The accuracy of the measurement is equivalent to that of a good standing-wave detector. The use of the three physical lengths to specify the unknown impedance can be facilitated with the aid of simple formulas or charts. Although, in the current version of the bridge described here, the fringing susceptance was eliminated for only one frequency, it should be possible by conventional broadbanding techniques to make the device quite broadband, so that the fringing susceptance is almost zero over a considerable range of frequency.

## III. Theory

To find the equivalent circuit for the structure shown in Fig. 1, it is necessary to set up suitable basis fields in the various arms such that any possible field can be represented as a linear combination of these. Then, by consideration of the symmetry properties of the structure, it is possible to get the relation between admittances in the various arms. This method is similar

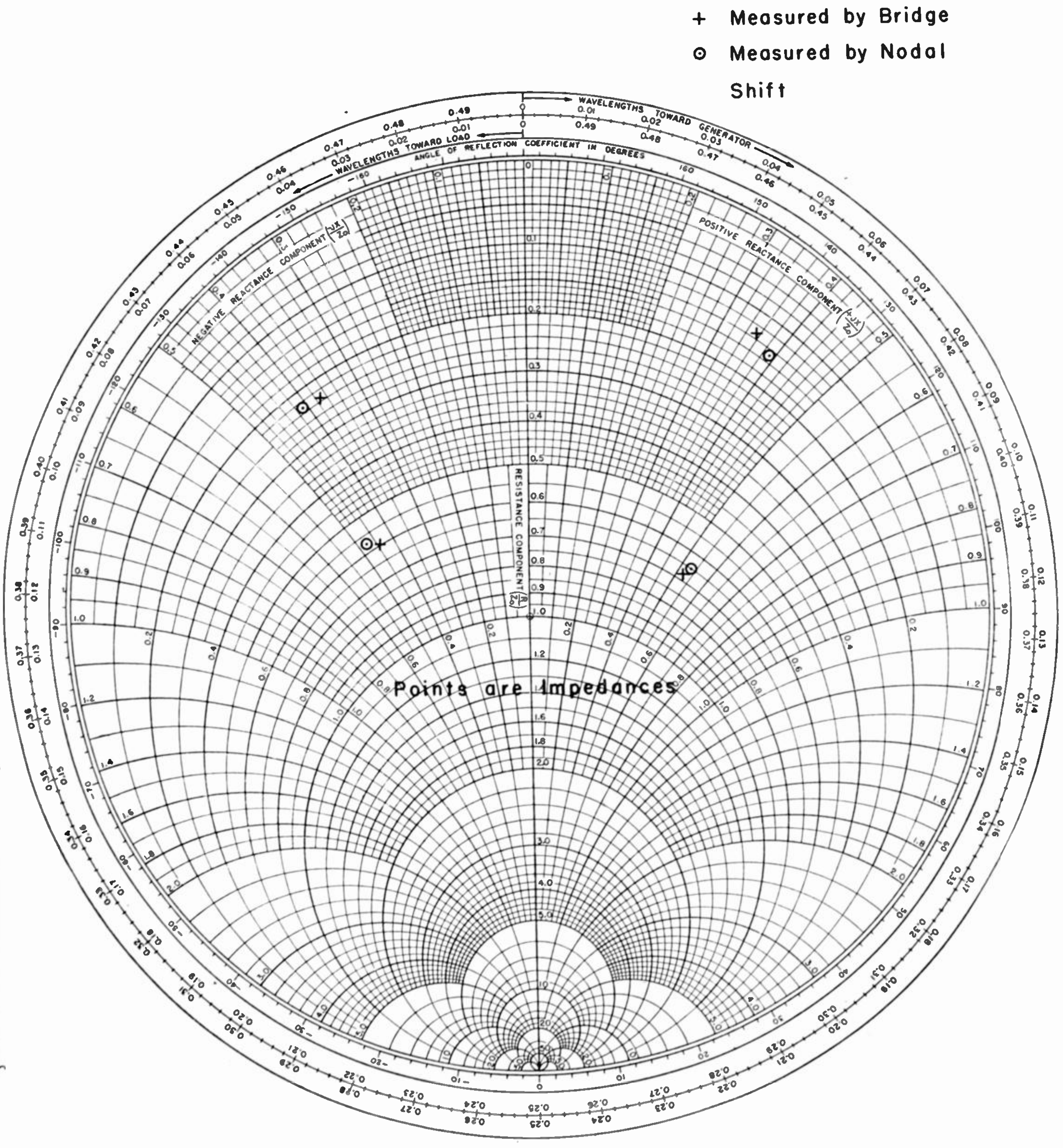


Fig. 6—Showing some bridge measurements compared to known impedances determined by nodal shift technique.

to that used by Slater[1] to obtain equivalent circuits for series and shunt tees and similar structures. In the cases treated by Slater it is possible to select the basic field distributions by inspection. The bridge discussed here has a more complicated symmetry, and the choice of a suitable set of basic fields is facilitated by use of the formal methods of group theory. The latter, however, are not indispensable.

The field in a waveguide can be represented in the following fashion (co-ordinates $u$, $v$, $z$):

$$E_u = \frac{2}{ab}\frac{\sin \pi v}{a} e^{j\omega t} V(z)$$

$$H_v = -\frac{2}{ab}\frac{\sin \pi v}{a} e^{j\omega t} I(z)$$

where $V(z)$ and $I(z)$ represent an equivalent voltage and current. It is possible to define an impedance

$$\frac{E_u}{-H_v} = \frac{V(z)}{I(z)}.$$

For a traveling wave in $\pm z$ direction,

$$\frac{V(z)}{I(z)} = \pm Z_0 = \pm 377 \frac{\lambda_g}{\lambda_0},$$

and

$$V = \pm IZ_0 = V_n e^{\pm(2\pi z/\lambda_g)j} = V_n e^{\pm \beta z j}$$

where $V_n$ is the amplitude and $\beta = 2\pi/\lambda_g$. Instead of traveling waves, it will be convenient to use standing waves expressed in terms of $I$ and $V$.

It is necessary to choose a reference plane for these standing waves and also a direction of the positive $E$ vector in each arm. Choose axes $x$, $x'$, $y$, $y'$, $z$, $z'$ along each arm as shown in Fig. 1, all having positive direction outward from the junction. The direction of the positive $E$ vector in each is also shown in the diagram. The functions will be expressed in terms of $\sin \beta z$ and $\cos \beta z$, etc., but these have no significance so far as evenness or oddness with respect to the junction is concerned. Since the $E$ vector in $z'$ waveguide and the $z$ waveguide are oriented at right angles to each other, there is no such simple symmetry property through the junction for the $E$ vector. Instead, the cosine and sine merely refer to the behavior of the function in its own particular waveguide, as referred to some arbitrary reference plane. These planes are assumed to be located at the same position in each guide. It turns out that the appropriate functions can be divided into two sets of three each. The three functions constituting each set are degenerate in the sense that one function describes a distribution in which there is a field in four of the guides, and the other two are obtained from it by permuting some of the indices. This will be somewhat clearer after we have written down the functions.

Listed below are the three functions comprising the first set:

| I(1) | II(1) | III(1) |
|---|---|---|
| $A \sin \beta z$ | $B \sin \beta z$ | $C \sin \beta x$ |
| $A \sin \beta z'$ | $-B \sin \beta z'$ | $C \sin \beta x'$ |
| $-A \sin \beta x$ | $B \sin \beta y$ | $C \sin \beta y'$ |
| $A \sin \beta x'$ | $-B \sin \beta y'$ | $C \sin \beta y$ |

It can be seen that these are independent, since it is not possible to get any one of them by linear combination of the other two.

The symmetry of this set can be illustrated by considering the function I(1). For the fields in the $z$ and $z'$ guides, the orientation is such that in the junction they can only have a resultant along the $y$ or $y'$ axis, and therefore cannot cause any fields in those wave-guides. The fields in the $x$ and $x'$ guides have the same properties. However, from the way we have chosen the signs and since the reference planes are at the same position in all of the guides, the resultant of the $x$, $x'$ function will be in the opposite direction from that of the $z$, $z'$ function. Another way of saying this is to consider the fields of I(1) as propagating unchanged from the waveguides into the junction region, i.e., as if the four waveguides concerned intersect each other without interfering with each other. In this case, at the midpoint, waves from all of these would cancel and give zero resultant.

For the next set there is a different type of symmetry. This set can be written as:

| I(2) | II(2) | III(2) |
|---|---|---|
| $D \cos(\beta z - \phi)$ | $E \cos(\beta z - \phi)$ | $F \cos(\beta x - \phi)$ |
| $D \cos(\beta z' - \phi)$ | $-E \cos(\beta z' - \phi)$ | $F \cos(\beta x' - \phi)$ |
| $D \cos(\beta x - \phi)$ | $-E \cos(\beta y - \phi)$ | $-F \cos(\beta y - \phi)$ |
| $D \cos(\beta x' - \phi)$ | $E \cos(\beta y' - \phi)$ | $-F \cos(\beta y' - \phi)$ |

The fact that this has been written as a cosine is of no significance as far as the evenness or oddness about the junction is concerned, and the phase angle in the cosine merely means that the zero amplitude for these functions is not necessarily 90° from the reference plane of the first set. In general, the zero plane for one set will be shifted some arbitrary amount with respect to that of the other set. This depends on the details of the discontinuity at the junction and cannot be determined by this type of theory. It will appear in the theory as a fringing susceptance which can be determined empirically, once a fixed reference plane is selected for each guide. The significant thing about this set can be found by examining the behavior of the vectors in each waveguide where they come together at the junction. It is found that they behave differently than the first set. For example, if we consider functions I(2), the resultant of the $z$, $z'$ functions will again be a vector along the $yy'$ axes. The $xx'$ functions also have a resultant vector along the $yy'$ axes, but in this case as opposed to the set (1) the two vectors due to $zz'$ set and $xx'$ set are in the

[1] J. C. Slater, "Theory of Symmetrical T's," Research Laboratory of Electronics, MIT, Technical Report No. 37.

same direction at the junction midpoint. If we again imagine as before that the waveguides extend right through the junction intersecting without interfering with each other, the resultant vector at the center of the junction would be the sum of all four, and would not cancel as it did in set (1). The fact that this set is of different symmetry than the previous set can be shown rigorously by means of group theory. The necessity for this symmetry as well as for the other can also be shown. The details of the proof are omitted. Briefly it may be stated that the structure described here has the symmetry of the tetrahedral group and the functions I and II transform under different irreducible representations of this group.

The requirement of the phase shift $\phi$ for these functions as opposed to the first set is not proved since there is no symmetry property which determines this. It is to be stressed, however, that even if by some accidental property $\phi$ should equal 90°, i.e., the functions of both sets (1) and (2) should be sine functions in each guide, the symmetry of these functions with respect to the whole structure is not the same. It would still be true that, with regard to the whole structure, it would not be possible to form one distribution by a linear combination of the other five.

In terms of these distributions, it is now possible to set down the general expression for the voltage and current in each arm.

$$V_x = -A\sin\beta x + C\sin\beta x + D\cos(\beta x - \phi) + F\cos(\beta x - \phi)$$

$$jI_xZ_0 = +A\cos\beta x - C\cos\beta x + D\sin(\beta x - \phi) + F\sin(\beta x - \phi)$$

$$V_x' = A\sin\beta x' + C\sin\beta x' - D\cos(\beta x' - \phi) + F\cos(\beta x' - \phi)$$

$$jI_x'Z_0 = A\cos\beta x' - C\cos\beta x' - D\sin(\beta x' - \phi) + F\sin(\beta x' - \phi)$$

$$V_y = B\sin\beta y + C\sin\beta y - E\cos(\beta y - \phi) - F\cos(\beta y - \phi)$$

$$jI_yZ_0 = -B\cos\beta y - C\cos\beta y - E\sin(\beta y - \phi) - F\sin(\beta y - \phi)$$

$$V_y' = -B\sin\beta y' + C\sin\beta y' + E\cos(\beta y' - \phi) - F\cos(\beta y' - \phi)$$

$$jI_y'Z_0 = B\cos\beta y' - C\cos\beta y' + E\sin(\beta y' - \phi) - F\sin(\beta y' - \phi)$$

$$V_z = A\sin\beta z + B\sin\beta z + D\cos(\beta z - \phi) + E\cos(\beta z - \phi)$$

$$jI_zZ_0 = -A\cos\beta z - B\cos\beta z + D\sin(\beta z - \phi) + E\sin(\beta z - \phi)$$

$$V_z' = A\sin\beta z' - B\sin\beta z' + D\cos(\beta z' - \phi) - E\cos(\beta z' - \phi)$$

$$jI_z'Z_0 = A\cos\beta z' + B\cos\beta z' + D\sin(\beta z - \phi) - E\sin(\beta z' - \phi).$$

If one selects as the plane of reference in each guide the plane where $\sin\beta x$, $\sin\beta x'$, etc. $=0$, that is, where the voltages of set (1) are identically zero, this simplifies the above equations considerably. One can then write the admittances at these planes; these admittances being measured looking away from the junction. These admittances are:

$$\frac{I_x}{V_x}jZ_0 = \frac{A-C}{D+F}\frac{1}{\cos\phi} - \tan\phi$$

$$\frac{I_{x'}}{V_{x'}}jZ_0 = \frac{A+C}{D-F}\frac{1}{\cos\phi} - \tan\phi$$

$$\frac{I_y}{V_y}jZ_0 = +\frac{B+C}{E+F}\frac{1}{\cos\phi} - \tan\phi$$

$$\frac{I_{y'}}{V_{y'}}jZ_0 = \frac{B-C}{E-F}\frac{1}{\cos\phi} - \tan\phi$$

$$\frac{I_z}{V_z}jZ_0 = -\frac{A+B}{D+E}\frac{1}{\cos\phi} - \tan\phi$$

$$\frac{I_{z'}}{V_{z'}}jZ_0 = \frac{B-A}{D-E}\frac{1}{\cos\phi} - \tan\phi.$$

Using the notation

$$y_x = \frac{I_x}{V_x} - j\frac{\tan\phi}{Z_0}, \text{ etc., or } y_x = Y_x + jb$$

and eliminating the coefficients $A$, $B$, $C$, $D$, $E$, and $F$, one obtains (3).

## IV. Four-Terminal Application

It should be pointed out that since this structure is the exact equivalent of a bridge, it can be used as a four-terminal lattice section in microwave filter design or in similar applications. Most microwave filter structures used to date have been the equivalents of ladder sections. As is well known, a lattice section is the most general form of four-terminal network, and offers much greater flexibility in design.[2]

For example, the image impedances and transfer constant can be separately adjusted. Also, any physical four-terminal network can be synthesized by means of lattice sections, while this is not always true for ladder sections.

[2] E. A. Guillemin, "Communication Networks," Vol. II, p. 378 ff., John Wiley and Sons, New York, N. Y.

# Closed- and Open-Ridge Waveguide*

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*Summary*—**Expressions are developed for the voltage-current and voltage-power impedance of closed-ridge and open-ridge waveguide, with the discontinuity capacitance taken into account. Approximate expressions for impedance are derived which are valid under given typical conditions.**

## I. Closed-Ridge Waveguide

In a recent paper, Cohn presented much information on the cutoff frequency and impedance of closed-ridge waveguide.[1] He noted excellent experimental checks of theoretical calculations for cutoff frequency; also for impedance, providing the ratio of guide height to width was small. However, when this ratio was 0.5, calculated impedances were approximately 25 per cent too high. He stated, correctly, that the reason for this discrepancy was the partial neglect of the discontinuity capacitance at the edges of the ridge in the derivation of the impedance equation. In the following calculation, the effect of this susceptance will be taken into account, giving expressions which are only slightly more complex than those derived by Cohn. In fact, Cohn's curves may be used to obtain a preliminary answer, which is then modified to include the total effect of the discontinuity.

Cohn's derivation of impedance was based upon the simplifying assumption that $E$ lines within the guide run vertically from top to bottom. At the edges of the ridge, voltage was made continuous by assuming a discontinuity in $E$ at that point. Impedance was defined as the ratio of voltage across the center of the guide to total longitudinal current on the top face. This current was calculated on the basis of the assumed distribution of $E$, thus neglecting the effect of the higher order modes

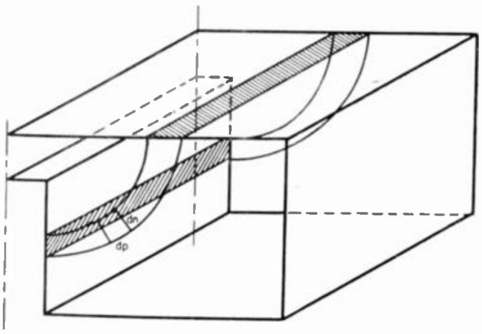


Fig. 1—$E$ lines terminating on ridge side.

which are necessary to satisfy the boundary conditions at the edges of the ridge. As a result, when the ridge step was of appreciable size, the omission of the contribution of current due to the capacitance of the ridge edges resulted in a theoretical impedance which was too high.

This discontinuity current is readily evaluated. Consider two banks of $E$ lines terminating on the side of the ridge in a ridge waveguide, as in Fig. 1. If their spacing is $dn$, the capacitance per unit length of the shaded conductors contained between the terminations of the $E$ lines may be written as

$$dC = \frac{\epsilon_0 dQ}{\int E dp} = \frac{\epsilon_0 E dn}{V}.$$

The longitudinal current in either conductor is given by the tangential component of transverse $H$, i.e.,

$$dI_z = H_n dn = \frac{1}{Z_{TE}} E dn$$

where $Z_{TE}$ is the ratio of transverse $E$ to transverse $H$ for a $TE$ mode,

$$\frac{E_t}{H_t} = Z_{TE} = \frac{120\pi}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}}.$$

Hence,

$$dI_z = \frac{1}{\epsilon_0 Z_{TE}} V dC. \quad (1)$$

Equation (1) states that current flow down the guide may be obtained by considering the product of voltage and capacitance. It is necessary to apply this equation only at the discontinuity, since the current in the rest of the guide walls has been determined by Cohn. In ridge guide there are two discontinuity capacitances at the edges of the ridge. Their presence, therefore, increases the total longitudinal current by the amount $2VC_d/\epsilon_0 Z_{TE}$, where $C_d$ is the familiar Whinnery and Jamieson discontinuity capacitance[2] given in Fig. 2 of this paper. Strictly speaking, this value of $C_d$ must be modified if the ridge edges are not far enough from the sides of the guide that reflection of higher-order modes is appreciable. However, with Cohn, these proximity effects are neglected in the determination of $C_d$.

The derivation of an expression for impedance may now be carried through in a manner similar to that of Cohn, except that his expression for total longitudinal

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[1] S. B. Cohn, "Properties of ridge waveguide," Proc. I.R.E., vol. 35, pp. 783–788; August, 1947.

[2] J. R. Whinnery and H. W. Jamieson, "Equivalent circuits for discontinuities in transmission lines," Proc. I.R.E., vol. 32, pp. 98–114; February, 1944.

current must be augmented by the discontinuity current, namely, $2C_dE_0b_2 \cos\theta_2/\epsilon_0 Z_{TE}$. Since his impedance has been based on the ratio of voltage to current,

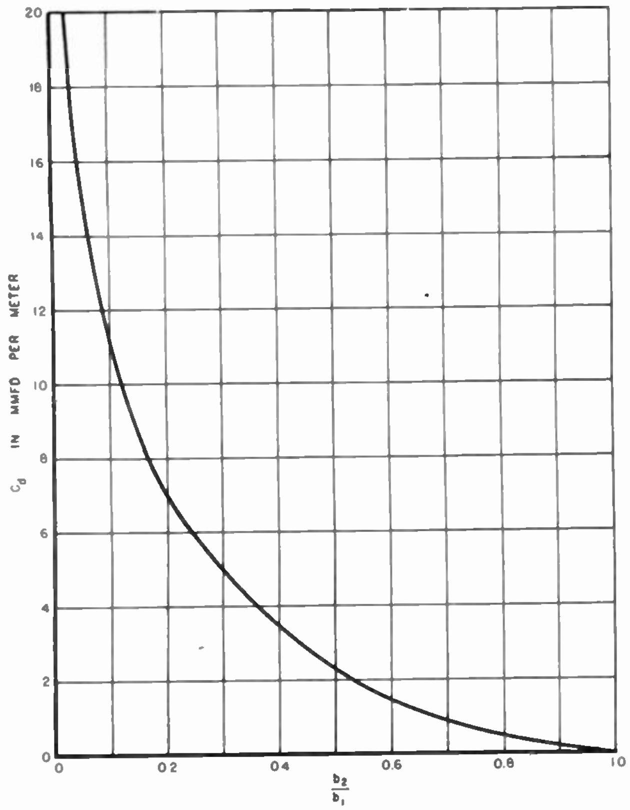



Fig. 2—Discontinuity capacitance as a function of ridge step.

it will be termed $Z_{vi}$. Accordingly, his equation (3) becomes

$$Z_{vi}^{\infty} = \frac{120\pi}{\dfrac{2C_d\cos\theta_2}{\epsilon_0} + \dfrac{\lambda_c'}{\pi b_2}\left[\sin\theta_2 + \dfrac{b_2}{b_1}\cos\theta_2 \tan\dfrac{\theta_1}{2}\right]}, \quad (2)$$

where

$$Z_{vi}^{\infty} = Z_{vi}\sqrt{1-\left(\frac{\lambda}{\lambda_c'}\right)^2}$$

$$\theta_1 = \left(1-\frac{a_2}{a_1}\right)\frac{\lambda_c}{\lambda_c'}\frac{\pi}{2}$$

$$\theta_2 = \frac{a_2}{a_1}\frac{\lambda_c}{\lambda_c'}\frac{\pi}{2}$$

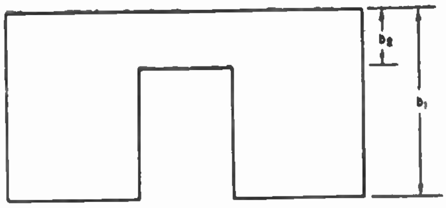
$\lambda_c$ = cutoff wavelength without ridge $= 2a_1$

$\lambda_c'$ = cutoff wavelength with ridge.

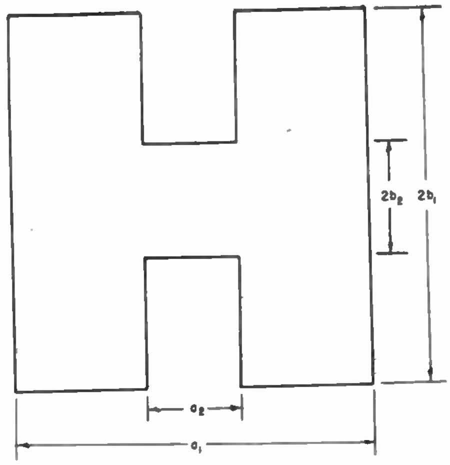
Equation (2) gives the impedance of single-closed-ridge guide as illustrated in Fig. 3(a). The impedance of the double-ridge structure is just twice that of the corresponding single-ridge guide.

If a narrow ridge is employed, $\theta_2 \ll 1$; $\cos\theta_2$ may be replaced by unity and $\sin\theta_2$ by $\theta_2$, giving the approximate expression

$$Z_{vi}^{\infty} = \frac{120\pi}{\dfrac{2C_d}{\epsilon_0} + \dfrac{a_2}{b_2} + \dfrac{1}{\pi}\dfrac{\lambda_c'}{b_1}\tan\dfrac{\theta_1}{2}}. \quad (3)$$




(a)




(b)

Fig. 3—Closed-ridge waveguide.

If $\lambda_c'/\lambda_c > 3$, little error results in replacing $\tan\theta_1/2$ by $\theta_1/2$, giving a very useful approximation which is independent of cutoff frequency

$$Z_{vi}^{\infty} = \frac{120\pi}{\dfrac{2C_d}{\epsilon_0} + \dfrac{a_2}{b_2} + \dfrac{1}{2}\dfrac{a_1}{b_1}\left(1-\dfrac{a_2}{a_1}\right)}. \quad (4)$$

While the voltage-current definition of impedance is useful in problems dealing with the matching of guides, it should be remembered that the definition of impedance is not unique when circuit dimensions are an appreciable fraction of a wavelength. If it is desired to find the voltage developed across a guide for a given power propagating down it, a power-voltage impedance must be defined and calculated. Let this impedance be termed $Z_{pv}$. Then

$$Z_{pv} = \frac{V_0^2}{\int_{c.s.} E \times H da}. \quad (5)$$

The effect of the discontinuity capacitance may again be taken into account, with the aid of Fig. 2, by noting that

$$d^2P = E \times H da = \frac{1}{Z_{TE}} E^2 dn dp = \frac{1}{Z_{TE}}(Edn)(Edp).$$

Since $E\,dn$ is constant between the two $E$ lines, the integration over $p$ may be carried out, giving

$$dP = \frac{V}{Z_{TE}} E dn$$

From which

$$dP = \frac{1}{\epsilon_0 Z_{TE}} V^2 dC.$$

Hence, the additional power flow due to $2C_d$ is $2V^2C_d/\epsilon_0 Z_{TE}$. Writing (5) as

$$Z_{pv} = \frac{V_0^2}{\dfrac{2V^2C_d}{\epsilon_0 Z_{TE}} + \dfrac{1}{Z_{TE}}\displaystyle\int_{c.s.} E^2 da}$$

Cohn's expression for $E$ may be inserted and the integration carried out, giving

$$Z_{pv}^{\infty} = \frac{120\pi^2 b_2}{2\pi b_2 \dfrac{C_d}{\epsilon_0}\cos^2\theta_2 + \lambda_c'\left\{\dfrac{\theta_2}{2} + \dfrac{\sin 2\theta_2}{4} + \dfrac{b_2}{b_1}\dfrac{\cos^2\theta_2}{\sin^2\theta_1}\left[\dfrac{\theta_1}{2} - \dfrac{\sin 2\theta_1}{4}\right]\right\}}$$

This simplifies, if a narrow ridge is assumed, to

$$Z_{pv}^{\infty} = \frac{120\pi}{\dfrac{2C_d}{\epsilon_0} + \dfrac{a_2}{b_2} + \dfrac{1}{\pi}\dfrac{\lambda_c'}{b_1}\dfrac{1}{\sin^2\theta_1}\left[\dfrac{\theta_1}{2} - \dfrac{\sin 2\theta_1}{4}\right]}. \quad (6)$$

If $\theta_1 < 30°$, an excellent approximation is

$$Z_{pv}^{\infty} = \frac{120\pi}{\dfrac{2C_d}{\epsilon_0} + \dfrac{a_2}{b_2} + \dfrac{1}{3}\left(\dfrac{a_1 - a_2}{b_1}\right)}.$$

## II. Open-Ridge Waveguide

The preceding expressions were developed primarily for determining the behavior of the open-ridge waveguide

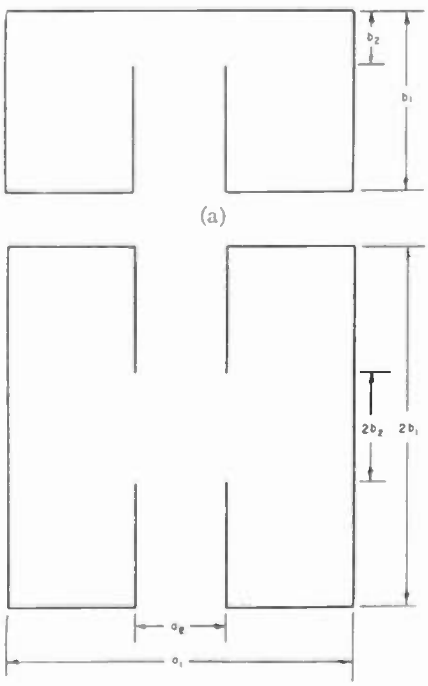



Fig. 4—Open-ridge waveguide.

of Fig. 4, an advantageous structure for certain applications. In this guide, the discontinuity susceptance is a major factor governing guide impedance, since the shunt-capacitance has been greatly reduced by removal of the ridge top. So far, the equations have been corrected for the effect of $C_d$. They must be modified further to include the change of geometry and the resulting loss of capacitance. Henceforth, it will be assumed that the ridge is narrow and represents a lumped capacitance subject to a voltage $V_0$. The discontinutiy capacitance is also lumped with the ridge-top capacitance. A major problem is to determine how much the ridge-top capacitance changes when the bridging conductor is eliminated. The simplest approach to this problem is the experimental approach. Measurements have been made to determine the difference in capacitance between a ridge with and without a top. It was found that this change of capacitance $\Delta C$ depends chiefly upon the ratio of the inside spacing of the fins $s$ to the ridge spacing $b_2$, providing this ratio is greater than unity. Experimental results for several fin thicknesses are presented in Fig. 5, from which it is evident that varying this parameter has little effect on $\Delta C$. This handily eliminates an unwanted variable.

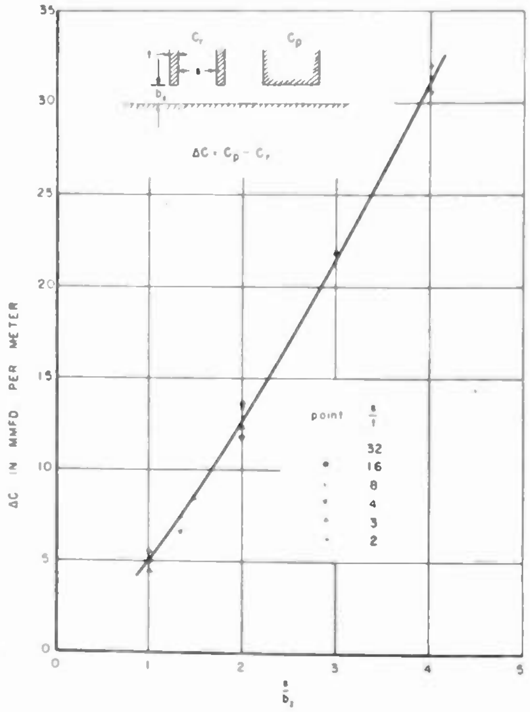



Fig. 5—Experimental determination of $\Delta C$ as a function of $s/b_2$.

Sufficient information is now available to calculate the cutoff wavelength and impedance of open-ridge waveguide. The expressions will again be developed for single-ridge guide, but apply equally well to double-ridge guide, providing the impedance is doubled. The shunt admittance of $C_d+\frac{1}{2}C_r$ must equal the admittance of a transmission line of length $\theta_1$ at cutoff, hence

$$\omega(C_d+\tfrac{1}{2}C_r) = Y_{01}\cot\theta_1.$$

Noting that $Y_{01}=1/120\pi b_1$, this may be written

$$\frac{\dfrac{b_1}{a_1}\left(\dfrac{2C_d+C_r}{\epsilon_0}\right)}{1-\dfrac{a_2}{a_1}} = \frac{\cot\theta_1}{\theta_1}. \tag{7}$$

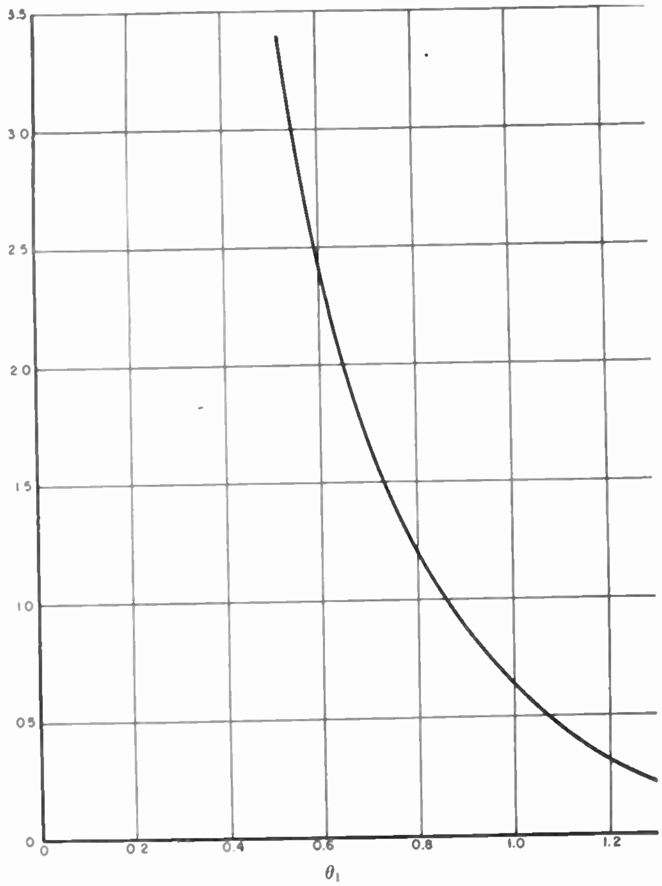



Fig. 6—Graphical solution of equation (7).

The solution of this equation may be found from the plot of Fig. 6, and $\lambda_c'/\lambda_c$ obtained from the definition of $\theta_1$,

$$\frac{\lambda_c'}{\lambda_c} = \left(1-\frac{a_2}{a_1}\right)\frac{\pi}{2}\frac{1}{\theta_1}. \tag{8}$$

Impedance may now be obtained from (3) or (6), except that $a_2/b_2$ must be replaced by $C_r/\epsilon_0$ as determined by

$$\frac{C_r}{\epsilon_0} = \frac{a_2}{b_2} - \frac{\Delta C}{\epsilon_0}. \tag{9}$$

Hence, for open-ridge guide,

$$Z_{vi}^{\infty} = \frac{120\pi}{\dfrac{2C_d+C_r}{\epsilon_0}+\dfrac{2}{\pi}\dfrac{\lambda_c'}{\lambda_c}\dfrac{a_1}{b_1}\tan\dfrac{\theta_1}{2}} \tag{10}$$

and

$$Z_{pr}^{\infty} = \frac{120\pi}{\dfrac{2C_d+C_r}{\epsilon_0}+\dfrac{2}{\pi}\dfrac{\lambda_c'}{\lambda_c}\dfrac{a_1}{b_1}\dfrac{1}{\sin^2\theta_1}\left[\dfrac{\theta_1}{2}-\dfrac{\sin 2\theta_1}{4}\right]}. \tag{11}$$

## III. Sample Calculations

Consider a closed-ridge guide with $b_1/a_1=0.5$, $b_2/b_1=0.133$, and $a_2/a_1=0.352$. Cohn's Fig. 3 gives $\lambda_c'/\lambda_c$ as 2.57 and a $Z_{vi}^{\infty}$ of 65 ohms. From Fig. 2 of this paper, $C_d$ is found to 9.5 $\mu\mu$f/meter for a step of 0.133. Since $\epsilon_0$ is 8.85 $\mu\mu$f/meter in mks units, the correction term may be evaluated, $2C_d\cos\theta_2/\epsilon_0=2.1$. Cohn's denominator must have been $377/65=5.8$, but this must be increased by 2.1. Hence, the impedance is $377/(5.8+2.1)=47.5$. Cohn reported experimental data on a similar guide, but with $b_1/a_1$ scaled down by a factor 0.472/0.500; its impedance was measured as approximately 50 ohms. Scaling 47.5 down by this factor, the corrected theoretical impedance is 45 ohms, a figure within 10 per cent of the experimental value. The approximate formula (4) gives 47 ohms.

Cohn gave a second example, a guide with $b_1/a_1=0.5$, $a_2/a_1=0.4$, and $b_2/b_1=0.1$. At a frequency one and one-half times cutoff, the uncorrected calculation gives 60 ohms, while including the discontinuity term reduces this to 46 ohms. Cohn reported 35 to 40 ohms experimentally. The approximate formula gives 45 ohms. Unfortunately, no further experimental data are available, but it is evident that the correction is in the right direction and of the right order of magnitude.

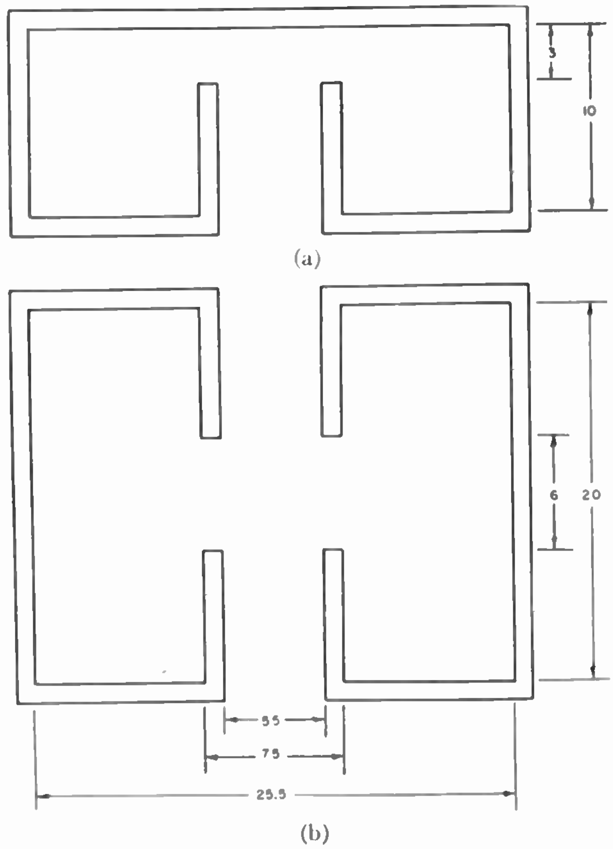



Fig. 7—Open-ridge guide for sample calculation.

As an example of open-ridge calculations, consider the guide of Fig. 7(a). Here $b_1/a_1=0.392$, $b_2/b_1=0.30$, and $a_2/a_1=0.294$. The discontinuity capacity for a step of 0.30 is given by Fig. 2 as 5.0 $\mu\mu f$/meter. The $\Delta C$ for a ridge with $s/b_2=5.5/3=1.83$ is given by Fig. 5 as 11 $\mu\mu f$/meter. Thus, from (9), $C_r=11$ $\mu\mu f$/meter. Substitution in (7) and use of Fig. 6 gives a $\theta_1$ of 0.77. This allows evaluation of $\lambda_c'/\lambda_c$ from (8), giving the value 1.44. Impedance is calculated directly from (10) and (11), giving $Z_{vi}^{\infty}=113$ ohms, and $Z_{pv}^{\infty}=124$ ohms.

No direct experimental check of these figures has been made; however, the above guide was set up in an equivalent form on a rectangular network analyzer designed and built at Stanford University under the direction of Spangenberg.[3] Measurement of $\lambda_c'/\lambda_c$ from the board gave the value 1.65. This is within 15 per cent of the theoretical value, the lack of agreement probably being due to the difficulty of representing thin reentrant fins on the board, since it makes use of discrete lumped elements.


### Acknowledgment

The writer wishes to thank E. L. Ginzton and M. Chodorow for their encouragement and guidance.


[3] K. Spangenberg and G. Walters, "An electrical network for the study of electromagnetic fields," Technical Report No. 1, ONR Contract N6-ori-106, Task III; May 15, 1947. This report describes a board designed for use with fields having cylindrical symmetry. Recently, however, another board has been constructed, one which allows solution of the two-dimensional wave equation in rectangular co-ordinates.

# Broad-Band Dissipative Matching Structures for Microwaves*

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***Summary*—Interpolation in the complex plane is employed to handle microwave network functions. This yields an approximating rational function over a specified bandwidth, and leads to a lumped-circuit approximation for the microwave structure, which is used as a basis for the synthesis of matching networks. In various problems involving dissipative devices, the poles of the rational approximating function may satisfy special conditions. In such cases the ideal lumped matching network has a simple realizable form, and may be transformed into a suitable microwave structure. Applications of this method and experimental results are given for the synthesis of a new type broad-band coaxial "chimney" attenuator.**

## I. Introduction

In order to insure uniformity of apparatus design in a composite transmission system, and, at the same time, maintain conditions for maximum power transfer, it has been the practice to match the terminal impedance of components to the characteristic impedance of the main transmission line. This paper considers methods of design of microwave structures which match the input impedance of transmission components over very wide frequency bands. The matching structures are analogous to the constant-resistance networks of low-frequency circuit theory, and hence have loss. Thus the methods are most directly applicable to devices in which dissipation can be tolerated, such as attenuators, terminating loads, certain types of broad-band probes, etc. Later in the paper, applications of the method to the design of fixed pad coaxial "chimney" attenuators will be given.

The design procedure consists of essentially three steps:

*Step 1:* Determine the input driving point impedance function of the microwave device when properly terminated, and approximate this over a specified frequency band by a suitable rational function and its equivalent lumped-parameter circuit representation.

*Step 2:* Determine an ideal lumped-parameter matching network which, in combination with the lumped-circuit approximation of Step 1, produces a constant-resistance input impedance.

*Step 3:* Transform the ideal matching network of Step 2 into the final microwave matching structure.

## II. Representation of Microwave Impedance Functions

The method employed here for approximating by a rational function the nonrational function which represents the input impedance of a specified microwave device, is that of interpolation in the complex plane. Algebraic methods of interpolation for the approximation of impedance functions have been used heretofore by Zobel,[1] but a more general method employing contour integration, described by Walsh,[2] appears to be little known to engineers. The basic theorem for interpolation in the complex plane may be stated as follows:

Let the function $f(s)$ be analytic in a region of the

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[1] Otto J. Zobel, U.S. Patent No. 1,720,777.

[2] J. L. Walsh, "Interpolation and Approximation by Rational Functions in the Complex Plane," Amer. Math. Soc. Colloq. Public Vol. XX, 1935, chap. VIII.

complex plane bounded by the contour $\Gamma$. Let the points $\beta_1, \beta_2, \cdots \beta_{n+1}$ lie inside $\Gamma$. Let $r(s)$ be a rational function of degree $n$ with preassigned poles $\alpha_1, \alpha_2, \cdots, \alpha_n$ and of the form:

$$r(s) = \frac{b_0 s^n + b_1 s^{n-1} + \cdots + b_n}{(s-\alpha_1)(s-\alpha_2)\cdots(s-\alpha_n)}. \qquad (1)$$

If $r(s)$ takes on the values of $f(s)$, i.e., interpolates to $f(s)$, at the points $\beta_1, \beta_2, \cdots, \beta_{n+1}$, then $r(s)$ is given by

$$r(s) = \frac{1}{2\pi j}\oint_\Gamma \left[1 - \frac{(s-\beta_1)(s-\beta_2)\cdots(s-\beta_{n+1})(t-\alpha_1)(t-\alpha_2)\cdots(t-\alpha_n)}{(t-\beta_1)(t-\beta_2)\cdots(t-\beta_{n+1})(s-\alpha_1)(s-\alpha_2)\cdots(s-\alpha_n)}\right]\frac{f(t)}{t-s}\,dt. \qquad (2)$$

Equation (2) may be applied in a variety of ways, and can yield much information on the general nature of the interpolation process. For example, suppose all the $\beta_k$ to be equal, i.e., a multiple point of interpolation. Then the integrand has a pole of order $n+1$ at $\beta_k$, and its evaluation by residues will involve the first $n$ derivatives of $f(s)$ at $s=\beta_k$. If, now, the $\alpha_k$ are chosen $\alpha_k=\infty$, (2) will then reduce to

$$r(s) = a_0 + a_1(s-\beta_k) + a_2(s-\beta_k)^2 + \cdots + a_m(s-\beta_k)^n \qquad (3)$$

where $a_k$ become the coefficients of a Taylor series; $r(s)$ is thus given by the first $n$ terms of a series expansion of $f(s)$ about the point $s=\beta_k$. This result indicates that the often-used practice of obtaining an approximation for a given function by using several terms of a power series expansion for that function is actually a form of interpolation in which all the interpolating points are identical.

In many cases the use of nonidentical points of interpolation distributed over the region in which an approximation is required gives better results than approximation by power series (i.e., repeated interpolation at a single point). A case in point is the approximation of nonrational impedance functions by lumped-parameter networks. For this, (2) may be applied if certain additional restrictions imposed by network realizability are added. A necessary condition that $r(s)$ be a network function is that the roots and poles of (1) be real or complex conjugate. It can be shown that, if this is to be true, then $r(s)$ may be determined by (2) in conjunction with the additional equations formed by setting the imaginary parts of the numerator coefficients $b_k$ respectively equal to zero.

It should be pointed out that although $r(s)$ will satisfy certain necessary conditions for realizability as a driving-point impedance, it may still not be realizable. However, $f(s)$ is itself an impedance function, though of a distributed-parameter structure, and is in many ways similar to rational network functions.[3] Therefore, in practical problems $r(s)$ will usually turn out to be positive real and hence realizable; and even when this is not the case, additional devices may be employed, such as a slight change in the choice of interpolating points, to produce a realizable rational function.

As a simple illustrative example, and one which will be useful in a design problem to be considered later in this paper, consider the case of two interpolating points, $\beta_1=0$, $\beta_2=jX_2$ where $X_2$ is a real linear function of either wavelength or frequency. Then $r(s)$ from (2) is

$$r(s) = \frac{-\alpha(s-\beta_2)f_1 - s(\beta_2-\alpha)f_2}{\beta_2(s-\alpha)}$$

where

$$f_1 = f_{1r} + jf_{1i} = f(\beta_1)$$
$$f_2 = f_{2r} + jf_{2i} = f(\beta_2).$$

If the imaginary part of the numerator coefficient in $s$ is set equal to zero then $\alpha$ is found:

$$\alpha = \frac{X_2 f_{2i}}{f_1 - f_{2r}} \qquad (4)$$

and

$$r(s) = \frac{s(f_{2r} - \alpha f_{2i}/X_2) - \alpha f_1}{s-\alpha}. \qquad (5)$$

With appropriate values given for the parameters, (5) may be realized as a suitable network.

As an example of the application of (5), consider the input impedance to a long, lossy coaxial transmission line, which may be written:

$$f(s) = Z_0\sqrt{1-s}, \qquad (6)$$

where

$Z_0$ = lossless characteristic impedance of the line
$s$ = complex wavelength variable with imaginary part $= X = R\lambda/2\pi Z_0$
$R$ = resistance per unit length of line
$\lambda$ = real wavelength.

A range of $0 \leq X \leq 2$ is applicable to the class of lossy coaxial lines encountered in resistive-film microwave attenuators,[4] and may be chosen as the range of approximation for this example. The two points of interpolation may be determined by a method of least squares described in detail elsewhere,[5] or by inspection. The latter method generally yields satisfactory results and has the considerable advantage of saving much computational time when applicable. Suppose, then, that $X_1=0$, $X_2=1.6$. Substituting in (4) and (5), the result-

[3] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., chap. 13, p. 298; New York, N. Y., 1945.

[4] H. J. Carlin and J. W. E. Griemsmann, "A bead supported coaxial attenuator," *1947 Proc. Nat. Electr. Conf.*, pp. 79–89.

[5] H. J. Carlin, Report R-139-47, PIB-90, Microwave Research Institute of the Polytechnic Institute of Brooklyn, p. 30; March, 1947.

ing function on the imaginary $s$ axis is given by

$$r(X) = \frac{5.34 - j3.42X}{5.34 - jX}. \tag{7}$$

The real and imaginary parts of this approximating function are compared with the exact results obtained from (6) in Fig. 1. Shown also in Fig. 1 is the lumped-circuit representation of (7).

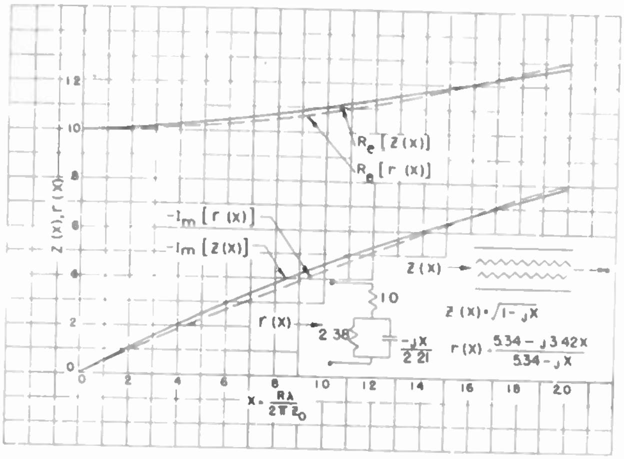


Fig. 1—Interpolation to the input impedance of dissipative coaxial line of infinite length.

Another example is plotted in Fig. 2. In this case the given distributed-parameter structure is a dissipative short-circuited coaxial line of length $l$, and resistance $R_{do}$. Three points of interpolation were chosen at $l/\lambda = 0.0$, $0.05$, and $0.20$, respectively, to cover the range $0 \leqq l/\lambda \leqq 0.25$.

The input admittance function of the line is given by

$$Y(s) = \frac{Y_0 \coth \sqrt{1 + \rho/s}}{\sqrt{1 + \rho/s}} \tag{8}$$

and the interpolating rational function found from (2) is then

$$r(s) = \frac{8.13s^2 + 4.30s + 22.0}{s^2 + 22.74s + 9.78}, \tag{9}$$

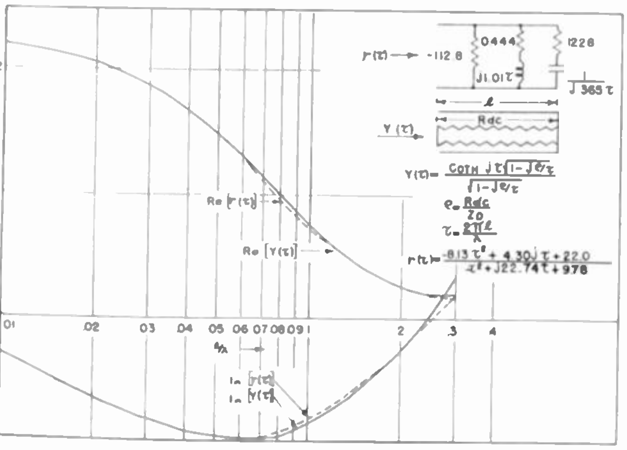


Fig. 2—Interpolation to the input impedance of a short-circuited dissipative coaxial line.

where

$s$ = complex frequency variable $= \sigma + j\tau$

$\tau = 2\pi l/\lambda$

$\rho = R_{do}/Z_0$.

The real and imaginary parts of $r(j\tau)$ are plotted on Fig. 2, which also shows a lumped-circuit representation of (9). The network shown contains a negative resistance, although it can be rigorously shown that (9) is a positive real function in $s$, and hence realizable by a finite assemblage of positive resistors and reactors. This type of structure with negative resistance is chosen since it leads to a simple form of ideal matching network. This is discussed further in the next section.

## III. Determination of Ideal Matching Networks

Having found a realizable lumped-circuit approximation for the given microwave structure, the next step in the design process is to find a lumped ideal matching network. Fano[6] has shown how to obtain a lossless network which approximately matches an arbitrary two pole, and has given the value of minimum reflection coefficient which may be obtained using an infinite ladder structure of reactance elements. The disadvantages of reactance matching networks are that they must result in a finite residual reflection, and are difficult to realize as distributed-parameter structures for use at microwaves. If loss can be tolerated, the constant-resistance networks discussed here are more suitable, since they give ideal reflectionless matching, and the relatively small number of elements they contain simplifies the problem of finding a final representation as a microwave structure.

In order to proceed with the determination of a lumped ideal matching network, use is made of a property of driving-point network functions given by Bode.[7] This may be stated as follows: If $Y(s)$ is a rational, realizable driving-point admittance of the minimum-susceptance type, then a partial fraction expansion gives: ($Y(s)$ has only simple poles),

$$Y(s) = \pm G_0 + \sum_i \left( \frac{a_i s^2 + b_i s + c_i}{d_i s^2 + e_i s + f_i} + \frac{A_i s + B_i}{C_i s + D_i} \right), \tag{10}$$

where each term of the first summation represents a physically realizable biquadratic admittance function with one pair of complex conjugate poles, and each term of the second summation represents a physically realizable bilinear admittance function with a single real pole. In order to insure that all terms under the summation sign are positive real, it may be necessary to make the constant term $G_0$ negative. The network shown in Fig. 3(b) is always a possible representation of (10) for, since none of the terms, by hypothesis, can have poles on the imaginary $s$ axis, a finite series resistance can be extracted from every shunt branch.[8]

[6] R. M. Fano, Technical Report No. 41, Research Laboratory of Electronics, MIT, January, 1948.

[7] See chap. 10, p. 200, of footnote reference 3.

[8] In case a term has a zero minimum value of conductance, it will be necessary to add some finite conductance extracted from $G_0$.

Thus the impedance of each branch is

$$Z_i = R_i + \bar{Z}_i. \tag{11a}$$

To form a constant-resistance network, a set of shunt circuits, as in Fig. 3(a), is added to the given structure. Each of these circuits has only positive elements and corresponds to one of the branches given by (11a) with an impedance

$$Z_{ic} = R_i + \bar{Z}_{ic} \tag{11b}$$

where $\bar{Z}_{ic}$ and $\bar{Z}_i$ are inverse networks;

$$\bar{Z}_{ic}\bar{Z}_i = R_i^2. \tag{11c}$$

The resultant structure has an input admittance which is a real positive constant, $G_T$, independent of the sign or magnitude of $G_0$. This follows from the fact that $Y(s)$ is a positive real function.

$$G_T = (\pm G_0 + \sum G_i) \geqq 0. \tag{12}$$

Other questions concerning the general case of Fig. 3(b), such as the possibility of eliminating mutual reactance, will not be treated here, but a special case which has found frequent application in design problems

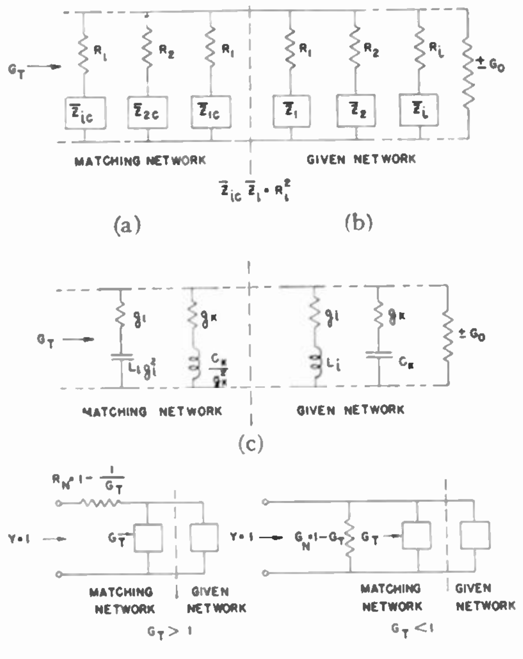


Fig. 3—General determination of ideal matching networks.

will be considered. If $Y(s)$ has only real poles, then the terms of the second degree in (10) are eliminated. The given network is then made up of shunt branches, each of which is of the series $RL$ or $RC$ type, plus one shunting resistor which may be negative. The input admittance is then of the form

$$Y(s) = \pm G_0 + \sum_i \left[\frac{1}{L_i}\,\frac{1}{s+\alpha_i}\right] + \sum_k \frac{g_k s}{s+\gamma_k} \tag{13}$$

where

$$\alpha_i = \frac{1}{g_i L_i}; \qquad \gamma_k = \frac{g_k}{C_k}. \tag{13b}$$

The matching procedure for this case is indicated in Fig. 3(c). If the impedances are all normalized to the line characteristic impedance, it is desired to have the ideal matching network produce a constant input resistance of 1.0. To do this, a single additional normalizing resistance $R_N$ is used. If $G_T > 1$, as in Fig. 3(d), $R_N$ is placed in series with the matching network and has a value

$$R_N = 1 - \frac{1}{G_T}. \tag{14a}$$

If $G_T < 1$, as in Fig. 3(e), $R_N$ is placed in shunt with the matching network and has a value

$$R_N = \frac{1}{1 - G_T}. \tag{14b}$$

Fig. 4 shows the matching procedure for the two interpolating networks whose characteristics were shown in Figs. 1 and 2.

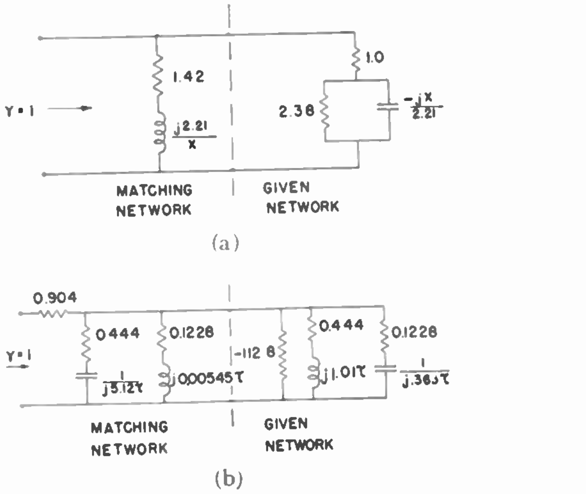


Fig. 4—Determination of ideal matching networks. (a) Infinite dissipative line. (b) Short-circuited dissipative line.

## IV. Determination of the Microwave Structure

The final step in the design procedure is to find a distributed-parameter structure which will approximate the impedance characteristics of the ideal matching network over a specified frequency band. This problem is essentially the inverse of the original interpolating procedure. To obtain an indication of the order of magnitude of the mismatch produced by an impedance which deviates from the ideal matching network, the input voltage-standing-wave ratio may be calculated as a function of the admittance deviation between the microwave network and the ideal matching network. The input reflection factor is given by

$$K = \frac{1-y}{1+y} \tag{15}$$

where $y$ is the normalized actual input admittance. But

$$y = y_a + y_e = y_a + y_m + \Delta y = 1 + \Delta y \tag{16}$$

where $y_a$ is the normalized admittance of the originally specified microwave device, $y_e$ is the admittance introduced by the final microwave equalizer, $y_m$ is the admittance of the ideal matching network, and $\Delta y = y_m - y_e$. Substituting in (15),

$$K = \frac{\Delta y}{2 + \Delta y} \doteq \frac{\Delta y}{2} \tag{17}$$

where the approximation is valid for $\Delta y \ll 2$. The VSWR, $\eta$, is then given by

$$\eta = \frac{1 + |K|}{1 - |K|} \doteq \frac{2 + |\Delta y|}{2 - |\Delta y|} \doteq 1 + |\Delta y|. \tag{18}$$

The final approximation of (18) may be taken as a fair indication of the mismatch whenever $\eta < 1.30$.

An important point indicated by (18) is that the microwave equalizer may deviate from the ideal matching network by an amount of the order of 30 per cent of $Z_0$ without introducing an inordinate reflection. Thus, if a VSWR of 1.30 is permissible, an accurate determination of the final microwave structure is not necessary, unless an extremely large amount of equalization is required.

The first step in finding the microwave matching network is to choose the form of structure which is to be used. A variety of structures will usually be possible, and experience, coupled with some ingenuity on the part of the designer, will help in arriving at a practical choice. In this phase of the problem, the ideal matching network is an invaluable aid in suggesting the type of distributed-parameter structure which should be chosen.

Fig. 5 shows two simple microwave transmission lines with loss, which are extremely flexible as circuit elements, and hence are applicable to many problems as matching networks. Fig. 6 gives typical input-admit-

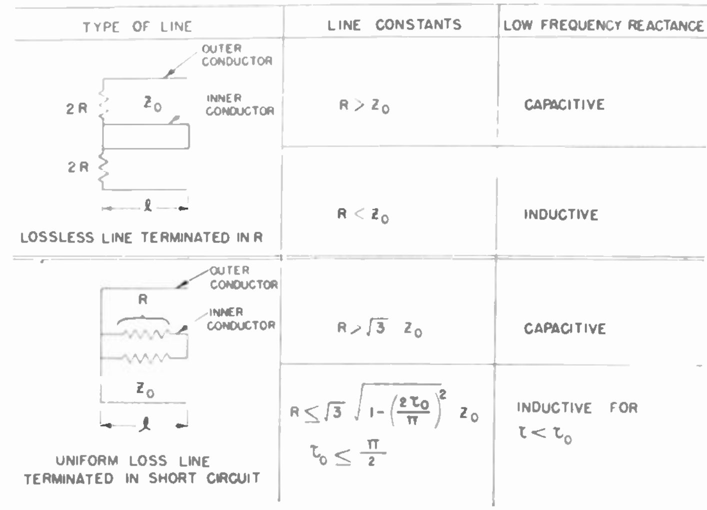


Fig. 5—Low-frequency response of coaxial lines.

tance curves of the short-circuited, uniformly dissipative line of Fig. 5 drawn for the entire frequency spectrum. It indicates quite clearly the wide range of admittance loci that it is possible to realize with this simple structure, and suggests how such a line may be used as a circuit element. More detailed characteristics are given by Crosby and Pennypacker.[9]

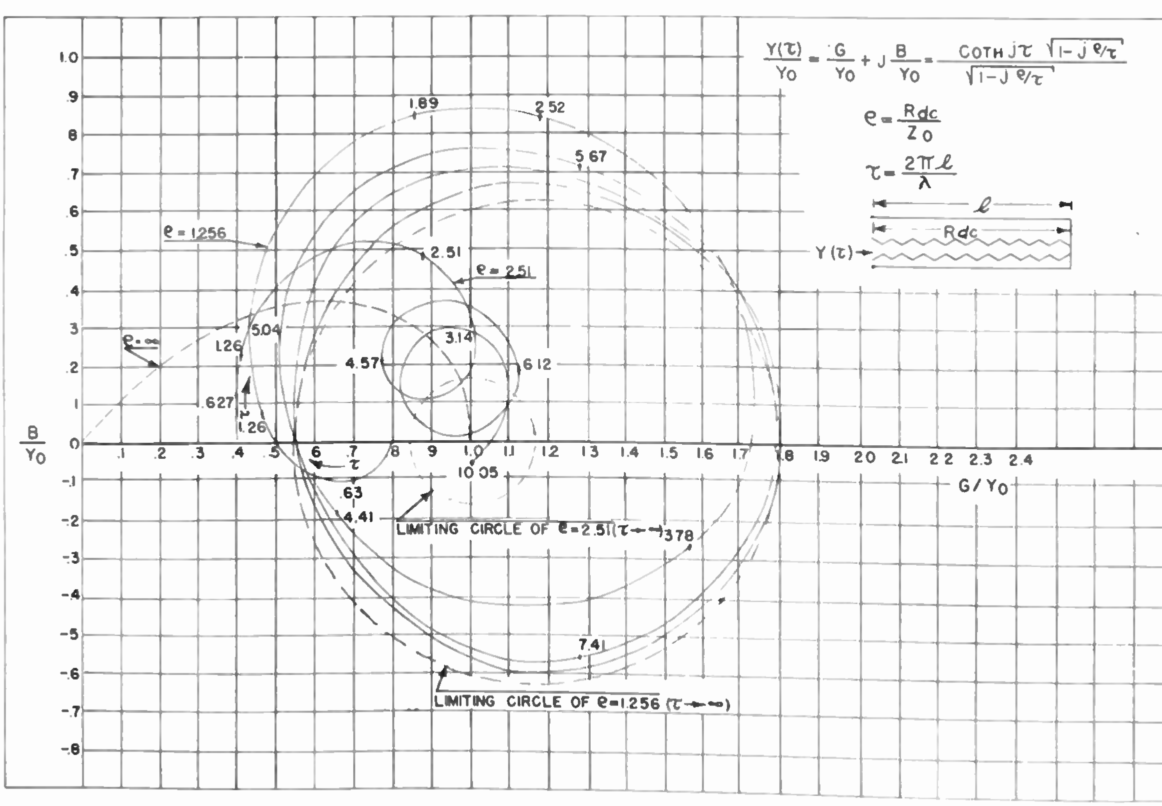


Fig. 6—Input admittance loci of short-circuited dissipative coaxial lines.

[9] D. R. Crosby and C. H. Pennypacker, "Radio-frequency resistors as uniform transmission lines," Proc. I.R.E., vol. 34, pp. 62–66; February, 1946.

The final problem is to determine quantitatively the onstants of the chosen microwave structure so as best o approximate the ideal matching network. If the in- )ut admittance of the microwave device is given by

$$Y_e(s) = F(s, \Phi_1, \Phi_2, \cdots, \Phi_n) \tag{19}$$

vhere the $\Phi j$ are the parameters such as termination, characteristic impedance, length, etc., which char- acterize the structure, then the solution for these param- eters may be formally written down in terms of the nethod of least squares. This can be done if it is noted rom (17) that the reflection factor $K$ is proportional to $\Delta y$, so that the solution which minimizes $|K|^2$ over a specified frequency band may be expressed in terms of the square of the magnitude of the admittance deviation $|\Delta y|^2$. The result is the set of $n$ simultaneous equations of the form

$$\int_{S_a}^{S_b}\left[(y_{er} - y_{mr})\frac{\partial F_r}{\partial \Phi_j} + [y_{ei} - y_{mi}]\frac{\partial F_i}{\partial \Phi_j}\right]dq = 0 \qquad (20)$$
$$j = 1, 2, \cdots, n$$

where the subscripts $r$ and $i$, respectively, denote real and imaginary parts, and $S_a$ and $S_b$ are the band limits.

A direct solution of equations (20) is generally quite difficult, but a great deal of numerical simplification is possible.[10] In particular, these equations can be converted into a set of simultaneous linear algebraic equations in the unknowns $\Phi j$, provided it is possible to make a reasonable initial estimate of these quantities. In the example given below, a graphical method is used to obtain the approximate values of the microwave network constants, and, because a high degree of accuracy is not generally needed, as discussed earlier in connection with (18), these values are sufficiently good to make the use of (20) unnecessary. This is found to be true in many design problems, particularly when the ideal matching network does not have a large number of branches.

## V. Application to the Design of Fixed Pad Coaxial Attenuators

As a typical example, the design procedure for a 20-db coaxial pad which is to operate over the frequency band 1000- to 4000-Mc (7.5 to 30 cm) will be described briefly. The given structure to be matched is chosen as a length of metallized glass whose resistance per unit length does not introduce excessive attenuation variation over the frequency band. This is easily accomplished but the details are not pertinent here and have been described elsewhere.[4] Initially the entire 20-db loss is assumed associated with this lossy line, but the finite loss of the matching network is taken care of in the final design by merely shortening the main attenuator insert by the requisite amount. Since the loss is high, the input impedance of the main attenuator insert is equal to its characteristic impedance, and is given by (6). The two-point interpolation of Fig. 1 is quite adequate, since, corresponding to the choice of frequency range and attenuator insert constants, the value of $X$ is well within the range of Fig. 1. The interpolating rational function is therefore given by (7), and the ideal matching network is formed according to Fig. 4(a).

A consideration of the ideal matching network, which is inductive, in conjunction with Figs. 5 and 6 suggests a short-circuited lossy stub line as the microwave matching structure. A graphical procedure is used to find the constants $R_{dc}$, $Z_0$, and $l$ of the stub line shown in Fig. 2. First, the admittance of the ideal matching network is plotted. This is a circular locus in the $G-B$

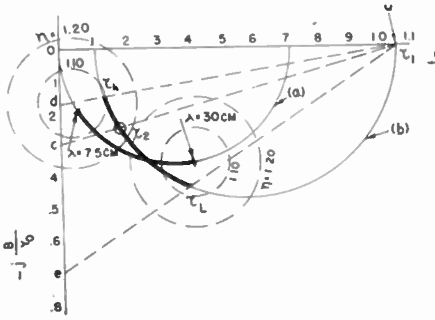

Fig. 7—Graphical determination of a microwave matching structure. Curve (a): Admittance locus of an ideal matching network. Curve (b): Approximate admittance locus of a microwave structure.

plane, and is shown as curve (a) of Fig. 7, with the terinal points corresponding to the band limits, $\lambda = 7.5$ cm

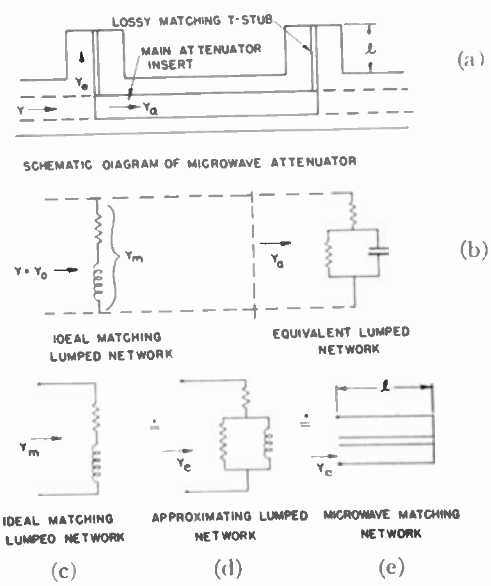


Fig. 8—Summary of design procedure for a 20-db "chimney" attenuator.

and $\lambda = 30$ cm, indicated. The constants of the stub line must now be chosen so that $y_e$, the admittance of the microwave network, approaches curve (a) over the specified band. To do this it is assumed that the stub line, whose admittance is given by (8), can be approximated over the 1000- to 4000-Mc frequency band, by a rational admittance function found from a two-point

[10] I. S. Sokolnikoff and E. S. Sokolnikoff, "Higher Mathematics for Engineers and Physicists," McGraw-Hill Book Co., New York, N. Y., 1934, p. 407.

interpolation process, i.e., a bilinear form similar to (7). This function is realized by a network of the type shown in Fig. 8(d), but the element values are unknown. However, a network of this type must have a circular admittance locus. Since (8) is a function of $\tau = 2\pi l/\lambda$, the length $l$ has no effect on the shape of this locus, but only on the frequency scale of the curve. Hence, by drawing several loci corresponding to a few chosen values of $R_{dc}$ and $Z_0$, the best locus curve can be found by comparing the resultant circles with curve (a). This gives two of the three required parameters, and, by graphically adjusting the scale, the third unknown, $l$, is found. The whole procedure is actually a very straightforward process, for the circular loci may be located and drawn very simply in the following way. Suppose specific values for $R_{dc}$ and $Z_0$ have been chosen. Two interpolating points are taken, one being $\tau_1 = 0$, the other some assumed value of $\tau_2$ in the neighborhood of $2\pi l/\lambda = 0.4\pi$; i.e., in order that $Y_e$ be inductive, $\tau_2$, according to Fig. 5, is chosen $\tau_2 < \tau_0 < (\pi/2)$. The values of $Y_e(s)$ are calculated from (7):

$$Y_e(s)\big|_{\tau=0} = \rho = \frac{R}{Z_0} \tag{21a}$$

$$Y_e(s)\big|_{\tau=\tau l} = g_{e_2} + jb_{e_2} \tag{21b}$$

where $g_{e_2}$ and $b_{e_2}$ will now have some definite numerical value. These two points are plotted, and, since the center of the circle must lie on the real axis, the entire circular locus may be immediately constructed. In this manner curve (b) of Fig. 7 was chosen (other circles are left out for the sake of clarity) from a number of circles; $R_{dc}$ and $Z_0$, associated with this curve, are thus determined. It is now necessary to fix the terminal points of the (b) locus curve. Only one of these can be chosen independently. In Fig. 7, the high-frequency terminal point $\tau_h$ is chosen on curve (b) to lie as close as possible to the point $\lambda = 7.5$ cm of curve (a). To find the value of $\tau_h$, a simple geometric property of the linear transformation is used.[11] Referring to the construction lines on Fig. 7,

$$\tau_h = \tau_2 \frac{\overline{Oc}}{\overline{Od}}, \tag{22a}$$

and $l$ is given by

$$l = \lambda \cdot \frac{\tau_h}{2\pi} = 7.5 \times \frac{\tau_h}{2\pi}. \tag{22b}$$

In a similar fashion, a point corresponding to any wavelength may now be located on curve (b). For example, for the low-frequency limit at 30 cm,

$$\tau = \tau_L = \tfrac{1}{4}\tau_h,$$

and the construction line $\overline{Qe}$, with $\overline{Oe} = 4\,\overline{Od}$, locates the point on curve (b) corresponding to $\tau_L$. On the basis of (18), the VSWR circles shown in Fig. 7 are drawn, and these indicate the mismatch which may be expected.

[11] C. W. Carter, "Graphic representation of the impedance of resistance networks containing two reactances," *Bell Sys. Tech. Jour.*, vol. 4, pp. 387–402; July, 1925.

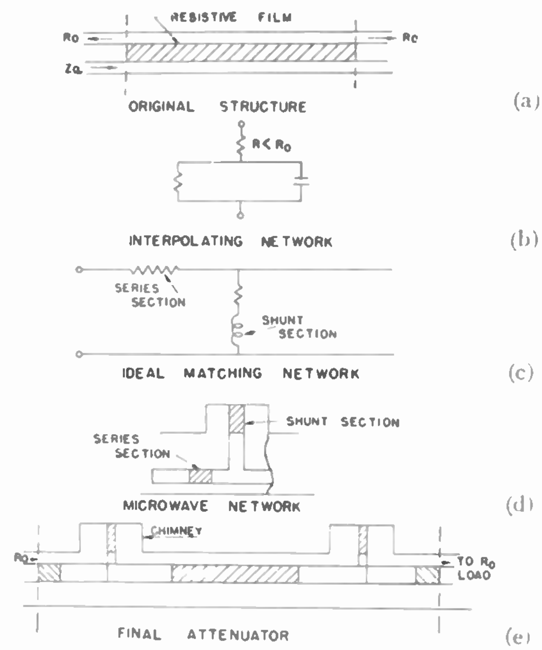


Fig. 9—Summary of design procedure for a 10-db "chimney" attenuator.

Fig. 8 summarizes the procedure described for the 20-db attenuator, and Fig. 9 shows the steps in the design of a 10-db unit for the frequency band 0 to 1000 Mc. Attenuators of this type have become known as "chimney attenuators" from the appearance of the stub lines. Although space does not permit discussion here, it is possible to utilize the matching networks for attenuation equalization as well as matching. The experimentally measured performance curves of Fig. 9 for a 10-db and 20-db chimney attenuator are for typical designs which have been matched as described in this paper, and designed as well for attenuation equalization.

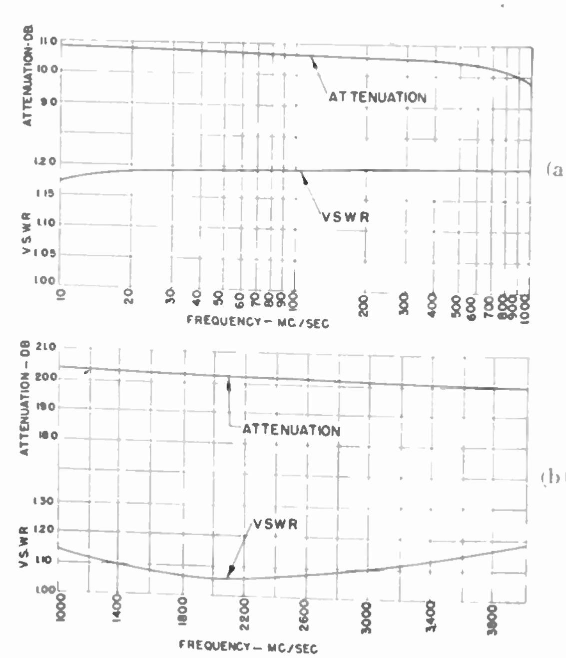


Fig. 10—Typical measured characteristics of chimney attenuators (a) 10-db, $\frac{7}{8}$-inch line unit, 0–1000 Mc. (b) 20-db, $\frac{7}{8}$-inch line unit, 1000–4000 Mc.

# Analysis of a Wide-Band Waveguide Filter*

SEYMOUR B. COHN†, MEMBER, IRE

*Summary*—**A rectangular-waveguide structure consisting of a series of constrictions and cavities is analyzed and shown to have the properties of a wide-band filter. The lower cutoff of the lowest-frequency pass band is due to the natural cutoff behavior of the waveguide itself. The upper cutoff of this band is due to the succession of discontinuities. Under the assumption of no dissipation in the filter sections, exact equations for the image parameters of the filter are derived. These equations take full account of the discontinuities and their interaction.**

**As a by-product of this analysis, formulas are obtained for the exact equivalent circuits of three rectangular-waveguide structures: (a) a cavity formed by two changes in height (Fig. 2(a)); (b) an increase of height followed by a short-circuiting wall (Fig. 2(b)); (c) the hypothetical case of an increase of height followed by an open-circuiting wall (Fig. 2(c)). It is shown how the analysis of structure (b) may be used to obtain an improved solution for ridge waveguide.**

**The writer will cover the more practical problems of filter design and experimental verification in another paper.**

THE FILTER structure that is analyzed in this paper is shown in Fig. 1.[1] Design relations are given for it in the literature,[2] for $b \leq l$. In this paper, the more useful case of $b$ unrestricted is treated. It is assumed that the conductance of the metallic walls is infinite.

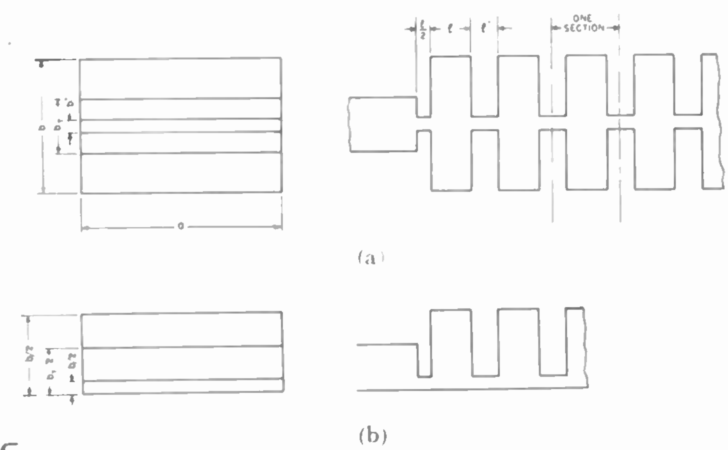


Fig. 1—The waveguide filter structure studied in this paper. Because of symmetry considerations, the theory developed for (a) applies to (b).

## Basic Principles

Waveguide is considered from the transmission-line standpoint in this paper.[3,4] The *impedance* at a given reference plane is defined as

$$Z = -C\frac{bE_y}{aH_x}, \tag{1}$$

where $C$ is an arbitrary constant, $a$ is the width, and $b$ the height of the guide, and $E_y$, $H_x$ are the transverse electric and magnetic field strengths at a given point in the reference plane. The *characteristic impedance* of the guide is

$$Z_o = \frac{Cb\eta}{a\sqrt{1-(f_c/f)^2}} \tag{2}$$

where $\eta$ is the characteristic impedance of free space (377 ohms if mks units are used), $f_c$ the cutoff frequency of the $TE_{10}$ mode, $f$ the frequency of operation, and $C$ the same constant as in (1).

The guide wavelength $\lambda_g$ is given in terms of the space wavelength $\lambda$ by

$$\lambda_g = \frac{\lambda}{\sqrt{1-(f_c/f)^2}}. \tag{3}$$

The image parameters of classical filter theory may be used with wave-guide filters, as explained in the literature.[5] In Fig. 1 it is shown how the waveguide filter may be split up into identical sections. Let $Y_{sc}$ and $Y_{oc}$ be the short-and open-circuit admittances of a filter half section (Fig. 2). Then the image admittance $Y_I$ and image transfer constant $\theta$ of a full section are given by

$$Y_I = \sqrt{Y_{sc}Y_{oc}} \tag{4}$$

$$\tanh\frac{\theta}{2} = \sqrt{\frac{Y_{oc}}{Y_{sc}}}. \tag{5}$$

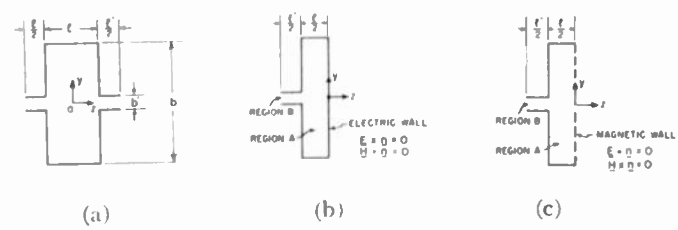


Fig. 2.—A single filter section (a); a short-circuited half section (b); an open-circuited half section (c).

For a nondissipative filter, $Y_I$ is real in pass bands and imaginary in stop bands. $\theta$ is imaginary in passbands and real in stop bands. $\theta$ is usually written

$$\theta = \alpha + j\beta, \tag{6}$$

where $\alpha$ is the the image attenuation constant in nepers, and $\beta$ the image phase constant in radians.

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[1] Filter configurations of this and other types have been proposed by numerous investigators before the war. In order to keep the length of this paper within reasonable bounds, a history of waveguide filters is omitted.

[2] Radio Research Laboratory Staff, "Very High-Frequency Techniques," McGraw-Hill Book Co., New York, N. Y., section 27-28; 1947.

[3] S. A. Schelkunoff, "The impedance concept and its application to problems of reflection, refraction, shielding and power absorption," *Bell Sys. Tech. Jour.*, vol. 17, pp. 17-49; January, 1938.

[4] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand, Inc., New York, N. Y., pp. 319-320, 490-494; 1943.

[5] See chapter 26 of footnote reference 2.

Fig. 3 shows the kind of $Y_{oc}$ and $Y_{sc}$ functions that one obtains for a waveguide filter half section. The independent variable is the inverse guide wavelength $1/\lambda_g$, which is analogous to frequency in ordinary transmission-line filters. The corresponding $Y_I$, $\alpha$, and $\beta$ functions for this case are sketched in Fig. 3, also. Note that there is a succession of pass bands, but the first and second pass bands are separated by a wide stop band. In an actual filter design, only the first pass band is usually used. The next pass band begins near the frequency for which $\lambda_g = b$.

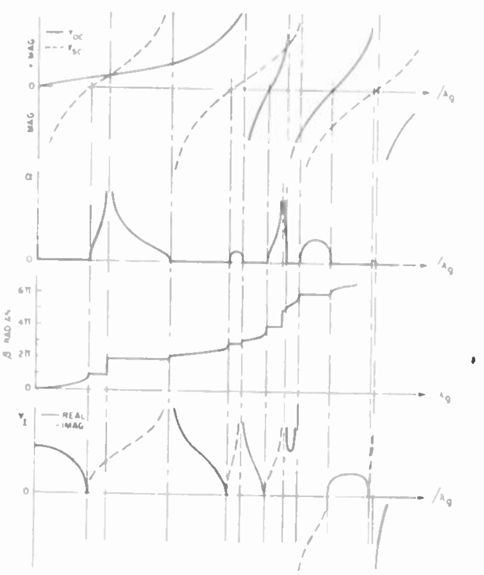


Fig. 3—Open- and short-circuit admittances of a typical half section filter. The image parameters of the corresponding full section are also shown.

## Theoretical Analysis

The field in each uniform region of the filter may be represented by a superposition of all the modes possible in that region. The boundary conditions at the conductors parallel to the $z$ axis will be met automatically by the use of the characteristic modes of the waveguide

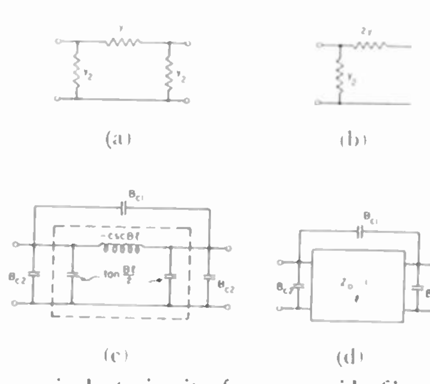


Fig. 4—The equivalent circuit of a waveguide filter section. The capacitances and inductances are, of course, functions of frequency.

cross section. On the conducting surfaces perpendicular to the $z$ axis (Fig. 2), and at the boundaries between the different regions, the boundary conditions are met by the proper choice of amplitude and phase for each mode. The electric and magnetic fields at each point inside the filter will then be expressed by an infinite series of mode terms. If these series are everywhere convergent, and satisfy all boundary conditions, then they will be the true solution of the field problem. The equivalent circuit or the image parameters of the filter may be calculated once the field is known.

There are several possible methods of carrying out such a solution for this filter. For the waveguide filter, Hahn's method[6] is easily applicable and fairly straightforward, and hence is used in this report. It gives an entirely adequate and sufficiently accurate solution of the filter problem. Schwinger's *variational* method might also have been used. This method is covered in Radiation Laboratory lecture notes,[7] and in a paper by Miles.[8]

Examination of Fig. 1 shows that, except for the ends, the filter consists of a periodic composition of two different shapes, one of length and height $l$ and $b$, the other of length and height $l'$ and $b'$. It may be shown that the discontinuities at the ends of the $l'$ lengths are virtually isolated from each other for any value of $l'/b'$. The interaction through the lengths $l'$ will therefore be neglected, especially since, in most filter designs, $l'/b'$ is greater than or equal to one. Because of this, the filter may be divided into sections as shown in Figs. 1 and 2(a). Each section is independent of the adjacent ones or of the terminating guide, as far as higher mode coupling is concerned.

A possible method of obtaining the filter-section parameters $Y_I$ and $\theta$ that immediately suggests itself is as follows: (a) calculate the field inside the filter by Hahn's method for the two cases of an open and short circuit at the plane $z=0$; (b) calculate the open-circuit and short-circuit admittances for the fundamental mode at the plane $z=-(l+l')/2$ through the use of (1); (c) then calculate $Y_I$ and $\theta$ by means of (4) and (5).

The waveguide structure of Fig. 1 is excited by a $TE_{10}$-mode wave, which has a half-sine-wave variation of field in the $x$ direction. Since the surface of the filter structure is cylindrical in the $x$ direction, with shorting planes at $x=\pm a/2$, there is nothing in the filter that can excite higher modes with more than a half-wave $x$-direction variation. The filter discontinuities will, however, excite modes having variations in the $y$ direction, although the $TE_{10}$ mode has none. Because of the $y$ axis symmetry, only modes which have $2n$ half-wave variations in this direction are excited. That is, only $TE_{1,2n}$ and $TM_{1,2n}$ modes can exist ($n=0, 1, 2, 3, \cdots$). The $TE$ higher modes are needed to provide at the $x$-$y$-plane

[6] W. C. Hahn, "A new method for the calculation of cavity resonators," *Jour. Appl. Phys.*, vol. 12, p. 62; 1941.

[7] D. S. Saxon, "Notes on Lectures by Julian Schwinger—Discontinuities in Wave Guides," MIT Radiation Laboratory Theoretical Group, February, 1945.

[8] J. W. Miles, "The equivalent circuit for a plane discontinuity in a cylindrical wave guide," Proc. I.R.E., vol. 34, pp. 728-743; October, 1946.

conducting surfaces an $H_z$ field which will cancel the $H_z$ component of the $TE_{10}$ mode. The $TM$ higher modes are needed in order to provide an $E_z$ component for the fringing electric field which must exist at these same conductors.

Consider the region of height $b$ and length $l$ in Fig. 2(a). The origin of the $x, y, z$ co-ordinate system is taken at the center of this region. Owing to reflections at both ends, there will be waves of all possible modes traveling in both the plus and minus $z$ direction. That is, there will be a set of modes multiplied by $e^{-\gamma_n z}$, and a set multiplied by $e^{+\gamma_n z}$. The total field may, therefore, be written as a set of modes multiplied by $\sinh \gamma_n z$, plus a set multiplied by $\cosh \gamma_n z$.

The problem can be simplified at this point if one recognizes that an $E_x$ component of field can not be excited by this filter structure. Since both the $TE_{1,2n}$ and $TM_{1,2n}$ modes have $E_x$ components for $n$ greater than zero, one must choose a linear combination of these modes for each value of $n$ such that $E_{xn}$ is canceled. When this is done, the following field relations result for the $n$th mode. The time dependence factor $e^{j\omega t}$ is understood.

$$E_{yn} = \cos k_x x \cos k_{yn} y\,(C_n \sinh \gamma_n z + D_n \cosh \gamma_n z) \quad (7)$$

$$_{zn} = \frac{j4\pi^2}{\eta k \gamma_n \lambda_g{}^2} \cos k_x x \cos k_{yn} y (C_n \cosh \gamma_n z + D_n \sinh \gamma_n z) \quad (8)$$

$$E_{zn} = \frac{k_{yn}}{\gamma_n} \cos k_x x \sin k_{yn} y (C_n \cosh \gamma_n z + D_n \sinh \gamma_n z) \quad (9)$$

$$H_{zn} = \frac{-jk_x}{\eta k} \sin k_x x \cos k_{yn} y (C_n \sinh \gamma_n z + D_n \cosh \gamma_n z) \quad (10)$$

$$H_{yn} = \frac{jk_x k_{yn}}{\eta k \gamma_n} \sin k_x x \sin k_{yn} y (C_n \cosh \gamma_n Z + D_n \sinh \gamma_n Z) \quad (11)$$

where $k_x = \pi/a$, $k_{yn} = 2\pi n/b$, $k = 2\pi/\lambda$, $\gamma_n = \sqrt{k_x^2 + k_{yn}^2 - k^2}$, $\lambda_g$ = guide wavelength of the $TE_{10}$ mode, and $\eta = 120\pi$ ohms.

Equations (7) to (11) give the complete field in the region bounded by $x = \pm a/2$, $y = \pm b/2$, and $z = \pm l/2$. The boundary conditions on the surfaces parallel to the $z$ axis are automatically satisfied for each value of $n$. On the transverse surfaces at $z = \pm l/2$, the boundary conditions will be met through the choice of a proper superposition of modes for all values of $n$ from zero to infinity. In addition, since $E_z = E_z = H_y = 0$ for $y = 0$, the same field that satisfies Fig. 1(a) will also satisfy Fig. 1(b).

The advantage of solving the half sections of Fig. 2 separately is now apparent, since for the short-circuit case, $D_n$ is zero, while for the open-circuit case, $C_n$ is zero.

It can be shown that if the total components $E_y = \sum E_{yn}$ and $H_x = \sum H_{xn}$ are matched at the boundary plane, then the other total components must also be matched.[9] In the Appendix the matching of $E_y$ and $H_x$ is partly carried out for the short-circuited half section. The open-circuit case is similar and the derivation is therefore omitted. The open- and short-circuit admittances at $z = -l/2$ normalized with respect to the guide of height $b$ are:

$$y_{oc} = j \tan \frac{\beta l}{2} + j \frac{2b}{\lambda_g} \left\{ \frac{S_0(\delta)}{\pi^2} + \sum_{n>0} \left[ \frac{\tanh \frac{n\pi l F}{b}}{F} - 1 \right] \frac{\sin^2 \pi n \delta}{n(\pi n \delta)^2} \right\} + j\epsilon_{oc} \quad (12)$$

$$y_{sc} = -j \cot \frac{\beta l}{2} + j \frac{2b}{\lambda_g} \left\{ \frac{S_0(\delta)}{\pi^2} + \sum_{n>0} \left[ \frac{\coth \frac{n\pi l F}{b}}{F} - 1 \right] \frac{\sin^2 \pi n \delta}{n(\pi n \delta)^2} \right\} + j\epsilon_{sc} \quad (13)$$

where

$$F = \sqrt{1 - \left( \frac{b}{n\lambda_g} \right)^2}. \quad (14)$$

In these equations, $\beta = 2\pi/\lambda_g$, $\lambda_g$ is the guide wavelength of the $TE_{10}$ mode, $\delta = b'/b$, $\epsilon_{oc}$ and $\epsilon_{sc}$ are correction terms. Zero dissipation is assumed. $S_0(\delta)$ is the Hahn function of zero order. It is tabulated by Whinnery and Jamieson.[10] Their value for $\delta = 0.05$ is in error, however, and should be $S_0(0.05) = 26.23$.

The evaluation of the correction terms $\epsilon_{oc}$ and $\epsilon_{sc}$ is carried out in reference 9, but is omitted here because it is tedious and lengthy. Fortunately, for the usual filter case of $\delta$ small, the correction terms are also small, and may be closely approximated by

$$\epsilon_{oc} \approx -0.09 \frac{b}{\lambda_g} \approx \epsilon_{sc}. \quad (15)$$

More accurate values are given by (42) and (43).

The first term on the right side of (12) is the input admittance of an open-circuited line of length $l/2$ and characteristic admittance unity. The second term is a capacitive susceptance effectively shunted across the input of this line. The same remarks apply to (13), except that here the first term is the admittance of a shorted line.

If in either (12) or (13), $l$ is made indefinitely large, the capacitive susceptance term becomes that of a sin-

[9] S. B. Cohn, "A Theoretical and Experimental Study of a Waveguide Filter Structure," Office of Naval Research, Cruft Laboratory, Harvard University, Report No. 39, April 25, 1948; also, same title, doctorate thesis, Department of Engineering Science and Applied Physics, Harvard University, 1948.

[10] J. R. Whinnery and H. W. Jamieson, "Equivalent circuits for discontinuities in transmission lines," Proc. I.R.E., vol. 32, pp. 98–115; February, 1944.

gle isolated step discontinuity. Several values of this normalized susceptance $B_c/Y_0$ have been calculated for $\delta=0.1$ and are compared with values calculated from the "Wave-Guide Handbook," formula.[11] Values of the quantity $B_c\lambda_g/Y_0 b$ are given in Table I. The agreement is very good even for $b/\lambda_g$ equal to 0.9, despite the fact that the series in (13) and in $\epsilon_{sc}$ do not converge in the close vicinity of $b/\lambda_g=1$. Even for $\delta$ as large as 0.5 (12) or (13) and (15) are only 3 per cent less than the exact value for $b/\lambda_g=0$ and $l$ infinite. If the more accurate correction-term formula of (42) or (43) is used, the error is reduced to one per cent.

TABLE I

| $b/\lambda_g$ | *WGHB* Equation | Equation (12) or (13) |
|---|---|---|
| 0 | 3.84 | 3.85 |
| 0.6 | 4.36 | 4.38 |
| 0.9 | 6.41 | 6.48 |

The equivalent circuit of the double discontinuity will now be developed. Any symmetrical two-terminal-pair network or waveguide structure has the equivalent circuit shown in Fig. 4(a). The half-structure equivalent circuit is shown in Fig. 4(b). From this one sees that

$$y_{oc} = y_2, \qquad y_{sc} = y_2 + 2y_1$$

or

$$y_2 = y_{oc}, \qquad y_1 = (y_{sc} - y_{oc})/2. \tag{16}$$

On applying these to (12) and (13), one obtains

$$y_2 = j \tan \frac{\beta l}{2} + jB_{C2} \tag{17}$$

$$y_1 = -\frac{j}{2}\left(\cot \frac{\beta l}{2} + \tan \frac{\beta l}{2}\right) + jB_{C1}$$

$$= -j \csc \beta l + jB_{C1} \tag{18}$$

where

$$B_{C2} = \frac{2b}{\lambda_g}\left\{\frac{S_0(\delta)}{\pi^2} + \sum_{n>0}\left[\frac{\tanh \frac{n\pi l F}{b}}{F} - 1\right]\frac{\sin^2 \pi n\delta}{n(\pi n\delta)^2}\right\} + \epsilon_{oc} \tag{19}$$

$$B_{C1} = \frac{2b}{\lambda_g}\sum_{n>0}\frac{\operatorname{csch}\frac{2\pi n l F}{b}}{F}\,\frac{\sin^2 \pi n\delta}{n(\pi n\delta)^2} + \frac{\epsilon_{sc} - \epsilon_{oc}}{2}. \tag{20}$$

For the range of guide wavelength of interest here, $B_{C1}$ and $B_{C2}$ are capacitive susceptances, and $\tan(\beta l/2)$ and $\csc\beta l$ are positive functions which are roughly proportional to $1/\lambda_g$ and $\lambda_g$, respectively. The equivalent circuit of the portion of Fig. 2(a) between the planes $z=\pm l/2$ may therefore be drawn as shown in Fig. 4(c). Note that all element values shown on this circuit are susceptances normalized with respect to the characteristic admittance of the waveguide line of height $b$. A complete filter section will have a short length of guide of height $b'$ and length $l'/2$ at each end. The principal effect of this is to add additional shunt capacitance to the equivalent circuit. The inverse guide wavelength for the waveguide filter is seen to be analogous to frequency for the equivalent low-frequency circuit. Since the circuit is that of an $m$-derived low-pass filter, one may expect the waveguide filter to have characteristics similar to its low-frequency equivalent, except for the change in the frequency scale. Note that without the bridging susceptance $B_{C1}$, the infinite-attenuation point found with the waveguide filter could not exist.

The portion of Fig. 4(c) within the dotted line is the exact equivalent of a transmission line of length $l$ and characteristic admittance unity. Hence the equivalent circuit of the double discontinuity may be drawn as in Fig. 4(d). (The last circuit is a symbolic one which must be redrawn as in Fig. 4(c) before it can be used.)

## The Filter Relations

The values of $y_{oc}$ and $y_{sc}$ given in (12) and (13) are for $l'=0$. If the symbols $y_{oc}$ and $y_{sc}$ are now applied to a filter half section having $l'$ greater than zero, and $y_{oc}'$ and $y_{sc}'$ to the case of $l'$ equal to zero, then the actual half-section open- and short-circuit admittances are

$$y_{oc} = jY_{02}\tan\left[\frac{\pi l'}{\lambda_g} + \tan^{-1}\left(\frac{Y_{01}y_{oc}'}{Y_{02}j}\right)\right]$$

$$y_{sc} = jY_{02}\tan\left[\frac{\pi l'}{\lambda_g} + \tan^{-1}\left(\frac{Y_{01}y_{sc}'}{Y_{02}j}\right)\right].$$

(These equations may be derived from King, Mimno, and Wing.[12]) But $Y_{01}/Y_{02}=b'/b=\delta$, and $Y_{01}$ was set equal to one in the last section. Therefore,

$$y_{oc} = \frac{j}{\delta}\tan\left[\frac{\pi l'}{\lambda_g} + \tan^{-1}\left(\frac{\delta y_{oc}'}{j}\right)\right] \tag{21}$$

$$y_{sc} = \frac{j}{\delta}\tan\left[\frac{\pi l'}{\lambda_g} + \tan^{-1}\left(\frac{\delta y_{sc}'}{j}\right)\right] \tag{22}$$

where $y_{oc}$ and $y_{sc}$ are normalized with respect to a guide of height $b$. $y_{oc}'$ and $y_{sc}'$ are obtained from (12) and (13). The image parameters of the filter section may now be calculated by means of (4), (5), (21), and (22).

The upper cutoff of the principal pass band occurs for the first zero of (22), since the image admittance changes from real to imaginary at that point. Let $\lambda_{g1}$ be the value of $\lambda_g$ at the upper cutoff frequency. Equation (22)

[11] "Wave-Guide Handbook, First Revised Edition," Radiation Laboratory Report No. 43, February 7, 1944. Also, N. Marcuvitz. "Wave-Guide Handbook," Radiation Laboratory Series, vol. 10, McGraw-Hill Book Co. New York, N. Y., to be published.

[12] 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., 1945.

is zero for

$$\tan \frac{\pi l'}{\lambda_{g1}} + \frac{\delta y_{sc}'}{j} = 0. \tag{23}$$

It is seen from (5) that the infinite-attenuation point occurs for $y_{sc} = y_{oc}$.

In Fig. 5 are shown typical image-admittance, phase-constant, and attenuation-constant curves. Inspection of the image-admittance curve shows that it follows the low-frequency constant-$k$ image-admittance function very closely in the pass band. The latter function is given by

$$y_I = y_{I0}\sqrt{1 - (f/f_1)^2}. \tag{24}$$

where $y_{I0}$ is the value of image admittance at zero frequency and $f_1$ is the cutoff frequency. The variable $f$ for the low-frequency filter is, of course, the analog of $b/\lambda_g$ for the waveguide filter, and $f_1$ corresponds to $b/\lambda_{g1}$.

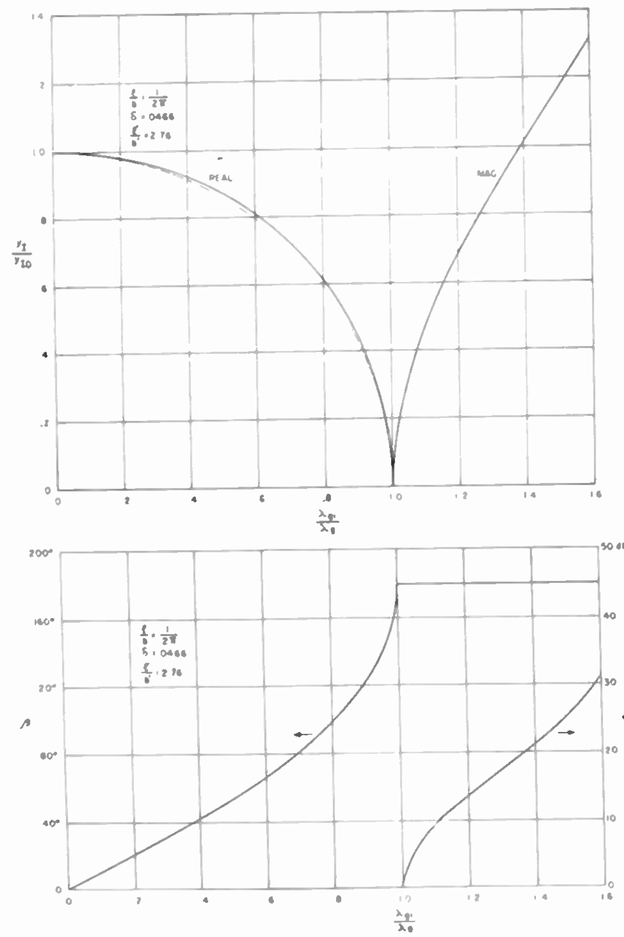


Fig. 5—Image parameters of a filter section. The image admittance of a low-frequency shunt-terminated constant-$k$ filter is shown as a dashed curve.

Because of the fact that the waveguide-filter image admittance is very nearly independent of the filter parameters, one may match dissimilar sections with very little error by merely making their cutoff frequencies and the zero-frequency image admittance referred to the terminating lines the same. This makes it possible to design a filter to have sections with different values of infinite-attenuation wavelength, so as to obtain a sharp cutoff characteristic, and maintain high attenuation far into the stop band. This procedure is common in the design of low-frequency filters.

In a paper now under preparation, the author will show how the above analysis may actually be used in the design of practical filters, and a set of design curves will be presented which will obviate the use of the complicated relations given here. Such matters as matching, spurious responses, etc., will be taken up. Tests will be described which have shown that the analysis and design procedure may be relied on to give cutoff and infinite-attenuation frequencies within one or two per cent.

## Application to Ridge Waveguide

The solution of the double discontinuity of Fig. 2(a) and of the two subcases of a single discontinuity in a waveguide terminated by an electric wall and by a magnetic wall (Figs. 2(b) and 2(c)) should find uses other than the application to the waveguide filter. One such use will now be described.

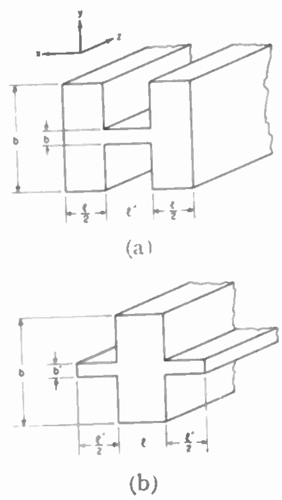


Fig. 6—Waveguide cross sections whose properties may be deduced from the waveguide-filter solution.

Fig. 6(a) shows the cross section of symmetrical ridge waveguide. The *approximate* cutoff frequency and characteristic impedance of ridge waveguide has been derived and presented graphically.[13] It will now be shown that the analysis in this article can be extended to give a more exact solution for ridge waveguide.

At the cutoff frequency, one may regard the field distribution in ridge guide as due to waves traveling transversely in the plus and minus $x$ direction. If the structure of Fig. 6(a) is compared with that of Fig. 2(b), it is seen that the ridge-guide cross section has the shape of two short-circuited filter half sections joined together. At the cutoff frequency of the fundamental mode in ridge waveguide, the ratio $H_z/E_y$ must be zero at the center of the cross section (see reference 13). Hence this

[13] S. B. Cohn, "Properties of ridge wave guide," Proc. I.R.E., vol. 35, pp. 783–789; August, 1947.

condition corresponds to $y_{s} = -H_x/E_y = 0$ for the filter half section, which is the condition for cutoff of the waveguide filter. The wavelength at which ridge wave guide cuts off is, therefore, the same as that at which the corresponding filter structure cuts off. This is given by the root of (23) with $\lambda_g$ set equal to space wavelength $\lambda$. Since the filter analysis can give the field components in ridge guide, it is also possible to calculate the exact characteristic impedance.

In a similar way, the exact cutoff frequency and characteristic impedance of the cross section shown in Fig. 6(b) might be derived from the open-circuit half-section case of Fig. 2(c).

## Acknowledgment

The writer wishes to acknowledge the many helpful suggestions of R. W. P. King and L. Brillouin.

## Appendix

For the short-circuit case of Fig. 2(b), set $D_n = 0$ in (7) to (11). Then, with $x = 0$ and $z = -l/2$, the $n$th-mode field in region $A$ is

$$E_{yn} = -C_n \cos k_{yn}y \sinh \gamma_{An}l/2 \tag{25}$$

$$H_{xn} = \frac{j4\pi^2 C_n}{\eta k \gamma_{An}\lambda_g^2} \cos k_{yn}y \cosh \gamma_{An}l/2. \tag{26}$$

The other field components are zero for $x = 0$.

The total region-$A$ field for $x = 0$ and $z = -l/2$ may be written as

$$E_{yA} = R_0 + \sum_{n>0} R_n \cos \frac{2\pi n y}{b} \tag{27}$$

$$H_{xA} = w_{A0}R_0 + \sum_{n>0} w_{An}R_n \cos \frac{2\pi n y}{b}. \tag{28}$$

where

$$R_0 = -jC_0 \sin(\beta l/2) \tag{29}$$

$$R_n = -C_n \sinh(\gamma_{An}l/2), \quad n > 0 \tag{30}$$

$$w_{A0} = jy_0 \cot(\beta l/2) \tag{31}$$

$$w_{An} = -y_{An} \coth(\gamma_{An}l/2) \tag{32}$$

$$y_0 = \lambda/\eta\lambda_g \tag{33}$$

$$y_{An} = \frac{j4\pi^2}{\eta k \gamma_{An}\lambda_g^2} \tag{34}$$

$$\beta = 2\pi/\lambda_g \tag{35}$$

$$\gamma_{An} = \sqrt{\left(\frac{\pi}{a}\right)^2 + \left(\frac{2\pi n}{b}\right)^2 - \left(\frac{2\pi}{\lambda}\right)^2}. \tag{36}$$

$y_0$ is the characteristic wave admittance of the $TE_{10}$ mode; $y_{An}$ is the characteristic wave admittance of the higher composite $TE$-$TM$ modes; $w_{A0}$ and $w_{An}$ are the negative of the wave admittances for the fundamental and higher modes at $z = -l/2$—that is, they are the ratio $+H_{xn}/E_{yn}$ for the region-$A$ fields at that plane.

For the $B$ region, one may write

$$E_{yB} = B_0 + \sum_{n>0} B_n \cos \frac{2\pi n y}{b'} \tag{37}$$

$$H_{xB} = w_{B0}B_0 + \sum_{n>0} y_{Bn}B_n \cos \frac{2\pi n y}{b'} \tag{38}$$

where

$$y_{Bn} = \frac{j4\pi^2}{\eta k \gamma_{Bn}\lambda_g^2} \tag{39}$$

$$\gamma_{Bn} = \sqrt{\left(\frac{\pi}{a}\right)^2 + \left(\frac{2\pi n}{b'}\right)^2 - \left(\frac{2\pi}{\lambda}\right)^2}. \tag{40}$$

$w_{B0}$ is the negative of the $TE_{10}$ wave admittance in region $B$, referred to the plane $z = -l/2$. Since this wave admittance is the short-circuit wave admittance $w_{sc}$ for the half-section, one may write

$$w_{sc} = -w_{B0}. \tag{41}$$

The characteristic admittance in guide $B$ is $\delta$ times that of guide $A$, and hence the normalized short-circuit admittance *referred to guide A* is $w_{s}/\delta y_0$, which will be defined as $y_{s}$.

$E_{yA}$ is next set equal to $E_{yB}$, and $H_{xA}$ is set equal to $H_{xB}$. From these relations, the mode amplitudes and $w_s$ may be determined. The method used is that of Hahn,[6] and the work is similar to that of Whinnery and Jamieson for a capacitive step in a parallel-plane transmission line.[10] For the sake of brevity, the remainder of this analysis is omitted. The derivation of $y_{oc}$ is similar to that of $y_{s}$, and is therefore omitted entirely.

The exact evaluation of $\epsilon_s$ and $\epsilon_{oc}$ requires the solution of an infinite set of linear equations having an infinite number of unknowns. Good approximations for $\epsilon_s$ and $\epsilon_{oc}$ are[9]

$$\epsilon_{sc} \approx -0.0417 \frac{b}{\lambda_g} \sum_{m=1} \frac{[L_m(\delta)]^2}{m^3} \tag{42}$$

$$\epsilon_{o} \approx -0.0417 \frac{b}{\lambda_g} \sum_{m=1} \frac{[T_m(\delta)]^2}{m^3} \tag{43}$$

where

$$T_m(\delta) = S_m(\delta) + \sum_{n=1} \left[\frac{b}{F}\tanh\frac{n\pi l}{F} - 1\right] \frac{\sin^2 \pi n\delta}{n\left[\left(\frac{n\delta}{m}\right)^2 - 1\right]} \tag{44}$$

$$L_m(\delta) = S_m(\delta) + \sum_{n=1} \left[\frac{b}{F}\coth\frac{n\pi l}{F} - 1\right] \frac{\sin^2 \pi n\delta}{n\left[\left(\frac{n\delta}{m}\right)^2 - 1\right]} \tag{45}$$

$S_m(\delta)$ is a Hahn function defined and tabulated in reference 6. For $\delta$ small, (15) is usually a satisfactory approximation for (42) and (43).

# Correspondence

## Notes on Digital Coding*

The consideration of message coding as a means for approaching the theoretical capacity of a communication channel, while reducing the probability of errors, has suggested the interesting number theoretical problem of devising lossless binary (or other) coding schemes serving to insure the reception of a correct, but reduced, message when an upper limit to the number of transmission errors is postulated.

An example of lossless binary coding is treated by Shannon[1] who considers the case of blocks of seven symbols, one or none of which can be in error. The solution of this case can be extended to blocks of $2^n-1$-binary symbols, and, more generally, when coding schemes based on the prime number $p$ are employed, to blocks of $p^n-1/p-1$ symbols which are transmitted, and received with complete equivocation of one or no symbol, each block comprising $n$ redundant symbols designed to remove the equivocation. When encoding the message, the $n$ redundant symbols $x_m$ are determined in terms of the message symbols $Y_k$ from the congruent relations

$$E_m \equiv X_m + \sum_{k=1}^{k=(p^n-1)/(p-1)-n} a_{mk} Y_k \equiv 0 \pmod{p}.$$

In the decoding process, the $E$'s are recalculated with the received symbols, and their ensemble forms a number on the base $p$ which determines univocally the mistransmitted symbol and its correction.

In passing from $n$ to $n+1$, the matrix with $n$ rows and $p^n-1/p-1$ columns formed with the coefficients of the $X$'s and $Y$'s in the expression above is repeated $p$ times horizontally, while an $(n+1)$ st row added, consisting of $p^n-1/p-1$ zeroes, followed by as many one's etc. up to $p-1$; an added column of $n$ zeroes with a one for the lowest term completes the new matrix for $n+1$.

If we except the trivial case of blocks of $2S+1$ binary symbols, of which any group comprising up to $S$ symbols can be received in error which equal probability, it does not appear that a search for lossless coding schemes, in which the number of errors is limited but larger than one, can be systematized so as to yield a family of solutions. A necessary but not sufficient condition for the existence of such a lossless coding scheme in the binary system is the existence of three or more first numbers of a line of Pascal's triangle which add up to an exact power of 2. A limited search has revealed two such cases; namely, that of the first three numbers of the 90th line, which add up to $2^{12}$ and that of the first four numbers of the 23rd line, which add up to $2^{11}$. The first case does not correspond to a lossless coding scheme, for, were such a scheme to exist, we could designate by $r$ the number of $E_m$ ensembles corresponding to one error and having an odd number of 1's and by $90-r$ the remaining (even) ensembles. The odd ensembles corresponding to two transmission errors could be formed by re-entering term by term all the conbinations of one even and one odd ensemble corresponding each to one error, and would number $r(90-r)$. We should have $r+r(90-r)=2^{11}$, which is impossible for integral values of $r$.

On the other side, the second case can be coded so as to yield 12 sure symbols, and the $a_{mk}$ matrix of this case is given in Table I. A second matrix is also given, which is that of the only other lossless coding scheme encountered (in addition to the general class mentioned above) in which blocks of eleven ternary symbols are transmitted with no more than 2 errors, and out of which six sure symbols can be obtained.

It must be mentioned that the use of the ternary coding scheme just mentioned will always result in a power loss, whereas the coding scheme for 23 binary symbols and a maximum of three transmission errors yields a power saving of $1\frac{1}{2}$ db for vanishing probabilities of errors. The saving realized with the coding scheme for blocks of $2^n-1$ binary symbols approaches 3 db for increasing $n$'s and decreasing probabilities of error, but a loss is always encountered when $n=3$.

MARCEL J. E. GOLAY
Signal Corps Engineering Laboratories
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TABLE I

| | $Y_1$ | $Y_2$ | $Y_3$ | $Y_4$ | $Y_5$ | $Y_6$ | $Y_7$ | $Y_8$ | $Y_9$ | $Y_{10}$ | $Y_{11}$ | $Y_{12}$ |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| $X_1$ | 1 | 0 | 0 | 1 | 1 | 1 | 0 | 0 | 0 | 1 | 1 | 1 |
| $X_2$ | 1 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 0 | 1 |
| $X_3$ | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 0 | 1 | 0 |
| $X_4$ | 1 | 0 | 1 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 0 | 0 |
| $X_5$ | 1 | 1 | 0 | 0 | 1 | 1 | 1 | 0 | 1 | 1 | 0 | 0 |
| $X_6$ | 1 | 1 | 0 | 1 | 0 | 1 | 1 | 1 | 0 | 0 | 0 | 1 |
| $X_7$ | 1 | 1 | 0 | 1 | 1 | 0 | 0 | 1 | 1 | 0 | 1 | 0 |
| $X_8$ | 1 | 1 | 1 | 0 | 0 | 1 | 0 | 1 | 0 | 1 | 1 | 0 |
| $X_9$ | 1 | 1 | 1 | 0 | 1 | 0 | 1 | 0 | 0 | 0 | 1 | 1 |
| $X_{10}$ | 1 | 1 | 1 | 1 | 0 | 0 | 0 | 0 | 1 | 1 | 0 | 1 |
| $X_{11}$ | 0 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 |

| | $Y_1$ | $Y_2$ | $Y_3$ | $Y_4$ | $Y_5$ | $Y_6$ |
|---|---|---|---|---|---|---|
| $X_1$ | 1 | 1 | 1 | 2 | 2 | 0 |
| $X_2$ | 1 | 1 | 2 | 1 | 0 | 2 |
| $X_3$ | 1 | 2 | 1 | 0 | 1 | 2 |
| $X_4$ | 1 | 2 | 0 | 1 | 2 | 1 |
| $X_5$ | 1 | 0 | 2 | 2 | 1 | 1 |

* Received by the Institute, February 23, 1949.

[1] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. 27, p. 418; July, 1948.

# Contributors to Proceedings of the I.R.E.

J. T. BOLLJAHN

J. T. Bolljahn (A'43) was born in Oakland, Calif., on June 10, 1918. He received the B.S. degree in electrical engineering from the University of California in 1941. 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 primarily with the development of aircraft antennas.

Since February, 1946, he has been a part-time graduate student and a member of the research staff of the Antenna Laboratory at the University of California.

JAMES G. BUCK

James G. Buck was born in 1918 in Torrington, Conn. He received the A.B. degree from Dartmouth College in 1940, and the Ph.D. degree from MIT in 1942. During the academic year 1940–1941, he was a fellow of the Cramer Foundation. From 1942 until 1946 he was a member of the staff at the Massachusetts Institute of Technology Radiation Laboratory. During part of 1944 and 1945, while on leave of absence from the Radiation Laboratories, he worked on the Manhattan project for the Westinghouse Electric Company.

In 1946, Professor Buck was appointed assistant professor of physics at the University of Notre Dame. He is a member of the American Physical Society, the Instrument Society of America, the American Association for the Advancement of Science, Phi Beta Kappa, and Sigma Xi.

For a photograph and biography of EDWARD L. GINZTON, see page 1002 of the August, 1948, issue of the PROCEEDINGS OF THE I.R.E.

For a photograph and biography of J. R. PIERCE see page 1003 of the August, 1948, issue of the PROCEEDINGS OF THE I.R.E.

# Contributors to Proceedings of the I.R.E.

Herbert J. Carlin (M'47) was born in New York, N. Y., in 1917. He attended Columbia University, receiving the B.S. degree in 1938, and the M.S. degree in 1940. In 1947 he was awarded the D.E.E. degree from the Polytechnic Institute of Brooklyn.

H. J. Carlin

From 1940 to 1945 Dr. Carlin was associated with the Westinghouse Company as a design engineer in the power-system relay section of the meter division, and has written several papers on power-system protection. He joined the Microwave Research Institute of the Polytechnic Institute of Brooklyn in 1945, and has made contributions in the field of microwave networks and microwave power measurements. He holds the position of research supervisor, and lectures in the Graduate School at the Institute.

Dr. Carlin is a member of the American Physical Society, the American Institute of Electrical Engineers, Tau Beta Pi, and Sigma Xi.

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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, as an assistant professor.

Marivn Chodorow

Dr. Chodorow is a member of the American Physical Society, and of Sigma Xi.

❖

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. He received 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.

Seymour B. Cohn

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.

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.

❖

E. A. Coomes (M'46–SM'46) was born in Louisville, Ky., on June 27, 1909. He received the B.S. in electrical engineering from the University of Notre Dame in 1931, and M.S. degree in 1933. In 1937 he was awarded the Sc.D. degree in applied physics from the Massachusetts Institute of Technology. From 1933 to 1935, and from 1937 to 1942, he was a member of the teaching and research staffs of the physics department at the University of Notre Dame.

E.A. Coomes

In 1942, Dr. Coomes was granted leave of absence to join the staff of Radiation Laboratory to carry out research on oxide cathodes for microwave magnetrons. He returned to Notre Dame as professor of physics and senior member of the Physical Electronics Group in 1945. He is a Fellow of the American Physical Society.

❖

Edwin Istvánffy was born in Parkany, Hungary, on January 4, 1895. He received the M.Sc. degree in 1922 from the Technical University of Budapest. Since 1923, he has been with the United Incandescent Lamp and Electrical Co. Ltd., Ujpest, Hungary. In 1928 when the Standard Electric Co. Ltd., Budapest, an International Telephone and Telegraph associate, was established, he joined this company, becoming the technical director in 1938. Most of his work is connected with radio development.

Edwin Istvánffy

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For a photograph and biography of T. G. Mihran, see page 1135 of the September, 1948, issue of the Proceedings of the I.R.E.

John F. Kane was born on January 18, 1922, in Brewer, Maine. He received the B.S. degree in engineering physics from the University of Maine in December, 1943. Following his graduation, he joined the Sperry Gyroscope Company as an assistant project engineer on the microwave development staff, where he remained until April, 1946. Since that time, Mr. Kane has been a candidate for the Ph.D. degree in physics at Stanford University, California, where he holds a teaching and research fellowship.

John F. Kane

Mr. Kane is a member of the American Physical Society, and an associate member of Sigma Xi.

❖

Laurence A. Manning (S'43–A'45) was born in Palo Alto, Calif., on April 28, 1923. He received the A.B. and M.S. degrees in electrical engineering from Stanford University in 1944 and 1947, respectively. From 1944 to 1945 Mr. Manning was engaged in broad-band microwave oscillator development at the Radio Research Laboratory, Harvard University. Returning to Stanford after the war, he served successively as a teaching assistant in physics, research associate in electrical engineering, and, since 1947, as an acting assistant professor in electrical engineering. He is currently engaged in a study of the effect of meteors upon the upper atmosphere.

L. A. Manning

Mr. Manning is a member of Sigma Xi, Phi Beta Kappa, and of Commission 3, the USA National Committee of the URSI.

❖

Allen S. Meier was born on March 9, 1911, at Windsor, Conn. He received the B.S. degree in physics and mathematics in 1932, and the M.S. degree in physics in 1933 from Trinity College, Hartford, Conn. He was associated with the Pratt and Whitney Machine Tool Co. and the Connecticut Mutual Life Insurance Co. before service in the Army in 1940. After serving with the 704th Military Police Battalion and the 242nd Coast Artillery Corps, he was transferred to the Signal Corps in 1942, and was assigned to the Radiation Laboratory

Allen S. Meier

# Contributors to Proceedings of the I.R.E.

of the Massachusetts Institute of Technology, where he was engaged in research and development.

From 1942 to 1946, he served as branch officer of the Antenna Branch, Special Projects Laboratory, Aircraft Radio Laboratories, at Wright Field, Ohio. Following release from the Army, he was employed as project engineer at the Antenna Laboratory, Ohio State University Research Foundation. During 1947, he was associated with the Kinsey Electric Products Corp., Fairburn, Ga., as secretary of the company, and has recently been engaged as a development engineer at Sargent and Co., New Haven, Conn.

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B. Y. Mills was born at Sydney, Australia, in 1920. He received the B.Sc. degree in 1940, and the B.E. degree in 1942, both from Sydney University. He immediately joined the staff of the Radiophysics Division of the Council for Scientific and Industrial Research, and until 1946 was engaged in research and development work in the radar field, particularly in regard to receiver and indicator systems. During the next two years he was interested in the application of radar techniques to linear accelerators, and was successful in developing a million-volt X-ray tube of this type. He has recently begun work in the field of radio astronomy.

B. Y. Mills

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Morton Nadler (A'45) was born in Brooklyn, N. Y., in 1921. He received the bachelor's degree from the College of the City of New York, and undertook subsequent graduate work at the University of Southern California and the Illinois Institute of Technology. During the war he was employed in various electronic instrument research programs under the office of Scientific Research and Development, at the University of Southern California. In the post-war period he worked in television research, first at Belmont Radio Corporation, and then as chief of television receiver design at Sonora Radio and Television Corporation.

Morton Nadler

At present Mr. Nadler is a member of the research staff of Tesla, the national electrical enterprise of the Czechoslovak Republic, in Prague, as director of the electronic applications section.

Greenleaf W. Pickard (M'12–F'15) was born on February 14, 1877, in Portland, Me. He has been associated with radio since 1901, when he became engineer for the American Wireless Telegraph and Telephone Company. In 1902, he was made chief engineer for the Federal Wireless Telegraph and Telephone Company. Later, he joined the American Telephone and Telegraph Company, remaining until 1907, when he organized the Wireless Specialty Apparatus Company, which became the R.C.A. Victor Company of Massachusetts. From 1942 to 1945 Mr. Pickard was director of research for the American Jewels Corporation. He is now associated with the Cosmic Terrestrial Research Laboratory, Massachusetts Institute of Technology, and with the firm of Pickard and Burns, consulting engineers.

G. W. Pickard

Mr. Pickard received the IRE Medal of Honor in 1926 for his "contributions as to crystal detectors, coil antennas, wave propagation and atmospheric disturbances." He also was the recipient of the Armstrong Medal of the Radio Club of America in 1940. He has served on numerous IRE committees, including the Board of Editors, Constitution and Laws, Wavelength Regulation, and Wave Propagation, and was actively associated with the organization of the Boston Section, in 1914. Mr. Pickard was a member of both the Wireless Institute and the Society of Wireless Telegraph Engineers, when these two organizations fused into the present IRE on May 13, 1912. He was President of the IRE in 1913.

Mr. Pickard is a Fellow of the American Academy of Arts and Sciences, as well as of the American Institute of Electrical Engineers and the Radio Club of America. He also holds membership in the American Geophysical Union, the American Meteorological Society, and is a life member of the *Société des Radoélectriciens.*

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Harlan T. Stetson (A'31) was born in Haverhill, Mass., on June 28, 1885. After graduating from Brown University, he received the Sc.M. degree from Dartmouth College in 1910, and the Ph.D. degree from the University of Chicago in 1915. He was associated with the physics department at Dartmouth for four years, and later taught astronomy and mathematics at Northwestern University. From 1916 to 1929, while he was an assistant professor of astronomy at Harvard University, Dr. Stetson became associated with Mr. Pickard in the investigation of the effect of sunspots on radio reception. From 1929 to 1934 Dr. Stetson was Perkins professor of astronomy at Ohio Wesleyan University and director of the Perkins Observatory, as well as lecturer at Ohio State University. He returned to Harvard in 1934 as a research associate in geophysics. He joined the Massachusetts Institute of Technology in 1936, and is now the director of the Cosmic Terrestrial Research Laboratory at Needham, Mass.

H. T. Stetson

Dr. Stetson is the author of numerous papers on solar activity, radio reception, and ionization of the upper atmosphere. He has been chairman of the Special Committee on Cosmic Terrestrial Relationships of the American Geophysical Union, National Research Council, since 1938. He is the author of "Man and the Stars," "Earth, Radio and the Stars," "Sunspots and Their Effects," and "Sunspots in Action."

❖

Willard P. Summers (S'45–A'47) was born on August 7, 1922, at Bellevue, Ohio. He attended Findlay College, received the B.E.E. degree from Ohio State University in June, 1946, and did graduate work in antenna theory from 1947 to 1948.

Willard P. Summers

From 1944 to 1945, Mr. Summers was an engineer at WOSU. He was a research associate at the Antenna Laboratory, Ohio State University, from June, 1946, to August, 1948.

At present, Mr. Summers is in charge of instrumentation for the Aeronautical Research Laboratory, Ohio State University. He is a member of Eta Kappa Nu and Sigma Pi Sigma.

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Nelson Wax was born on April 2, 1917, at Philadelphia, Pa. He received the B.S. degree in 1937 and the M.S. degree in 1938, from the Moore School of Electrical Engineering, University of Pennsylvania, and the Ph.D. degree in 1942 from the Ohio State University. Dr. Wax was with the Bell Telephone Laboratories, Inc., from 1942 to 1948. Since February, 1948, he has been an assistant professor of electrical engineering at the University of Illinois.

Nelson Wax

# Institute News and Radio Notes

## Technical Committee Notes

The **Standards Committee** held an annual meeting on March 8, 1949. The new chairman is J. G. Brainerd of the University of Pennsylvania; A. G. Jensen of BTL and L. G. Cumming of IRE Headquarters are vice-chairmen. Standards on *Radio Aids to Navigation: Definitions of Terms, 1949;* and *Railroad and Vehicular Communications: Methods of Testing, 1949* have been approved by the Committee and will shortly be published. Mr. Jensen reported for the Definitions Co-ordinating Subcommittee, which has two purposes: to save work and time for groups writing definitions by calling attention to any overlapping in work, and to establish a more efficient transfer of information between the IRE and other organizations. It is hoped that the Subcommittee will receive definitions while they are in a tentative stage, so that they may be referred to ASA C-42 Committee for agreement prior to adoption by the IRE. The **IRE/AIEE Co-ordinating Committee on Definitions** has agreed to set up a group to work on the term "Noise." It is advisable that no other committee work on this subject until the group's task is completed. The meeting was addressed by the Institute's Standards Co-ordinator, W. R. G. Baker, who offered the IRE's thanks to A. B. Chamberlain for his fine work as Chairman of the Standards Committee, and also thanked the chairmen of all the technical committees for their accomplishments during the year. He spoke of a new method of reproducing Standards by photo-offset, which will require strict adherence to the rules governing preparation of drafts of standards. . . . A meeting of the **Antennas and Wave Guides Committee** was held on March 30 in Washington, D. C., and L. C. Van Atta was chosen chairman for the coming year. A vote of thanks was given to the retiring chairman, P. S. Carter. . . . J. G. Brainerd, retiring chairman of the **Circuits Committee**, introduced the new chairman, W. N. Tuttle, at the March 9 meeting. The Committee is continuing its work on definitions, and subcommittees will be formed for this activity. . . . L. R. Nergaard will assume the duties of chairman of the **Committee on Electron Tubes** and Solid State Devices, it was announced on March 7. The Committee plans to have a standard on definitions of terms ready for publication shortly. . . . At the meeting of the **Receivers Committee** on March 10, R. F. Shea was introduced as the new chairman. W. O. Swinyard is the retiring chairman. . . . The **Research Committee** held a meeting on March 9, during which D. E. Chambers led the Committee in a discussion of its scope. The Committee agreed to secure for publication interpretative or review papers and original papers, to promote a "Letters to the Editor" page in the PROCEEDINGS, to prepare abstracts, and to report on technical meetings. . . . The **Video Techniques Committee,** under the chairmanship of J. E. Keister, met on March 9. Three new subcommittees have been formed: Definitions, headed by R. H. Daugherty, Jr.; Methods of Measurement and Test for Video Systems and Components, headed by W. J. Poch; and Methods of Measurement and Test for Video Signal Transmission, headed by L. W. Morrison. Besides the material which these new subcommittees are preparing for standardization, the Subcommittee on Methods of Measurement and Test for Utilization, Including Video Recording, under the chairmanship of Vernon J. Duke, is preparing six tutorial reports for publication. . . . At the meeting of the **Wave Propagation Committee** on March 9, it was announced that C. R. Burrows would head the Committee for the coming term. The Committee reviewed the *Glossary of Terms Used in Radio Propagation,* which had been circulated for comment by the British Standards Institute. It also discussed its activities in connection with the JTAC. . . . The **Committee on Professional Groups** met on March 7. The chairman W. L. Everitt, announced that three Groups have been formally organized and approved: Audio, Broadcast Engineers, and Antennas and Wave Propagation. Six others are under consideration: Circuit Theory, Nuclear Science, Vehicular and Railroad Radio Communications, Receivers, Transmitters, and Electron Tubes. The Committee shall have primary jurisdiction on policy, which will include the actual approval of Groups and their objectives. Operation of groups will be under the control of the Executive Committee through the Standards Co-ordinator. Minor revisions of the Professional Groups Manual were authorized, and a plan was adopted for stimulating action in the Sections for forming Section Groups, the Committee acting as a clearing house for the Sections in supplying information, speakers, papers, etc. It was decided that all members of Professional Groups, national or sectional, must be members of the Institute. All Groups shall hold one technical conference of national scope each year.

### Calendar of COMING EVENTS

Central Section Regional Meeting, Society of Motion Picture Engineers, Toledo, Ohio, June 10

**IRE Conference on Electronic Devices, Princeton, N. J., June 20 and 21**

AIEE Summer General Meeting, Swampscott, Mass., June 20–24

NBS Symposium on Construction and Applications of Conformal Maps, Los Angeles, Calif., June 24–25.

NBS Symposium on Probability Methods in Numerical Analysis. Los Angeles, Calif., June 27–29.

AIEE Pacific General Meeting, San Francisco, Calif., August 23–26

**1949 IRE West Coast Convention, San Francisco, Calif., August 30–September 2**

**1949 National Electronics Conference, Chicago, Ill., September 26–28**

National Radio Exhibition, Olympia, London, England, September 28 to October 28

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21

**1950 IRE National Convention, New York, N.Y., March 6–9**

## Military Standards Agency Reorganized

Under the terms of a charter signed by the assistant secretaries of the three military services, the government body organized in December, 1943, as the Army-Navy Electronic Standards Agency has been reconstituted as the Armed Services Electro Standards Agency. The change provides for official participation by the Air Force, which became a separate component of the National Military Establishment under the provisions of the National Security Act of 1947.

The purpose of the Standards Agency is to reduce the number of styles and types of electronic components used in the manufacture of military equipment of all kinds, to insure their quality and dependability, and to designate approved sources of supply.

## British "Radiolympia" Exhibit to be Held in Fall

Television, as perfect technically as it can be demonstrated in all stages from studio to receiver, will be a feature of the sixteenth National Radio Exhibition ("Radiolympia") to be held at Olympia, London, England, from September 28 to October 8.

Other displays will include an enlarged section for communications equipment, navigational aids, and electronics in industry, and exhibits for the first time since the war by the Royal Navy, the Army, and the Royal Air Force. The Ministry of Supply, the Department of Scientific and Industrial Research, and the GPO will also exhibit.

## Constitutional Amendment Ballot

In accordance with Article IX, Section 2 of the Institute Constitution, the ballots on amending Article II, Section 1(c) and Article II, Section 2(c) of the IRE Constitution were counted on May 3, 1949, under the supervision of the Tellers Committee, J. L. Callahan, Acting Chairman. The final result of the ballot count is as follows:

| | *Yes* | *No* | *Total* |
|---|---|---|---|
| Total | 812 | 2334 | 3146 |
| Per cent | 25.8 | 74.2 | 100 |

Total ballots mailed. . . . . . . . . . . . . 7,015
Total ballots received. . . . . . . . . . . 3,189
—45.4 per cent of ballots mailed
Total ballots void. . . . . . . . . . . . . . . 43
— 0.6 per cent of ballots mailed
Total ballots counted. . . . . . . . . . . . 3,146
—44.8 per cent of ballots mailed

# Report of the Secretary—1948

To the Board of Directors,
The Institute of Radio Engineers

Gentlemen:

The Secretary again submits a report, required annually as stipulated in the By-laws, of the status of Institute affairs resulting from activities during 1948.

It is especially gratifying to note that Institute growth continues unabated both in membership and activities. Membership is spread over 63 countries. Student Branches have increased over 100 per cent from 27 to 59. Sections increased from 43 to 48 or about 11 per cent. Additional data and the usual charts are included herewith. (Figs. 1–4).

The past year has seen the inauguration of the Regional Director plan, with a reduced number of Board of Directors meetings. The attendance at these was excellent, and not only were regional problems more fully dealt with, but members of the various Sections enjoyed a closer relationship with the Institute through their ability to express views through the Regional Directors. These Directors carry to their Regions authentic information resulting from their participation in the operations of the Board. A plan was adopted to render limited financial assistance for the handling of approved Regional meetings, which should go far in promoting activities in these several far-flung areas that have area interests.

Further encouragement toward better Section activities was provided through the increase in Section membership rebates, which took effect at the first of the year.

The year 1948 saw the commencement of a new conception created by the Board, the Professional Group system. This plan was evolved in an effort to develop a method whereby members having common interests in a given specialized field can give expression to those special interests and develop related activities. The rapid growth of the general field of radio and electronics and of this Institute dictated the need for something of this kind. The response seems to justify the plan, three Professional Groups having been organized; and it appears that the needs of these particular highly specialized fields will be fulfilled thereby.

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The Institute, by very careful management, met its budget, as can be noted from the accompanying statement. However, the Institute's income is inadequate to provide many desired services, to publish more of the available technical papers in the Proceedings, and to increase the surplus, which is not keeping pace with Institute growth. The advent of 1949 finds the Institute management with a better understanding of its problems, and it is hoped that the new activities introduced during 1948 will render real service to the membership.

Respectfully submitted,

Haraden Pratt

Haraden Pratt, *Secretary*

March 24, 1949

Table I—Membership Distribution by Grades

| Grade | As of Dec. 31, 1948 Number | % of Total | As of Dec. 31, 1947 Number | % of Total | As of Dec. 31, 1946 Number | % of Total |
|---|---|---|---|---|---|---|
| Fellow | 259 | 1.1 | 239 | 1.1 | 218 | 1.2 |
| Senior Member | 2,192 | 9.4 | 2,068 | 9.8 | 1,763 | 9.7 |
| Member | 3,334 | 14.2 | 3,017 | 14.4 | 2,330 | 12.8 |
| Associate | 11,713*** | 50.0 | 12,079** | 57.4 | 11,591* | 63.9 |
| Student | 5,939 | 25.3 | 3,634 | 17.3 | 2,252 | 12.4 |
| Totals | 23,437 | | 21,037 | | 18,154 | |

* Includes 1,701 Voting Associates.
** Includes 1,490 Voting Associates.
*** Includes 1,260 Voting Associates.

Table II—Five-Year Analysis of U. S. and Foreign Membership

| | 1948 | 1947 | 1946 | 1945 | 1944 |
|---|---|---|---|---|---|
| Total | 23,437 | 21,037 | 18,154 | 15,779 | 13,137 |
| U.S. and Possessions | 21,048 | 18,723 | 15,898 | 14,053 | 11,596 |
| Foreign (including Canada) | 2,389 | 2,314 | 2,256 | 1,726 | 1,541 |
| Per Cent Foreign | 10.2 | 11.0 | 12.4 | 10.9 | 11.7 |

## Membership

At the end of the year 1948, the membership of the Institute, including all grades, was 23,437, an increase of 2,400, or 11 per cent over the previous year. The 2,400 member increase in 1948 compares favorably with 2,375 and 2,883, the increases for 1946 and 1947, respectively. The percentage increase was 16 per cent in 1946, and 15 per cent in 1947. The membership trend from 1912 to date is shown graphically in Fig. 2.

The distribution of members in the various grades for the years 1945, 1946, 1947, and 1948 is shown in Fig. 3. Actual figures for 1946, 1947, and 1948 are shown in Table I. Note that the loss of Associates was due largely to the loss of Voting Associates because of transfer to Member grade. Of the 10,453 nonvoting Associates, 2,680 have been in that grade for more than five years. 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 and 1948, a very satisfactory trend. Note also that the drop in percentage of higher grade members was due to the very substantial increase of Student members.

Table II shows an analysis for the past five years of the distribution of members at home and abroad. It may be noted that the foreign membership increased rapidly at the end of the war. Monetary exchange difficulties since the end of the war have contributed to the reduced increase in foreign membership during the past two years.

It is with deep regret that this office records the death of the following members of the Institute during the year 1948:

*Fellows*

Louis Cohen (F'15)
Harry Diamond (A'26–M'30–F'43)
Kenneth B. Warner (A'18–M'22–F'36)
William Wilson (M'26–F'28)

*Senior Members*

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Ralph G. McCurdy (M'42–SM'43)
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George H. Rockwood, Jr. (A'31–SM'46)
McMurdo Silver (SM'43)
Frank N. Waterman (A'12–M'15–SM'43)

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Richard R. Syrdal (M'46)

*Associates*

Robert L. Anderson (S'40–A'41)
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Ralph S. Clarke (A'21–VA'39)
R. Douglas Clerk (A'37–VA'39)
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George B. David (A'43)
Frank H. Fay (A'38–VA'39)
Edwin W. Hamlin (A'40)
Alexander M. Haubrich (A'37–VA'39)

(Continued on page 663)

# Institute News and Radio Notes

## Technical Committee Notes

The **Standards Committee** held an annual meeting on March 8, 1949. The new chairman is J. G. Brainerd of the University of Pennsylvania; A. G. Jensen of BTL and L. G. Cumming of IRE Headquarters are vice-chairmen. Standards on *Radio Aids to Navigation: Definitions of Terms, 1949;* and *Railroad and Vehicular Communications: Methods of Testing, 1949* have been approved by the Committee and will shortly be published. Mr. Jensen reported for the Definitions Co-ordinating Subcommittee, which has two purposes: to save work and time for groups writing definitions by calling attention to any overlapping in work, and to establish a more efficient transfer of information between the IRE and other organizations. It is hoped that the Subcommittee will receive definitions while they are in a tentative stage, so that they may be referred to ASA C-42 Committee for agreement prior to adoption by the IRE. The **IRE/AIEE Co-ordinating Committee on Definitions** has agreed to set up a group to work on the term "Noise." It is advisable that no other committee work on this subject until the group's task is completed. The meeting was addressed by the Institute's Standards Co-ordinator, W. R. G. Baker, who offered the IRE's thanks to A. B. Chamberlain for his fine work as Chairman of the Standards Committee, and also thanked the chairmen of all the technical committees for their accomplishments during the year. He spoke of a new method of reproducing Standards by photo-offset, which will require strict adherence to the rules governing preparation of drafts of standards. . . . A meeting of the **Antennas and Wave Guides Committee** was held on March 30 in Washington, D. C., and L. C. Van Atta was chosen chairman for the coming year. A vote of thanks was given to the retiring chairman, P. S. Carter. . . . J. G. Brainerd, retiring chairman of the **Circuits Committee**, introduced the new chairman, W. N. Tuttle, at the March 9 meeting. The Committee is continuing its work on definitions, and subcommittees will be formed for this activity. . . . L. R. Nergaard will assume the duties of chairman of the **Committee on Electron Tubes and** Solid State Devices, it was announced on March 7. The Committee plans to have a standard on definitions of terms ready for publication shortly. . . . At the meeting of the **Receivers Committee** on March 10, R. F. Shea was introduced as the new chairman. W. O. Swinyard is the retiring chairman. . . . **The Research Committee** held a meeting on March 9, during which D. E. Chambers led the Committee in a discussion of its scope. The Committee agreed to secure for publication interpretative or review papers and original papers, to promote a "Letters to the Editor" page in the PROCEEDINGS, to prepare abstracts, and to report on technical meetings. . . . The **Video Techniques Committee,** under the chairmanship of J. E. Keister, met on March 9. Three new subcommittees have been formed: Definitions, headed by R. H. Daugherty, Jr.; Methods of Measurement and Test for Video Systems and Components, headed by W. J. Poch; and Methods of Measurement and Test for Video Signal Transmission, headed by L. W. Morrison. Besides the material which these new subcommittees are preparing for standardization, the Subcommittee on Methods of Measurement and Test for Utilization, Including Video Recording, under the chairmanship of Vernon J. Duke, is preparing six tutorial reports for publication. . . . At the meeting of the **Wave Propagation Committee** on March 9, it was announced that C. R. Burrows would head the Committee for the coming term. The Committee reviewed the *Glossary of Terms Used in Radio Propagation*, which had been circulated for comment by the British Standards Institute. It also discussed its activities in connection with the JTAC. . . . **The Committee on Professional Groups** met on March 7. The chairman W. L. Everitt, announced that three Groups have been formally organized and approved: Audio, Broadcast Engineers, and Antennas and Wave Propagation. Six others are under consideration: Circuit Theory, Nuclear Science, Vehicular and Railroad Radio Communications, Receivers, Transmitters, and Electron Tubes. The Committee shall have primary jurisdiction on policy, which will include the actual approval of Groups and their objectives. Operation of groups will be under the control of the Executive Committee through the Standards Co-ordinator. Minor revisions of the Professional Groups Manual were authorized, and a plan was adopted for stimulating action in the Sections for forming Section Groups, the Committee acting as a clearing house for the Sections in supplying information, speakers, papers, etc. It was decided that all members of Professional Groups, national or sectional, must be members of the Institute. All Groups shall hold one technical conference of national scope each year.

### Calendar of COMING EVENTS

Central Section Regional Meeting, Society of Motion Picture Engineers, Toledo, Ohio, June 10

**IRE Conference on Electronic Devices, Princeton, N. J., June 20 and 21**

AIEE Summer General Meeting, Swampscott, Mass., June 20–24

NBS Symposium on Construction and Applications of Conformal Maps, Los Angeles, Calif., June 24–25.

NBS Symposium on Probability Methods in Numerical Analysis. Los Angeles, Calif., June 27–29.

AIEE Pacific General Meeting, San Francisco, Calif., August 23–26

**1949 IRE West Coast Convention, San Francisco, Calif., August 30–September 2**

**1949 National Electronics Conference, Chicago, Ill., September 26–28**

National Radio Exhibition, Olympia, London, England, September 28 to October 28

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21

**1950 IRE National Convention, New York, N.Y., March 6–9**

## Military Standards Agency Reorganized

Under the terms of a charter signed by the assistant secretaries of the three military services, the government body organized in December, 1943, as the Army-Navy Electronic Standards Agency has been reconstituted as the Armed Services Electro Standards Agency. The change provides for official participation by the Air Force, which became a separate component of the National Military Establishment under the provisions of the National Security Act of 1947.

The purpose of the Standards Agency is to reduce the number of styles and types of electronic components used in the manufacture of military equipment of all kinds, to insure their quality and dependability, and to designate approved sources of supply.

## British "Radiolympia" Exhibit to be Held in Fall

Television, as perfect technically as it can be demonstrated in all stages from studio to receiver, will be a feature of the sixteenth National Radio Exhibition ("Radiolympia") to be held at Olympia, London, England, from September 28 to October 8.

Other displays will include an enlarged section for communications equipment, navigational aids, and electronics in industry, and exhibits for the first time since the war by the Royal Navy, the Army, and the Royal Air Force. The Ministry of Supply, the Department of Scientific and Industrial Research, and the GPO will also exhibit.

## Constitutional Amendment Ballot

In accordance with Article IX, Section 2 of the Institute Constitution, the ballots on amending Article II, Section 1(c) and Article II, Section 2(c) of the IRE Constitution were counted on May 3, 1949, under the supervision of the Tellers Committee, J. L. Callahan, Acting Chairman. The final result of the ballot count is as follows:

| | *Yes* | *No* | *Total* |
|---|---|---|---|
| Total | 812 | 2334 | 3146 |
| Per cent | 25.8 | 74.2 | 100 |

Total ballots mailed . . . . . . . . . . . . . 7,015
Total ballots received . . . . . . . . . . . . 3,189
—45.4 per cent of ballots mailed
Total ballots void . . . . . . . . . . . . . . 43
— 0.6 per cent of ballots mailed
Total ballots counted . . . . . . . . . . . . 3,146
—44.8 per cent of ballots mailed

# Report of the Secretary—1948

To the Board of Directors,
The Institute of Radio Engineers

Gentlemen:

The Secretary again submits a report, required annually as stipulated in the By-laws, of the status of Institute affairs resulting from activities during 1948.

It is especially gratifying to note that Institute growth continues unabated both in membership and activities. Membership is spread over 63 countries. Student Branches have increased over 100 per cent from 27 to 59. Sections increased from 43 to 48 or about 11 per cent. Additional data and the usual charts are included herewith. (Figs. 1–4).

The past year has seen the inauguration of the Regional Director plan, with a reduced number of Board of Directors meetings. The attendance at these was excellent, and not only were regional problems more fully dealt with, but members of the various Sections enjoyed a closer relationship with the Institute through their ability to express views through the Regional Directors. These Directors carry to their Regions authentic information resulting from their participation in the operations of the Board. A plan was adopted to render limited financial assistance for the handling of approved Regional meetings, which should go far in promoting activities in these several far-flung areas that have area interests.

Further encouragement toward better Section activities was provided through the increase in Section membership rebates, which took effect at the first of the year.

The year 1948 saw the commencement of a new conception created by the Board, the Professional Group system. This plan was evolved in an effort to develop a method whereby members having common interests in a given specialized field can give expression to those special interests and develop related activities. The rapid growth of the general field of radio and electronics and of this Institute dictated the need for something of this kind. The response seems to justify the plan, three Professional Groups having been organized; and it appears that the needs of these particular highly specialized fields will be fulfilled thereby.

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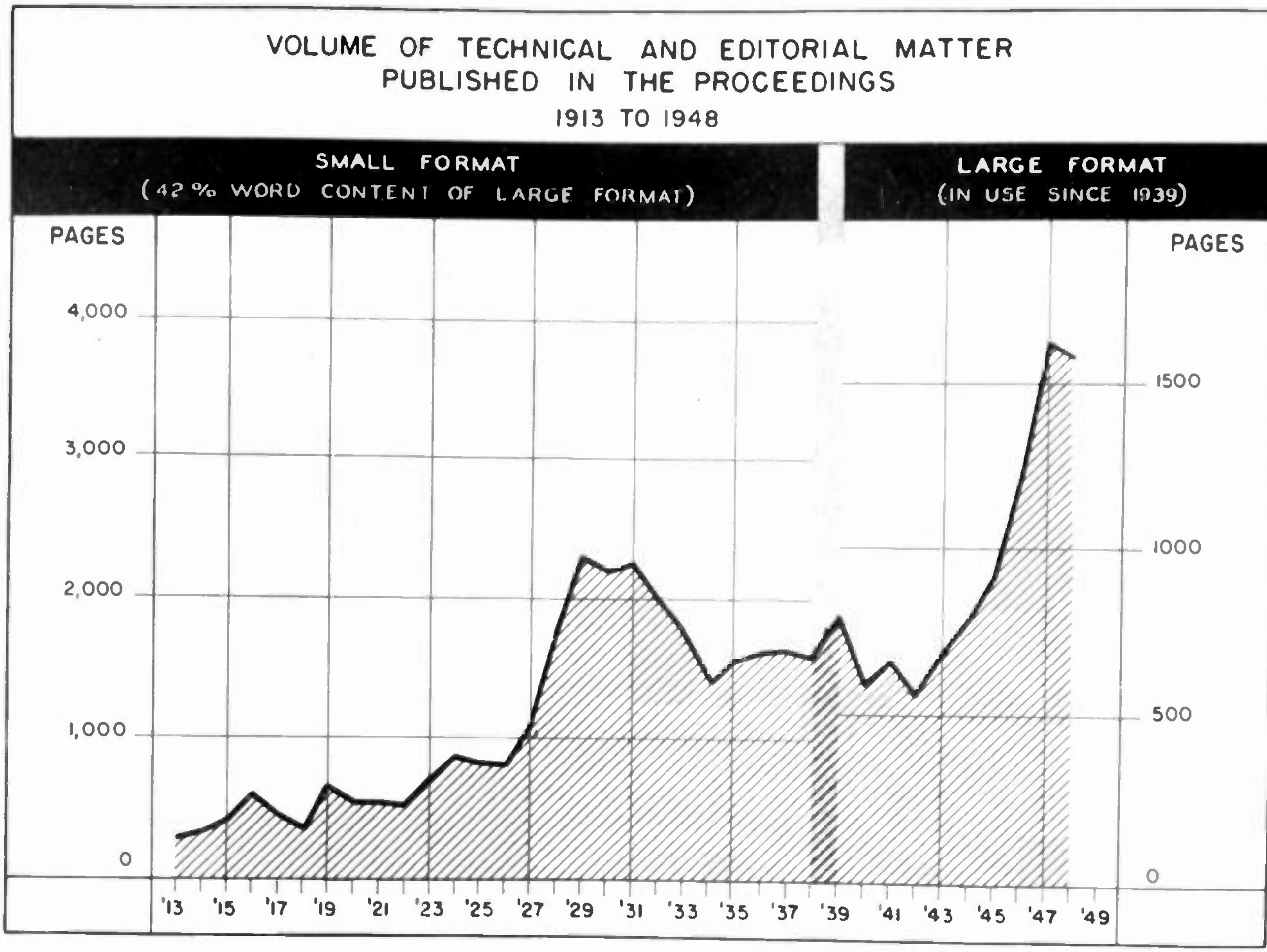
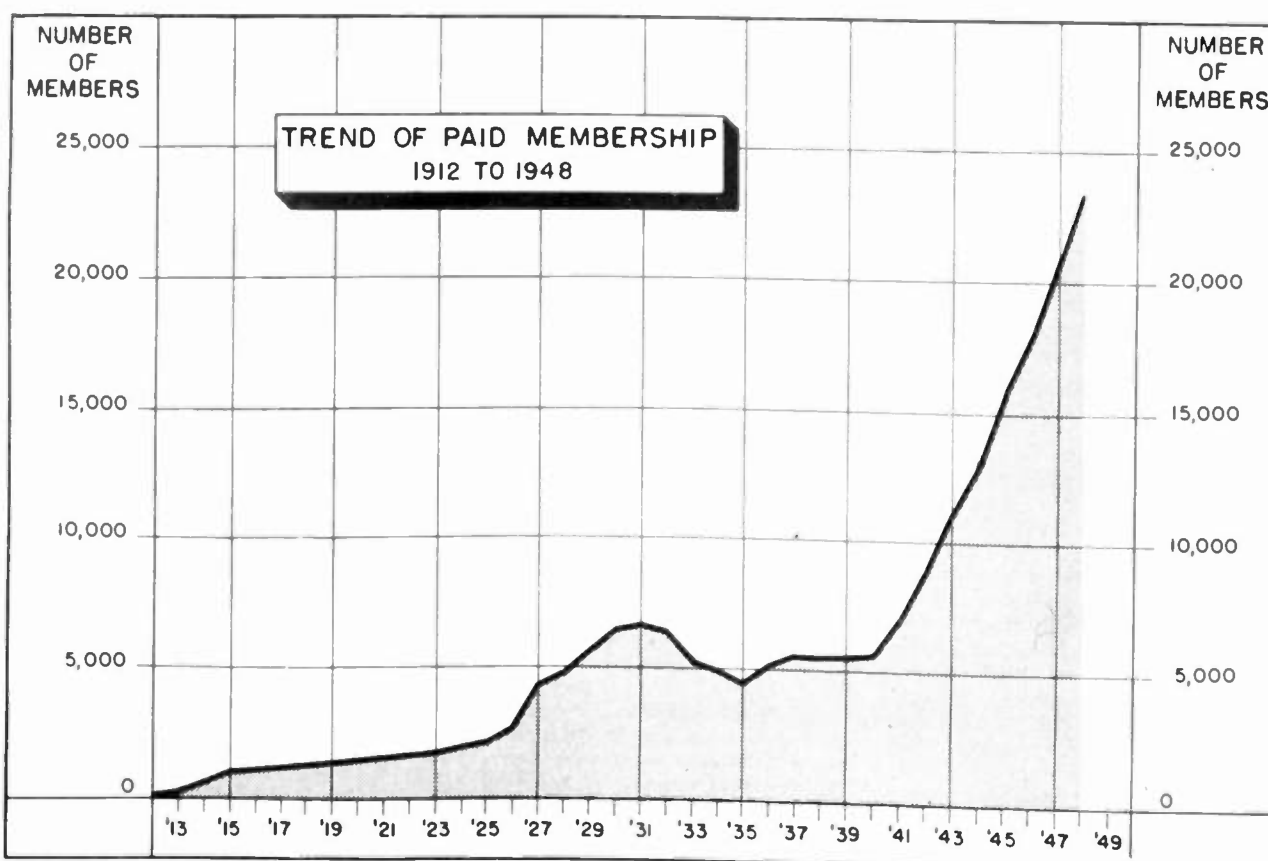
(Continued on page 663)



VOLUME OF TECHNICAL AND EDITORIAL MATTER
PUBLISHED IN THE PROCEEDINGS
1913 TO 1948
SMALL FORMAT
(42% WORD CONTENT OF LARGE FORMAT)
LARGE FORMAT
(IN USE SINCE 1939)
PAGES
PAGES
4,000
3,000
2,000
1,000
0
1500
1000
500
0
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49


Fig. 1



NUMBER OF MEMBERS
NUMBER OF MEMBERS
TREND OF PAID MEMBERSHIP
1912 TO 1948
25,000
20,000
15,000
10,000
5,000
0
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49
CHART-FACTS, N Y C


Fig. 2

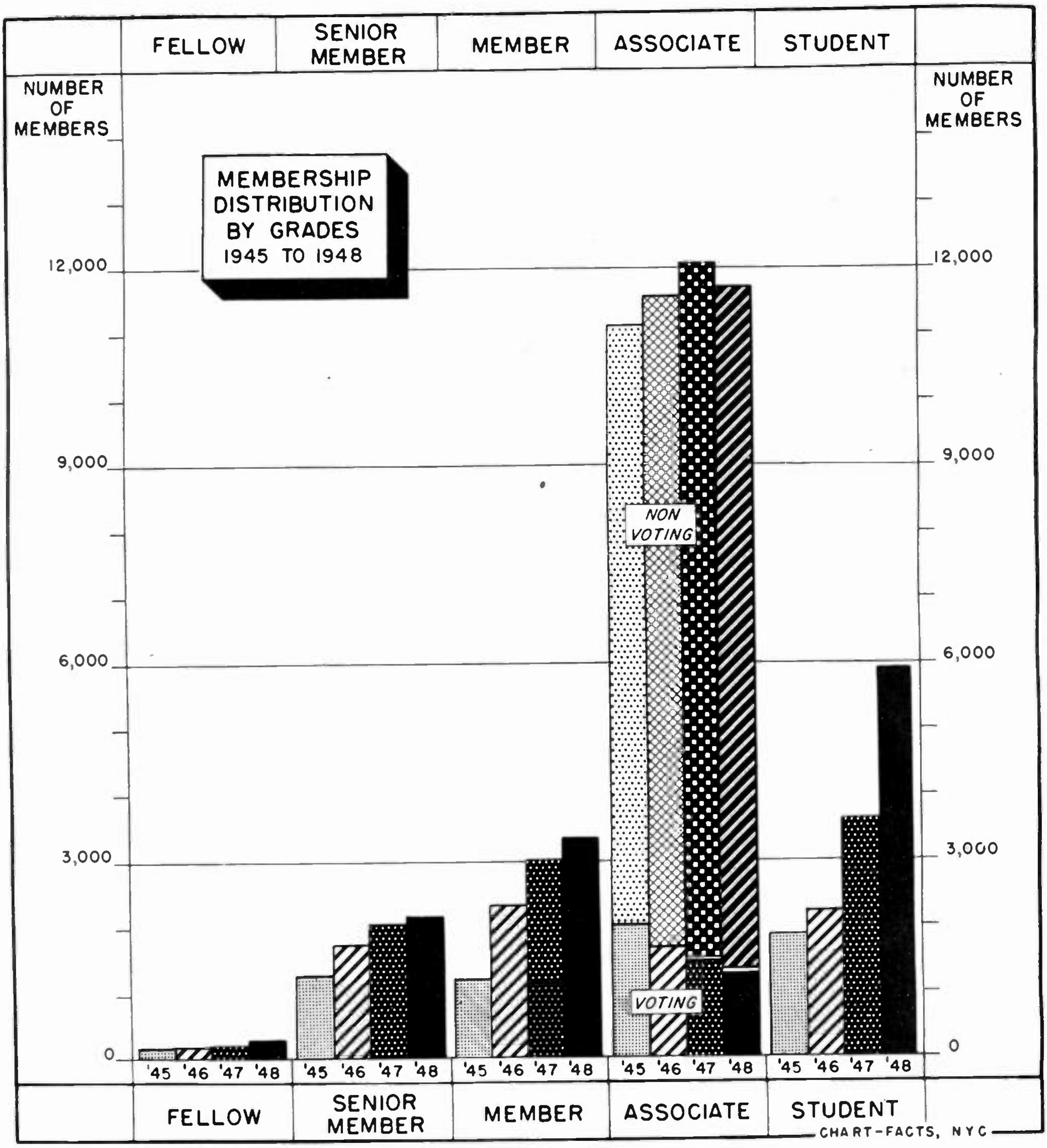



Fig. 3

(continued from page 661)

Frank R. Hillyer (A'45)
Albert E. Johnson (A'46)
Charles J. Lemieux (A'38–VA'39)
Virgil L. Miller (A'48)
Alfred J. Mills (A'44)
Howard R. Molton (A'45)
Irvin Nevins (A'45)
Willis B. Plummer (S'36–A'40)
George E. Rueppel (A'43)
Jack M. Sadowsky (A'37–VA'39)
Kenneth L. Scott (A'21–VA'39)
William J. Sharpe (A'44)
Monroe H. Sheppard (S'43–A'45)
Allen C. Stecher (A'44)
Llewelyn L. B. Summers (A'40)
Charles W. Taussig (A'22–VA'39)
Mervin J. Ullman (A'46)
William J. Yahnke (A'46)

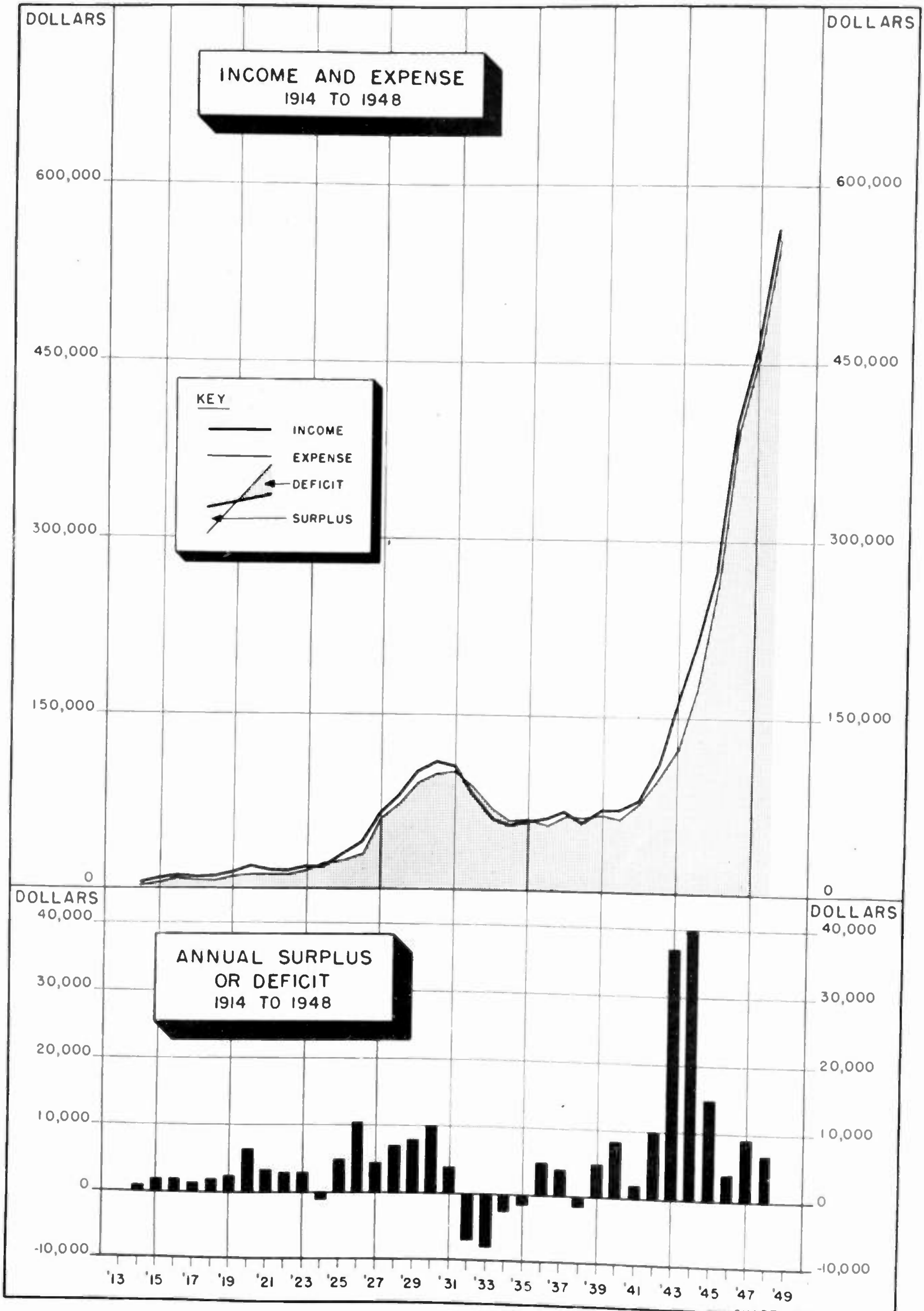
## Editorial Department

The year 1948 again found the Editorial Department faced, on the one hand, with an ever-increasing inflow of submitted papers, and on the other with rigid space limitations imposed by stern budgetary realities. The problem, while it could not be said to have been solved as of the end of the year, was being brought under partial control by drastic editorial selection and compression.

In 1948 there were published in the PROCEEDINGS a total of 2,452 pages, including covers. Of these, 1,592 were editorial pages and 860 advertising pages. Of the editorial pages, 1,151½ were devoted to technical papers, (including discussions and correspondence), 166½ to Abstracts and References, and 274 to nontechnical material (including covers, contents pages, and annual index). Of the advertising pages, 757⅓ represented paid advertising, while the remaining contained useful editorial filler in the form of Institute Membership and Section-meeting lists, and news of new products and the industry in general.

The total of 2,452 pages published during the year compares with 2,576 in 1947 and 2,240 in 1946. The number of editorial pages (1,592) compares with 1,636 in 1947, and 1,256 in 1946. Advertising pages numbered 940 in 1947 and 984 in 1946. The number of editorial pages published each year since 1913 is shown in Fig. 1.

Technical papers totaling 173 were published in 1948, as against 172 in 1947. Authorship of these papers was by 241 individuals, of whom 176, or 73 per cent, were



DOLLARS
INCOME AND EXPENSE
1914 TO 1948
600,000
450,000
300,000
150,000
0
KEY
INCOME
EXPENSE
DEFICIT
SURPLUS
DOLLARS
40,000
30,000
20,000
10,000
0
-10,000
ANNUAL SURPLUS
OR DEFICIT
1914 TO 1948
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49
CHART-FACTS, N Y C


Fig. 4

members of the Institute. In 1947, 205 of the 266 authors, or 77 per cent, were members.

The backlog of papers on hand in the Editorial Department at the end of 1948 included 141 papers, totaling an estimated 981 PROCEEDINGS pages, of which 69 papers representing 442 pages had been accepted, the remainder being under review. At the start of the year the backlog was only 95 papers with 651 pages, having been reduced from the postwar peak of 171 papers totaling 1,225 pages by the "expanded publication program" executed during the last part of 1947. Under this program the size of issues was increased from an average for the first eight issues of 1947 of 109 pages of editorial content to 144 pages in September, 1947, 176 pages in October, 224 pages in November and December, respectively, and concluding with 176 pages in the January, 1948 issue. The average number of pages for the remaining issues of 1948 was 128 pages, or 19 more than the average for the "regular" issues of 1947.

Despite this increased regular allotment, the volume of material received was so great as to make it impossible to hold entirely the gains on the backlog which had been achieved by the expanded publication program.

The volume of papers submitted for publication continues to increase markedly. During 1947, 201 papers totaling an estimated 1,298 PROCEEDINGS pages were submitted, or an average of 16.7 papers and 108.2 pages per month. During 1948 these receipts increased to 289 papers of 2,043 pages, or 24 papers and 170 pages per month.

To aid in keeping the backlog within manageable dimensions and to enable more effective employment of the editorial budget allocated by the Executive Committee, new procedures were set up within the Editorial Department and in the Editorial Administrative Committee. In consequence, even more stringent standards have been established for the acceptance of papers, and maximum compression feasible is required in all accepted papers.

In connection with the related problems of printing costs and lack of space, studies were made of possible methods of achieving printing production economies. A particularly detailed study was made of the possibilities of space and cost saving by the use of smaller typeface sizes. The study revealed that no significant economies could be realized by such a course without using measures which would result in degradation of legibility and appearance.

With the approval of the Executive Committee, modifications in format and treatment of nontechnical material were made to conserve space for technical papers. This has been accomplished with slight impairment or reduction of content, and even with some improvement in readability and discernability. Publication of the Abstracts and References Section of the PROCEEDINGS continued during the year. The question of its merit relative to the great need for additional space for technical papers was given careful consideration, and the membership at large was asked to express opinions. More than 100 responses were received, the opinions being overwhelmingly favorable for its continuance. The Editorial Department was finally instructed by the Executive Committee to continue publication of the section, at least throughout 1949.

An innovation in the printing production procedures of the Institute was the decision to reproduce the editorial section of the 1949 Yearbook by the photo-offset method of reproduction, the membership lists being composed by the varitype method. It is expected that appreciable economies will be effected by the use of these methods, and that production will be considerably expedited, with tolerable impairment of appearance and legibility, which is regarded as of negligible importance because of the nonpermanent, reference nature of the book.

Several Standards reports were produced or were in production during the year, which are listed under the subject of Technical Activities in this report.

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 administration, both functioning through Technical Editor Clinton B. DeSoto. An effective staff assists these officers. It has been greatly aided by the helpful advice and assistance unstintingly given by the 78 members of the Board of Editors, the 92 members of the Papers Review Committee, the Papers Procurement Committee, and the Editorial Administrative Committee.

## Section Activities

We were glad to welcome five new Sections into the Institute during the past year. They are as follows:

| | | |
|---|---|---|
| San Antonio | (March) | 1948 |
| New Mexico | (April) | 1948 |
| Toledo | (April) | 1948 |
| Omaha-Lincoln | (Sept.) | 1948 |
| Salt Lake | (Oct.) | 1948 |

The total number of Sections is now 48. 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. In addition, during the year there has been formed the following group, unofficially designated as a Sub-section:

Center County (Emporium)

## Student Branches

The Institute's program with respect to Student Branches, reactivated in 1947, continued to flourish in 1948. The number of Student Branches formed during 1948 was 32, 14 of which operate as joint IRE-AIEE Branches. The total number of Student Branches is now 59, 26 of which operate as Joint IRE-AIEE Branches. This unprecedented increase of student interest was accompanied by a large increase in student members. Additional applications for the establishment of Branches are now in process. This work, under the direction of E. K. Gannett, Assistant Secretary, is very productive in building the foundations upon which future Institute progress will depend.

Following is a list of the Student Branches formed in 1948:

Alabama Polytechnic Institute; *University of Arizona; Polytechnic Institute of Brooklyn; California State Polytechnic College; *Columbia University; University of Dayton; University of Detroit; *University of Florida; Illinois Institute of Technology; *Iowa State College; John Carroll University; Lehigh University; Manhattan College; *University of Maryland; *Massachusetts Institute of Technology; *Michigan State College; *Missouri School of Mines and Metallurgy; Newark College of Engineering; University of New Mexico; *University of North Dakota; *University of Notre Dame; Oregon State College; *Princeton University; *University of South Carolina; South Dakota School of Mines & Technology; St. Louis University; *Agricultural & Mechanical College of Texas; *University of Toledo; Utah State Agricultural College; University of Virginia; University of Wisconsin; and University of Wyoming.

## Technical Activities

*Technical Committees.* During 1948, 20 technical committees, with their subcommittees and task groups, held a total of 112 meetings at Institute Headquarters, 21 meetings at the Hotel Commodore during the National Convention, and 25 meetings in other localities, or a total of 158 meetings, a substantial increase over 1947.

The following standards prepared by the IRE technical committees were published during the year:

- Standards on Abbreviations, Graphical Symbols, Letter Symbols, and Mathematical Signs, 1948
- Standards on Antennas: Methods of Testing, 1948
- Standards on Antennas, Modulation Systems, and Transmitters: Definitions of Terms, 1948
- Standards on Television: Definitions of Terms, 1948
- Standards on Television: Methods of Testing Television Receivers, 1948

In addition, publication of *Standards on Radio Receivers: Methods of Testing Frequency-Modulation Broadcast Receivers*, 1947, was completed in January, 1948.

In addition to the standards published, approval of the Board of Directors was secured for *Standards on Receivers: Methods of Testing Amplitude-Modulation Broadcast Receivers.* Approval was also secured for a large portion of the revision to the standards on Definitions of Terms relating to Electron Tubes and to the standards on Methods of Testing Electron Tubes. These standards will be available during 1949. The Standards Committee prepared Proposed Standard Frequency-Band Designations for publication in the PROCEEDINGS. It will be considered for standardization in 1949. The Committee on Electroacoustics, in co-operation with ASA, prepared a glossary of acoustical terms which has been published and distributed as a preliminary ASA standard. It will be adopted as a final standard by IRE and ASA during 1949.

The Technical Committee on Annual Review completed a survey, "Radio Progress During 1948." Thirty-six pages in the March, 1949, issue of PROCEEDINGS are devoted to publication of this material.

* Joint IRE-AIEE Student Branches.

The Nuclear Studies Committee procured a series of papers in its field for publication in the PROCEEDINGS. Four papers have appeared in 1948 issues, and one paper per month is available and scheduled for publication in 1949.

A Definitions Co-ordinating Subcommittee of the Standards Committee was established during the year. Its function has been to correlate definitions of all terms under consideration within the technical committee structure of the IRE and, in addition, to take cognizance of the definitions of outside agencies to eliminate duplication of man hours and conflict. Work was started at Headquarters toward the compilation of a master index of all terms defined by IRE and other organizations having definitions in the electronics field. This index will be made continually available to all IRE committees, and, upon request, to other professional societies.

Two major symposia were sponsored by technical committees during 1948; the Electron Tube Conference, and the Conference on Electronic Instrumentation in Nucleonics and Medicine. The Electron Tube Conference, sponsored by the Committee on Electron Tubes, was held on June 28 and 29 at Cornell University, and was attended by over 200 scientists. The Nucleonics Conference, sponsored by the Nuclear Studies Committee in co-operation with AIEE, was held in New York, N. Y., on November 29 and 30, and December 1. It was attended by 550 scientists. Both conferences are considered to have been outstanding contributions to electronic engineers.

*Professional Group System.* Two groups, namely the "Audio Group," established in June, and the "Broadcast Engineers Group," established in July, were becoming active by the end of 1948. The formation of Professional Groups in "Nuclear Science" and "Antennas and Wave Propagation" was under way. In addition, interest was developing for the formation of Professional Groups in the fields of Circuit Theory, Railroad and Vehicular Communications, Radio Receivers, Radio Transmitters, Telemetering, and Quality Control.

*Joint Technical Advisory Committee.* The first task undertaken by the JTAC resulted from a request by the FCC to study the problem of uhf television development and to provide information for the FCC hearing scheduled for September 20. The JTAC provided this information in its Proceedings, Volume 1, entitled "Utilization of Ultra-High Frequencies for Television," Docket 8976, September 20, 1948. This report was followed by a second volume, "Allocation Standards for vhf Television and FM Broadcasting," Docket 9175, December, 1948.

*General Joint Technical Activities.* During 1948 there was an appreciable increase in the co-operative technical activities of the Institute with other professional societies and federal agencies. Sixty-five committee meetings were held jointly with other organizations during the year, in addition to the 158 meetings of technical committees previously mentioned. Work on this total of 223 committee meetings was accomplished by 580 committee members, of whom approximately 180 were engaged in the joint committee activities. These meetings represent co-ordination of technical activities with the American Standards Association, American Institute of Electrical Engineers, Joint Technical Advisory Committee, Radio Technical Committee on Aeronautics, National Research Council, Army-Navy Electrical and Electronics Standards Agency, and its successor, Armed Services Electro Standards Agency, International Scientific Radio Union, Research and Development Board of the National Military Establishment, Atomic Energy Commission, Civil Aeronautics Administration, and International Civil Aviation Organization.

Technical activities were carried on under the guidance of the Standards Coordinator W. R. G. Baker, Technical Secretary L. G. Cumming, and a competent staff.

## Fiscal

A condensed summary of income and expenses for 1948 is shown in Table III, and a balance sheet for 1948 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, 1948

| | | |
|---|---|---|
| *Income* | | |
| Advertising* | $157,501.64 | |
| Member Dues and Conventions | 339,177.92 | |
| Subscriptions | 33,679.55 | |
| Sales Items, Binders, Emblems, Standards, etc. | 18,229.77 | |
| Investments Income | 10,337.60 | |
| Miscellaneous Income | 3,296.27 | |
| Total Income | | $ 562,222.75 |
| *Expense* | | |
| PROCEEDINGS Editorial Pages | $142,372.35 | |
| Advertising Pages | 81,979.41 | |
| Yearbook | 31,423.27 | |
| Section Rebates | 21,228.76 | |
| Sales Items | 20,991.33 | |
| Miscellaneous Printing | 4,807.77 | |
| General Operations | 157,139.73 | |
| Convention Cost | 89,802.44 | |
| Total Expense | | $ 549,745.06 |
| Surplus | | 12,477.69 |
| Reserve for Depreciation | | $ 5,880.59 |
| Net Surplus | | $ 6,597.10 |

TABLE IV—BALANCE SHEET—DECEMBER 31, 1948

| | | |
|---|---|---|
| *Assets* | | |
| Cash and Accounts Receivable | $211,827.25 | |
| Inventory | 12,061.91 | |
| Total Current Assets | | $ 223,889.16 |
| Investments at Cost | 361,023.50 | |
| Building and Land at cost | 356,687.30 | |
| Furniture and Fixtures at cost | 67,394.92 | |
| Other Assets | 64,661.38 | |
| Total | | 849,767.10 |
| Total Assets | | $1,073,656.26 |
| *Liabilities and Surplus* | | |
| Accounts Payable | 7,371.38 | |
| Federal Taxes on emblems, etc. | 1,979.78 | |
| Total Current Liabilities | | $ 9,351.16 |
| *Deferred Liabilities | | 261,635.13 |
| Total Liabilities | | $ 270,986.29 |
| Reserve for Depreciation | | 16,665.72 |
| Surplus—Donated | 596,468.83 | |
| Surplus—Earned | 189,535.42 | |
| Total Surplus | | 786,004.25 |
| Total Liabilities and Surplus | | $1,073,656.26 |

* 1949 items, PROCEEDINGS for members and subscribers, Advertising, and Convention service.

# Industrial Engineering Notes[1]

### NBS Develops Absolute Values

Recent experimental work at the National Bureau of Standards has resulted in developments of importance to the design of cathode-ray tubes and many other electronic devices.

The absolute value of the magnetic moment of the proton has been determined with unprecedented accuracy, and the value of the basic constant $e/m$—electric charge to mass ratio of the electron—with even greater accuracy.

A complete report on the development was published in the May issue of *The Technical News Bulletin*, monthly publication of the NBS, copies of which may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

### Lack Succeeds Sarnoff as AFCA President

RMA Director Frederick R. Lack was elected to succeed David Sarnoff as president of the Armed Forces Communications Association, which met in Washington on March 28 and 29. Close co-operation and co-ordination between government and industry in the communications operating and manufacturing fields were emphasized by speakers on both sides during the third annual meeting of the association.

### New Export Schedule Published By OIT

A new comprehensive export schedule (No. 27), containing all export regulations in effect on March 31, 1949, was published by the Office of International Trade, U. S. Department of Commerce.

The new edition supersedes No. 26, published last fall, and covers all changes in regulations previously described in Current Export Bulletins 485 to 517, inclusive. Both the new Comprehensive Export Schedule and all Current Export Bulletins are now published in loose-leaf form to provide in a single volume all new regulations, as well as changes in older ones.

### WAA Completes Disposal of Electronic Surplus

The War Assets Administration has completed its disposal of surplus radio and electronic equipment, except for the processing of documents and the completion of final auditing reports of agents and distributors.

Out of a total surplus electronic-radio acquisition of $816,307,000, at original cost, the WAA disposed outright of electronic equipment and parts valued at $584,310,000. Approximately $63,672,000 was received in cash. About $300,000,000 worth of radio and electronic equipment was distributed among state, city, and county governments and educational institutions without charge.

[1] The data on which these NOTES are based were selected, by permission, from "Industry Reports," issues of March 18 and 25, and April 1, 8, and 15, published by the Radio Manufacturers' Association, whose helpful attitude in this matter is hereby acknowledged.

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# Books

### Elements of Electromagnetic Waves, by Lawrence A. Ware

Published (1949) by the Pitman Publishing Corp., 2 W. 45 St., New York, N. Y., 200 pages+3-page index+x pages. 69 figures. $6\times 9\frac{1}{4}$. $3.50.

The purpose and level of this book is stated by the author as follows:

"This text has been prepared to meet the need for an elementary introduction to the basic ideas of electromagnetic theory. It is intended as first-course material for either juniors or seniors, and the only prerequisites are calculus and the fundamentals of alternating current theory. . . . It is assumed that the reader is familiar with the circuited law of magnetism. . . . A knowledge of Faraday's emf law in line-integral form is also assumed."

It is quite revealing to read this book and learn how many inconsistencies and errors in physics, mathematics, logic, and concepts can be made in a little volume of 200 pages. It is also amusing to observe that the author somehow managed to arrive at correct answers most of the time, in spite of the numerous mistakes in many of the derivations. The following examples represent a small fraction of the errors presented in this book.

First of all, the author does not distinguish among electrostatic, magnetostatic, quasistationary, and electromagnetic problems. For example, on page 41, after stating

$$\oint E\cdot ds = E = -\frac{d\phi}{dt},$$

the author goes on to say that "The analysis of electric and magnetic fields are better carried out in terms of the gradients of the two scalar potentials

$$H = -\nabla \text{mmf} \quad \text{and} \quad E = \nabla\phi,"$$

on the same page. Still on the same page, one finds equation 3.5:

$$B = \mu H = \frac{\phi}{A},$$

a new form of vector analysis where a scalar quantity can be divided by a vector quantity.

As an illustration of mathematical laxity, one finds on page 65:

$$\int_S \nabla \times H \cdot da = \int_S J \cdot da.$$

The two integrals are taken over the same area so that, necessarily, the integrands must be equal, as follows:

$$\nabla \times H = J.$$

The result happens to be correct, provided the $J$ is interpreted correctly.

"In preliminary form," the author states, "the text presented here has been successfully used through two semesters." The assumption might be, consequently, that the footnotes in the book are the results of questions raised by the students. In one footnote (page 125) the author manages to prove that "the real part of $\epsilon^{-\gamma z}\epsilon^{j\omega t}$ is the same as $\epsilon^{-\gamma z}\cos\omega t$" where $\gamma$ is a complex quantity! The reviewer is reminded of his school days, when he managed in a quiz to prove a theorem in a dubious way, and was proud of his accomplishment. The only difference in the above case is that the theorem is incorrect.

The publishers are to be commended for the excellent appearance of the book, and the fine line drawings.

L. J. Chu
Massachusetts Institute of Technology
Cambridge, Mass.

### Photofact Television Course, edited by B. V. K. French

Published (1949) by the Howard W. Sams Co., Indianapolis 7, Ind. 200 pages+5-page index+6-page glossary+5-page bibliography. 250 figures. $8\frac{1}{2}\times 11$. Bound in boards. $3.00.

This book, based on a series of lectures by Albert C. W. Saunders, describes television principles, operation, and practice for those already familiar with radio techniques. Using the cathode-ray tube as a point of departure, the book is divided into three sections, covering beam formation and control, beam deflection systems, and beam modulation and synchronization. Under those headings, cathode-ray tube construction, camera tubes, voltage supplies, sawtooth generators, sync circuits, control functions, receiving antenna circuits, rf input tuning systems, if systems, video amplification, and many other aspects of the field are analyzed, with diagrams.

### Tables of Bessel Functions of Fractional Order, Volume II

Published (1949) by the Columbia University Press, Morningside Heights, New York 27, N. Y. 365 pages+xviii pages. $8\times 10\frac{1}{2}$. $10.00.

This volume, compiled by the Computation Laboratory of the National Applied Mathematics Laboratories, National Bureau of Standards, is devoted to the tabulation of $I_\nu(x)$ for $\pm\nu = \frac{1}{4}, \frac{1}{3}, \frac{2}{3}, \frac{3}{4}$, and is a sequel to the volume containing $J_\nu(x)$ for the same orders. The functional values in both volumes are given either to ten decimal places or to ten significant figures. Tables of the Everett interpolation coefficients, a list of constants, and a table for facilitating interpolation with respect to $\nu$ are also included.

# The Scintillation Counter*

J. W. COLTMAN†

*Summary*—**The scintillation counter is a relatively new instrument for the detection and measurement of nuclear and atomic radiations. Light produced in a fluorescent material by the incident particle or quantum is picked up by a photomultiplier tube. The resultant photoelectric emission is amplified in the photomultiplier by a series of secondary-emission stages.**

**This paper is a review of the present state of knowledge regarding the characteristics of the counter and the details of its operation.**

THE FLASHES of light released when high-energy alpha particles strike a fluorescent screen constituted one of the earliest means of detecting nuclear particles. With such means, several of the classic experiments in radioactivity were performed. As other more convenient and sensitive detectors became available, this early method fell into complete disuse. Recently, however, it has become possible to replace the eye of the observer with an extremely sensitive and rapidly responding device, the multiplier phototube. The resulting detector is one which responds to the individual particles or quanta of all high-energy ionizing radiations, and which thus has a field of application similar to that of the Geiger counter.[1] For many purposes, the higher counting rate of the scintillation detector, its tremendous range of sensitivity, and its high intrinsic efficiency make it superior to the Geiger counter.

An important advantage is that the fluorescent material is a highly absorbing solid, rather than a gas. The phosphor is in an exposed position external to the vacuum envelope of the phototube, and requires no difficult thin-window technique with easily absorbed radiations like alpha rays, beta rays, and soft X rays. Even for very penetrating radiations, the position of the detecting layer is well defined. The new detector is rugged and stable, shows no evidence of temperamental qualities, and appears to have an indefinitely long life.

The detector consists of a photomultiplier tube, such as the RCA 931A,[2] and a fluorescent material mounted external to the envelope. Some system for collecting as much of the light as possible is desirable, and in the arrangement of Coltman and Marshall[3] this takes the form of a hemispherical mirror. Fig. 1 shows the phototube with the fluorescent screen attached, and Fig. 2 the completed detector, with the light-gathering mirror in place. Fig. 3 shows diagrammatically the construction and operation of the detector. The radiation—for example, an alpha particle—enters the window and causes a scintillation within the fluorescent screen. The rays of light from this flash are collected by the light-gathering mirror and focused upon the sensitive area of the photocathode of the multiplier tube. Since the incident quanta or particles of all the radiations for which the detector is used have large energies in comparison with photons of light, and since the energy efficiency of the fluorescent screen in converting these radiations into light is high, each quantum of absorbed radiation results in a very large number of photons of light. Despite light losses in the screen, incomplete collection of light by the mirror, and a yield of only about one photoelectron per ten photons of light at the photoelectric surface, an appreciable number of photoelectrons are ejected from the photocathode for each quantum of detected radiation, thus forming a signal pulse of photoelectrons. This pulse is then amplified by the familiar secondary-emission system of the multiplier tube. Each photoelectron is swept to the first dynode by a potential difference of about a hundred volts, and ejects four or five secondary electrons. These, in turn, are swept to the second dynode and similarly multiplied by secondary-emission amplification. After nine such stages, an avalanche of a million electrons, more or less, appears at the output of the multiplier tube as a result of each initial photoelectron. The output signal pulses are sufficient to operate directly the amplifier of an oscilloscope, though it is usually preferable to introduce one or two intermediate stages. Fig. 4 shows the detector in a compact unit, including power supply and auxiliary stages, ready for operation of an oscilloscope, speaker, or discriminator-scaling-counting system.

The excellent properties of the scintillation detector are due in large part to the unique character of the amplification afforded by the multiplier tube, by which the signal is lifted to a high output level before being released to external circuits. Not only is the response time of the multiplier amplifier extremely short, but the amplification is almost totally free of drift, stray pickup, and resistor noise. One may think of the intermediate step in the new detector, the step by which the original radiation is trans-

* Decimal classification: 621.375.2×535.3. Original manuscript received by the Institute, January 14, 1949.

† Westinghouse Research Laboratories, East Pittsburgh, Pa.

[1] Fitz-Hugh Marshall and J. W. Coltman, "The photomultiplier radiation detector," *Phys. Rev.*, vol. 72, p. 528; September, 1947.

[2] R. B. Janes and A. M. Glover, "Recent developments in photo tubes," *RCA Rev.*, vol. 6, pp. 43–54; July, 1941.

[3] J. W. Coltman and Fitz-Hugh Marshall, "Photomultiplier radiation detector," *Nucleonics*, vol. 1, pp. 58–64; May, 1947.

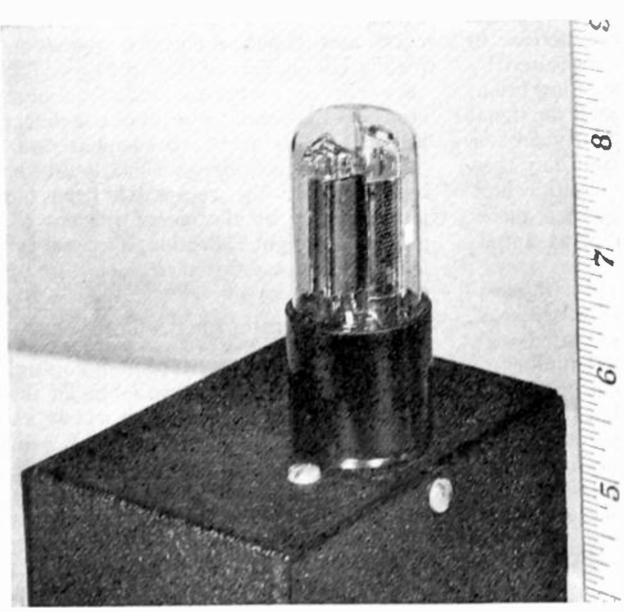

Fig. 1—A photomultiplier scintillation counter without mirror.

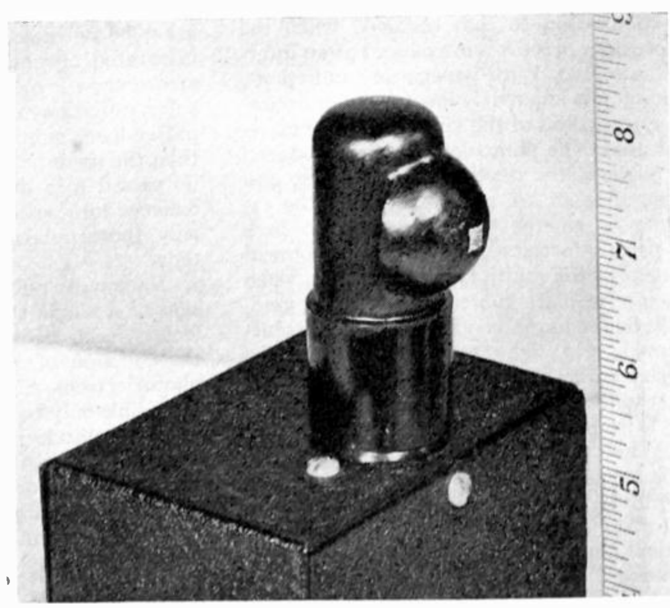

Fig. 2—Complete counter with mirror, window, and light-proof coating.

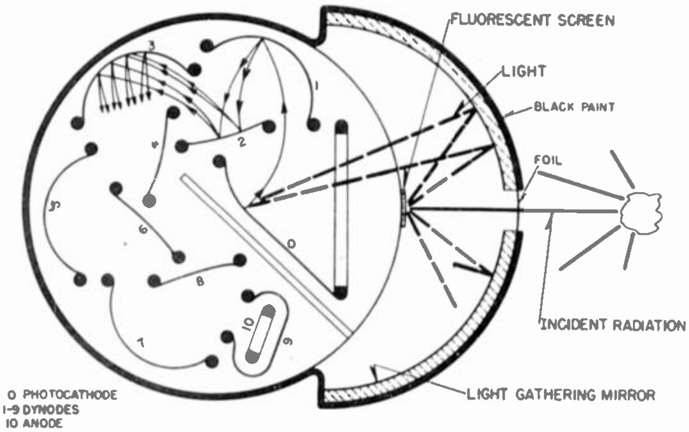


Fig. 3—Schematic diagram of operation of the scintillation counter.

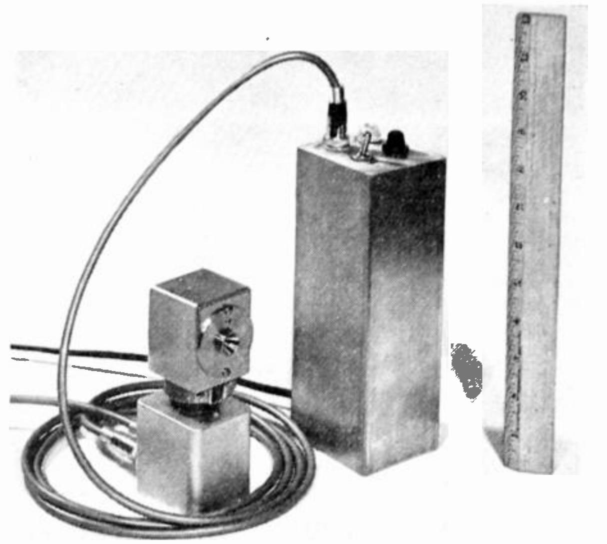

Fig. 4—A scintillation counter with built-in preamplifier and power supply.

formed to light before it is used for release of electrical charge, as being a means by which the multiplier type of amplification can be introduced with a minimum of difficulty. Radiation which may or may not be sufficiently penetrating to traverse a vacuum envelope is transformed to light, which definitely can. Thus a commercial sealed-off multiplier tube is practical, and the pressure of the surrounding atmosphere is thereafter of no importance to the operation of the detector. The question often arises as to whether penetrating radiations might better be allowed to eject photoelectrons directly, without the intermediate step of producing light in the fluorescent screen. With the fluorescent screen omitted from the detector, some response due to direct photoelectric excitation is actually observed with X rays, but only a few of the absorbed quanta give rise to signal pulses, and the number of photoelectrons in these pulses is small. With any known photoelectric material, the efficiency of such a method would be low in comparison to that obtained when the fluorescent process is introduced as an intermediate step. With penetrating radiations, efficiency is inherently low for direct excitation, regardless of the kind of cathode material used. The photoelectrons are produced throughout the depth of the cathode, and most of them are absorbed within the cathode or emerge with energies too high to permit focusing by the electron-optical system of the multiplier. In contrast, when the intermediate fluorescent process is used, radiation energy is effective when transformed into light anywhere within the thickness of the solid fluorescent screen, at least to the extent that the fluorescent material is transparent to light.

While the advantages of multiplier amplification are beyond dispute, the multiplier phototube brings with it two problems of its own. These are dark-current noise and statistical fluctuation in gain for elemental processes. The dark-current noise appears as pulses which tend to overlap the signal pulses in amplitude, determining in this way the low-intensity limit of sensitivity of the detector, at least for radiations characterized by weak signal pulses. The electrode surfaces, having a low work function, as is necessary for good sensitivity to light and to electron bombardment, tend also to release electrons whenever they acquire the higher thermal velocities associated with molecular agitation even at room temperature.[4] Thus a dark current of 10,000 to 100,000 thermal-emission electrons per second, depending on the particular tube, is emitted from the photocathode and from each dynode. Only those emitted from the photocathode receive the full amplification of the tube, and are of primary importance in contributing to noise. On an oscilloscope at the output of the multiplier tube, these noise pulses may be observed individually. They intermingle with any radiation signal pulses which may be present, and tend to obscure them. Only at radiation intensities so high that the signal-pulse rate approaches the noise-pulse rate can the presence of radiation be detected as a general increase in integrated output current as measured by an ordinary dc meter. But even at low intensities, radiation may be detected if the signal pulses have amplitudes sufficiently greater than the noise pulses. The output may then be passed to a discriminator circuit which delivers to a scaling-counting-circuit meter only those pulses above a selected amplitude.

Each noise pulse is associated with emission of a single thermal electron, whereas, ideally, each signal pulse is associated with the emission of an appreciable number of photoelectrons. Since signal-pulse detection is possible only to the extent that the signal-pulse amplitude is greater than the noise-pulse amplitude, it is exceedingly important that the detector be made as efficient as possible in the production of photoelectrons at the photocathode. To this end, the number of photoelectrons per radiation quantum absorbed must be made as great as possible by the proper choice of fluorescent screen, by the use of reasonably efficient light-collecting systems, and by the use of a selected multiplier phototube. Also, the distinction between signal and noise pulses must be maintained as carefully as possible by the use of a proper external circuit. These precautions are desirable for all radiations, and are essential to good low-intensity sensitivity for radiations yielding weak pulses. The output amplitudes of both signal and noise pulses vary widely, because of the statistical nature of the amplification at the various stages of the multiplier and other factors. Since the noise pulses are so numerous and the spread of amplitudes so great, the tendency to overlap extends to much greater signal-pulse amplitudes than might, at first, be expected.

Several experimenters have resorted to cooling the photomultiplier tube with dry ice or liquid air, at which temperatures the dark-current pulses may be almost completely eliminated.[4] While this procedure is very effective, and particularly useful in studying the characteristics of the detector, the method is very inconvenient and, in most cases, unnecessary. Signal pulses have been successfully separated from noise pulses simply by the use of efficient phosphors, good light collection, and proper attention to the associated circuits.

### Photomultiplier Tubes

Only a very few types of photomultiplier tubes are commercially available. Of these, the RCA 931A has been most widely used. It has a photocathode surface of cesium antimony, with a spectral response curve as given in Fig. 5.

The 1P21 tubes are identical with the 931A in construction, but have been especially selected by the manufacturer for better over-all sensitivity, which is considerably above that of the average 931A.

The 1P28 has the same internal construc-

[4] R. W. Engstrom, "Multiplier phototube characteristics: application to low light levels," *Jour. Opt. Soc. Amer.*, vol. 37, p. 420; June, 1947.

tion as the 931A, but is enclosed in an envelope of ultraviolet-transmitting glass which extends the spectral response of the tube well into the ultraviolet. The 1P22 utilizes a cesium-bismuth photosurface which has a spectral response much further into the red, but has greatly reduced total sensitivity.

Recently RCA has made an experimental photomultiplier, designated the C7132, which departs radically from the construction of the other types. The photocathode is transparent and deposited on the inside of the end of the cylindrical envelope. It is $1\frac{1}{2}$ inches in diameter, an area some 150 times that of the sensitive area of the 931A, and is almost directly accessible to the exterior, so that good light collection is assured. Though no reports are available as to its performance as a scintillation counter, the geometry is certainly excellent for this purpose.

All of these tubes have one characteristic in common—they are extremely variable from one sample to the next. Because it is very difficult to control accurately the properties of the photoelectric and secondary-emission surfaces, and because any factor which affects the sensitivity of all the surfaces is raised to the 10th power, it is little wonder that the over-all sensitivity varies so much from tube to tube. In addition, the amount of noise current generated in the tube varies over a wide range. For use as a scintillation counter, the gain and sensitivity are of little importance, for the output of the tube is well up to where simple amplifiers may be used. The important characteristic is the signal-to-noise ratio, i.e., the response to available amounts of light as compared to the dark current. Since some of the dark current occurring at the anode is simply dc leakage, it is important that only the ac component (shot noise) of the dark current and signal currents be compared.

Marshall, Coltman, and Bennett[6] have made comparative measurements on fifty-three 931A tubes bought from unused war surplus stock, using a wide-band ac voltmeter connected to the output of the photomultiplier. The voltmeter reading with no light admitted to the phototube served as a measure of noise. The increase of voltmeter reading with the introduction of a standard feeble beam of light upon the photocathode served as a measure of signal. The appropriate ratio of these two values then provided a relative signal-to-noise ratio which served satisfactorily for arranging tubes in the order of their merit. The standard light beam used was crudely collimated, was smaller in cross section than the photosensitive area, and was directed perpendicularly through the electrostatic shielding grid to the position on the cathode showing maximum response. Measurements were made quite rapidly. Once set, the geometry was not varied as successive tubes were introduced into the rigidly mounted socket, the position of maximum response having been found not to vary appreciably from tube to tube. The light intensity was set so low that the increase in meter reading for light was never more than a few times the reading for noise alone. The actual ratios are shown as abscissas in Fig. 6. Unfortunately, the optical system used for reducing the light intensity involved a number of elements and could not be specified accurately enough to permit duplication by others for the purpose of comparing single tubes with their results.

The histogram of Fig. 6 shows the results of the measurements. Most of the tubes are seen to be very poor in comparison to those in roughly the upper 20 per cent which proved to be satisfactory when made into detectors. No correlation whatever was found between signal-to-noise ratios and output sensitivity as indicated by the signal values obtained in these tests. When installed in radiation detectors, the tubes have been found to be effective in the order expected from the histogram within the limits of simple oscilloscopic comparison.

An ac meter was used for these measurements so as to eliminate the dc component of the dark current arising from resistive leakage in the output stage of the multiplier tube. This leakage, which does not contribute measurably to noise, is partly internal along the coating of condensed cesium between the elements of the tube, and it cannot always be eliminated or easily measured. The meter used was a Ballantine Model 300 voltmeter, which has a bandwidth of 150 kc. In this instrument the signal is amplified and passed through a full-wave rectifier, so that the meter indicates the average of absolute values of fluctuations impressed upon it.

It is fairly evident that signal-to-noise ratios obtained in these tests are independent of the gains of later stages of the multiplier tube. Instead, they tend to be related only to the very-early-stage pulse events, which are thus of primary concern. The obvious practical success of the method arises from the fact that the signal-to-noise ratios have to do essentially with photocathode

[6] Fitz-Hugh Marshall, J. W. Coltman, and A. I. Bennett, "Photo-multiplier radiation detector," *Rev. Sci. Inst.*, vol. 19, pp. 744–771; November, 1948.

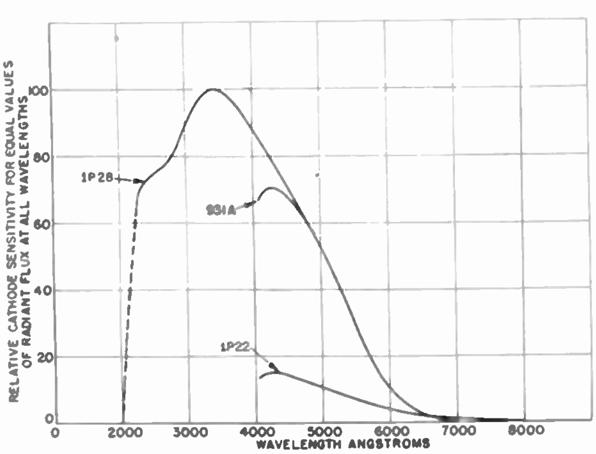


Fig. 5—Relative cathode response of various photomultiplier tubes as a function of the wavelength of the received light.

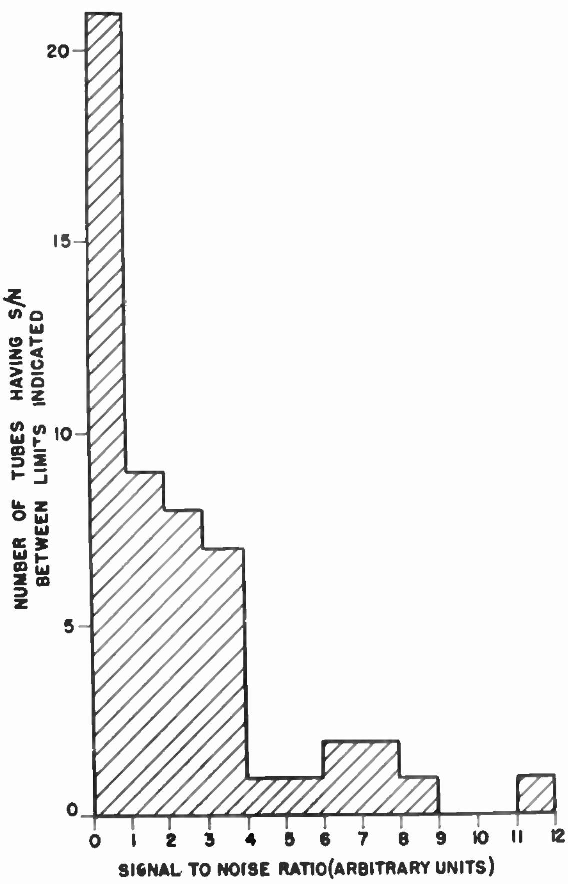


Fig. 6—Distribution of fifty-three 931-A tubes with respect to signal-to-noise ratio.

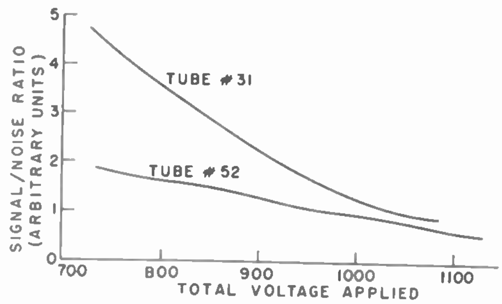


Fig. 7—Variation of signal-to-noise ratio with dynode voltage for two typical 931-A's.

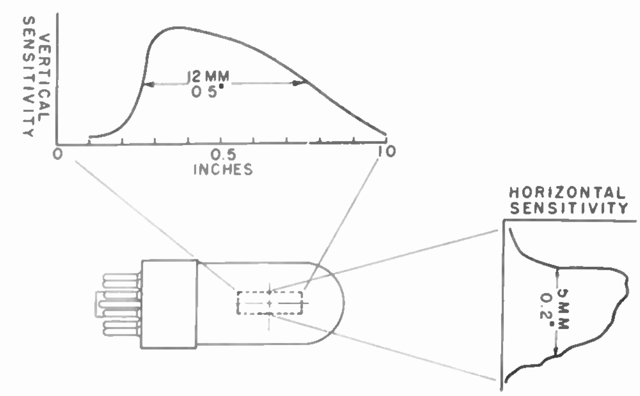


Fig. 8—Variation of sensitivity across the width and length of the photocathode of a 931-A tube.

sensitivity to light and with cathode dark current. Beyond this point there is some question as to exactly what is measured, and even as to what is most desired. For low pulse rates, as in the measurement of low thermal-emission currents, the pulses are probably largely resolved by the meter amplifier, and the meter reading tends to be proportional to the pulse rate. At high pulse rates the pulses may be poorly resolved, so that the reading tends toward proportionality to the square root of the pulse rate. Such a scale distortion may weight the signal-to-noise ratios in a way to favor tubes with unusually low cathode noise-pulse rates. It is questionable whether this is strictly desirable.

The variation in signal-to-noise ratio as a function of applied dynode voltage is shown for two typical tubes in Fig. 7. The curves suggest that it is generally advisable to operate at the lowest voltages that will yield sufficient gain; about 80 volts per stage is usually satisfactory. While the majority of tubes tested show a drop in signal-to-noise with increased voltage, a very few exhibit a slight improvement in signal-to-noise ratio as the voltage is increased.

Some difficulty is occasionally experienced with noise due to external electrical leakage. The chief source is corona leakage in the socket into which the tube is inserted, between the cathode terminal, which is operated at a high negative potential, and the adjacent anode terminal, which is operated near ground potential. This source of noise can be completely eliminated by removing the anode terminal lug from the socket, drilling out the socket, and inserting a new tubular pin grip surrounded completely by a polystyrene cylinder which is shielded, in turn, by a thin grounded metal tube. There is also leakage in the tube base itself, but this leakage seems to be purely resistive and does not contribute measurably to noise. While of no importance in the pulse detector, variations in this leakage are sometimes annoying in dc measurements; they can be reduced by removing the anode pin from the base of the tube, drilling a rather large hole through the side of the base for easy access, and connecting a polyethylene-insulated wire directly to the anode lead at the base of the glass envelope of the tube. Noise is also introduced when grounded conductors are permitted to contact the glass envelope of the tube, so that shielding or light-proofing coatings must either be insulated from ground or kept from touching the glass. The cause of this phenomenon has not been determined.

Many factors associated with noise, gain, and cathode photosensitivity in the multiplier phototube are poorly understood. For example, there is no indication whether the large variations in signal-to-noise ratio shown in Fig. 6 are primarily attributable to variations in cathode photosensitivity or in cathode noise. As another example, the increase in noise with dynode potential, as shown in Fig. 7, is not fully understood. Engstrom[4] mentions that the increase of noise with dynode potential may be associated with ionic feedback from the anode to the cathode. Also, it is difficult to explain Richardson plots determined experimentally for the photosurface, and to correlate them with work functions as indicated by the observed long-wavelength cutoff.[4]

The results of the signal-to-noise measurements, together with the performance of these tubes as scintillation counters, indicate that only about one tube out of five will be satisfactory for counter use. It appears that tubes should be selected on the basis of signal-to-noise ratio, a requirement which is not met by existing procedure in photomultiplier production. Though the 1P21 is a selected tube, the manufacturer has in the past based the selection on overall sensitivity, which would appear to bear little or no relation to the signal-to-noise ratio.

A characteristic of the photomultiplier which must be considered is the area and location of the sensitive portion of the photocathode. Though the entrance grid on the 931A is approximately 1 inch by $\frac{1}{2}$ inch, the useful cathode area is very much smaller, because only photoelectrons ejected from a certain rather narrow region will be carried to the first dynode by the focussing field.

The size and shape of the useful photosensitive area on the photocathode has been measured[5] for a number of 931A tubes. To simplify the process of measurement of a large number of tubes, an arrangement has been used whereby a line along the cathode was scanned rapidly with a 13-mil beam of light directed perpendicularly through the electrostatic shielding grid, and the response of the tube observed on a synchronized oscilloscope. Typical central traces along the width and length of the cathode are shown in Fig. 8. All tubes tested exhibited the asymmetry of vertical sensitivity shown in the figure. The shape of these curves, of course, is primarily determined by the electron-optical focusing properties of the electrode system, and does not indicate true variation of photosensitivity over the cathode. The potential difference per stage was 100 volts. No tests were made of change of size of photosensitive area with applied potential or of variation in sensitivity with angle of incidence of the light upon the cathode surface.

The tubes were found to be quite consistent, the width at half-maximum sensitivity varying not more than $\pm 10$ per cent. Despite the asymmetry in the vertical curve, the sensitive area, measured to points of half-maximum sensitivity, is approximately centered in the vertical direction midway between the mica spacers at the top and bottom of the cathode. In the horizontal direction the sensitive area is approximately centered along the projected midline of the entrance grid.

## Phosphors

The search for suitable fluorescent materials for use in scintillation counters at present still proceeds at a rapid pace. Since the desired characteristics for this use differ markedly in certain respect from those required in ordinary applications, many materials which are usually considered inferior have proved to give better results than the more usual phosphors.

For highly penetrating radiations, such as gamma and beta rays, the phosphor must be relatively thick if high absorption (i.e., good counting efficiency) is desired. In order that the fluorescent light may reach the photocathode, it is, of course, necessary that the material be relatively transparent to the light wavelengths emitted. For very thick sections, the usual finely divided crystalline materials are quite opaque and are thus at a considerable disadvantage, even though they may have a relatively high intrinsic conversion efficiency.

The attention of most experimenters has turned toward materials which can be obtained in large clear pieces, and some effort has been expended in growing single crystals of various materials for this use. However, for readily absorbed radiations, such as alpha particles, X rays, and low-energy beta rays, the materials in powder form can be used quite effectively, especially if some optical system such as the mirror of Fig. 3 is employed.

It is very difficult to compare in a rigorous manner the effectiveness of the various phosphors. Ideally, one would like to know the intrinsic efficiency of the luminescent process, i.e., the light energy generated by the crystal for a given energy absorbed from the exciting radiation. In addition, the absorption of the crystal for the exciting radiation, the efficiency of transfer of the fluorescent light from the crystal to the photosurface, and the spectral distribution of the emitted light should be known. Unfortunately, it is quite difficult to determine any of these factors.

Almost all observations taken consist simply of measuring the number of counts obtained with a given source of radiation. Usually a distribution curve of pulse heights is also taken, which may sometimes enable one to get a measure of the relative pulse heights for different materials. This appears to be a straightforward-enough method; yet it is beset with many uncertainties. First, results from different experimenters are almost impossible to compare unless they have each measured a common sample; for the geometries, radiation sources, and electronic equipments are bound to differ, often in important respects. Second, it is not easy to interpret the data taken on different materials, even with the same apparatus. As will be enlarged upon later, the distribution curves of pulse heights from a single sample have a shape somewhat similar to a hyperbola, so that the extrapolation to zero pulse height is often very uncertain. Without a knowledge of total number of pulses, the true average pulse height cannot be obtained. In spite of this, there is very great temptation to estimate pulse heights from visual observations. This is a dangerous procedure, for the *apparent* average pulse height is greatly affected by the pulse rate, which is a function of the absorbing power of the material as well as its intrinsic efficiency. In a few clear-cut cases, the differential bias curves (distribution of pulses according to height) show sufficient difference in slope when plotted on a logarithmic scale to enable meaningful statements as to the relative efficiencies of various phosphors to be made.

In spite of these drawbacks, the tests usually performed have the virtue that they represent the practical performance of a given sample, at least under the conditions of the test. What is perhaps an even worse state of affairs is the notorious variability of fluorescent substances. The fluorescent properties of many of the interesting materials depend on the presence of minute amounts of a foreign substance, or a very slight stoichometric excess of one of the constituents. It is possible to manufacture phosphors with reasonably uniform properties only when extreme measures are taken to control rigidly the purity of the materials, and all the steps in the processing. Naturally occurring materials are, of course, likely to vary widely in their fluorescent properties. Until more work is done on separating factors responsible for the behavior of a given phosphor sample, it will be difficult to correlate results of experiments, and a certain amount of disagreement between experimenters is to be expected.

The first materials successfully used for counting gamma and beta rays were silver activated zinc sulfide[1] and naphthalene.[6] The first material has a very high intrinsic efficiency (X-ray measurements indicate roughly 50 per cent), but its finely divided form prevents its being used in thick sections, and its relatively long decay time (20 microseconds or more) is prone to give rise to multiple counting. Most of the experimenters have compared their materials with naphthalene, and so it has, in a sense, become a standard of comparison. However, there are no published data on the variations of naphthalene from sample to sample, nor is there a generally accepted procedure for making comparisons, so that no real standard exists as yet. Moon[7] published results on twenty-eight inorganic materials tested with 1P28 and 1P21 photomultipliers using radium gamma rays as a source. The list is divided into categories of excellent, fair, medium, weak, and no "response." Listed as excellent are $CaWO_4$ (Scheelite) $CaF_2$ (Fluorite) $LiAlSi_2O_6$ (Spodumene) $Al_2O_3$ (synthetic sapphire) and LiF (synthetic). $CaWO_4$, in the form of a large, clear natural crystal about 7 by 10 by 1.1 mm, was said to be by far the best, and yielded about three times the gamma counting rate obtained with a much larger (11 by 16 by 45 mm) naphthalene crystal. No pulse-height measurements are reported.

Bell[8] has reported that anthracene, in the form of large, clear pieces crystallized from the melt, is much more effective than naphthalene. His pulse-height distribution curves show the total numbers of pulses (extrapolating to zero pulse height) as equal for the two materials, yet the anthracene pulses are about three times as high using gamma radiation, and five times using alpha particles.

Collins and Hoyt[9] report that anthracene, naphthalene, diphenyl, and a $BaSO_4$ +6 per cent $PbSO_4$ phosphor are effective with radium beta rays, but do not ascribe an order to their effectiveness.

Hofstadter[10] reports some work with crystals of KI and NaI activated with small amounts of thallium. The NaI appears to be a much more efficient phosphor than naphthalene, although the distribution curves do not afford a quantitative measure of pulse height or counting rate, for the distributions appear to be quite different, and the choice of bias setting (which determine the minimum pulse height counted) greatly affects the relative counting rate.

A private communication from Hofstadter indicates that NaI(Tl) gives about 1.5 times the pulse height of anthracene, and that apparently this material is the most effective known at the present time.

Before the true status of the various materials can be properly evaluated much work remains to be done. The NaI(Tl) appears to have a relatively high intrinsic efficiency, but it is doubtful if it approaches that of ZnS(Ag). Under X rays, a natural $CaWO_4$ crystal is quite noticeably weaker in visible fluorescent light yield than a good X-ray screen of synthetic $CaWO_4$. The intrinsic efficiency of the latter is known[11] to be 5 per cent for 80 kv X rays, while that of ZnS(Ag) is probably about 50 per cent. Though no direct comparison of NaI(Tl) with $CaWO_4$ is at hand, one would guess from the above references that NaI(Tl) is not more than three times as efficient as the natural $CaWO_4$. It would appear, then, that there is at least theoretically room for considerable improvement.

It should be emphasized again that conditions of an experiment will greatly affect its results. For example, the spectrum of emitted light from naphthalene is believed[6] to extend well into the ultraviolet, so that experiments with the 931A, which cuts off at about 3500A°, may favor other materials with a spectrum nearer the visible but with less total output. Again, when the decay time of the phosphor is not very small with respect to the time constants of the amplifier circuit, the spreading of the pulse in time lowers the relative height, even though the area (representing total light in the pulse) may be large.

The experiments have, however, turned up several materials which give excellent results, and the place of the photomultiplier scintillation counter as a detector seems to be assured, together with promise of considerable improvement in the near future.

## Optical Systems

For greatest possible discrimination between signal pulses and dark-current pulses, attention must be given to collecting as much light as possible from the phosphor. Rough computations based on the variation of sensitivity over the cathode (Fig. 8) and its geometrical configuration indicate that, for a small screen mounted directly on the envelope, the response is only 1/30 of that which would be obtained if all the light were directed to the most sensitive spot. These calculations assumed a Lambertian distribution of light from the screen, but did not take into account possible variation of photosensitivity with angle of incidence of the light.

When relatively large crystals are used for the phosphor, it is evident that only a

[6] H. Kallmann, *Natur und Technik*, July, 1947.

[7] Robert J. Moon, "Inorganic crystals for the detection of high-energy particles and quanta," *Phys. Rev.*, vol. 73, p. 1210; May 15, 1948.

[8] P. R. Bell, "The use of anthracene as a scintillation counter," *Phys. Rev.*, vol. 73, pp. 1405–1406; June 1, 1948.

[9] George B. Collins and Rosalie C. Hoyt, "The detection of beta rays by scintillations," *Phys. Rev.*, vol. 73, p. 1259; May 15, 1948.

[10] Robert Hofstadter, "Alkali halide scintillation counters," *Phys. Rev.*, vol. 74, pp. 100–101; July 1, 1948.

[11] J. W. Coltman, E. G. Ebbighausen, and W. Altar, "Physical properties of calcium X-ray screens," *Jour. Appl. Phys.*, vol. 18, pp. 530–544; June, 1947.

small fraction of the light will be collected from each scintillation. By wrapping the crystal in bright aluminum foil except for the face adjacent the tube, an improvement can be obtained which may be of the order of two times, depending on the geometry and transparency of the crystal.

For low-penetrating radiations such as soft beta rays, X rays, and heavy particles, a thin phosphor layer can be used; and, if the reduced area can be tolerated, the mirror system of Coltman and Marshall[12] provides the gain which is almost essential for low-energy radiations.

The dimensions of the sensitive cathode area and the magnification of the mirror system determine the maximum useful size of the fluorescent screen. Measured to points of half-maximum sensitivity in Fig. 8, the dimensions of the sensitive cathode area are about 5 mm by 12 mm. The mirror has a magnification of about two. For this mirror therefore, a fluorescent screen size of 2.5 mm by 6 mm gives an image which approximately covers the sensitive area on the cathode.

The mirror was made from the hemispherical bottom of a Corning T-10 glass bottle, cut off with a wet carborundum wheel, and ground to the dimensions shown in Fig. 9. The inside surface is aluminized by vacuum evaporation.

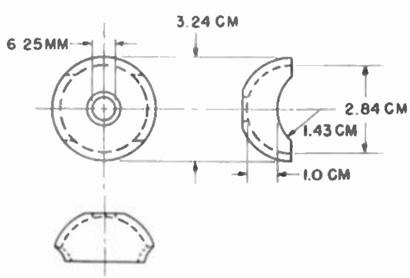


Fig. 9—Dimensions of an effective light-gathering mirror.

In the final assembly, the position of the mirror on the tube envelope is adjusted individually for each unit. The phosphor is first strongly excited by daylight, causing a faint phosphorescence which persists for several minutes after the tube is taken into the dark. With a vacuum-tube voltmeter used as a dc microammeter to measure anode current, the position of the mirror for maximum response is then located. At the same time the phosphorescing screen provides a means for measuring the effectiveness of the light-gathering mirror. Response with the mirror in place has been found to be about five times that from the phosphor alone without a mirror.

It is necessary, of course, to exclude external light from the tube when in operation. For most purposes this requires the use of either a light-proof coat of black laquer, or a small enclosure around the tube, and, for most radiations, an opaque thin window. Perhaps the most satisfactory window material is beaten aluminum leaf. This can be obtained as thin as 0.00003 inch (0.18 mg/cm²) from M. Swift and Sons, 10 Love Lane, Hartford, Conn. Because of small pinholes, it is best to use two thicknesses, which still have a very small stopping power, even for alpha particles.

The enclosure shown in Fig. 4 has the mirror attached to the inside wall of the box rather on the phototube, and the window mount includes a convenient provision for simultaneously punching out a round disk of the aluminum leaf and securing it in position.

[12] J. W. Coltman and Fitz-Hugh Marshall, "Photomultiplier radiation detector," *Nucleonics*, vol. 1, pp. 58–64; November, 1947.

## Circuit Considerations

The type of amplifier used with the scintillation counter will be dictated to some extent by the intended use of the counter. For alpha particles, where the response is extremely high, it is possible to operate a pair of headphones directly from the anode of the photomultiplier.[13] For operating discriminating circuits, some form of preamplifier is desirable, and its design will depend on the maximum counting rate desired, and on the fluorescent decay characteristics of the phosphor.

While decay times of naphthalene and anthracene phosphors are apparently extremely short, other useful phosphors have measurable decay times ranging from 0.6 to 20 microseconds.[14] The time constants of the amplifier will have a distinct influence on the ratio of signal pulse heights to dark current pulse heights, and it is worth while to design the amplifier for optimal performance, at least in the case of the lower-energy radiations. The considerations governing amplifier design will be discussed in some detail, not only for their practical importance, but because of the insight which they afford into the detection process itself.

The resolving time of the multiplier tube itself is of the order of $10^{-8}$ second, this limit being set by variations in the $10^{-7}$ second total transit time for electrons through the successive multiplier stages. Approximately stated in round figures, each electron ($10^{-19}$ coulomb) leaving the cathode gives rise to an avalanche of $10^6$ electrons ($10^{-13}$ coulomb) at the anode, spread over a time of $10^{-8}$ second. With the anode capacitance reduced to the minimum of $10^{-11}$ farad, and with the anode output resistance set at $10^3$ ohms to provide an $RC$ time constant matching the resolving time of the tube, such a pulse might appear at the anode as a potential swing rising to $10^{-2}$ volts and persisting for $10^{-8}$ second. The dark-current noise of a selected 931A tube is made up of perhaps 20,000 such pulses a second, randomly distributed in time. These pulses vary considerably in amplitude because of large statistical fluctuation in gain in the early stages of the multiplier, where the number of electrons carrying the pulse is still small. If an oscilloscope with the necessary sensitivity and resolving time were available, the dark-current noise might appear as in Fig. 10(a). In this figure and in those which follow, a single extraordinary

[13] R. Scherr, "Scintillation counter for the detection of α-particles," *Rev. Sci. Instr.*, vol. 18, pp. 767–771; October, 1947.

[14] Fitz-Hugh Marshall, "Microsecond measurement of the phosphorescence of X-ray fluorescent screens," *Jour. Appl. Phys.*, vol. 18, pp. 512–519; June, 1947.

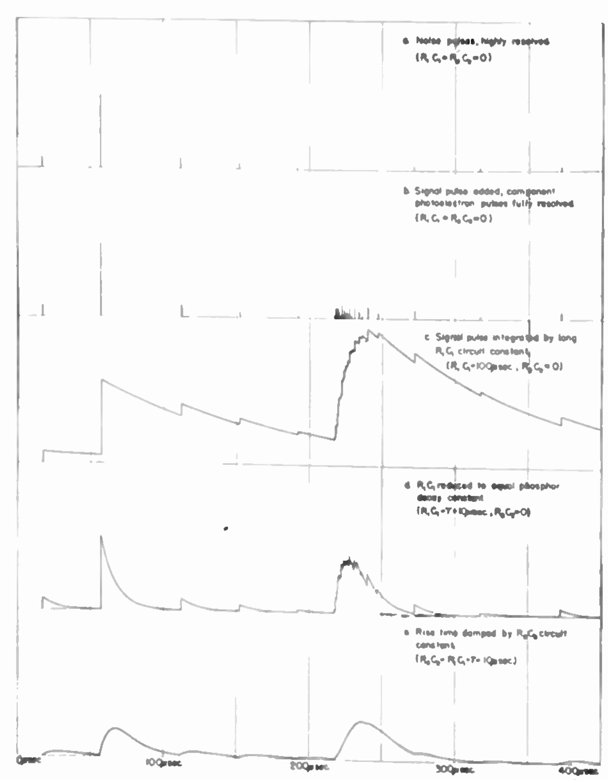


Fig. 10—Effect of circuit constants on noise and signal pulses.

large noise pulse is included at the left to represent one which can be discriminated against only with difficulty; otherwise, the distribution of amplitudes and spacing is typical.

Let us now consider a signal pulse, from a beta particle, say, which results in the emission of 20 photoelectrons at the photocathode. Although the stimulation of the fluorescent screen by the beta particle is essentially instantaneous, emission of light is spread over an appreciable interval of time because of phosphorescent persistence of the screen. The intensity is a maximum at the instant of stimulation, and then drops off according to the decay curve of the phosphor; for the present example we may think of this curve as exponential, with a decay time constant of ten microseconds. Such a signal pulse, as it might appear at the phototube anode, is shown in Fig. 10(b), superimposed upon the pattern of noise pulses carried over from the previous figure.

It now becomes clear that a circuit time constant of $10^{-8}$ second would be an unfortunate choice, for it would resolve a signal pulse into its component photoelectron pulses, as illustrated in Fig. 10(b). The component photoelectron pulses average no more in amplitude than noise pulses. A discriminator set to miss the high noise pulse at the left in the figure would also completely miss the signal pulse, even though the total charge associated with the latter happens to be greater than that of the large noise pulse. This difficulty can be avoided by using a circuit time constant long enough to integrate the signal pulse.

Fig. 10(c) shows the effect of a high resistance across the output of the multiplier tube such that the anode $RC$ time constant is 100 microseconds. Since this is large in comparison to the decay constant of the phosphor, integration of the total charge associated with the signal pulse is almost as complete as it would be with an infinite circuit time constant, and the signal pulse is properly shown rising well above the amplitude of the large noise pulse at the left, permitting easy discrimination on the basis of amplitude. However, this time constant is too long; the trace does not let down sufficiently between noise pulses, which occur on the average once ever 50 microseconds, and the maximum signal-pulse counting rate is unduly low.

In Fig. 10(d) the other extreme is illustrated. The anode circuit time constant is set at 10 microseconds, equal to the decay constant of the screen. Though the response drops down almost completely between noise pulses, and though the counting rate is about as good as the decay time of the phosphor will permit, a new difficulty has arisen. The peak amplitude of the signal pulse in relation to that of the large noise pulse has begun to fall off appreciably and no longer approximates the sum of the componet photoelectron pulses. However, if we should choose to retain this shortened circuit time constant for purposes of fast counting or otherwise, the loss in signal-pulse amplitude can be approximately balanced by the introduction of another circuit device by which the noise pulse amplitude is also reduced.

Fig. 10(e) shows the effect of adding a second circuit time constant, also of 10 microseconds duration, designed primarily to control the rate of *rise* of potential. This damps out the initial fast-rising swing of the noise pulse, without greatly affecting the more slowly rising signal pulse. At the same time, it increases somewhat the pulse duration. With the initial sharp peak of the noise pulse cut off by this second circuit, the distinction originally inherent between the signal and noise pulses by virtue of their total charges may again be closely realized, even though the time constants used may approach the decay time of the fluorescent screen.

The basic circuit for introducing these two time constants is shown in Fig. 11. $R_1C_1$ is the anode-circuit time constant which controls the rate of fall so as to integrate the signal pulse. $R_2C_2$ is the constant introduced to control the rate of rise. As will be shown in the next section, these constants are best set approximately equal to the decay constant of the fluorescent screen. The simple $RC$ networks chosen to adjust the frequency characteristics of the system are not necessarily the most effective networks for the purpose of obtaining optimal signal-to-noise pulse discrimination. However, as will be shown, the results obtainable are quite close to the optimum.

The features pointed out above in the discussion of signal and noise pulse characteristics are strikingly demonstrated in the high-speed oscillogram of Fig. 12, which was made under approximately the conditions assumed in Fig. 10(c), though with a shorter circuit time constant. In this oscillogram time increases from left to right, but each successive trace appears at a higher position than the last. The time for one complete sweep is 200 microseconds, the gap near the middle of each trace being a 100-microsecond time marker. Signal pulses from $P^{32}$ beta rays are included in the upper part of the figure, the largest shown being well below the maximum amplitude observed. The component photoelectron pulses making up these signal pulses are clearly resolved as steps, the statistical variations in amplitude and spacing being quite evident. The decay of the Patterson Type-D fluorescent screen is not exponential, and persists longer than might be expected from the initial decay rate. The middle part of the figure shows only background noise pulses, being a fairly typical section of the recording made while the beta rays were blocked off. The bottom part of the figure also shows only noise pulses, but it is selected to include an exceptionally large noise pulse. Although this large pulse is distinctively different in appearance from signal pulse, being characterized, as always, by a sharp rise and an unbroken exponential fall, it has a greater amplitude than many of the signal pulses recorded; with a reasonable discriminator setting, it would ordinarily be counted as a signal pulse.

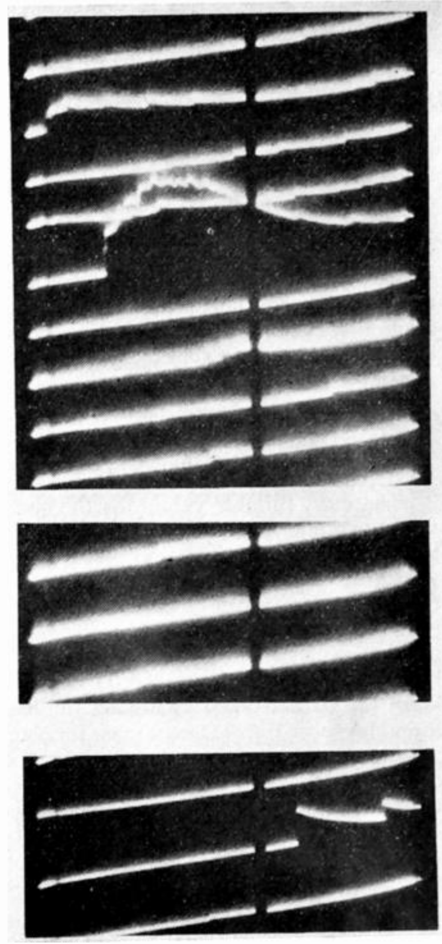

Fig. 12—Oscillograms of a beta-ray pulse (upper), noise pulses (middle), and an exceptionally large noise pulse (lower).

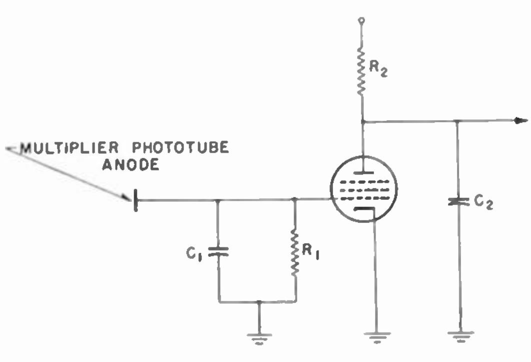


Fig. 11—Basic circuit for introducing $R_1C_1$ and $R_2C_2$ time constants.

The oscillogram of Fig. 12 brings out the somewhat unusual situation that arises in the multiplier anode circuit with respect to resolving time. The $R_1C_1$ time constant applies a resolving time only during the fall of the response. The resolving time for the rise is essentially instantaneous, and is unaffected by the $R_1C_1$ constant. This instantaneous rise has some interesting consequences. It makes possible the resolution of the individual photoelectron pulses in the signal pulses in the oscillogram. On the other hand, the $R_2C_2$ time constant later in the circuit is required if the rise rate is to be slowed down to match the response time of the fluorescent screen.

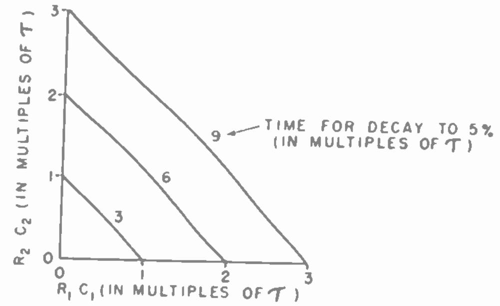


Fig. 14—Contours of time required for a noise pulse to subside to 5 per cent of maximum amplitude, as a function of the circuit time constants.

## Derivation of Optimal Circuit Constants

The most suitable values for the circuit time constants $R_1C_1$ and $R_2C_2$ may be determined by mathematical analysis. For convenience, the phosphorescent decay of the fluorescent screen can be assumed to be exponential with a decay time constant. This approximation is justifiable in view of the noncritical nature of the results obtained. With the circuit taken as having the basic form shown in Fig. 11, equations for signal pulses and noise pulses at the pentode anode can be derived, and from these the ratio of their maximum amplitudes can be obtained. The variation of this ratio may be plotted for various values of $R_1C_1$ and $R_2C_2$ to permit selection of the optimal values. The analysis proceeds as follows.

Let the total charge $Q$ due to a signal pulse be delivered to the multiplier anode in Fig. 11, according to the exponential decay law

$$I_g = \frac{Q}{\tau} e^{-t/\tau} \tag{1}$$

where $I_g$ is the multiplier anode current or pentode grid current at the time $t$ after the start of the signal pulse, and $\tau$ is the decay time constant of the excited phosphor. Since the current $I_g$ divides between the resistance $R_1$ and the capacitance $C_1$,

$$\frac{Q}{\tau} e^{-t/\tau} = \frac{E_g}{R_1} + C_1 \frac{dE_g}{dt} \tag{2}$$

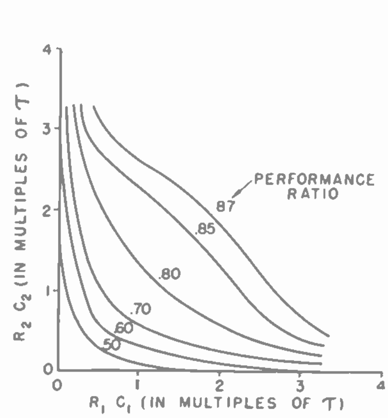


Fig. 13—Contours of circuit performance ratio (signal pulse amplitude versus noise pulse amplitude for the same total charge) as a function of the circuit time constants.

where $E_g$ is the grid potential. The solution of this differential equation is

$$E_g = \frac{Q}{C_1} \frac{R_1C_1/\tau}{1 - R_1C_1/\tau} \left(e^{-t/\tau} - e^{-t/R_1C_1}\right). \tag{3}$$

The plate current of the pentode is then $I_p = g\,E_g$, where $g$ is the transconductance of the tube. Since the plate current divides between the resistance $R_2$ and the capacitance $C_2$,

$$g \frac{Q}{C_1} \frac{R_1C_1/\tau}{1 - R_1C_1/\tau} \left(e^{-t/\tau} - e^{-t/R_1C_1}\right) = \frac{E_p}{R_2} + C_2 \frac{dE_p}{dt} \tag{4}$$

where $E_p$ is the plate potential. The solution of this differential equation is

$$E_p = g \frac{Q}{C_1} R_2 \frac{R_1C_1/\tau}{1 - R_1C_1/\tau} \left[ \left( \frac{R_1C_1}{R_1C_1 - R_2C_2} - \frac{1}{1 - R_2C_2/\tau} \right) e^{-t/R_2C_2} + \left( \frac{1}{1 - R_2C_2/\tau} \right) e^{-t/\tau} - \left( \frac{R_1C_1}{R_1C_1 - R_2C_2} \right) e^{-t/R_1C_1} \right]. \tag{5}$$

Given $\tau$, this equation for the emerging signal pulse as a function of time can be maximized and the maximum pulse height thus determined.

For a noise pulse, let the total charge again be $Q$. We may represent the fact that this charge arrives instantly at the multiplier anode by letting $\tau = 0$. Then (5) simplifies for the noise pulse to

$$E_p' = g \frac{Q}{C_1} R_2 \frac{R_1C_1}{R_1C_1 - R_2C_2} \left(e^{-t/R_1C_1} - e^{-t/R_2C_2}\right) \tag{6}$$

The ratios of the maximum amplitudes of $E_p$ and $E_p'$, which incidentally do not occur at the same values of time $t$, may be determined by graphical methods for various values of $R_1C_1$ and $R_2C_2$ and are plotted as contours in Fig. 13. The curves are symmetrical with respect to the axes. Since the same total charge $Q$ was assumed for both types of pulse, the best possible value for the ratio plotted in the figure is unity. As might be expected, the circuit fails to maintain this perfect ratio for finite values of $R_1C_1$ and $R_2C_2$; however, the performance ratio approaches unity for reasonable values of the circuit constants.

The effect of $R_1C_1$ and $R_2C_2$ upon the resolving time of the circuit must be taken into account in making a choice of circuit constants. In Fig. 14 contours are plotted which show the time required for a noise pulse to decay to $e^{-3}$, or 5 per cent of its maximum value. This is a more complete decay than is necessary as an indication of resolving time, but its choice simplifies computations from (6). Comparison of Figs. 13 and 14 shows that in the likely working range the best resolving time on each performance ratio contour occurs at $R_1C_2 = R_2C_2$. Considering only such values and keeping in mind that a short resolving time is preferable, we see that with these circuit time constants increased only to the value of the phosphorescent decay constant $\tau$, the performance ratio is about 75 per cent; the inherent ratio of charge in a signal pulse to charge in a dark current pulse is brought out three-fourths as well as would be possible if the resolving time were increased to infinity. Since further improvements in the performance ratio cannot be large, and since they entail disproportionately large increases in resolving time, optimal values of the circuit constants would appear to be approximately $R_1C_1 = R_2C_2 = \tau$. As will be seen, however, the final choice of constants may be varied somewhat according to the conditions and requirements of the application.

## Fluorescent Screen Decay Constant

Since the time constants of the circuit following the multiplier tube do not have to be set with high accuracy, an approximate value for the phosphorescent decay constant of the fluorescent screen will suffice. Satisfactory measurements can be made by direct study of the signal pulses on an oscilloscope, according to any of several circuit arrangements. Even with radiations characterized by weak signal pulses, the wide distribution of pulse amplitudes assures the occurrence of occasional pulses comprising a sufficient number of photoelectrons to permit estimation of the decay rate. When the yield of photoelectrons per signal pulse is very large, the decay curve of the phosphor, representing light intensity as a function of time during a signal pulse, may be directly observed on an oscilloscope having the necessary short resolving time. Since the curve observed on the oscilloscope represents merely the interaction of the desired decay curve on the circuits which follow the multiplier tube, it is essential that the resolving time of all parts of the circuit be kept short in comparison to the decay time of the phosphor or that appropriate corrections be introduced. Thus, in a circuit like that shown in Fig. 11, $R_1C_1$ and $R_2C_2$ must be small. With this method of observation, the sweep may be triggered by the signal pulse itself. Successive signal pulses then appear at approximately the same starting position along the time axis. Though they vary statistically in amplitude, they form an umbrella of curves, the lower edge of which tends to be sharply defined by pulses having just enough amplitude to trigger the sweep, these being the pulses which occur with greatest frequency. Good visual measurements can be made, or the umbrella may be photographed with a simple time exposure. With a more elaborate photographic arrangement for registering high writing speeds, individual pulse traces may be photographed, the shutter time being adjusted to admit on the average only a few pulses.

With weak signal pulses, the method of the last paragraph is less suitable. Each signal pulse is composed of only a few component photoelectron pulses, which vary statistically in amplitude and position. The trace partially resolves these and appears very irregular in shape. The trace may be smoothed somewhat by using an $R_1C_1$ circuit constant which is not a great deal smaller than the decay constant of the phosphor, but the effect of this circuit constant on the apparent decay curve must be taken into account. Such an increase of $R_1C_1$ also increases the rise time, the result being that several of the few available photoelectrons are used up in the initial rise before the decay process begins to register. For weak pulses, therefore, the component photoelectrons are more economically utilized and the results are perhaps less questionable if one goes to the other extreme of introducing a very large multiplier-anode circuit constant $R_1C_1$, $R_2C_2$ still being kept small, so as to obtain an oscilloscopic trace similar to the response illustrated in Fig. 10(c). Under these conditions the resolving time for the rise is essentially instantaneous, and all the information inherent in the original signal pulse tends to be preserved. For a very large $R_1C_1$, the *slope* of the trace corresponds to the light intensity from the phosphor; except for an exponential decay, the curve does not strictly represent an inverted decay curve. Though the signal pulses may be used to trigger the oscilloscope sweep, no sharply defined envelope is formed, and accurate visual estimates are difficult.

Greater accuracy with weak signal pulses can be obtained by a logical extension of the method of the last paragraph, which requires, however, that individual traces be photographed. Though the method will be illustrated by the oscillographic recording of Fig. 12, which involved rapid motion by the film, a satisfactory oscillogram could have been obtained with the film at rest and with the exposure limited by an appropriate shutter time to one or two self-triggered signal pulses. From the oscillogram, the time of occurrence of each component photoelectron pulse in the signal pulse is first tabulated. A new curve is then drawn which is like the original one, except that all photoelectron pulses are plotted as steps of identical amplitude and the effect of the $R_1C_1$ capacitor discharge is omitted. A curve derived in this way from Fig. 12 is shown in Fig. 15. On still another curve the actual decay of the phosphor is plotted as in Fig. 16, the ordinates being slopes taken from the curve of Fig. 15; these slopes are proportional to the instantaneous light intensity or rate of occurrence of photoelectron pulses. The primary improvement of this method arises from the elimination of statistical variation in amplitudes of the photoelectron pulses comprising the signal pulse, the variation in this factor being much greater than that in the time-distribution of the photoelectron pulses. The method also eliminates all dependence upon circuit constants.

In Fig. 16 a curve for Patterson Type-D screen from a separate experiment is included for comparison, this curve having been obtained with microsecond-pulsed 30 kv X rays.[14] On the basis of these closely agreeing curves, the circuit time constants $R_1C_1$ and $R_2C_2$ in the final detector should be set at about 17 microseconds for this phosphor. Though the nonexponential decay drops to $e^{-1}$ of the initial intensity in about 10 microseconds, the $e^{-2}$ point is at about 35 microseconds, and the $e^{-3}$ point is at about 100 microseconds. Thus the 17-microsecond circuit constant choice corresponds somewhat arbitrarily to the decay constant measured through the $e^{-2}$ point; performance studies suggest that larger circuit constants would be better at low counting rates. This choice is primarily appropriate only to the detection of low-amplitude signal pulses, such as are characteristic of X rays, beta rays, and gamma rays. For alpha rays in fast-counting applications, considerably shorter circuit constants could be used.

For a phosphor having a nonexponential decay, the decay curve depends upon the type of excitation. This is easily observed when large signal pulses from alpha particles are compared with weak signal pulses from beta particles. The decay time of the Patterson Type-D screen for pulses from $Po^{210}$ alpha particles is observed to be about one-fourth that for $P^{32}$ beta particles. The density of excitation with alpha particles is much higher than with beta rays, due to the short range of the alpha particles. The shorter alpha-pulse decay time may be simply the consequence of the nonexponential decay curve, the active centers being excited to a higher point on the decay curve where the slope or decay rate is higher. However, the sharp inflection of the decay curve taken with pulsed X rays[14] occurring before the 1-microsecond point, suggests the presence of two distinct fluorescent processes of different decay times. If this is so, the short decay time obtained with high-density alpha-particle excitation might be explained on the basis of saturation of the fluorescence having the longer decay time. Regardless of the decay mechanism involved, it is interesting to note that as an alpha pulse subsides to the level of a beta pulse, its curve may be observed to fall off at about the same rate as that of a beta pulse.

The close agreement of the two curves for beta rays and X rays in Fig. 16 is rather surprising in view of the nonexponential nature of the decay. Two different types of radiation are involved, in one case X ray

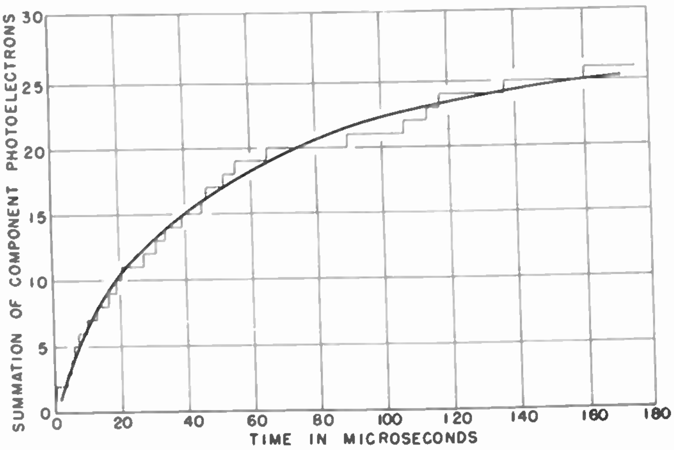


Fig. 15—Curve of integrated response for a beta-ray signal pulse, replotted from the large pulse in the oscillogram of Fig. 12.

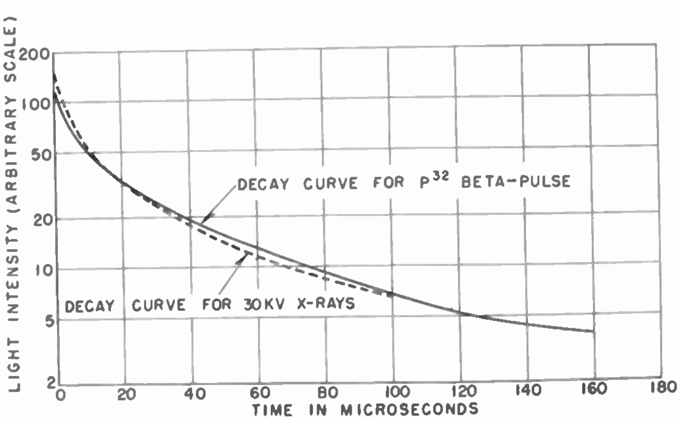


Fig. 16—Phosphorescent decay curves of zinc-sulfide (Patterson "D" X-ray screen) with excitation; (a) by a single beta particle, and (b) by microsecond irradiation with 30-kv X rays.

quanta having energies up to 30 kev, and in the other case a beta particle with an energy which can be as high as 1.7 Mev. Further, the excitation time might be expected to be different; a microsecond interval of "continuous" irradiation was used with the X rays, as compared with the essentially instantaneous excitation of the single beta particle. However, the excitation level is not necessarily different in the two cases. From estimates in the X-ray case of the number of phosphor particles in the screen and the number of X-ray quanta absorbed by the screen during each irradiation interval, it is found that the average number of X-ray quanta absorbed per phosphor crystal was considerably less than one; it is unlikely that any crystal received excitation from more than one quantum during a single observation interval. This is also obviously true for the beta-ray case, where a single event was observed. It is also known that excitation is not, in general, transmitted from one excited crystal to its neighbors. Thus, in each case, the level of excitation would appear to depend largely on the energy imparted to the screen by a single quantum or particle. Calculation of the energy lost by a beta particle in traversing the screen shows that it loses only about 30 kev, which is roughly the energy supplied by an X-ray quantum. One concludes that the actual excitation conditions for the two cases are quite similar, despite what appear at first glance to be very different conditions.

Several measurements of the decay time of phosphors suitable for photomultipliers have been published. Hofstadter reports a decay constant of $\frac{1}{4}$ microsecond for NaI(Tl). $CaWO_4$ is known[14] to have an exponential decay with a time constant of 6 microseconds, and $BaSO_4$–$PbSO_4$ a decay time of 0.6 microsecond. Wouters[15] has estimated the decay constant of anthracene as being in the neighborhood of 0.05 microsecond, though a more sensitive determination by Collins[16] gives a value of $1.3 \pm 0.2 \times 10^{-8}$ seconds. Collins also reports decay constants for naphthalene ($5.7 \pm 0.5 \times 10^{-8}$ seconds (and phenanthrene ($0.9 \pm 0.2 \times 10^{-8}$ seconds).

In the general field of phosphorescence many fundamental problems remain which have not yet been satisfactorily attacked. The methods discussed in this section may offer an additional experimental approach by which the mechanism of phosphorescence can be studied. Aside from techniques for studying ultra-short phosphorescent events, a chief merit of these methods is that measurements may be made upon individual quantum events.

## Preamplifier Circuits

The output signal from the photomultiplier circuit is in the form of a charge, which produces a corresponding voltage on the capacitance of the anode circuit. The anode resistor serves merely to return this capacitor voltage to zero, and if the signal pulse duration is short compared to the time constant of this circuit, the peak anode voltage is independent of the resistance, and depends only on the charge and the capacitance. It is thus advantageous to keep the capacitance of the anode circuit small and constant in value. For this reason it is preferable to use a preamplifier head to which the photomultiplier tube is directly connected. The anode-circuit capacitance is then the grid capacitance of the first preamplifier tube, the anode capacitance of the multiplier itself, and the strays due to sockets and wiring. Compact construction will result in a total capacitance of about 10 $\mu\mu$f. The anode resistance is then chosen to give the required resolving time.

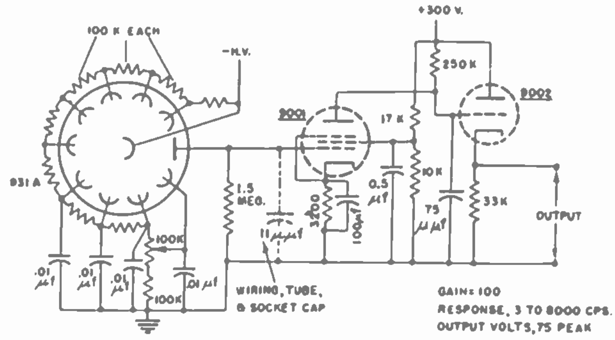


Fig. 17—The preamplifier circuit diagram.

A practical preamplifier circuit is shown in Fig. 17. This circuit was designed in accordance with the principles outlined under the section on circuit considerations for use with a ZnS:Ag phosphor which has a relatively long (17-microsecond) decay time. For use with other phosphors the photomultiplier anode resistor and the 75-$\mu\mu$f interstage filter capacitor should be modified to keep the corresponding *RC* time constants close to the decay time of the phosphors.

The power supply for the circuit is not shown, for it will vary in form according to the requirements of stability, compactness, and portability. A negative voltage of 700 to 1,000 volts is required for the dynode potential divider. It should preferably be variable over this range, since photomultiplier tubes show wide variations in gain and noise characteristics. The divider shown provides equal voltage steps for the various dynodes, although Wouters[15] has shown that greatly increased over-all gain can be obtained by running the later stages at higher voltages, without affecting the signal-to-noise ratio. It is likely that individual tubes behave differently in this respect, and it is probably necessary to adjust the voltage distribution for each tube used. The power supply need not be highly filtered, but, since the gain varies rapidly with dynode voltage, and the pulse-height distribution curves are quite steep, a certain amount of stabilization is desirable. A simple stabilizing device which is adequate for most purposes consists of a series of small neon tubes connected across the high-voltage supply, with a suitable series resistor to limit the current to a few milliamperes.

The 0.01-$\mu\mu$f capacitors across the bleeder at each of the last four dynode stages of the multiplier tube may be necessary to handle very large instantaneous currents during large signal pulses. A gain control in the last dynode stage is also provided to permit lowering of the amplification so as not to overload the 9001 pentode amplifier tube when large pulses are being observed. The latter part of the circuit in Fig. 17 comprises a direct-coupled cathode follower stage to drive at low impedance the cable leading to the discriminator and counting circuits. The output to the cable is positive and is of the order of 40 volts for beta pulses, which is adequate for precise operation of biased discriminators. The entire circuit may be built in a small box on the top of which the multiplier phototube detector may be mounted in an exposed position.

Discussion of discriminator, scaling, and counting circuits is outside the scope of this paper. These circuits are in general similar to those used with pulse ionization chambers, proportional counters, and Geiger counters, except that extremely fast circuits are necessary if full advantage of the high resolving power of the scintillation counter is to be realized.

Coincidence techniques, using two photomultipliers with a single phosphor, have been used[6,17] to obtain a considerable improvement in discrimination of signal over noise pulses.

## Performance

The scintillation counter is, as yet, in the earlier stages of its development. Of the rather sizeable number of publications so far released, only one[18] describes the use of such a counter for an actual experiment in nuclear physics; all the rest are concerned with reports of its characteristics using various materials as phosphors under a wide variety of experimental conditions. From these reports it is possible to formulate a general picture of the performance of the tube, but as yet each experimenter who plans an application for the counter will be obliged to determine for himself the performance of the particular arrangement chosen.

[15] Louis F. Wouters, "Pulse characteristics of anthracene scintillation counters," *Phys. Rev.*, vol. 74, pp. 489–490; August 15, 1948.

[16] George B. Collins, "Decay times of scintillations," *Phys. Rev.*, vol. 74, pp. 1543–1544; November 15, 1948.

[17] Martin Deutsch, "High efficiency, high-speed scintillation counters for beta and gamma rays," (abstract), *Phys. Rev.*, vol. 73, p. 1240; May 15, 1948.

[18] S. DeBennedetti, F. R. McGowan, and J. E. Francis, Jr., "Self-delayed coincidences with scintillation counters," *Phys. Rev.*, vol. 73, pp. 1404–1405; June 1, 1948.

The efficiency of detection and the ability to detect small amounts of radiation are important characteristics of the detector. These are determined partly by the absorbing power and light yield of the phosphor, but are also affected in practice by the extent to which signal pulses and noise pulses overlap in amplitude. For this reason, the pulse height distribution curves constitute important information as to the behavior of the counter.

When the counting rate is measured as a function of discriminator bias voltage, integral-bias curves such as those shown in Fig. 18(a) are obtained. In this figure a curve for background noise is included along with curves for two radiation intensities differing by a factor of 10; for the two signal-pulse curves the background counting rate has been subtracted before plotting. The bias-voltage scale will vary with over-all amplification in the multiplier and later amplifiers. The distribution of pulse heights is shown in Fig. 18(b), in which the slopes of the curves in Fig. 18(a) are plotted.

When a thin zinc-sulfide phosphor is used, no upper limit to the counting rate is ordinarily observed as the bias voltage is reduced toward zero, either in the bias integral curve or in the pulse-amplitude distribution curve. This is especially surprising with signal pulses, for which a finite limit must logically exist. At low radiation intensities, the signal curves cannot be traced back close to zero bias because the signal-pulse rate is soon masked by the rapid increase in noise-pulse rate. Yet, even at high radiation intensities, the signal-pulse rate for X rays, gamma rays, or beta rays appears to rise indefinitely as zero bias is approached. In the case of signal, this continued rise near zero bias does not appear to represent a true increase in pulse rate. It is caused by multiple counting of statistical fluctuations near the end of signal pulses and of late photoelectrons associated with long-period, low-intensity phosphorescence of the fluorescent screen. The dashed branches of the signal curves in Fig. 18 are believed to represent the true pulse rates at low bias.

The most striking feature of the curves in Fig. 18 is the extremely broad distribution of pulse amplitudes. As far as evidence from a routine data set is concerned, the curves for signal pulses and noise pulses are similar to hyperbolas in that they appear to extend indefinitely in both directions. Even if the portion of the signal curves below 4 volts bias that is believed to be spurious is disregarded, the distribution is very broad. As the radiation intensity is lowered, complete discrimination of signal pulses against noise pulses becomes impractical with this phosphor. Complete discrimination can be approximated only by raising the bias until a large proportion of signal pulses are also rejected, thus reducing the efficiency with which the signal quanta are detected. The difficulty is greatly increased if a tube having a poor signal-to-noise ratio is used, or if any other factor in the construction of the detector acts so as to expand the noise curve to the right in relation to the signal curves.

Since curves such as these do not intersect the axes at well-defined points, it is often not possible to make satisfactory comparisons of average heights of pulses of various types from curves as ordinarily taken. The difficulty is the lack of a zero-bias intercept. Another aspect of this difficulty is illustrated by the signal-pulse curves for different radiation intensities in Fig. 18. It is seen that the general *upward* shift associated with an increase in radiation intensity is readily confused with an *outward* shift; that is, with an apparent increase in pulse amplitude. Without a knowledge as to where the termini of the curves lie, very little information as to relative pulse heights may be obtained. This effect is also very pronounced with visual observation on an oscilloscope. The eye sets artificial termini to the distribution curve, ignoring very large pulses because of their infrequency and very low ones because of their small size. Though the average height may seem fairly well defined, the impression of increasing pulse heights is very strong as the intensity of the radiation is increased. Thus visual estimates of relative pulse heights are in general to be mistrusted.

The broad distribution of noise-pulse amplitudes arises from statistical variation in gain per stage in the multiplier, and from the fact that noise pulses may originate from any of the stages of the multiplier. Not only do pulses vary a great deal in amplitude for thermal electrons emitted from the photocathode, but they overlap pulses from thermal electrons emitted from the first dynode, which overlap in turn pulses from later dynodes. Thus, it is quite reasonable for the noise pulse rate to increase continuously all the way to zero bias. Of course, only noise pulses originating at the photocathode receive full amplification and are of primary importance.

For signal pulses the spread in amplitudes is similarly contributed to by statistical variations in amplification per stage, although for large signal pulses this factor is reduced by the larger number of electrons initially ejected from the photocathode. More important is the variation in position in the phosphor of the exciting event. While certain geometries with easily stopped particles may permit rather definite localization of the fluorescence, gamma or X rays may generate only a very few secondary electrons at widely scattered points. Some of the secondaries may escape from the phosphor before imparting much of their energy. The light produced in various locations will be collected with different efficiencies by the photocathode. If a relatively small number of photoelectrons is produced at that cathode, there will be of course a spread expected in this number. Upon these factors are superimposed variations in light produced by different energy particles when nonmonchromatic radiation is used. In this connection it should be remarked that, if the phosphor is thin, the light produced by penetrating particles tends to be *inversely* proportional to their energy.

Although a considerable spread in signal pulse amplitude is thus to be expected, it is not reasonable for the distribution curve to rise indefinitely as zero bias is approached. For one thing, there should be no signal pulses smaller than the range associated with individual first-electrode photoelec-

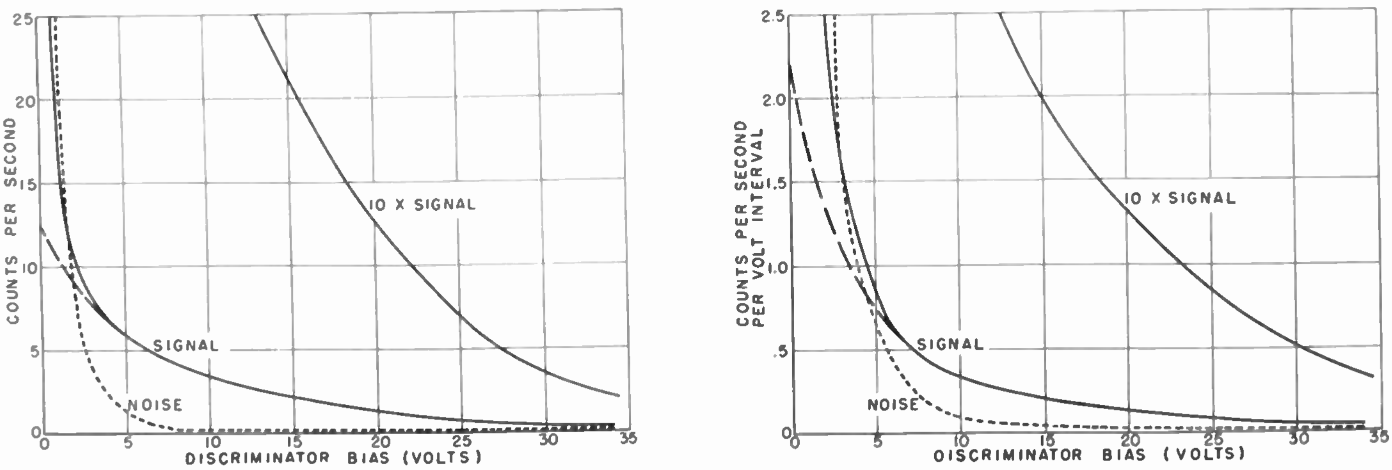


Fig. 18—(a) Sample integral bias curves, and (b) the corresponding pulse-height distribution curves.

trons. It has been definitely established that, in the case of the ZnS:Ag phosphor, this rise is due to multiple counting at low bias settings. The reason for this is apparent from oscillograms of the pulses. The individual photoelectron pulses which make up a signal pulse vary statistically in amplitude and position along the time axis. These component pulses are poorly integrated near the end of the signal pulse, where the average interval between photoelectrons begins to be an appreciable fraction of the time constant of the circuit. Thus, as the main pulse ends, the trace often bobs up and down one or more times. After this, there is an additional sprinkling of individual electrons which can sometimes be followed on the film with a degree of certainty for a millisecond or so. These photoelectrons are mostly late arrivals associated with the long period, low intensity phosphorescent persistence of the fluorescent screen. Thus multiple counting appears to arise in two ways; from poor integration at the end of the main pulse, and from late photoelectrons caused by long persistence in the phosphor.

Though ZnS is peculiar among the important phosphors in having a long period phosphorescence, multiple counting has also been observed with $CaWO_4$, which has an exponential decay. This is apparently entirely due to poor integration at the end of the main pulse. Such multiple counting is of course inherent in the nature of the detector; if resolving times are pushed close to the decay time of the phosphor, a certain amount of multiple counting will ensue. Multiple counting has also been observed to some extent with naphthalene and anthracene, even though these phosphors have apparently very short decay times.

It is evident from the above that under many conditions efficiency measurements are difficult to carry out with accuracy. With the localized fluorescence and very large pulse sizes obtained from alpha particles, the efficiency can be quite accurately measured; with less readily absorbed radiations which give smaller pulses, the count obtained will depend very much on the bias setting, and as this setting is reduced in an attempt to count smaller and smaller pulses, multiple counting may vitiate the results.

It appears in general, however, that the efficiency of the device is high. Deutsch[17] reports that, using naphthalene, substantially every electron is counted if it spends 0.15 Mev in the screen, when the tube is used at room temperature. At dry ice temperatures the minimum energy expenditure for which counting is highly efficient is reduced to 0.05 Mev. He also finds an efficiency of 20 per cent for 1.2 Mev gamma radiation, using naphthalene screens several centimeters thick. Most experimenters find that efficiencies are near 100 per cent for radiation which is effectively absorbed in the phosphor. Fast neutrons may be detected from the recoil protons produced in anthracene with an estimated efficiency of 10 per cent for one arrangement.[7] Using a zinc sulfide phosphor substantially 100 per cent efficiency is obtained for 8 kev X rays.[3] All of these measurements are rough, but indicate that with care very good efficiencies can be obtained for a wide variety of types of radiation.

## Comparison With Other Detectors

Under certain conditions the ionization chamber, the Geiger counter, the proportional counter, and the photomultiplier radiation detector may each approximate perfect efficiency of quantum detection, responding to all the information available in the incident radiation quanta. One may compare these detectors by considering the ease with which these conditions are obtained, the range of practical operation over which they apply, and the secondary characteristics which make one detector or another specially suited for certain applications.

In comparison with conventional detectors, the photomultiplier detector differs chiefly in that the primary detecting area may be quite small, but the absorption is relatively high. On the one hand, this restricts the new detector to narrow-beam applications if full sensitivity is to be realized. On the other hand, the small area of the fluorescent screen is an advantage in narrow-beam applications. It results in a greatly reduced sensitivity to general background radiation; the contribution of background radiation to the response of the new detector being so slight that it may be disregarded in most considerations. In its place, of course, one must consider the entirely different type of background arising from dark-current noise.

The new detector can hardly be compared with the ionization chamber, for the special advantages of the two instruments are entirely different. The ionization chamber tends to be much less stable and much more difficult to operate at its highest sensitivity. The photomultiplier detector, like the Geiger counter, is quite easily operated at full sensitivity. With the present phosphors the new detector does not share with the ionization chamber those fundamental properties which, according to the design of the chamber, permit measurements linearly related either to the absorbed radiation energy or to the roentgen units of incident radiation, over a wide range of radiation quantum energies. However, it is likely that linear dependence upon absorbed energy can be obtained with greater attention to the phosphor.

The photomultiplier detector is more easily compared with the proportional counter. Both detectors are capable of very high counting rates. There is a further similarity in that the pulse amplitude in the new detector tends to be dependent upon the type of radiation and the particle or quantum energy of the radiation. It is evident that the new detector may be used to discriminate between signal pulses which differ considerably in amplitude. The dependence of amplitude upon the energy of the radiation particle has not been investigated carefully. Other things being equal, the spread of pulse heights for particles of identical energy might be expected to be greater for the new detector than for the proportional counter. Perhaps cases peculiarly suited to discrimination by one detector or the other will be found. For high energy particles, however, complete absorption should be approximated better in the solid fluorescent screen than in the gas of the proportional counter. The photomultiplier detector tends to be more stable, more free of stray pickup, and less voltage sensitive than the proportional counter. Because of this the new detector would appear to be preferable for many applications not requiring strict proportionality of pulse amplitude to particle energy.

In many ways the photomultiplier detector is more nearly similar to the Geiger counter than any other conventional detector. Though technically limited by the decay time of the phosphor, the counting rate of the new detector effectively has almost no upper limit, inasmuch as integrated dc current measurements may be used with very high intensities. The speed with which measurements of a given accuracy can be completed with the new detector tends to be determined only by the available radiation intensity, rather than by an arbitrary upper limit of counting rate as in the Geiger counter. In addition to the ability to handle large amounts of radiation, the detector has an extremely high resolution which is determined by the decay characteristics of the phosphor. With certain phosphors, events separated by as little as 0.1 microsecond may be distinguished. The life time of the photomultiplier detector is much longer than that of a fast self-quenching Geiger counter, and the working characteristics appear to vary somewhat less from week to week. For penetrating radiations, the efficiency of the scintillation counter is superior to that of the Geiger counter, and for very soft radiations the absence of any window which might interfere with the measurement is a distinct advantage.

Perhaps the most promising feature of the new detector is the fact that the radiation is absorbed in a solid. Many of the coming advances in nuclear physics will arise from the study of the behavior of extremely energetic particles to be produced by the huge accelerators now being built. Since these particles are highly penetrating, the absorbing properties of the scintillation counter will place it at a considerable advantage with respect to other types of detectors. Thus one may expect to see a rapid development of this counter, together with a considerable widening of its field of application, in the near future.

# A Standard Diode for Electron-Tube Oxide-Coated Cathode-Core-Material Approval Tests*

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*Summary*—A diode has been designed and used for testing various samples of cathode material in several plants and laboratories during recent years. Several criteria have been used for evaluating the emissive power of the various materials tested. To simulate the usual space-charge-limited emission test commonly used on receiving tubes, a cathode-temperature versus emission characteristic has been taken on each test lot. Temperature-limited emission has been examined under both low-field, low-temperature conditions and normal-temperature, high-field conditions. Results indicate that this method has several important advantages over the present approved method.

## I. Introduction

THE ROLE played by the core material in the emission of electrons from oxide-coated nickel base cathodes has been the subject of considerable conjecture and controversy almost since the discovery of the oxide-coated cathode. Various hypotheses have been proposed from time to time to account for the apparent superiority of one core material over another. These hypotheses were not always in complete agreement, and attempts to resolve their differences frequently led to inconclusive results. The main difficulty seems to be in the complexity of the emission phenomena. Emission from oxide-coated cathodes is subject to time changes and fatigue effects, and to a very pronounced degree is dependent upon the composition, method of application, and processing of the materials used in the tube assembly. So far, it has been very difficult to isolate the single variable of core material for comprehensive studies.

Despite the fact that the mechanism of emission is not clearly understood, tube manufacturers have been producing, for many years, good oxide-coated-cathode tubes at relatively low cost. However, the possibility that a low-emission epidemic may suddenly shut down his production lines has caused every tube manufacturer to spend considerable time and money trying to immunize his product from all known carriers of this disease. Since the source of emission is the cathode, the material from which it is made obviously is one of the first places to look for trouble. For this reason, it has been standard practice for many years in radio-tube plants to check the performance of each new melt of nickel cathode material on one or more production tube types before accepting it for production use.

In the past, the type of tube chosen for such tests, the conditions under which the test was made, the acceptance criteria, and the interpretation of test results has varied widely from company to company, and even from division to division of the same company. Since it was obviously impossible for a company to test each new melt of nickel on all production tube types under all conceivable operating conditions, most manufacturers tested two or three tube types as representative of various classes of tube service, such as rectifiers, oscillators, and power-output types. Sample quantities of each new cathode-nickel melt in a tubing size to fit these types would be ordered with the fervent hope that the type chosen would be in production when the samples arrived. In the happy but infrequent event that all test results, both initial and life-test performance on the various types run for approval, indicated that the new melt was inferior to, equal to, or superior to the comparison melt, the course of action presented no problem. In the more usual case, when test results disagreed, it was sometimes necessary to resort to occult practices to effect a decision. It is not particularly surprising, therefore, that new melts of core material accepted by one company or division would be rejected by another.

Nickel for seamless-cathode electron-tube use is produced in melts of about 10,000 pounds. Prior to World War II, usage of cathode nickel in the electron-tube industry was such that it was possible for a cathode supplier to hold a melt of this size for the period of approximately three months required to obtain melt approval from the various tube manufacturers. However, this situation was at best only tolerable, as it involved some financial risk on the part of the cathode supplier. If all companies should reject the melt, its value depreciated to that of Commercial Grade A material, this loss being sustained by the cathode supplier, who would then channel this material into less critical usage. If some companies accepted and some rejected the melt (the normal situation), the cathode supplier would purchase another melt to submit to the companies rejecting the first melt, and so on until all companies were satisfied. This method required considerable juggling of melts on the part of the vendor, who assumed the risk of supplying the proper-size cathode cylinder of the approved melt to each customer.

By 1944, the usage of seamless cathode material had reached a new high, and, in addition, new applications for tubing, particularly in the aircraft industries, placed an added drain on the cathode vendor and the raw material supplier. It became increasingly difficult to supply

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each radio tube manufacturer with the particular nickel melt which met his melt approval requirements.

In January, 1945, a meeting was called by one of the cathode base material vendors to discuss this problem. Representatives of all cathode-tubing vendors and users were invited. It was the consensus of this group that the problem could best be studied through the medium of four committees, whose membership would be drawn from the various companies represented at this meeting. One committee was to propose referee methods for the chemical analysis and sampling of nickel melts. A second committee was to outline metallurgical test methods and a program for metallurgical development. A third committee was to prepare a specification for a standardized diode test to serve as a starting point for a laboratory test method. The fourth and last committee was to propose methods for presenting data on the multitube production test method of melt approval. These four committees were organized and later incorporated as Section A, Committee B4, Subcommittee VIII of the American Society for Testing Materials. This paper is a report on the work of the Diode Committee.

## II. The Standard-Diode Approval-Test Method

The primary purpose of the diode structure, as conceived by the committee members, was to provide a standard tool for the investigation of the various cathode-material melts supplied for production use. It was felt that by providing a single structure which could be run conveniently in every tube plant, many of the variables introduced by the previously mentioned multitube tests could be eliminated. A secondary purpose of the diode was to investigate the effect of variations in the additives in the nickel melt and variations in processing and testing methods. In deference to the primary purpose of the diode, that of melt approval, it was agreed that the structure should consist of standard receiving-tube parts, so that standard receiving-tube equipment and techniques could be employed in processing them, thus insuring that a melt accepted for production use on the basis of diode tests would have been tested under conditions prevalent in the factory in which the melt would be used. It was realized that a structure utilizing receiving-tube parts and techniques might limit the effectiveness of the diode as a sensitive instrument for the investigation of melt additives and process variations.

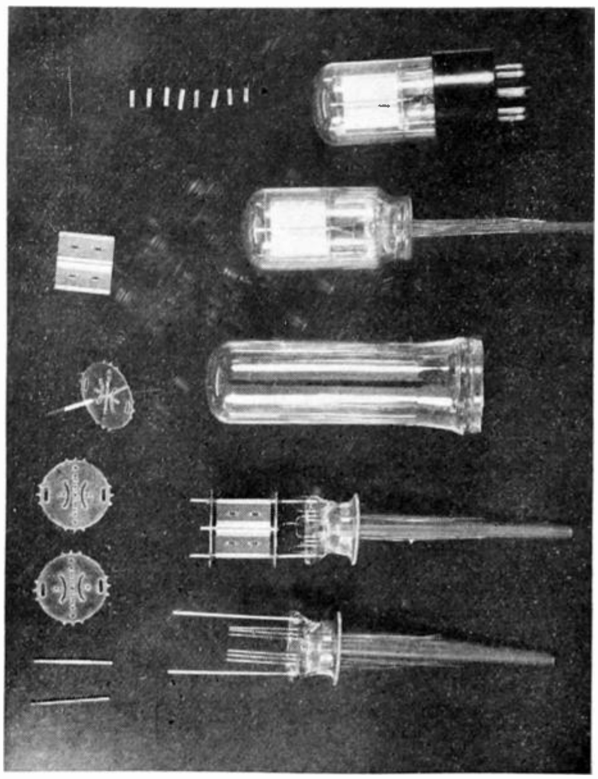

Fig. 1—Structure of the standard diode designed by the Diode Committee.

The structure designed by the Diode Committee is shown in Fig. 1. Table I lists the important tube dimensions. The heater is designed for 6.3-volt, 300-ma operation. Under these conditions, the cathode operating temperature is approximately 1050°K. The anode was designed with an oval barrel to permit the use of a grid, should the occasion arise.

TABLE I

Standard Diode—Construction Details

| | |
|---|---|
| *Anode* | |
| Material | 0.005 inch Grade A Nickel |
| Dimensions | Inside diameter oval cross section: 0.132 ×0.185 inches |
| | Barrel length: 0.75 inches |
| *Cathode Cylinder* | |
| Material | Electronic Grade A seamless nickel tubing |
| Dimensions | Outside diameter: 0.045 inch |
| | Length: 1.06 inches |
| *Cathode Coating* | |
| Material | ASTM barium-strontium double-coprecipitated carbonates |
| Dimensions | Coated length: 0.475 inch (centrally located on cylinder) |
| | Coating weight: 3.40 ±0.4 milligrams |
| *Stem* | |
| Material | Lead glass (G12) |
| | Nickel-dumet-copper welds |
| *Getter* | |
| Material | Barium-cored iron wire |
| *Heater* | |
| Material | Cleaned and straightened tungsten wire |
| Dimensions | Wire weight: 12.0 milligrams ±1½ per cent per 200 mm |
| | Wire cut length: 7.05 ±0.04 inch |
| Form | Six-leg folded |
| Insulation Coating | Alundum |
| *Bulb* | Lime glass; T9 |
| *Base* | Standard octal |
| *Spacers* | Mica 0.011 ±0.001 inch thick |

The diode melt-testing method has several advantages over the multitube test method. It is a simple structure, easily produced on equipment available in all tube plants. Standard parts have been made available to all companies interested in diode testing, so that, regardless of where the tube is produced, uniformity of parts is assured. It is necessary for the cathode-material vendor to produce only one size of cathode cylinder for melt-approval use if all testing is done on the diode. A comparison melt has been set aside for diodes which provides a uniform standard of comparison. Standardization of design, processing techniques, test methods and controls are possible, and changes found to improve the accuracy or reliability of the test method by one laboratory can be applied readily in other laboratories.

The general procedure used in conducting melt-approval tests on the diode is as follows: Test lots, as a rule, are run in groups consisting of at least ten "control" diodes employing the standard control-melt cathodes (Melt 66), and at least ten each of diodes employing cathodes of the melt or melts under test. All parts other than cathodes are obtained from a common inventory. Identical parts-processing and tube-processing schedules are used on all lots within the group. In general, these processes are standardized for a particular plant; i.e., all diode-test lots in one plant are processed under the same schedules. Some of these processes have been proposed as industry-wide diode standards by the Chemical Committee; e.g., parts cleaning, cathode-coating preparation, and certain testing schedules. Each plant has been allowed to set its own exhaust and aging schedules, because the approval of melts on the basis of diode tests implies the acceptance of the melt for production use, and the tube manufacturer is most interested in knowing how the melt will perform under his particular plant conditions.

## III. Testing of the Standard Diode

To evaluate the emissive power of a cathode, it was assumed by the Diode Committee that some method of observing its total emission current (not limited by space charge) was necessary. If one attempts to test the diode structure, or any receiving-tube structure, for total or temperature-limited emission current, at its normal cathode temperature under static conditions, the rated wattage dissipation of the anode is usually exceeded long before temperature-limited current is drawn. Anode overheating usually results in gas evolution and subsequent cathode poisoning.

To avoid this difficulty, tubes may be tested for total emission by reducing the cathode temperature so that the temperature-limited current drawn does not exceed the rated anode dissipation. With this in mind, the Diode Committee fixed the anode potential for emission test at 40 volts, which, at normal cathode temperature, would give a space-charge-limited emission current in the neighborhood of 50 ma in the diode structure. This corresponds to a current density of about 120 ma per square centimeter, which is somewhat higher than is required by conventional emission tests on receiving-tube types. The full cathode-temperature versus emission-current curve at this fixed 40-volt anode potential was then taken on each tube under test and the position of the "knee" of the curve (that is, the transition point between temperature-limited emission and space-charge-limited emission) was taken as a measure of the relative emission capabilities of the cathode. Fig. 2 shows this curve on a normal tube and on a tube deficient in emissive power. It will be noted that, at rated cathode temperature, both tubes have the same space-charge-limited emission, but that on tube B the temperature-limited emission is somewhat lower than on tube A.

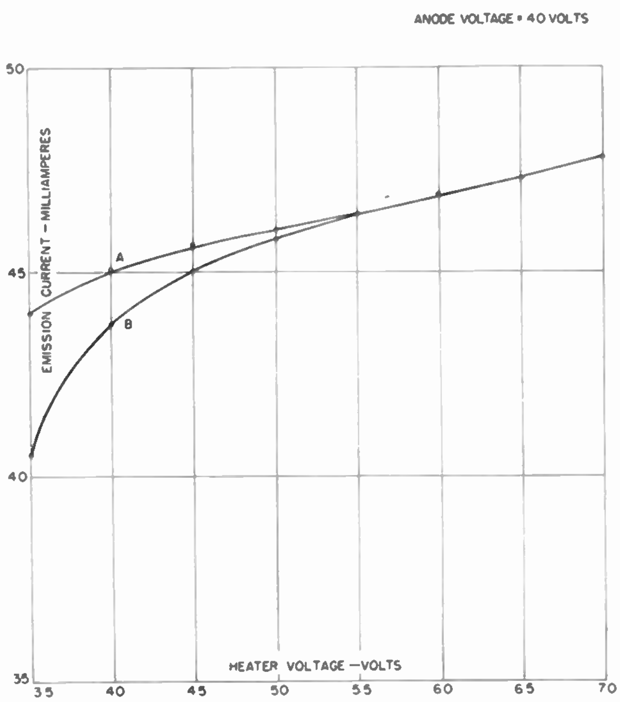


Fig. 2—Emission-current versus heater-voltage characteristic, American Society for Tetsing materials standard diode.

It was this sort of data that the Diode Committee set out to obtain on each new melt of cathode nickel submitted for test. Practically all conclusions reached to date in this program are based on cathode performance under this type of test. However, it was found that this cathode-temperature versus emission test has at least one serious limitation; the emission currents observed under these conditions at very low cathode temperatures were apt to be unstable. This may be due to the relatively high anode potential used on the test. It is well known that certain oxides and chlorides, which are on the surface of the anode as a result of the cleaning and processing of the tube parts and structure, tend to break down when the anode voltage exceeds about 4.5 volts. Oxygen or chlorine are released, and may poison the cathode surface. This effect should be particularly noticeable at low cathode temperatures, where the rate of barium production within the cathode is low. If the rate of poisoning exceeds the rate of barium production, slumping emission will result.

To avoid this difficulty, data have recently been taken under low field conditions, with an anode potential of 4.0 volts, which should be safely below the breakdown voltage of the oxides and chlorides likely to be found on the anode surface. To obtain temperature-limited emission at this low anode voltage on the diode, it was necessary to lower the heater input voltage to 1.75 volts, which corresponds to a cathode temperature of about 640°K. Emission currents obtained under these conditions are relatively stable.

During the late war, there was considerable interest in the pulse-emission properties of cathodes, and, since many receiving tube types were being used in pulsed applications, the Diode Committee thought it desirable to examine new cathode melts for their pulsed-emission properties. Peak emissions as of high as 100 amperes per square centimeter were reported on nickel-base cathodes by observers at the MIT Radiation Laboratory,[1] the Bartol Foundation,[2] and elsewhere. The pulsed-emission testing techniques developed by these laboratories were borrowed and utilized for diode pulse testing by one of the companies involved in the diode program. The pulse technique has one significant advantage over the two static emission tests mentioned previously, in that it permits examination of temperature-limited emission at rated cathode temperatures without exceeding the rated anode dissipation of the diode.

### *Life Tests*

The electron-tube manufacturer is interested not only in the initial performance of a cathode melt, but also in the performance against time under operating conditions. The Diode Committee has incorporated a life test as part of the melt approval procedure. The conditions of this test have been modified as data were accumulated under various drain conditions. The present life test schedule consists of a minimum of 500 hours operation at rated heater voltage ($Ef=6.3$ volts) and a drain of approximately 100 ma per square centimeter.

## IV. Diode-Test Results

From the time of its inception, the Diode Committee has been conducting diode tests of three general classes:

1. Tests directed toward improving the structure and standardizing processing and testing procedures
2. Melt-approval tests on commercial cathode nickel
3. Tests directed toward examining the effects of variations in chemical composition of the cathode nickel.

Tests in the first class have been and are being conducted in considerable volume. Results from this class of tests are the basis for the formulation of diode structure and process specifications.

Melt-approval tests are represented in the summary in Table II, by the compilation of test results on commercial melts 70 through 76.

[1] E. A. Coomes, "Pulsed properties of oxide cathodes," *Jour. Appl. Phys.*, vol. 17, pp. 647–654; August, 1946.
[2] M. A. Pomerantz, "Magnetron cathodes," Proc. I.R.E., vol. 34, pp. 903–910; November, 1946.

Melt-composition tests have been conducted on about fifty different nickel-base materials. Test results on only a few of these lots are shown in Table II, for comparison with the 220-alloy commercial melts. The test results reported here represent only a very small portion of the total diode tests run during the three-year period of diode investigation, but serve to illustrate the effectiveness of this structure in differentiating between the emission capabilities of cathodes as a function of core material. The data reported are based solely on the information obtained from the 40-volt static characteristic curves (see Section III). Low-field-emission test, pulsed-emission test, and life-test results are not included in this report, because these data are still incomplete. Since it is difficult to characterize the 40-volt static characteristic test data by a complete heater-voltage versus emission-current curve, a "figure of merit" rating based on the difference in slope of these curves in the region between heater 4.5 and 6.5 volts has been substituted. This "figure of merit" was determined in the following manner:

$$\text{F.M.} = \left[\frac{I_{S6.5} - I_{S4.5}}{2}\right]_{\text{control}} - \left[\frac{I_{S6.5} - I_{S4.5}}{2}\right]_{\text{test}}$$

where

$I_{S6.5}$ = emission current with 6.5 volts applied to heater
$I_{S4.5}$ = emission current with 4.5 volts applied to heater.

Thus, for example, a "figure of merit" of −0.5 for melt 70 (Company "2") indicates that the slope of the $Ef$ versus $I_s$ curve for the average of the ten tubes made with melt 70 cathode material was 0.5 ma/volt steeper in the region between 4.5 and 6.5 volts heater voltage than its control lot (melt 66).

## V. Discussion of Diode Test Results

The seven melts, 70 through 76, whose "figure of merit" ratings are given in Table II, are 220-alloy seamless tubing material submitted to the radio tube indus-

TABLE II

Summary of Initial Test Results on Diode Tests of Commercial Cathode Material Melts Seamless Tubing

"Normal" 220 Alloy Material Figure-of-Merit Rating

| Melt Number | Company "1" | Company "2" | Company "3" | Company "4" | Company "5" | Company "6" | Test Description |
|---|---|---|---|---|---|---|---|
| 70 | | *−0.5 | −2.1 | −0.3 | +0.7 | | |
| 71 | | −1.8 | −1.0 −1.9 | −1.9 | +0.1 | | Initial sample |
| 71C | | +0.1 | −0.2 | | | | Initial sample cleaned |
| 71R | +0.5 | −0.1 | +0.1 | −0.3 | +0.5 | | Second sample, melt 71 |
| 72 | −0.4 | −0.4 | −0.2 | | −0.1 | −0.4 | |
| 73 | −0.8 | −0.1 | −0.2 | | −0.1 | −0.2 | |
| 74 | −0.5 | −0.3 | −0.4 | | −0.1 | +0.5 | |
| 75 | −0.5 | −0.2 | −0.5 | | 0 | −0.1 | |
| 76 | −0.2 | 0 | +0.4 | | +0.1 | −0.3 | |
| *Lockseam Tubing* | | | | | | | |
| W | −9.6 | −7.5 | −15.9 | | | | "Passive" material, 499 alloy |
| K | −10.0 | −11.4 | −15.4 | | | | "Passive" material, 707 alloy |
| X | −0.2 | +0.1 | −1.3 | | | | "Active" material, 599 alloy |
| M | −0.4 | 0 | −1.3 | | | | "Active" material, 766 alloy |

try for approval testing. Most of the data obtained by the Diode Committee so far has been on this alloy, although approval tests have also been conducted on other alloys in lockseam tubing. For comparative purposes, "figures of merit" are shown in Table II for the three grades of cathode tubing commonly employed by the receiving-tube industry. These three grades are termed "normal," "active," and "passive" alloy base materials by the ASTM.[3] This classification refers to the relative barium-producing potential of the alloy, as determined by the percentage of reducing agent present in the melt. "Active" alloys contain a relatively high percentage of reducing agents, such as silicon and magnesium; "passive" alloys contain relatively low percentages of reducing agents; and "normal" alloys contain a moderate amount of reducing agents. Typical chemical analyses of these three grades of material are listed in Table III.

TABLE III

TYPICAL CHEMICAL COMPOSITIONS OF "NORMAL," "ACTIVE" AND "PASSIVE" CATHODE BASE MATERIALS†

| | "Normal" | "Active" | "Passive" |
|---|---|---|---|
| Carbon | 0.15%* | 0.12%* | 0.10%* |
| Copper | 0.20 | 0.04 | 0.04 |
| Iron | 0.20 | 0.05–0.01 | 0.05 |
| Magnesium | 0.01–0.1 | 0.0 –0.15 | 0 |
| Manganese | 0.20 | 0.10 | 0.02 |
| Sulphur | 0.008 | 0.005 | 0.005 |
| Silicon | 0.01–0.05 | 0.15–0.25 | 0.01 |
| Nickel and Cobalt | 99.0 min. | Balance | Balance |

† All figures are percentages by weights. Compositions are maxima unless a range is indicated.

* Carbon content is generally lower as wall thickness is reduced through successive annealings. At 0.0022 inch wall. carbon is 0.02 per cent nominal, while at 0.005 inch wall, carbon may exceed 0.05 per cent nominal.

Certain minor constituents may be found in varying amounts according to melting practice for the alloys. Such elements may be aluminum, titanium, boron, etc.

The differences observed in "figure-of-merit" ratings for melts 70 through 76 as reported by the six companies reporting diode-test results were considered by the diode committee to be random variations within the sensitivity of the test method, with one exception, the initial sample of melt 71. It was felt that some significance could be attached to the fact that three of the four companies reporting diode results on this melt found it somewhat inferior to the control melt 66. Company "3" repeated the test on their initial sample, and again found it poor. Investigation showed that the three companies reporting poor results on the initial sample had obtained their melt-approval samples from one run of cathode tubing, while the company reporting good results, Company "5," on the initial sample had obtained its melt-approval sample from a later drawing of the material. Surface contamination of the initial sample was immediately suspected. Companies "2" and "3" therefore tried surface cleaning of their initial sample. The results of these tests, indicated as melt 71C on Table II, indicated that this suspicion might be well founded. Company "3" found a definite improvement in emission performance from the suspected melt after etching the bare sleeves. Company "2," employing a different cleaning procedure, found the performance of this suspected lot of tubing could be brought above that of the control melt by this means. The original sample processed along with this test uncleaned, was still very poor.

[3] A specification covering Radio Tube Core Material classifications is now being prepared by Section A, Committee B4, Subcommittee 8, ASTM.

As a result of these findings, the cathode-tubing supplier prepared a new sample of melt 71, with special care being taken to avoid all possible contaminants in the drawing operations. These samples were submitted to the same companies who had reported on this melt previously. Results on this second sample are shown in Table II as melt 71R. All companies testing this sample found it better in emission performance than its control melt 66. To substantiate the conclusion that the initial sample of melt 71 was bad, two comparative tests were run later by Company "2" on a production tube type under normal factory conditions. In both cases, a high initial shrinkage due to poor emission was obtained on the initial sample, whereas normal results were obtained on the comparison melt run at the same time.

Results on a few melts of "active" and "passive" alloys are also tabulated on Table II. Melts W and K show the effect on emission of reduction in silicon and magnesium content of the base material. The "active" alloy tests X and M show performance characteristics very similar to the "normal" alloy tests. It should be borne in mind however, that these lots represent results on only single samples of these alloys, and should not be not be taken as necessarily representative of the average emission performance characteristics of these materials.

## VI. CONCLUSIONS

An inventory of the results achieved in the three years of diode testing shows a very favorable balance. Correlation between test results reported by the various plants testing cathode-core-material melts has shown marked improvement with diode testing. The study of testing methods has brought about a better understanding of the effects of tube design and processing upon emission. This has resulted in improved tube-to-tube and test-to-test uniformity. Work on processing, testing, and materials standards has been started, and in most cases completed. The effects on emission of various additives in core materials, and the significance of modifications in core material melting, drawing, and cleaning methods has been evaluated. In some cases, the cause of poor activity in a particular test sample has been isolated.

Full understanding of the emission problem still seems remote. However, because active interest in the problem has been aroused and crystallized into a joint effort, the immediate goal, that of a satisfactory melt-approval test method, now seems attainable.

# Testing Cathode Materials in Factory Production*

J. T. ACKER†

*Summary*—**The paper deals with the methods of testing radio-tube cathode materials in factor production, and especially with a comparison of several specific lots of materials of variable content. It is believed that this is the first time the electron-tube industry has made mass tests on a well-controlled engineering basis of cathode materials which vary in singe component elements.**

THE CHAOTIC conditions under which the manufacturers of electron tubes have been testing cathode-nickel tubing for their use have been described by McCormack.[1] These costly tests, together with prove-in tests on other components of electron tubes, such as alkaline-earth carbonates, getters, and anodes, are performed to insure good manufacturing yields. The lack of fundamental knowledge concerning thermionic emission makes these acceptance tests necessary.

Each company in the industry accepts or rejects a new melt on the basis of the final test results and life performance of electron tubes containing cathodes made from the new melt, as compared with tubes being currently made using cathodes from the previously accepted melt. The type of tube used for this comparison is usually one of those in current production. The results of these tests, as reported by the entire industry to the suppliers of cathode tubing, were contradictory and full of inconsistencies. Through one of the supplier's efforts, a meeting of the representatives of the tube manufacturers was called in January, 1945, to discuss the situation and try to formulate better methods of evaluating cathode nickel tubing. A section of Subcommittee VIII of Committee B4 of the American Society for Testing Materials, was formed for this purpose, and possibly to improve the understanding of thermionic-emission phenomena and the type of nickel used.

The intent of this paper is to report on the progress made by the so-called "Data Subsection" in co-ordinating the methods of test and interpretation of results so that an industry-wide evaluation of each melt may be made to compare with evaluations based on the use of simple and much cheaper diode structures and ultimately chemical and/or metallurgical tests on the nickel.

One of the first steps taken to reduce the factory prove-in tests to a common denominator was to establish a standard melt of cathode nickel against which all tube manufacturers would rate new melts. One of the major suppliers of cathode tubing agreed to set aside for this purpose a 200-pound portion of melt 66, which is a 220-grade nickel melt of typical composition and normal 10,000-pound size, and which had given apparently normal test results in electron tubes. This quantity of material is sufficient to last for many years for melt-approval control purposes exclusively. Sample cathodes of appropriate size from this melt are sent along with the same size cathodes from a new melt to each factory when a new melt is to be tested for approval. The prove-in factory runs are made on cathodes from the new melt and melt 66 simultaneously and under identical processing conditions.

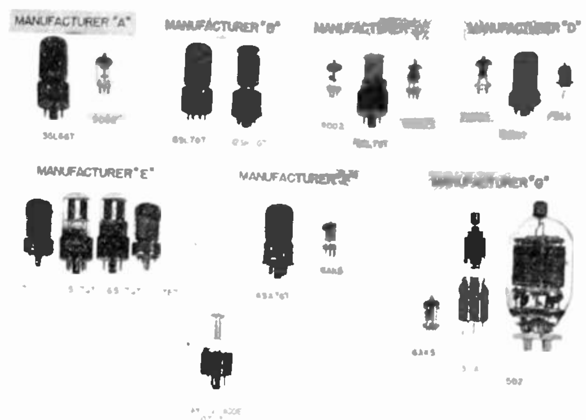


Fig. 1—Variety of electron tubes formerly used in cathode testing.

A second step in the direction of greater uniformity of test conditions was to strive to have the participating companies choose similar types of tubes for prove-in factory runs. It had been the practice to use any type of tube that happened to be in production at the time of the melt-approval test. The accompanying Fig. 1 shows the great variety of types of tubes formerly used by the industry for this purpose. Fig. 2 shows the pentodes that are currently used in these tests.

The committee found that the industry was evaluating the melts on the basis of initial shrinkage, initial tube characteristics, and life-performance tests. A study was made of the early test results in order to formulate a pattern or form that could be used by each company in reporting its test results, regardless of the parameters, for purposes of evaluation.

Factory prove-in runs of 50 to 200 tubes each are made using cathodes from the test melt and the standard melt 66. Variations in geometry and processing are

* Decimal classification: R720. Original manuscript received by the Institute, May 25, 1948. Presented, 1948 IRE National Convention, March 24, 1948, New York, N. Y.

† Western Electric Company, Allentown, Pa.

[1] R. L. McCormack, "A standard diode for electron-tube oxide-coated cathode-core-material approval tests," Proc. I.R.E., this issue, pp. 683–687.

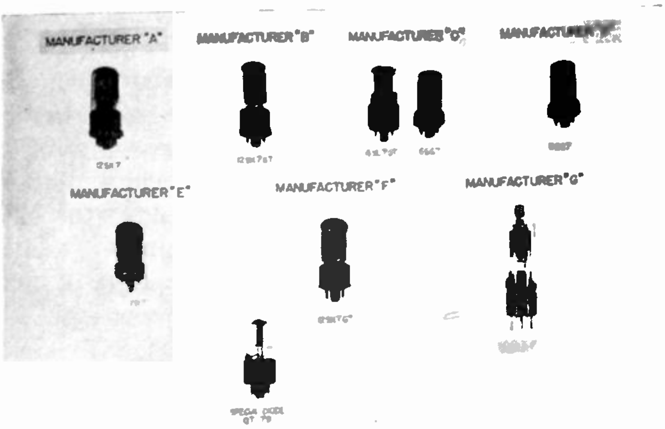


Fig. 2—Electron tubes used in recent cathode testing.

minimized by simultaneous manufacture under shop conditions. The shrinkage is counted only for causes related to the cathode. The initial tube characteristics are taken either on all of the tubes or on a representative sample of the run. Life tests are carried out for 500 hours on 5 to 10 tubes selected at random from the lots. In each category the comparison is expressed as a ratio of results for the test melt to the control melt. In order to avoid zero values in the shrinkage ratios, the per cent yield rather than the per cent shrinkage is used. It has been agreed that shrinkage and life are more important than initial characteristics, so that, in arriving at the over-all figure of merit, they are rated as twice as important. The sum of all the ratios so adjusted gives the over-all figure of merit for the industry. Figs. 3, 4, and 5 show the results obtained by eight plants, on melts 72 to 76 inclusive, when compared with melt 66. Melt 72 is a normal 220-grade melt of seamless tubing, while melts 73 to 76, inclusive, were made from a single 10,000-pound melt of nickel, half of which was poured normally with additional silicon added to the second half before pouring. Each half was split between seamless and weldrawn tubing as shown in Table I.

**INITIAL CHARACTERISTIC DATA**

| MELT NO. | MANUFACTURER A | B | C | | D | E | F | G | POINT TOTAL |
|---|---|---|---|---|---|---|---|---|---|
| | 12SK7gt | 12SK7 | 6SL7gt | 6SS7 | 6SS7 | 7B7 | 12SK7gt | 311A | |
| | AV. OF Gm 10 TUBES | AV Ia + AV. Gm OF 5 TUBES FIG. USED HERE IS AV OF AV Ia AND AV. Gm | 72-74: AV Gm OF 20 TUBES 75: AV Gm OF 10 TUBES | 72-74: AV. Gm AT 5.5 Ef OF 10 TUBES 75,76 AV Gm OF 32 TUBES AND 10 TUBES RESP. | 72,74,76: AV. Gm OF 8 TUBES 73,75 AV. Ia OF 8 TUBES | 72-76 AV. Gm OF 5 TUBES | AV Gm + AV. Ia OF 10 TUBES 672-576 FIG. USED HERE IS AV. OF AV. Gm + AV. Ia | AV Gm OF 72-64 TUBES 73-47, 74-51, 75-53, 76-35 | NOT WEIGHTED |
| 72 | 1.03 | 0.990 | 0.985 | 0.97 | 1.025 | 1.02 | 0.963 | 0.968 | 7.979 |
| 73 | 1.01 | 1.068 | 1.01 | 0.98 | 0.919 | 0.93 | 0.972 | 0.994 | 7.800 |
| 74 | 1.03 | 1.038 | 1.05 | 0.94 | 1.013 | 0.96 | 0.942 | 0.969 | 7.942 |
| 75 | 1.00 | 1.009 | 0.988 | 0.97 | 0.900 | 1.00 | 1.02 | 0.958 | 7.863 |
| 76 | 0.99 | 1.061 | 0.966 | 1.00 | 0.953 | 0.99 | 1.008 | 0.978 | 7.933 |

Fig. 4—Factory prove-in tests on cathode nickel melts.

TABLE I

| | Normal Half—0.02 per cent Si | Special Half—0.09 per cent Si |
|---|---|---|
| Seamless | Melt 73 | Melt 74 |
| Weldrawn | Melt 75 | Melt 76 |

It will be noted from Fig. 5 that the over-all figure of merit would be 40 for a melt of identical test results with melt 66. The spread of these five melts is not very large—only 2.9 per cent. This is not surprising, in view of the fact that these melts are all essentially identical with the exception of the variation in silicon content.

**INITIAL SHRINKAGE DATA**

| MELT NO | MANUFACTURER A | B | C | | D | E | F | G | POINT TOTAL |
|---|---|---|---|---|---|---|---|---|---|
| | 12SK7gt | 12SK7 | 6SL7gt | 6SS7 | 6SS7 | 7B7 | 12SK7 | 311A | |
| | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | 66, 72-74 RUN AS UNIT 66,75,76 RUN AS UNIT | WEIGHTED BY FACTOR OF 2 |
| 72 | 1.00 | 1.02 | 1.00 | 1.00 | 1.092 | 1.00 | 1.00 | 0.912 | 16.048 |
| 73 | 1.00 | 0.96 | 1.00 | 1.00 | 1.18 | 1.00 | 0.80 | 0.982 | 15.888 |
| 74 | 1.00 | 1.046 | 0.949 | 0.91 | 1.092 | 0.96 | 0.93 | 0.948 | 15.658 |
| 75 | 1.00 | 1.019 | 0.994 | 0.99 | 1.00 | 0.96 | 1.00 | 1.048 | 15.950 |
| 76 | 1.00 | 1.020 | 1.06 | 1.00 | 1.00 | 0.96 | 1.00 | 0.917 | 15.894 |

Fig. 3—Factory prove-in tests on cathode nickel melts.

**LIFE TEST DATA**

| MELT NO | MANUFACTURER A | B | C | | D | E | F | G | POINT TOTAL | GRAND TOTAL |
|---|---|---|---|---|---|---|---|---|---|---|
| | 12SK7gt | 12SK7 | 6SL7gt | 6SS7 | 6SS7 | 7B7 | 12SK7gt | 311A | | |
| | AV. Gm OF 5 TUBES 504 HR. LIFE | AV. Gm OF 5 TUBES 220 HR LIFE | AV. Gm OF 10 TUBES 72-74 AT 500 HR. LIFE | AV Gm 5.5 Ef OF 5 TUBES 72-74 AV Gm AT 6.3 Ef OF 10 TUBES 75,76 | AV. Gm OF 5 TUBES 500 HR LIFE | AV Gm OF 5 TUBES 500 HR LIFE | AV Gm OF 5 TUBES, 500 HR. LIFE 72, 73,74 RATIO CALC AT 300 HR PT | AV. Gm OF 10 TUBES 1,000 HR. LIFE 75, 76 RATIO CALC AT 500 HR. | WEIGHTED BY FACTOR OF 2 | |
| 72 | 1.08 | 1.000 | 1.11 | 1.06 | 0.978 | 1.03 | 1.00 | 1.013 | 16.488 | 40.513 |
| 73 | 1.01 | 0.973 | 1.163 | 1.04 | 0.981 | 1.07 | 1.008 | 1.006 | 16.502 | 40.188 |
| 74 | 1.02 | 0.970 | 1.11 | 1.04 | 0.989 | 0.989 | 0.824 | 1.005 | 15.870 | 38.470 |
| 75 | 1.02 | 0.989 | 1.08 | 1.18 | 1.012 | 1.04 | 0.998 | 0.997 | 16.472 | 40.308 |
| 76 | 0.98 | 0.932 | 1.07 | 1.180 | 1.032 | 0.960 | 0.997 | 0.978 | 16.126 | 38.793 |

Fig. 5—Factory prove-in tests on cathode nickel melts.

However, the individual variations in each melt among the various companies are still very large. The shrinkage results are +11.6 per cent and −18.3 per cent from the standard, the tube characteristics vary +8.5 per cent and −18.1 per cent from the standard, and the life results are +16.3 per cent and −17.6 per cent from the standard. There still are instances, as there were formerly, before this work started, where one company gets the best results for a given melt in a particular comparison, and another company finds it poorest. The maximum variations for a given melt are of the order of 25 per cent. Despite these difficulties, the over-all figure of merit seems to give a true picture of the quality of the cathode nickel. This is substantiated by the fact that the melt which shows up to be the worst was the only one of this series rejected by more than one company.

It is recognized that the cathode sleeve is, perhaps, of secondary importance to the coating applied to the cathode and the processing of the tube. That is why it is so important to keep these and other factors constant when making cathode-nickel tests. In the course of th work it has been found also that surface contamination of the cathode sleeve may mask the effect of the base-metal composition on thermionic emission.

It is felt that this work has put melt-approval tests on at least a semiquantitative basis, and that a melt substantially different from the usual variety will show a figure of merit higher or lower by several per cent. An instruction manual for factory testing by this method is being prepared in considerable detail so that even better control of factory tests will be made in the future, and the results are expected to be more reliable.

# A Method of Measurement of the Internal Series Resistance of a Capacitor Under Surge Conditions*

BEN S. MELTON†, SENIOR MEMBER, IRE

*Summary*—**Recent application of capacitors as energy-storage devices for low-impedance loads, such as short-duration light sources and electric detonators, has emphasized the importance of examination of capacitor efficiency from this standpoint. Ballistic measurements, using a vacuum thermocouple and galvanometer, are shown to afford the desired evaluation.**

## Theory

Recent application of capacitors to spark photography, radar, and the firing of low-resistance electric detonators has involved their use as energy-storage devices in connection with circuits of very low impedance. Consequently, it becomes pertinent to inquire about the effectiveness of the capacitor in delivering its stored energy under these conditions.

Evaluation of the low internal series resistance in question, not readily made by the usual bridge methods, is important in application of the capacitor to some low-impedance surge circuit, as will be shown below.

The energy stored in a capacitor is

$$J = \frac{CE^2}{2}$$

where $J$ is the work or energy in joules, $C$ is the capacitance in farads, and $E$ the potential in volts. It is convenient to discuss the percentage of this energy made available to the external load as a function of the load resistance and the internal series resistance of the capacitor. If we assume that the capacitor and load are represented by the simple circuit of Fig. 1, showing the capacitor

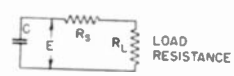


Fig. 1

of capacitance $C$, charged to a potential $E$ and having an internal series resistance $R_S$, connected to a load resistance $R_L$, the expression for the energy transfer must be

$$J = \frac{CE^2}{2} = R_S \int_0^\infty i^2 dt + R_L \int_0^\infty i^2 dt. \tag{1}$$

The first and second terms on the right-hand side of this expression represent the work converted into heat in the resistors $R_S$ and $R_L$, which can be represented by $J_S$ and $J_L$, respectively, so that

$$J = \tfrac{1}{2}CE^2 = J_S + J_L. \tag{2}$$

As the energy in the series circuit divided proportionally to the resistances,

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$$\frac{J_S}{J_L} = \frac{R_S}{R_L}. \qquad (3)$$

Then, combining (2) and (3),

$$J_L = \frac{J}{1 + \dfrac{R_S}{R_L}}. \qquad (4)$$

Thus, the energy delivered to the load can be expressed as the fraction of the energy of the charged capacitor by evaluating the fraction

$$\frac{1}{1 + \dfrac{R_S}{R_L}}.$$

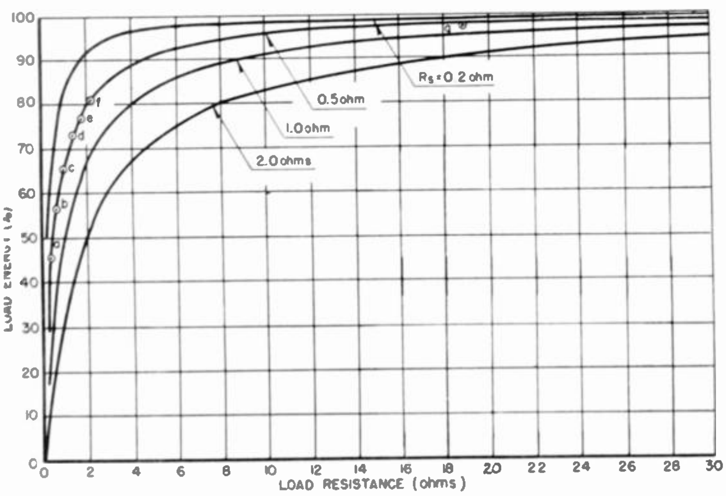


Fig. 2—Load energy versus load resistance, expressed as per cent of $\frac{1}{2}CE^2$ for several values of $R_S$.

Theoretical values have been obtained for capacitors whose internal series resistance $R_S$ is 0.2 ohm, 0.5 ohm. 1.0 ohm, and 2.0 ohms, and the resulting theoretical curves are plotted in Fig. 2. Of these curves, the one for $R_S = 1.0$ ohm can be considered as basic, so that, if the load resistance $R_L$ is expressed as a factor times $R_S$, the per cent of charge energy to the load can be derived.

Measurement of the internal series resistance can be made by varying the load resistance $R_L$ over a range of values, noting the corresponding variation of energy delivered to the load, and fitting the values thus obtained to the nearest theoretical curve. However, since the equations so far introduced contain only first-order terms, the theoretical curves may be replaced by straight lines and the computations correspondingly simplified.

Rewriting (4) as

$$\frac{J}{J_L} = 1 + R_S \frac{1}{R_L} \qquad (4')$$

and plotting $J/J_L$ against $1/R_L$ for various values of $R_S$, we obtain the straight solid lines of Fig. 3, which represent the same four cases shown by the curves of Fig. 2.

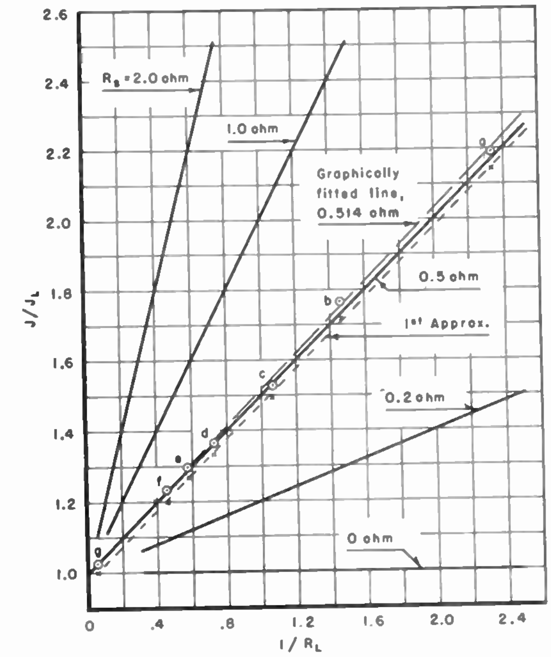


Fig. 3—$J/J_L$ versus $1/R_L$ for several values of $R_S$.

## Experimental Arrangement and Analysis of Reading Method

Since only total energies are involved in the foregoing analysis, some method of measurement of these energies is indicated, and it has been found that a simple ballistic method is satisfactory, provided adequate precautions are taken. Fig. 4 shows a circuit arrangement

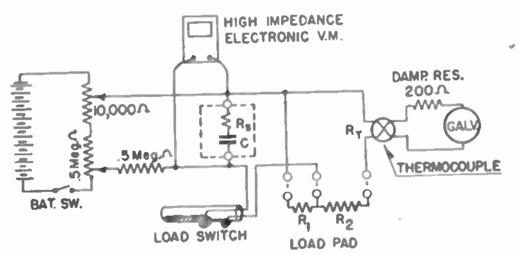


Fig. 4

which includes a small vacuum thermocouple and long-period galvanometer to measure the energy delivered to the load; a "load pad," which consists of a plug-in resistor combination; and a "load switch," whose function is to discharge the capacitor into the load when a galvanometer reading is to be taken. It is to be noted that the capacitor-charging circuit is of sufficiently high impedance so that no measurable current will be delivered directly from the battery to the thermocouple.

Resistors $R_1$ and $R_2$ of the load pad shown in the circuit, together with the heater resistance $R_T$ of the thermocouple, constitute the load resistance $R_L$ previously discussed, the relationship being

$$\frac{1}{R_L} = \frac{R_1 + R_2 + R_T}{R_1(R_2 + R_T)} \qquad (5)$$

where the reciprocal form is obtained for convenience in using the chart of Fig. 3.

In practice, because of the number of variables involved, it was found most convenient to proportion the values of $R_1$ and $R_2$ by setting up the experiment on a trial basis, reading galvanometer deflections as obtained from a 1-$\mu$f mica capacitor having a 1-ohm resistor connected in series. Resistance values were adjusted by trial until the galvanometer swings were of the same order. This precaution was taken to avoid errors due to nonlinearity of the thermocouple voltage with temperature. In order to translate the galvanometer scale reading into a value which can be used in conjunction with the curves of Fig. 2 or Fig. 3, it must be multiplied by a factor which expresses the ratio of the energy taken by the thermocouple heater to that of the entire load $J_L$. On the basis of instantaneous power and total energy considerations, it can be shown that

$$J_L = kD \frac{\left(1 + \frac{R_2}{R_T}\right)(R_1 + R_2 + R_T)}{R_1} \qquad (6)$$

where $D$ is the galvanometer deflection and $k$ is some calibration constant which need not be evaluated in the practice of this method. For convenience, we can now call the fraction in (6) the "load energy factor," writing

$$\text{load energy factor} = \frac{\left(1 + \frac{R_2}{R_T}\right)(R_1 + R_2 + R_T)}{R_1}.$$

## Computational Procedure

The computations are divided into two parts. The first of these covers the factors determined by the values of the load pads, and need not be repeated for other capacitors unless other load pads are constructed. The second part of the procedure is performed quickly by using successive approximations.

Table I includes sample calculations made on a 0.12-$\mu$f General Electric Pyranol capacitor (Cat. 26F699) rated at 15,000 volts dc. The first six columns may be set up once and for all when the load pads have been constructed. The values of $R_S$ sought for a particular capacitor are finally obtained in column 12, while the last column has been used here simply to record the "$RC$ constant" as a means of indicating the rate of discharge or equivalent frequency components.

It has been seen that, when load pad $g$ is used, nearly all of the energy goes to the external load. Therefore, if we set $J_L = J$ for this load, making $J/J_L$ equal to unity, first approximation values of $J/J_L$ can be obtained for the other pads, since these values are inversely proportional to the number of load energy units computed for each pad. Such first approximation values are shown in column 10 of Table I and have been plotted in Fig. 3, the dotted line being drawn to make the best fit possible.

Such a line will be seen to intercept the $J/J_L$ axis at a point slightly less than unity, about 0.975 in this case. Since the intercept must be at unity, we now raise the left end of the line arbitrarily. This shows that the ordinate for load pad $g$ should be increased to about 1.025. Using this corrected value of $J/J_L$ for load pad $g$, we now recompute the values of $J/J_L$ for the other load pads, obtaining column 11 as a second approximation. While it can be seen from the foregoing that a better estimate of the value of $J/J_L$ for the high-resistance load pad $g$ could have been made on the basis of expected values for $R_S$ for commercial capacitors, it can also be seen that this assumption is not necessary, since one or two approximations at the most will give rapid convergence to an accurate answer.

The values of column 11 have been plotted as shown in Fig. 3, and the graphically fitted line, 0.514 ohm, is seen to intercept the $J/J_L$ axis very close to unity. The values of $R_S$ given in column 12 are computed from (4), rewritten for convenience as

TABLE I

Load-Pad Characteristics and Calculations from Experimental Readings Taken on a 0.12-$\mu$f Capacitor

| Load-Pad Data | | | | | | Experimental and Calculated Values | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 | 11 | 12 | 13 |
| Load Pad | $R_1$ (ohms) | $R_2$ (ohms) | $R_T$ (ohms) | $1/R_L$ (ohms) | Load Energy Factor | $E$ (volts) | Galv. Defl. (cm.) | Load Energy Units | $J/J_L$ (first approx.) | $J/J_L$ (second approx.) | $R_S$ (ohms) | $RC$ (micro-sec.) |
| g | ∞ | 16.46 | 2.44 | 0.0529 | 7.75 | 150 | 2.89 | 22.4 | 1.00 | 1.025 | (0.473) | 2.33 |
| f | 3.76 | 2.77 | 2.44 | 0.458 | 5.09 | 150 | 3.65 | 18.60 | 1.20 | 1.233 | 0.509 | 0.32 |
| e | 2.96 | 1.79 | 2.44 | 0.575 | 4.21 | 150 | 4.21 | 17.70 | 1.27 | 1.296 | 0.515 | 0.27 |
| d | 2.28 | 1.00 | 2.44 | 0.729 | 3.54 | 150 | 4.76 | 16.84 | 1.33 | 1.364 | 0.500 | 0.22 |
| c | 1.42 | 0.31 | 2.44 | 1.068 | 3.31 | 150 | 4.54 | 15.04 | 1.49 | 1.526 | 0.492 | 0.17 |
| b | 0.953 | 0 | 2.44 | 1.46 | 3.56 | 150 | 3.65 | 13.00 | 1.72 | 1.766 | 0.524 | 0.15 |
| a | 0.519 | 0 | 2.44 | 2.34 | 5.70 | 150 | 1.84 | 10.48 | 2.14 | 2.190 | 0.508 | 0.11 |

$$R_S = \frac{\frac{J}{J_L} - 1}{1/R_L}. \tag{4''}$$

From the graph of Fig. 3 it is seen that only two load pads are absolutely necessary. If low and high values of $R_L$ are chosen, as obtained through load pads *a* and *g*, quite satisfactory measurements can be made to an adequate order of accuracy for subsequent calculations of energy deliverable by the capacitor to an external circuit.

## Discussion

The most important single factor in successful application of this method of measurement is the use of a "perfect" switch for closing the circuit. This switch, shown as the "load switch" in Fig. 4, must close the circuit through the load in such a manner that no electrical energy will be expended in the switch during the closing process. In practice, this means that the contacts must neither spark nor bounce, that they must not be required to break down an oxide film upon closing, and that the entire switch circuit, including leads, must have, when closed, a resistance low in comparison to the resistance being measured.

A number of switches have been tried, and no solid-contact switch, including one of the vacuum type, has proved satisfactory. The most successful switches seem to be those of the mercury-to-mercury contact type, provided they are well made, with very clean mercury and clean metal electrodes dipping into the mercury.[1]

The galvanometer used in the measurements described was a Rubicon (Cat. No. 3412-S), having a resistance of 39 ohms and a current sensitivity of 0.01 microamperes per mm. This instrument has a period of 4.5 seconds, and was used at slightly less than critical damping. It should be realized that use of a galvanometer of higher sensitivity will impose less rigid requirements on the load switch, inasmuch as the required capacitor-charging potential is lowered.

The vacuum thermocouple used was a G.E. Type K A-91×536 with a heater element rated at 2.5 ohms (actually, 2.44 ohms). These thermocouples are available with heater elements having a resistance as low as 1 ohm (Type K A-91×537), and there would be some advantage in the use of the lower resistance value in delivering more energy to the galvanometer when measuring with the lower resistance loads.

[1] Switches which have been found satisfactory—all of the mercury-to-mercury contact type—include the Minneapolis-Honeywell Type H-168 "Con-Tac-Tor," the G.E. Code Nos. 4-24 KR1 and 40-40 KR1 "Kon-Nec-Tors," and a magnetically operated mercury-plunger relay now available commercially from the Champion Mfg. Co., 509 Fifth Ave., New York, N. Y.

It is, of course, advisable in these measurements to keep lead length down and to eliminate errors due to contact resistance. In this connection, it would be preferable to provide four connectors for the load pads instead of three, in order to avoid inclusion of one contact resistance in the common load and thermocouple circuit. However, no detectable error due to this common contact has been discovered in the method as presented here. A rough calculation has been made of the resistances of the leads involved in the capacitor-discharge circuit, on the basis of transients of $10^{-7}$ seconds, with the result that the total resistance is estimated to be in the neighborhood of 0.05 ohm or less.

## Conclusion

The method of measurement here discussed seems to be the simplest approach to the problem in practice. It should be noted that, when factors such as dielectric hysteresis and polarization are considered, the circuit which is an accurate equivalent of a capacitor at all frequencies must have upwards of four branches, involving perhaps seven or more components of resistance and capacitance, and that any method of measurement which assumes simpler conditions should closely duplicate the conditions under which the capacitor will be used. Furthermore, it is generally difficult to measure very low values of resistance with measurement circuits of much higher resistance, such as would be involved with usual electron-tube devices. There are, of course, circuit arrangements by which some time constant might be evaluated, but no practical way is seen of devising the necessary low-impedance switching electronically.

Admittedly, there are involved here some implicit assumptions which may not be rigidly true. Variation of the load changes the discharge time, and probably thereby alters the path of current flow in the capacitor. Also, the internal losses in the dielectric may depend somewhat upon the discharge time. Nevertheless, the method closely parallels the use of the capacitor as an energy-storage device for low-impedance loads.

While it is realized that with this measurement method the charging voltage of the capacitor is quite low, it appears that there is no reason to expect a change in series-resistance characteristics of the capacitor at higher voltages.

## Acknowledgment

The author wishes to express his thanks to C. N. Warfield for his thoughtful criticism and suggestion of the linear relationships, and to other members of the Staff of The Johns Hopkins University, Applied Physics Laboratory, for their aid in preparing these notes for publication.

# Experimental Ultra-High-Frequency Multiplex Broadcasting System*

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*Summary*—**A definite indication that the need for broadcast facilities in the near future will far exceed the available channels is recognized. A feasible expansion appears to be in the ultra-high-frequency spectrum.**

**To produce the maximum number of channels at ultra-high frequencies, a multiplex system of broadcasting has many advantages: A large number of broadcast programs are transmitted at an assigned frequency from a single location and using only one transmitter and antenna system. Effective use is thus made of an optimum transmitting site, a considerable advantage at very- and ultra-high frequencies where line-of-sight limitation of transmission exists.**

**An experimental eight-channel high-fidelity multiplex broadcasting system has been developed and operated during the past year. Multiplex operation is achieved by time-sharing pulse-time modulation. The operating frequency is 930 Mc.**

**The discussion includes a description and operating characteristics of the major components, including modulator, transmitter-antenna system, and receiver.**

## I. Introduction

Even with the opening of the new FM broadcast band, there is definite indication that the need for broadcast facilities will, in the not-distant future, far exceed the available channels. To relieve the situation, it seems safe to assume that further expansion of the broadcast spectrum will become necessary. Such further expansion appears feasible only at ultra-high frequencies, or beyond.

The development here reported was, therefore, directed to the determination of how broadcasting could best be achieved at ultra-high frequencies. The actual frequency used was 930 Mc, within the 920 to 940-Mc range assigned by the Federal Communications Commission for experimental broadcasting. The system developed, however, could be reproduced anywhere in the ultra- or even the super-high-frequency range.

It should be emphasized that this development is not intended to replace either the present AM or FM broadcasting facilities, but rather to supplement both when there is demand for more channels.

To produce the maximum number of useful broadcast channels at ultra-high frequencies, it appears more desirable and economical to broadcast in multiplex with a large number of channels per assigned frequency, rather than one frequency assignment per broadcast channel, as is conventional at lower frequencies. This, of course, does not reduce bandwidth requirements per channel, but it does reduce or eliminate wasteful "guard" frequency separation between adjacent stations to avoid the damaging effects of possible carrier drift. In addition, there is a more potent and fundamental reason for choosing a multiplex form of broadcasting at ultra-high frequencies, rather than the conventional simplex system used at lower frequencies. This has to do with the radiation and propagation characteristics of these waves.

At ultra-high frequencies, a relatively high antenna power gain (concentration of radiated energy in desired directions) is easily and economically achieved. Since propagation is essentially limited to line of sight, tall building tops and other elevated spots offer the best locations for transmitting antennas. In most localities, there are only one or two such effective transmitting sites, and each will accommodate about one transmitter. In a multiplex broadcasting system, a substantial number of transmissions share each optimum site. This advantage is even more pronounced from the receiving standpoint especially for television.

For each transmitting station, there are a multitude of receiving sites. At ultra-high frequencies, antenna gain can be obtained cheaply, even for reception. A very small and simple array, a dish, or a horn will provide a large amount of power gain and directivity. When, because of a large number of transmitting sites, the programs are received from many directions, a single highly directive receiving antenna is not useable unless fitted with an expensive steering mechanism. With a multiplex broadcasting system, the same receiving-antenna orientation not only will give optimum response on all channels but may be adjusted to reduce interfering signals and noise. These important factors provide substantial reasons for using a multiplex form of broadcasting in the ultra-high-frequency spectrum.

A multiplex system, for broadcasting as well as for telephony and general communication applications, may take many forms. Amplitude modulation of several mutually interrelated subcarriers, finally modulating the ultra-high-frequency carrier, would be one possibility. Frequency modulation of subcarriers, which in turn modulate the final carrier frequency, might also be used. Pulse systems involving amplitude, time, width, etc., modulation also are important possibilities. Various combinations of the above might offer even greater advantages than any single type of modulation.

In view of recent advances in pulse techniques and, particularly, since high-power pulse-type rf generators (magnetrons and some triodes) are available for most, or all, of the ultra-high-frequency band, a pulse system

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f modulation with a time-sharing form of multiplexing vas agreed on for the first development of this system.[1]

## II. Pulse-Time-Modulation Multiplex System

To demonstrate as effectively as possible the versatility and advantages of the system, it was decided to multiplex eight broadcast-quality channels. For each channel, 24,000 pulses per second are used to sample each program. This rate corresponds approximately to three times the highest af to be transmitted. In addition to the channel pulses, which are varied in timing to provide the necessary modulation for each channel, a fixed marker pulse is also transmitted at a 24-kc rate, one for each train of eight pulses corresponding to the eight channels. This marker pulse provides the necessary reference point for demodulation at the receiver.

Fig. 1 shows a block diagram of the experimental system recently completed and demonstrated. Each of the eight af signals is used to vary the timing of pulses approximately 0.5 microsecond wide. The resulting video-frequency signal is then used to pulse an rf oscillator, in much the same manner as in radar applications.

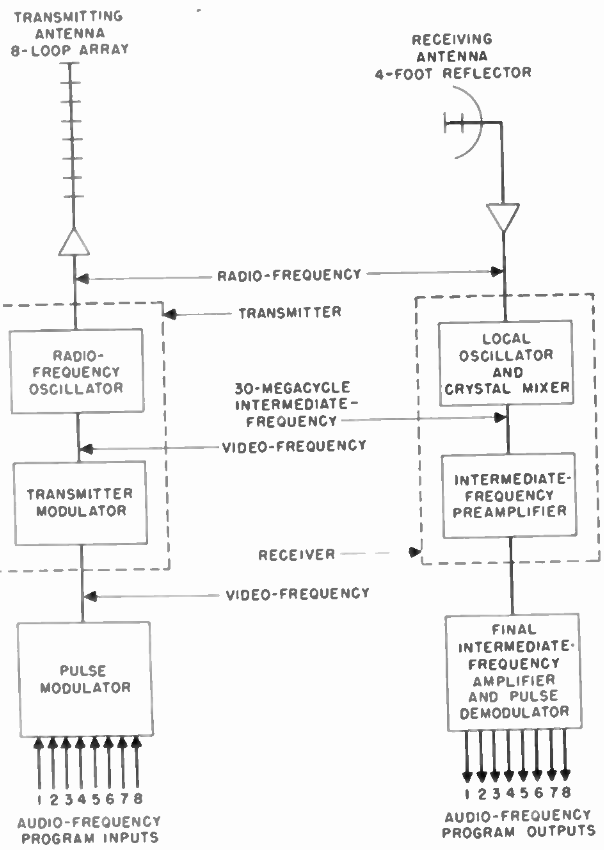


Fig. 1—Block diagram of the multiplex broadcasting system.

For reception, a conventional local oscillator and mixer produce the if signal voltage; this is amplified and detected to provide the video-frequency signal, which is then used to select channels and demodulate each program. Details of the various components making up the complete system are given in the following sections.

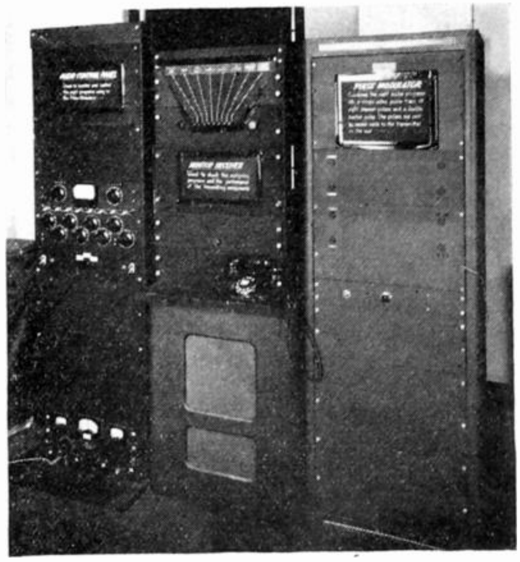

Fig. 2—Experimental multiplex broadcasting system, showing af control panel, pulse modulator, and monitor receiver.

The transmitting station during all tests on this system was located on the roof of 67 Broad Street, approximately 485 feet above ground in downtown New York City.

Fig. 2 shows a view of the experimental transmitter equipment exclusive of the transmitter modulator and rf generator; these were located close to the antenna. Fig. 3 shows the experimental omnidirectional transmitting antenna.

Most of the receiving tests were made at the Nutley Laboratories, 11 miles away, where the receiving antenna was located on a roof, approximately 25 feet above the surrounding terrain. Comparable receiving

Fig. 3—Omnidirectional eight-loop transmitting antenna for the experimental multiplex broadcasting system, located at 67 Broad St., New York, N. Y.

[1] D. D. Greig, "Multiplex broadcasting," *Elec. Commun.*, vol. 23, pp. 19–26; March, 1946.

tests have also been made in the surrounding countryside with portable equipment installed in a station wagon.

Typical services carried simultaneously during various tests and demonstrations included rebroadcasting of standard AM and FM programs and operation of stock ticker (Dow Jones), teletype (Mackay Radio), facsimile (New York Times), and photo transmission. The receiver, shown in Fig. 4, is quite conventional in appearance. It differs in operation from more conventional receivers only in that no rf tuning is required. The channel selection is done entirely at video frequencies by timing circuits.

Fig. 4—Master receiving console.

Where more than one program is desired simultaneously, for example, in separate rooms of the same building, it is not necessary to duplicate the complete receiver but only a portion of the video-frequency system and an af circuit for each service. These latter, called "satellite" receivers, are quite simple in construction but incapable of operation in the absence of the main receiving unit.

## III. Pulse Modulator

The pulse modulator consists of the following four main sections: af amplifier-limiter, pulse generator and delay line, pulse-time-modulation units, and finally a section for mixing the eight channels and marker to form a complete train of time-division time-modulated pulses.

Programs from the eight different sources are supplied to the af amplifier-limiters for modulating purposes. These amplifiers, which are gain controlled, have a response within 1 db from 30 to 10,000 cps. For full modulation (±1 microsecond), an input level of 0 db is required (1 mw in 600 ohms). The modulation excursions must be limited, however, so that any channel does not overmodulate and cause cross talk on any other channel by moving so far in timing that it appears in a position that might be held by an adjacent-channel pulse. To prevent this overmodulation, limiters are placed across the af input-transformer secondary. These consist of germanium crystals with a fixed bias voltage. The af output is then used to modulate a sawtooth wave obtained in proper sequence and delay from the pulse generator.

The pulse generator and delay line make up a single compact unit. The electron-coupled oscillator in this unit determines the base repetition rate of the pulse frequency per channel and is, therefore, designed for very stable operation. In this particular system, that rate is 24 kc. The sine-wave output is then shaped into a sawtooth having a 6-microsecond rise time, and this travels along the delay line from a cathode follower. This delay line is so designed that it will pass the sawtooth without deteriorating the slope and with minimum attenuation. Taps at every 4.63 microseconds are connected to cathode followers. Each cathode follower is connected by a short piece of coaxial cable to the modulator input. This now supplies each channel modulator with a wave form having proper delay, a linear rise time, and the required amplitude so that it may be operated on to obtain time modulation.

This is done by a circuit called the channel modulator.[2] As shown in Fig. 5, this unit takes a slice of the incoming sawtooth wave and converts it to a square pulse. By changing the bias on this double clipper at an af rate, a width-modulated pulse is obtained. Differentiating and using the leading edge produces a time-modulated pulse to be further shaped and processed. The channel

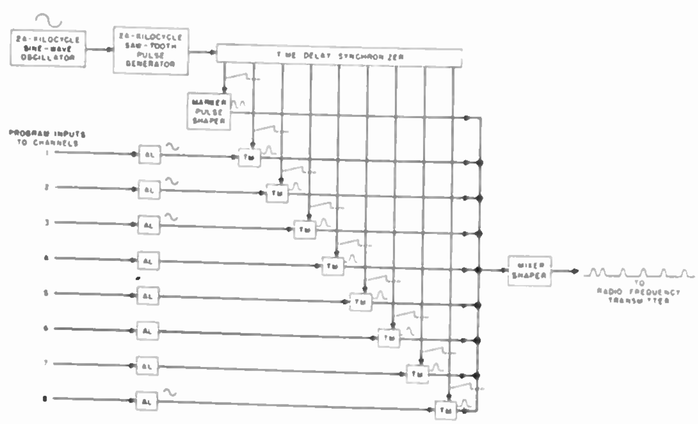


Fig. 5—Block diagram of the modulation system, showing conversion of af channels to mixed train of time-modulated pulses. *AL* are amplifier limiters, and *TM* are time modulators.

[2] D. D. Grieg and A. M. Levine, "Pulse-modulated multiplex radio relay system—terminal equipment," *Elec. Commun.*, vol. 23, pp. 159–178; June, 1946.

pulses are 0.5 microsecond wide at an amplitude 10 per cent above the base, and require 0.15 microsecond to build up to 0.9 or decay to 0.1 of maximum amplitude.

These channel pulses are clipped so that any amplitude modulation is removed, and they are then electronically mixed in such a manner that the pulse train is delivered from a cathode follower to a 70-ohm coaxial cable. Besides the channel pulses, a marker pulse is supplied to the mixer. This marker consists of two pulses, identical in shape to the channel pulses, but spaced only 1.3 microseconds apart. No other two pulses in the entire train, at even the remotest positions of modulation, ever come this close together. This then gives an identifying factor to the pulse train so that a selector device is able to distinguish the program channels. To produce such a double pulse, a single pulse is generated in an identical manner to the channel pulses and passed down an open-circuited delay line. This line has a delay of 0.65 microsecond, thus requiring 1.3 microseconds for the pulse to travel along the line, reflect at the open-circuited end, and return to the starting point. The resulting double pulse is mixed with the eight channel pulses, and the full train then goes to the transmitter modulator.

## IV. Transmitter Equipment

The transmitter proper, shown in Fig. 6, consists essentially of two parts: the transmitter modulator and the rf oscillator, both mounted on a single 72-inch relay rack.

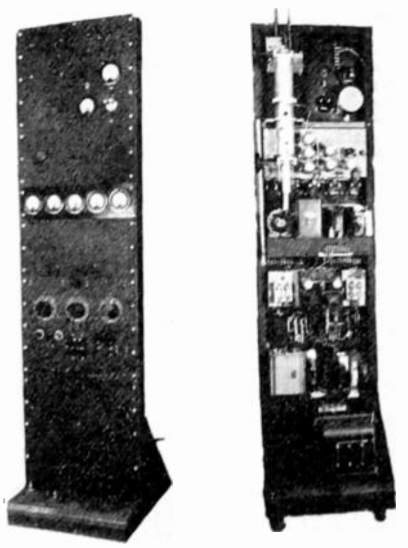

Fig. 6—Front and back view of the transmitter unit.

The transmitter modulator must amplify the low-amplitude pulse train coming from the pulse modulator and deliver signals at approximately 300 volts over a 50-ohm circuit to the grid of the oscillator. The output stage of the transmitter modulator employs two 807 tubes in parallel as cathode followers. The design follows that of a conventional video-frequency amplifier with the same precautions for linear phase shift and flat frequency response of both the low and high end of the video-frequency spectrum. In this particular case, 3 Mc was chosen as the highest frequency and 400 cps as the point where the low-frequency response was down 3 db. Both of these points were chosen for reasons of cross talk and maintenance of pulse characteristics for a signal-to-noise improvement[3] over AM of about 20 db. Because of the type of modulation, linearity of amplification is unimportant and the tubes may be operated for maximum efficiency.

In the experimental installation, it was convenient to locate the pulse-modulator unit (Fig. 2) several floors below the roof on which the transmitter and antenna were located. This necessitated a good resistive termination (accurate matching) of the cable carrying the pulse signals between the two units; any mismatch at this point would produce reflections that would travel back down the cable and cause objectionable cross talk.

For the rf generator, both triodes and magnetrons were considered. More power was possible with available magnetrons and, had the operating frequency been higher in the ultra-high-frequency range, a magnetron would have been chosen. However in the 1,000-Mc region, triode circuitry is simpler and more flexible for an initial experimental unit. A triode oscillator of the tuned-plate tuned-cathode variety using a type 2214B tube was, therefore, used. The circuit is shown schematically in Fig. 7. An adjustable grid "bell," fitting coaxially over the cathode line, provides the necessary feedback. Both plate and grid modulation are feasible, though the latter is somewhat simpler. The experimental unit produced a peak output of 500 watts or more, with an average power output not exceeding 50 watts.

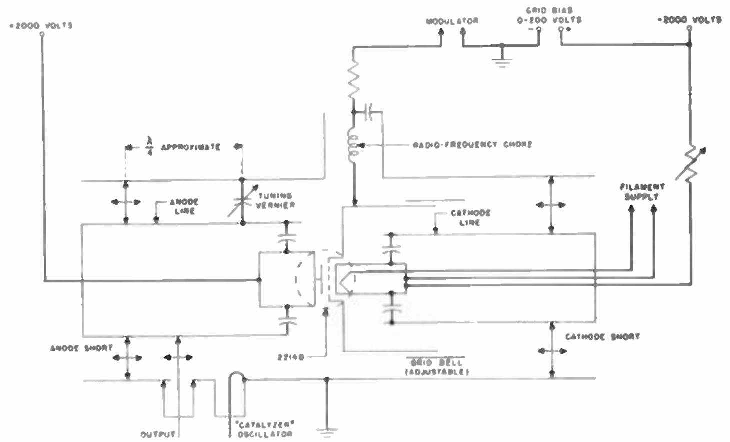


Fig. 7—Transmitter oscillator schematic.

An important consideration of transmitter design for this type of application is the question of frequency stability. An automatic-frequency-control system, to hold the center frequency stable with high precision, is feasible for both magnetron or triode oscillators. A crystal

[3] E. M. Deloraine and E. Labin, "Pulse time modulation," *Elec. Commun.*, vol. 22, n. 2, pp. 91–98; 1944.

frequency-control system with a series of multipliers to the final operating frequency, or more simply a master-oscillator power-amplifier combination, are also feasible through the 1,000-Mc region. In the present case, however, a frequency-stabilizing system did not appear to contribute sufficiently to the initial experimental tests to justify the additional equipment and complication.

Tests indicated that the inherent frequency instability, drift, etc., of the rf oscillator, described above, under all the various operating conditions did not exceed $\pm 0.4$ Mc; this is quite tolerable since the total rf bandwidth is nearly 6 Mc. For any final equipment design of a multiplex broadcasting system, a frequency-stabilizing circuit could be readily incorporated.

In a pulse-time system of modulation, a very important consideration is the uniformity of the starting time of the individual oscillations each time the oscillator is pulsed. Experiments show that, even when the pulses are very accurately timed, there is likely to be a considerable random variation of this factor. The reason for this is, apparently, the fact that a random-noise signal of a certain minimum amplitude is necessary to initiate the rf oscillations when the dc pulse is applied to the tube. This random variation in timing of the pulses is only a small portion of 1 microsecond (which represents full modulation), but produces a noise signal that may be only 30 db below the full-modulation signal. By optimum adjustment of the oscillator feedback circuit, it is possible to minimize this source of noise. A more practical solution is to couple to the transmitter oscillator cavity the output of a small continuous-wave rf oscillator of a type normally used for receiver local oscillators. This auxiliary oscillator—called the catalyzer oscillator—provides sufficient signal level to insure a starting time of oscillations substantially independent of any random-noise variation. Experimentally, it was determined that the catalyzer oscillator frequency could deviate considerably from the pulsed-oscillator frequency without appreciable change in performance.

In addition to the above, the transmitter has the usual monitoring equipment, including an echo-box-type cavity for checking frequency and a reflectometer-detector to monitor rf power output as well as the standing-wave ratio on the output transmission line to the antenna. An rf detector coupled to this transmission line plus a pulse-time-modulation demodulator allows complete over-all monitoring of the programs on the eight separate channels.

## V. Transmitting Antenna

As in the majority of broadcast applications, the radiated transmitter power must be omnidirectional in the horizontal plane to give uniform coverage within the broadcast service area. Also, in line with conventional practice, horizontal polarization was chosen. The basic radiator that provides omnidirectional horizontally polarized radiation is the horizontal loop antenna, used in a large number of applications in the very-high-frequency range, including aerial navigation and FM broadcasting.

To reinforce the radiation in the horizontal plane and reduce wasteful high- and low-angle radiation, a vertical stack of eight square loops (shown in Fig. 3) is used. This compresses the vertical radiation pattern to approximately 9° between the half-power points, and thus provides a power gain of approximately 8 compared to a half-wave dipole. Larger gain than this is feasible, and for a final design, a stack of 16 or more loops may be used.

The transmission-line problem is quite important in the 1,000-Mc region. Most reliable are standardized solid-dielectric cables. However their power-handling capacity is limited, and the amount of attenuation, if long lengths are to be used, is quite substantial. Available air-dielectric lines do not have these limitations, but have disadvantages of their own, namely: discontinuities due to beads and connectors. Furthermore, gassing or air-drying equipment is necessary to keep moisture out for operation over any period of time. Waveguides would perhaps be practical for a final installation, but were considered to be too bulky for an initial experimental setup. In the present case, only 30 feet of line was necessary and RG-17/U solid-dielectric cable was used. The attenuation is less than 1.5 db, which is quite tolerable. Solid-dielectric cable is also used for the feeders within the antenna array. The antenna and transmission-line system has given satisfactory operation under all weather conditions encountered during summer and winter.

## VI. Receiving Equipment

The receiving equipment installation at Nutley consisted of antenna cable, one master receiver, and several satellite receivers to reproduce simultaneously at least four of the eight broadcast channels.

Since in a multiplex broadcasting system all programs originate from the same transmitting site, it is evident that a highly directive receiving antenna oriented along the direction of propagation may be used. This means not only a great increase in the signal strength received, but also a substantial decrease in interfering signals. In the ultra-high-frequency range, a great deal of directivity may be obtained quite economically. In the present case, a 4-foot dish with a simple dipole giving an antenna power gain of approximately 50 was used. A more compact, though less directive, horn antenna was used for portable reception in field

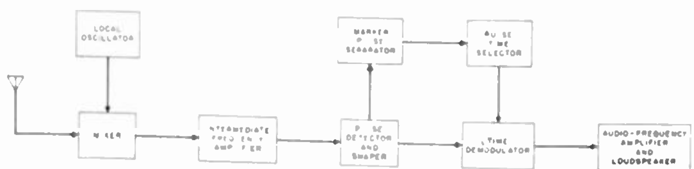


Fig. 8—Block diagram of the pulse-time-modulation receiver.

tests with a station wagon. Solid-dielectric RG-8/U cable connects the antenna to the master receiver.

An over-all block diagram of the complete receiver is shown in Fig. 8. The superheterodyne circuit uses a 2C40 local oscillator and a 1N28 crystal mixer with if output at 30 Mc. No tuning of any rf circuit is necessary for channel selection, because this is done entirely at video frequencies. A schematic of the local oscillator and mixer circuit is shown in Fig. 9. No rf amplification is used since, at these frequencies, the crystal-mixer input circuit provides a somewhat better over-all receiver noise-figure performance than would a tube-amplifier input circuit. However, this is only true if the if input circuit is properly designed. Fig. 10 shows a schematic of the if preamplifier, which follows the mixer.

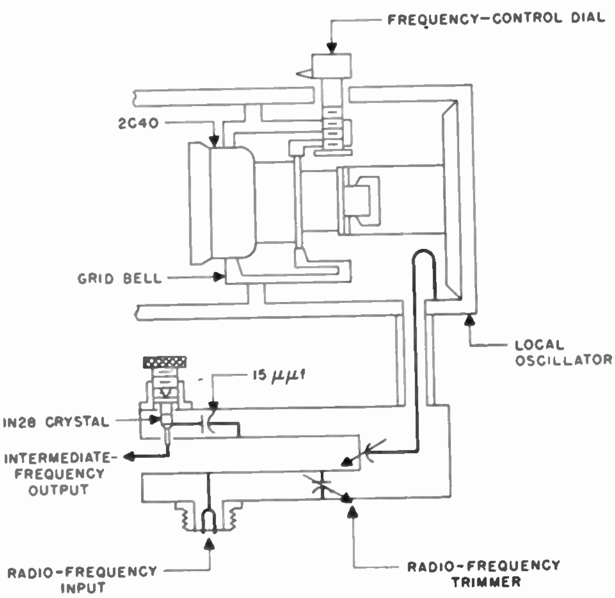


Fig. 9—Schematic diagram of the local oscillator and mixer unit for receiver.

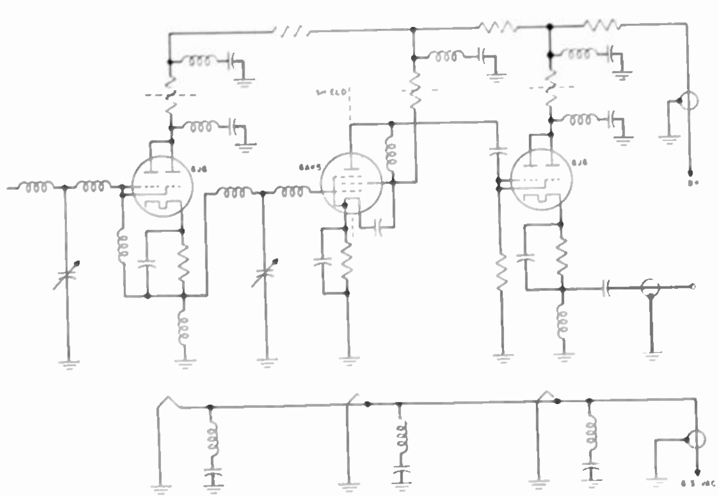

Fig. 10—Schematic circuit of the low-noise if preamplifier unit.

The preamplifier circuit has a noise figure of 2.5 db—that is, the actual circuit noise is only 2.5 db greater than the ideal limit. This is achieved by the use of triode tubes and unusual design of the input network, as well as the assembly and adjustment of the mixer and preamplifier in a single unit.

Referring to Fig. 10, the output from a high-mutual-conductance triode 6J6 used as a cathode-follower, passes to the grid of a 6AK5 through a matching network that is equivalent to a step-up transformer. The degeneration of the cathode follower allows increased input impedance, and permits the use of a triode while still providing power gain. A conventional pentode stage follows, amplifying the signal sufficiently to override the noise of following stages, and a cathode follower provides a low-impedance output circuit.

Since the total noise contributed by the tubes is considerably less than the input thermal noise, the design of the input circuit becomes a major factor. If the input impedance is adjusted to match the source, the resistance noise of the input circuit is added to the equal resistance noise of the source, doubling the noise and limiting the noise factor to not less than 3 db. For the cathode-follower circuit used, optimum noise occurs with a mismatch of about 2:1 between mixer and preamplifier.

Due to the wide bandwidth, unequal primary and secondary $Q$'s, and large coefficient of coupling required, it was found convenient to replace the conventional input transformer by a two-section tapered filter, analogous to a tapered transmission line. This permitted the use of simple elements of inductance and capacitance, eliminating mutual inductance. Assembly of the mixer and preamplifier as a single unit eliminated the usual connecting cable; this was found highly desirable in view of the mismatch required at this point. Loading resistors are, of course, undesirable because of their thermal noise, and the loading in this case is provided almost entirely by the mixer, the amplifier input having an unloaded $Q$ of nearly 50.

Since a cathode follower may oscillate where the sign of the cathode load (reactance) is similar to that of the grid-cathode reactance and opposite to that of the grid-ground reactance, it was necessary to select configurations for the input and output networks satisfying certain added requirements of reactance versus frequency. It was also found that an inductance approximately resonating the grid-cathode capacitance of the 6J6 above the center of the frequency band (32 Mc), improved performance by its effect on the loading reflected from the cathode circuit.

Fig. 11 shows the experimental local oscillator, mixer, and if preamplifier unit.

The main if unit is stagger-tuned, having a gain of about 75 db. Each stage is fixed-tuned, the coil wound on the basis of an average tube capacitance. No trouble was found in alignment, and change in tubes had negligible effect on band-pass characteristics. The if pre-amplifier operates directly into a 50-ohm resistor so that there is no mistuning of any input circuit. There is a manual gain control in the first stage

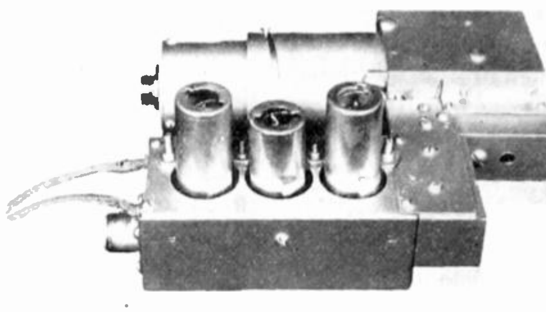

Fig. 11—Experimental local-oscillator, mixer, and if preamplifier unit.

but this, once adjusted, is left untouched. The output of the last stage operates into a plate detector for added gain and one stage of video-frequency limiting follows this. A dual cathode follower then supplies a train of pulses to the video-frequency demodulator, which has 70 ohms impedance.

Although most of the main if unit is conventional, a great deal of care must be exercised in the design and operation of the limiters for best signal-to-noise ratio. Some form of automatic volume control must be used to insure that the if stages themsleves do not limit improperly. Unlike FM (Fig. 12(a)), pulse-time-modulation requires "one-sided" limiting as shown in Fig. 12(b). Therefore, good avc action must take place. If this were not true and an if stage began limiting, the output signal-to-noise might become worse than the incoming signal-to-noise. This is shown in Figs. 12(c) and 12(d).

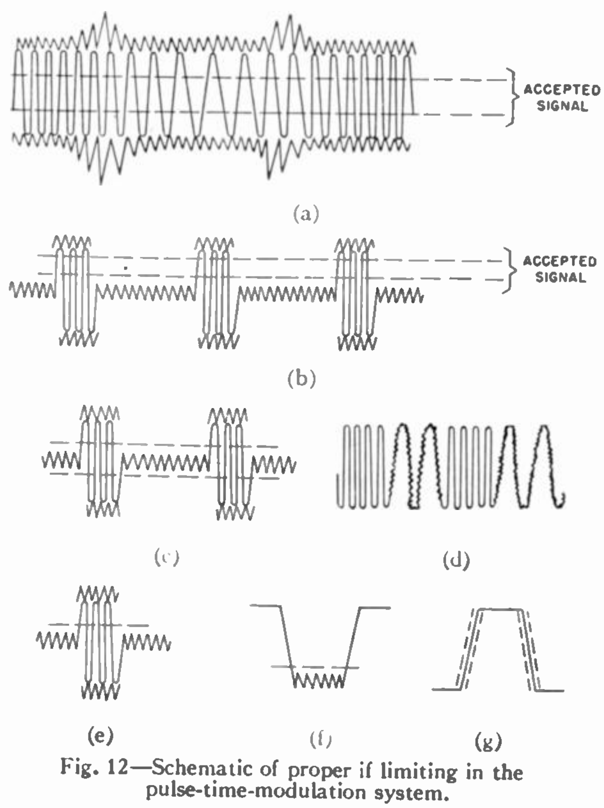


Fig. 12—Schematic of proper if limiting in the pulse-time-modulation system.

Fig. 12(c) shows what the signal may look like when applied to an amplifier that limits, and at Fig. 12(d) can be seen what appears at the output. Proper pulse-time-modulation limiting can no longer take place, as the AM signal-to-noise ratio has been degraded. In this particular if design, the type of avc was simple but effective. The last 3 stages of the strip preceding the detector contain "peak-riding" circuits in the grid with sufficiently long time constants so that no demodulation takes place. The plate detector is so biased as to remove noise on the base line. The negative output pulse is coupled to a video-frequency clipper, with dc restoring of the base line, so that top clipping of the pulse results (Figs. 12(e), 12(f), and 12(g)). Now the signal-to-noise improvement ratio of $K(\Delta Ts/t\gamma)$ can properly take place.[4] $\Delta Ts$ is the pulse deviation from midposition in microseconds, $t\gamma$ is the time of pulse rise in microseconds, and $K$ is a constant that includes the noise crest factor and ratio of pulse repetition rate to af band used. The only noise remaining is that which has been translated into time modulation. These limited time-modulated pulses are then passed to the video-frequency demodulator or to the satellite receivers.

Two types of pulse demodulators are employed, one using conventional components and another type using a special single-channel Cyclophon.[5] Both sets use the double marker for obtaining a synchronizing pulse to identify the desired channel. The marker is separated by the same process as that used in the modulator, namely, by applying the train of pulses to an open-circuited delay line. Because only the marker pulses are 1.3 microseconds apart, the first pulse after traversing the delay line is superimposed on the second pulse, producing a combined pulse larger than all others.[1] This marker pulse is selected by amplitude from the train of channel pulses and performs two functions, channel separation and channel demodulation.

In the first type of receiver, a delay line with taps in the same time relation as that used at the transmitter, is employed to isolate each desired channel. The separated marker pulse, as shown in Fig. 13, is widened, shaped, and transmitted along the delay line. This marker pulse is then applied to the control grid of a pentode biased to cutoff, while at the same time the entire series of pulses is applied to a second control grid. The 6AS6 is ideal for this purpose. The tube is keyed on at the appropriate time by the deblocking marker pulse, and the channel pulse is converted to amplitude modulation by "riding" the sloped top of the shaped marker pulse. Following this tube, a filter removes the pulse-frequency components and only af passes on to the speech amplifier. This method of demodulation proved simple and reliable, requiring no adjustments after the receiver was aligned

[4] B. Trevor, D. E. Dow, and W. D. Houghton, "Pulse time division radio relay," *RCA Rev.*, vol. 7, pp. 561–575; December, 1946.

[5] D. D. Grieg, J. J. Glauber, and S. Moskowitz, "The Cyclophon: a multipurpose commutator tube," Proc. I.R.E., vol. 35, pp. 1251–1257; November, 1947.

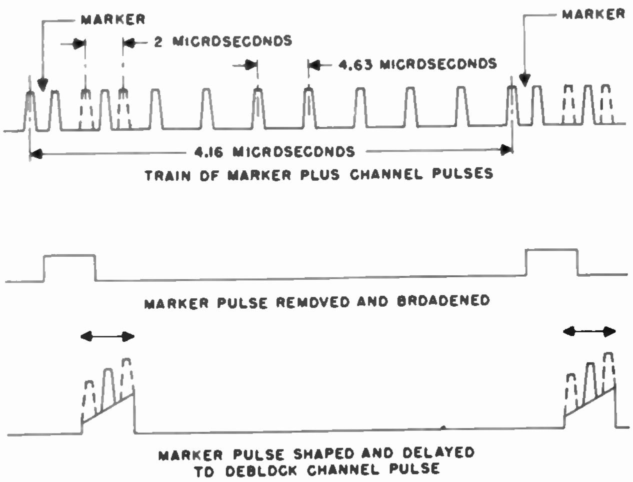


Fig. 13—Operation of marker as channel separator and demodulator.

with the transmitter. The harmonic distortion at 400 cps for the over-all system using this receiver was 2.5 per cent. The signal-to-noise ratio at 100-per cent modulation was 55 db.

A second method of channel separation and demodulation is unique in that it employs a single-channel Cyclophon. This tube, shown in Fig. 14, is comparable in size to the metal 6L6. Use of this tube eliminates a great many circuit components, including the delay line, and saves much time in receiver alignment. Although the effectiveness of the tube depends largely on stable operating voltages, little trouble has been experienced. The method of operation is straightforward. A sawtooth wave, generated by the marker pulse in a single triode, is applied to the two deflecting plates of this cathode-

Fig. 14—Single-channel Cyclophon compared with a 6L6 metal receiving tube.

ray-type tube and sweeps the beam across a plate at the end of the tube. At the same time, all the channel pulses are applied to the control grid of the Cyclophon, and the beam is on only while a pulse is present. The beam is so focused and positioned for the selected channel that the spot striking the end plate is half on and half off for zero modulation position. For all other channels, the beam is entirely away from the plate and produces no output. The plate current, flowing through the output load resistance, is proportional to the position of the beam as determined by the timing of the pulse.[5] Hence, the output amplitude is proportional to the time modulation of the keying pulse. Channel selection takes place by merely shifting the centering voltage on the deflecting plates of this tube. Fig. 15 illustrates this process.

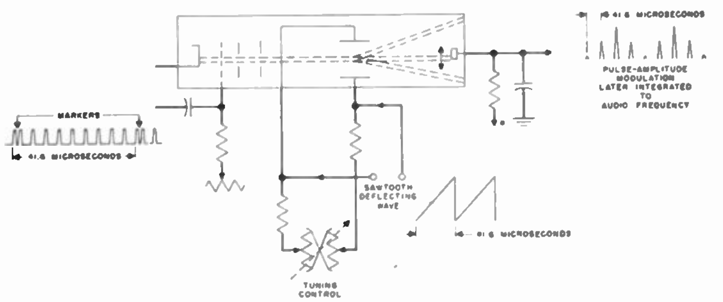


Fig. 15—Cyclophon demodulation process.

## VII. Conclusions

The equipment described above was developed and operated for over a year, during which frequent tests and demonstrations were made to show its performance and versatility. From this experience the following conclusions are possible:

A. Multiplex broadcasting appears feasible anywhere within the ultra-high-frequency spectrum.

B. Over-all performance equivalent to or better than that required in standard broadcasting can be obtained at many locations. More comparative propagation data at ultra-high-frequencies are necessary before more precise conclusions can be drawn.

C. The inherent advantages of high-efficiency and high-gain antenna systems at ultra-high frequencies can be easily and economically realized in this system.

D. Because of the antenna gain, the transmitter size and power may be considerably reduced in comparison with equipment for lower frequencies.

E. Multiplex broadcasting receivers are little, if any, more complicated than standard AM or FM receivers of comparable quality. However, for any widespread use of multiplex broadcasting, the economics of the receiver is the major problem to be faced. The possible use of an inexpensive multiplexing attachment for a standard AM or FM receiver should be considered carefully.

## VIII. Acknowledgment

As is true of most projects of this type and magnitude, many more people presented useful ideas and comments than can be listed here. Among those who contributed most substantially to the development of the various components that made up the complete system were R. T. Adams, D. D. Grieg, A. Horvath, A. Lesti, S. Moskowitz, and B. Parzen, all of Federal Telecommunication Laboratories.

# Pulse Modulation*

E. M. DELORAINE†, FELLOW, IRE

*Summary*—**The broad significance of pulse modulation is dealt with from its original concept through the various methods of attaining it. Its application to time-division multichannel systems is considered. More recent developments in pulse-count-modulation systems, and also potential applications to switching problems, are described.**

## 1. The Origin of Pulse Modulation

THE CLASSICAL CONCEPT of transmission of speech by electrical means involves converting the sound vibrations into currents, one characteristic of which, amplitude or phase, is made to vary according to the sound intensity as a continuous curve versus time and an inverse operation at the receiving end. In the last few years, however, other methods have been used, in which the transmission of any speech channel occupies but a fraction of the time, the information being transmitted by samples or current pulses recurring at a suitable rate. These methods are known under the general name of "pulse modulation."

The early publications on pulse modulation are found mostly in the form of patents.[1] A review of these indicates that up to 1935 the thinking of the authors was very often in terms of improving the power efficiency of the transmission system, either at the transmitter or in the link.

Since 1935, however, more and more consideration has been given to pulse modulation as a method of multiplexing a number of telephone channels.

The accepted technique in multichannelling has been to take each individual telephone frequency band and translate it to successive positions in the spectrum of frequencies by means of modulators and filters, transmit this information on a single path, and perform the inverse operation at the other end of the link. The bandwidth utilized on the link increases, of course, in direct proportion to the number of channels. A typical layout of a multichannel transmission system by frequency division is shown in Fig. 1.

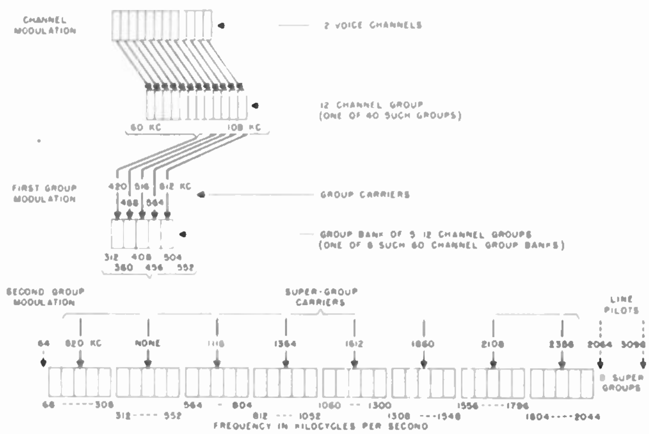


Fig. 1—Typical layout of a multiplex transmission system by frequency division.

The technical literature on these subjects at this time is very abundant. These elements have reached a high degree of perfection and practicability.

Terminal equipments were thus realized which were capable of reproducing the speech in any individual channel with high fidelity. The other parts of the transmission system had to be improved accordingly, and rather severe requirements had to be placed on them as to limitation of nonlinear distortion to avoid crosstalk between channels, stability of characteristics, and noise level.

While fully acknowledging the remarkable success of the frequency-division transmission system, Reeves and others[2] took an interest in 1936 and the following years in the possible merits of multichannel systems based on the concept of time distribution, translating into the telephone art some of the well-established practices of multichannel telegraphy. The attractive feature of the idea was that such methods might prove especially useful where it is difficult or expensive to attain the quality of characteristics required for reliable multichannel transmission on a frequency-division basis.

## 2. Application of Pulse Modulation to Time Multiplexing

The basis of pulse-modulation time multiplexing is the fact that it is not necessary to transmit the complex wave form of a speech wave in its entirety. It is sufficient to take successive "samples" of the amplitude in the channel at separate time intervals and transmit this in-

* Decimal classification: R148.6. Original manuscript received by the Institute, September 27, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 24, 1948.

† International Telephone and Telegraph Corp., New York, N. Y.

[1] Among such patents can be cited:

| *Name* | *United States Patent Number* | *Filing Date* |
|---|---|---|
| R. A. Heising | 1,655,543 | April 18, 1924 |
| R. D. Kell | 2,061,734 | September 29, 1934 |
| K. Posthumus and C. G. A. VonLindern | 2,161,087 | December 17, 1935 |
| A. M. Nilcolson | 2,021,743 | June 13, 1930 |
| R. M. Ranger | 1,873,786 | September 29, 1928 |
| A. S. Riggs | 2,048,081 | April 29, 1933 |
| R. E. Shelby | 2,171,150 | September 14, 1936 |
| D. G. C. Luck | 2,113,214 | October 29, 1936 |
| G. Bozzi (Italian Patent) | 348,656 | February 8, 1937 |

[2] E. M. Deloraine and A. H. Reeves, United States Patent number 2,262,838, filed, November 8, 1938.

formation as a series of pulses carrying the information, as indicated in Fig. 2. If the number of samples per

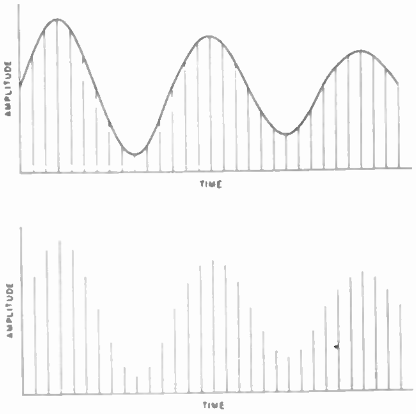


Fig. 2—Sampling of speech.

second exceeds twice the highest frequency to be transmitted in the voice band, the telephone channel can be correctly reproduced at the distant end.

The bandwidth required in the transmission medium through which the pulses are propagated depends on the shape of these sampling pulses, and the method adopted for carrying the information on pulses.

A number of incoming telephone channels can be sampled at time intervals that are displaced in the time scale just enough to avoid the superposition in time of any two pulses. The successive samples of telephone channel number 1 will be followed by the successive samples of telephone channel number 2 and so on until the whole time interval between two samples of channel number 1 is completely filled, as shown in Fig. 3.

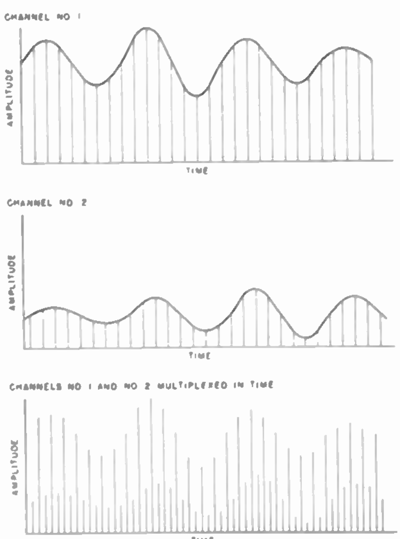


Fig. 3—Pulse-time multiplexing.

It is clear, in consequence, that the number of channels that can be transmitted over a single link increases as the individual samples are made shorter. If the time is used in the most efficient manner, the bandwidth required for the transmission of $n$ channels is close to $n$ times the bandwidth of an individual channel, as is the case for multichannel systems utilizing frequency division.

## 3. Methods of Pulse Modulation

### *Pulse-Amplitude Modulation*

The samples of speech amplitude can be transmitted as such. This is pulse-amplitude modulation. This method is efficient in utilization of bandwidth. The requirements on the link for nonlinear distortion are, however, less stringent than for the frequency-division method, since, at a given time, one channel only is present and crosstalk cannot be introduced in this manner, though it can reappear in a different manner if the pulses are lengthened by the transmission link so as to overlap each other to a certain extent. The requirements as to link stability and low noise are not changed materially.

It is the consideration of signal-to-noise-ratio improvement, particularly for radio links, that drew attention to other possible methods of pulse modulation.

### *Pulse-Width Modulation*

One of the first methods to be considered was pulse-width modulation, which inherently requires more bandwidth than pulse amplitude modulation, but provides a means to trade that increase in bandwidth for an improvement in the signal-to-noise ratio. Pulse-width modulation is not, however, economical or efficient in power insofar as the significant information is, in fact, the timing of the beginning and the end of individual pulses, and this led to the next method.

### *Pulse-Time Modulation*

It was suggested that pulses of constant amplitude, form, and duration should be used, these pulses being displaced in time from a uniform spacing to an extent corresponding to the amplitude of the sample. This is more efficient than transmitting the beginning and end of pulses modulated in width, since the amplitude is translated in time displacement of one pulse, from a reference position which, however, is not transmitted except as a marker for all channels. This is usually referred to as pulse-position or pulse-time modulation.[3] As compared to pulse-amplitude modulation, the number of channels that can be transmitted over a given bandwidth is reduced. If we assume, for instance, that the individual pulse is time-modulated only to the extent of its own duration, the number of channels is divided by two. But a major advantage is that the link need not have linear amplitude characteristics. The repeaters operate as triggers, with the requirement, however, that they must trigger accurately in time. The possible variations of level in the link are compensated

[3] A. H. Reeves, French Patent 833,929, filed, June 18, 1937; and addition 49,159, filed, July 5, 1937; and United States Patent number 2,266,401, filed, June 9, 1938.

by the trigger action, since the output level of the repeaters is independent of the input level between wide limits. This method provides a means to trade bandwidth for a signal-to-noise ratio comparable with that obtained with frequency modulation. Noise is introduced in the system to the extent only that it will change the timing of the pulse, so that the influence of a given noise amplitude in the link decreases as the pulses become sharper and also as the time displacement due to modulation is increased. This method of improving the signal-to-noise ratio has been found desirable in applications to radio links, though it is paid for by a reduction in the number of channels for a given transmission bandwidth.

### 4. Pulsed Frequency Modulation

Another pulse-modulation method for radio links in which the pulses remain constant in time position, as well as in amplitude and duration can be designated as pulsed frequency modulation.

The information is transmitted by means of a change in the carrier frequency of each individual pulse, which is made to vary proportionally to the amplitude sample at the corresponding time. The frequency modulation brings in a signal-to-noise improvement, which involves additional requirements in bandwidth. The utilization of the total bandwidth improves, however, as the number of channels is increased, because the additional band required for frequency modulation and a given signal-to-noise improvement becomes a smaller fraction of the bandwidth required for the correct transmission of the pulse shape.

### 5. Pulse-Count (or Code) Modulation

The assimilation of telephone transmission to telegraph transmission becomes still closer in a method suggested by Reeves.[4] Not only is the speech curve sampled, but the amplitude samples are submitted to two successive transformations: quantization, and counting or coding.

By quantization is meant the process by which the amplitude sample is not sent with its exact value but is approximated to one of a finite number of discrete levels, as shown in Fig. 4. It is clear that, when such a series of quantized pulses is sent into a low-pass filter cutting off at the maximum speech frequency which it is thought proper to reproduce, the result will show some distortion due to the effect of the difference between the quantized and exact amplitude samples. This distortion obviously will decrease when the number of quanta is increased for any given maximum amplitude.

At the receiving end, a method which discriminates against amplitudes that do not correspond to a quanta is used. This is capable of eliminating forms of speech impairment introduced in the transmission systems, if they can be distinguished from the quanta.

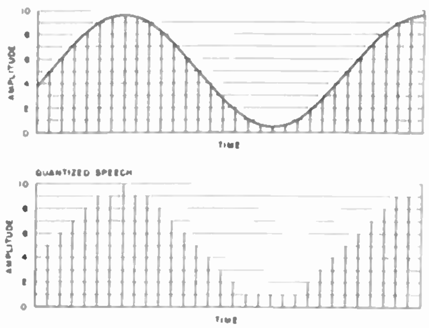


Fig. 4—Quantization of sampled speech amplitudes.

While the advantages of speech quantization do not necessarily apply only to pulse modulation, it is of interest to combine the two processes. However, a third step can be taken with advantage.

Instead of transmitting the series of quantized amplitudes directly over the line or through space, a third transformation is performed, which is counting or coding. The information, that it is really necessary to communicate is the number of quanta present in any particular amplitude sample; this can be achieved by the transmission of a group of pulses, $n$ for instance, each one having $s$ possible characteristics. The total number of distinguishable levels will be $q=s^n$, and such a mechanism will thus allow the counting of $q$ different levels. Thus, a group of five pulses of the on-or-off type can be used to count $2^5=32$ quantized levels in a manner similar to what is done in teleprinter signal transmission. In this instance, where $s=2$, the system is a binary one. Other values for $s$ can obviously be utilized, and ternary systems, for instance, have been considered particularly for wire transmission.

The fundamental advantage of binary or ternary systems is that the demodulation process consists in merely acknowledging the presence or absence of a pulse of a definite type at a definite time position. In the case of the binary code, the number of errors will fall to an insignificant value provided the pulse amplitude remains sufficiently larger than the noise rms amplitude. For instance, for a binary code and pure random noise a ratio of 18 db in rms signal-to-noise ratio will be sufficient to provide a signal-to-noise ratio in the order of 70 db after demodulation.

Because of the fact that noise in the demodulated signal originates in the random distribution of noise bursts above its rms value, the signal-to-noise improvement follows a very different law from that which it does in pulse-time modulation or frequency modulation.

In all cases, the relation between demodulated and input signal-to-noise ratios can be determined theoretically with a sufficient simplicity in the case when the signal amplitude is noticeably larger than the noise rms value. Above that threshold, the two ratios vary in direct proportion for pulse-time modulation and fre-

[4] A. H. Reeves, French Patent number 852,183, filed, October 3, 1938; and United States Patent number 2,272,070, filed, November 22, 1939.

quency modulation. For pulse-count modulation, however, the proportionality holds between input signal-to-noise in power ratio and demodulated signal-to-noise ratio in db.[5] Thus, lowering the input power ratio by 3 db brings the output ratio down to 35 db, and increasing it by 3 db brings the output ratio up to 140 db.

The effect of noise in the latter case thus becomes entirely negligible. This advantage is paid for by the distortion that quantization brings in and the additional bandwidth that the counting process necessitates. If $f$ is the maximum frequency that it is proposed to transmit by voice channel, and $n$ is the number of pulses per group, then the bandwidth per channel should be somewhat larger than $n$ times $f$.

## 6. Time Versus Frequency Translation Multiplexing

The technique of modern telecommunication equipments shows a definite trend toward systems that permit of multiplexing a great number of telephone channels. These channels in turn can be used for telegraph or facsimile communication. Time multiplexing by means of pulse modulation is a solution that presents some interesting features.

One common advantage belonging to all pulse methods is the comparative ease with which any particular channel can be dropped and reinserted at any repeating point. This necessitates a complicated and formidable array of modulating and filtering apparatus when frequency division is utilized. However, in pulse transmission this may be accomplished simply by gating out every $n^{th}$ pulse or group of pulses following the marker signal. Each channel keeps its individuality throughout the link, whereas in frequency-division technique, the economic considerations lead to a definite grouping of the channels.

Pulse transmission eases up the requirement for linearity in repeater equipments. This advantage is less marked on cables where very linear amplifiers are available and where propagation characteristics can be kept nearly constant and large signal-to-noise ratios maintained even for transmission through a great number of repeaters. It is definitely favorable for radio transmission where such conditions are not encountered. In this respect, pulse-count modulation would seem to be specially advantageous, since regenerative relays can be used in this case, reshaping the pulses at any repeating point. A very small signal increase is sufficient to provide for the required signal-to-noise ratio at the end of the chain.

The more general aspect of the integration of radio-relay links into networks needs also to be considered. Should microwave radio demonstrate its ability to carry a large volume of intercity communications economically, then those transmission methods that are best adapted to radio might be extended to certain other parts of the network if justified.

[5] A. G. Clavier, P. F. Panter, and W. Dite, "Signal-to-noise-ratio improvement in a PCM system," Proc. I.R.E., vol. 37, pp. 355–359; April, 1949.

## 7. Switching

One interesting feature of pulse transmission in this respect is the possibility of switching channels. A number $n$ of incoming telephone channels, after going through a simultaneous time-sampling operation, can be multiplexed in any desired time sequence by the insertion of suitable delays in the individual channels. Time scanning of the group could connect the individual incoming channels on outgoing terminals in an order determined by the value of the individual channel delay. By changing the value of the time delay introduced in the individual channels, the channel could be effectively switched between incoming and outgoing terminals in any desired manner. A typical example of switching by time displacement is shown in Fig. 5.

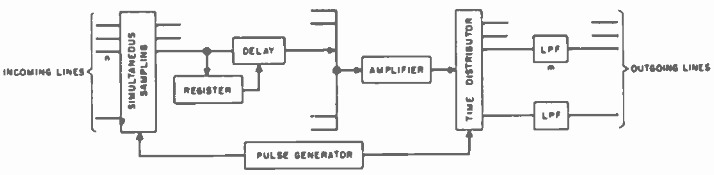


Fig. 5—A typical example of switching by time displacement.

If a multiplex pulse-transmission system is utilized, the switching operation also can be performed, but the individual pulse channels must first be isolated by a gating operation and passed through individual delay lines, and then reinserted in the multiplex system. At the receiving end, these channels would appear in an order that is determined by the value of the delays inserted.

The delays, which determine interconnection of channels between the ends of the system, may be actual time delays or the equivalent. It is possible to conceive that the delay may be introduced by counting units of time after the arrival of the incoming channel pulses. Such a counting operation can be performed by gas tubes designed for the purpose, the units of time being supplied by a common control element.

Many other methods to attain the same results in this comparatively new field of switching pulse-modulated channels can be visualized, but it is very difficult, at the present time, to evaluate their relative merits.

Pulse, being of the nature of telegraph technique, obviously offers possibilities in the telegraph field; and as switching, as described above, is effected by means which are essentially of a telegraph nature, it can play its part too.

If the use of the imagination is allowed, rather than established technical and economic facts, it is possible to visualize methods of transmission and switching using pulse modulation, which may permit a useful combination of these two fields which have so far been separate, with an ultimate gain in the complete network.

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J. T. Acker was born in Brooklyn, N. Y., in November, 1902. He is a graduate of the College of the City of New York, and also holds the S.M. degree in chemical engineering practice from the Massachusetts Institute of Technology. He was associated with the Bell Telephone Laboratories, Inc., in New York, N. Y., for eighteen years, engaged in research and development work on electrochemical and corrosion problems. In 1943 Mr. Acker transferred to the Western Electric Company to supervise the chemical operations in electron-tube manufacture. He is now assistant superintendent of manufacturing engineering in the Electronics Shops, located in Allentown, Pa.

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John W. Coltman was born in Cleveland, Ohio, in 1916. He was graduated from the Case Institute of Technology in 1937 with the B.S. degree in physics, and was later awarded the degrees of M.S. and Ph.D. from the University of Illinois. In 1941 he was given a Research Fellowship at the Westinghouse Research Laboratories. When wartime needs gave radar development top priority, Dr. Coltman turned to microwave research, specializing in magnetron development. In 1948 he won recognition in connection with the development of a unique X-ray "telescope" that provides a 500-times brighter image for X-ray fluoroscopic screens.

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E. M. Deloraine (M'25–F'41) was born in Paris, France, on May 16, 1898. He received the B.S. degrees, the Certificat de Mathematiques in 1918, and the Engineur Deplomié de L'ecole de Physique et Chimie, a branch of Paris University, in 1920.

In 1917 he joined the French Army Signal Corps, and later engaged in research work at the Eiffel Tower. He became associated with the London engineering staff of the International Western Electric Co., in 1921, becoming European technical director of the company in 1933. Later he directed experiments in connection with automatic radio compasses for aircraft.

Mr. Deloraine came to the United States in 1941 to take charge of the organization of the laboratories unit for the Federal Telephone and Radio Corp. In 1945 he was appointed president of International Telecommunication Laboratories, Inc., under the sponsorship of IT&T, where he became technical director in 1946, as well as vice-president of International Standard Electric Corp.

Mr. Deloraine was made a Chevalier of the Legion of Honor in 1938, and he was elected vice-president of the French Institute of Radio Engineers in 1939. He has been a member of the International Consultative Committee of Long Distance Telephony since 1927, and is also a member of the French Astronomical Society and a fellow of the American Institute of Electrical Engineers. He was Vice-President of the IRE in 1946.

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Armig G. Kandoian (S'35–A'36–SM'44) was born in Van, Armenia, on November 28, 1911. He received the B.S. degree in 1934 and the M.S. degree in electrical communication engineering in 1935, both from the Harvard University. Since 1935, Mr. Kandoian has been with the International Telephone and Telegraph Corporation and associated companies. His work has been primarily developments dealing with antennas, radiation, measurements, link communication, and air navigation. He is at present head of the radio and radar components division of Federal Telecommunication Laboratories.

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❖

Ben S. Melton

Ben S. Melton (SM'47) was born in Dallas, Tex., on January 2, 1904. He received the B.S. degree in electrical engineering from the Rice Institute in 1925. Following a student engineering course at the General Electric Co. and a short time as engineer with the Gulf States Utilities Co. at Port Arthur and Beaumont, Tex., he was employed by the General Exploration Co. of Houston, Tex., to do research and development work in electromagnetic prospecting for oil, during the years 1928 to 1930. Late in 1930 he entered the field of seismograph prospecting with Geophysical Service, Inc., of Dallas, Tex., and except for a short period remained in that work until 1942, becoming associated with the National Geophysical Co., from 1937 until April, 1942.

In 1942 Mr. Melton joined the staff of the Applied Physics Laboratory of The Johns Hopkins University as a radio engineer, working on the proximity-fuze project which was then at the Department of Terrestrial Magnetism of the Carnegie Institution of Washington. Following assignment to several associated projects, he entered the research program on supersonic aerodynamics which was initiated toward the close of the war, being most recently engaged in development of optical and other methods for examining supersonic flow. In December, 1948, he accepted an appointment with the Department of the Air Force, and is now attached to the 1009th Special Weapons Squadron, Washington, D. C.

Mr. Melton is a member of the Society of Exploration Geophysicists, the American Geophysical Union, and the Philosophical Society of Washington.

# Abstracts and References

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## ACOUSTICS AND AUDIO FREQUENCIES

**534** **1251**
Sound—E. G. Richardson. (*Rep. Progr. Phys.*, vol. 9, pp. 228–253; 1942–1943. Bibliography, pp. 253–255.) A general discussion, covering propagation in free media, ultrasonics, vibrating systems, acoustic impedance and properties of materials, speech and hearing, etc.

**534** **1252**
Sound—E. G. Richardson. (*Rep. Progr. Phys.*, vol. 10, pp. 120–128; 1944–1945.) Discussion of auditorium acoustics, metallurgical applications of ultrasonics, musical scales, and tuning. See also 1251 above.

**534** **1253**
**Some Recent Developments in Applied Acoustics**—R. G. Bolt and L. L. Beranek. (*Atti del Congresso internazionale della Radio* (Rome), pp. 225–251; September and October, 1947. In English.) Adequate progress has already been made with the understanding of (a) the basic principles of many electroacoustic instruments, (b) speech and hearing, though many physiological and psychological phenomena are still unexplained, and (c) the control of sound by porous absorptive materials. Further work is needed on (a) the transient response of loudspeakers and complicated coupled systems, (b) the study of all kinds of distortion, and (c) the design of studios for music.

**534.001.8** **1254**
**Generation of Sonic and Ultrasonic Waves in Liquids for Industrial Purposes**—Janovsky and Pohlman. (*See* 1442.)

**534.321.9:621.391.63** **1255**
**On the Modulation of Light at Radio Frequencies by means of Ultrasonic Waves**—A. Giacomini. (*Atti del Congresso internazionale della Radio* (Rome), pp. 302–311; September and October, 1947. In Italian.) See also 1847 of 1948.

**534.44** **1256**
**Electrical Methods of Sound Analysis**—H. Koschel. (*Fernmeldetech. Z.*, vol. 1, pp. 237–244; December, 1948.) An outline of the following methods: (a) Fourier analysis of oscillograms, (b) use of octave band-filters with oscillograph, (c) sonic-frequency spectrometer with 27 filters in parallel, each covering ⅓ octave, (d) search-tone analysis, (e) analysis by means of an acoustic diffraction grating, (f) analysis with an optical raster, and (g) visible-speech methods.

**534.6** **1257**
**The Double Crystal Acoustic Interferometer**—W. J. Fry. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 17–28; January, 1949.) A one-dimensional theory of operation is discussed. This is sufficiently general to include both resonance and off-resonance operation, and any amount of acoustic loading of the crystals.

**534.6:621.396.611.21** **1258**
**Low-Loss Crystal Systems**—Fry. (*See* 1326.)

**534.756** **1259**
**Hearing: Part 1—The Cochlea as a Frequency Analyzer**—T. Gold and R. J. Pumphrey. (*Proc. Roy. Soc. B*, vol. 135, pp. 462–491; December 14, 1948.) Possible methods of sensory appreciation of sounds are briefly surveyed. Information must, in general, be lost unless peripheral frequency analysis occurs under certain conditions which include proportionality between the frequency and the selectivity of the resonant elements. Experimental evidence is submitted which shows that this selectivity is very much higher than has been supposed and is roughly proportional to frequency, and that only the Helmholtz resonance theory, suitably interpreted, is consistent with observation. Evidence of high cochlear damping is criticized. The ear appears to be a perfect frequency analyzer up to 1 kc; at higher frequencies, it is imperfect because of the limited number of resonant elements and nerve cells available. Part 2: 1260 below.

**534.756** **1260**
**Hearing: Part 2—The Physical Basis of the Action of the Cochlea**—T. Gold. (*Proc. Roy. Soc. B*, vol. 135, pp. 492–498; December 14, 1948.) An attempt is made to explain observed results theoretically. The values of $Q$ for a string immersed in water (to simulate the basilar membrane immersed in its liquid) are much lower than those obtained for the human ear. For this reason, the hypothesis that the cochlea behaves like a regenerative radio receiver is proposed. Part 1: 1259 above.

**534.78** **1261**
**Speech Communication under Conditions of Deafness or Loud Noise**—W. G. Radley (*Jour. IEE* (London), part I, vol. 95, pp. 544–545; December, 1948.) Discussion on 2690 of 1948.

**534.78** **1262**
**Statistical Measurements of Vocal Intensity**—G. Sacerdote. (*Atti del Congresso internazionale della Radio* (Rome), pp. 389–401; September and October, 1947. In Italian.)

**534.782** **1263**
**On a Model of an Artificial Voice for Electroacoustic Technique**—P. Chavasse. (*Atti del Congresso internazionale della Radio* (Rome), pp. 273–279; September and October, 1947. In French.) A sound source whose intensity is nearly uniform with frequency is provided by an emf from a neon tube suitably polarized, associated with an amplifier and a loudspeaker. Any desired quality of voice can be obtained by adding a system of filters.

**534.844/.845** **1264**
**The Absorption Coefficient of Resonators**—G. G. Sacerdote. (*Alta Frequenza*, vol. 17, pp. 217–219; October, 1948. In Italian, with English, French, and German summaries.) Curves show the reverberation times as a function of frequency for a room (a) without and (b) with a cellular lining consisting of tubular resonators. From these curves, the absorption coefficient of a partition formed of the resonators is derived for the frequency range 300 to 2,000 cps.

**534.844.1** **1265**
**The Practical Determination of the Reverberation Time of a Room with the Cathode-Ray Oscillograph**—A. Moles. (*Radio Franç.*, pp. 4–8; February, 1949.) Reverberation curves can be photographed directly on slowly moving film (4 to 10 cm per second) without using the cro timebase, or on a stationary film, in which case, the necessary slow sweep is obtained by means of a rotary potentiometer of about 5,000Ω, which is shunted across a 150-volt source and supplies a variable voltage to the deflecting plates. The potentiometer is driven by a small motor of the type used for car windscreen wipers. Typical oscillograms thus obtained are given and the method of calculating the reverberation time, taking account of oscillogram irregularities, is explained in detail with numerical examples.

**534.86** **1266**
**Listeners' Sound-Level Preferences**—T. Somerville and S. F. Brownless. (*BBC Quart.*, vol. 3, pp. 245–250; January, 1949.) A sound-level meter incorporates a weighting network which simulates the response of the ear. Members of the public of varying ages, BBC engineers and musicians were tested with this device; results obtained are shown graphically.

The preferred maximum sound level of the four types of program chosen for the tests appears to decrease as the age of the listener increases.

534.861:621.396.621 1267

**The Acoustics of Broadcasting Receivers**—V. A. Govyadinov. (*Radiotekhnika* (Moscow), vol. 3, pp. 88–95; November and December, 1948. In Russian.)

621.395.61 1268

**A New Moving-Coil Microphone**—H. J. Griese. (*Fernmeldetech. Z.*, vol. 1, pp. 227–232; December, 1948.) Details of an instrument with a sensibly flat characteristic up to 6 kc for sound incident normally on the diaphragm and up to 10 kc for lateral incidence. The directional diagram is spherical up to about 1,500 cps; for higher frequencies, it is slightly pear-sahped. Sensitivity is good and nonlinear distortion low.

621.395.61/.62:621.315.612 1269

**Application of Activated Ceramics to Transducers**—H. W. Koren. (*Jour. Acous. Soc. Amer.*, vol. 21, p. 62; January, 1949.) Summary of Acoustical Society of America paper. Titanate ceramics can be made to have piezoelectric properties suitable for application to transducers if (a) the Curie point is well above the working temperature, (b) the material is polarized by means of an external voltage source within certain limits of potential gradient and time duration. Multilayer strips have been developed to provide the low mechanical impedance required in phonograph pickups, microphones, etc.

621.395.625.6 1270

**Volume Compressors for Sound Recording**—W. K. Grimwood. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 49–76; January, 1949.) Discussion of the desirability of volume compression, compressor characteristics, classification, performance measurement, and the relative merits of various types.

621.395.667 1271

**On Lower- and Upper-Register RC Equalizers for Audio Frequencies**—W. Daudt. (*Funk und Ton*, vol. 3, pp. 33–42 and 86–92; January and February, 1949.) A general treatment, with design formulas and curves, and detailed discussion of three particular cases. A design is given for a resistance amplifier with correction at both ends of the af range.

621.395.667 1272

**Simple Tone Control Circuit**—E. J. James. (*Wireless World*, vol. 55, pp. 48–50; February, 1949.) Adequate reduction or increase of bass or treble is obtained for normal requirements. Only resistors and capacitors are required. The complete circuit diagram for including the control in an amplifier is given.

## ANTENNAS AND TRANSMISSION LINES

621.3.091:621.315.2 1273

**Attenuation in Air-Space Cables for Centimetre Waves**—H. J. Wegener and O. Zinke. (*Frequenz*, vol. 2, pp. 203–207; August, 1948.) Diagrams show directly the reduction of attenuation, compared with that of a coaxial cable for *TM* and *TE* waves, a function of frequency and cable diameter. From other diagrams, the optimum diameter for a given value of attenuation can be found for any wavelength in the range 1 to 15 cm. For wavelengths of 2.5, 5, 7.5, and 10 cm, the dependence of attenuation on cable diameter is shown for all types of air-space cables. For wavelengths <2.5 cm the attenuation for $H_{11}$ waves is less than for $H_{10}$ waves.

621.315.012:518.3 1274

**R. F. Transmission Line Nomographs**—P. H. Smith. (*Electronics*, vol. 22, pp. 112–117; February, 1949.) 10 abacs for calculating characteristic impedance, high-frequency resistance, current-phase relationship, voltage gradient, SWR, etc.

621.315.2+621.394/.395 1275

**French Telecommunication Networks**—Mailley; Gastebois. (*See* 1492.)

621.315.2+621.394/.395].66 1276

**Construction of Cables and Loading Coils**—R. Belus. (*Câbles and Trans.* (Paris), vol. 3, pp. 31–47; January, 1949. With English summary.) Developments in technique from 1924 to 1949.

621.315.2 1277

**The Laying of Long-Distance Cables**—M. Troublé. (*Câbles and Trans.* (Paris), vol. 3, pp. 48–65; January, 1949. With English summary.) Historical review of French methods.

621.315.212:621.392.5 1278

**Coaxial Cable with High Characteristic Impedance**—J. A. Hodelin. (*Radio Franç.*, no. 2, pp. 23–24; February, 1949.) A formula is derived for the impedance of a cable with a helix of small diameter for the central conductor. Calculated and measured values of impedance for such delay lines are in good agreement. See also 2009 of 1947 (Zimmermann).

621.315.212:621.397.5 1279

**The London-Birmingham Television Cable: Part 1—General System and Electrical Requirements**—H. Stanesby and W. K. Weston. (*P.O. Elec. Eng. Jour.*, vol. 41, part 4, pp. 183–188; January, 1949.) The cable has two 0.975-inch and four 0.375-inch coaxial tubes. It is designed to transmit very-high-definition or color television, 405-line television and broadband telephony simultaneously. The large tubes may ultimately transmit frequencies up to 30 Mc or more with repeaters at 3 mile intervals.

621.392.2 1280

**Study of Reflections in Transmission Lines at U.H.F.**—M. Bouix. (*Onde Élec.*, vol. 29, pp. 35–43; January, 1949.) The reflections caused by one, and next by two obstacles in the line are considered; an obstacle of a given length is equivalent to a unique concentrated obstacle. The case of many obstacles is then examined. In practice, the frequency is only known approximately, and, even in the case of magnetron oscillations, the frequency may jump or split into two. These differences from the nominal frequency produce phase variations of the waves reflected by obstacles, thus causing variations of the amount of the stationary waves. Assuming such phase variations to be distributed at random, the probability of good matching being obtained is considered. A practical matching method is described.

621.392.2 1281

**Propagation of Plane Electromagnetic Waves in a Multi-Layer Medium of Periodic Structure**—M. L. Levin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1399–1404; November, 1948. In Russian.) The effect of equidistant dielectric spacers on the propagation of electromagnetic waves in cables and waveguides is usually considered in terms of the coefficient of reflection from a single spacer. This procedure is unsatisfactory since the equidistant spacers form a periodic structure causing dispersion of the signal and attenuating certain frequency bands. Accordingly, a dispersion equation (11) is derived, and analyzed in detail for the cases where the wavelength is (a) long, and (b) short in comparison with the distance between the spacers. Methods for determining the frequency bands attenuated are also indicated.

621.392.21 1282

**Simplified Procedure for Computing Behavior of Multiconductor Lossless Transmission Lines**—S. Frankel. (*Elec. Commun.* (London), vol. 25, pp. 286–290; September, 1948.) Summary noted in 2443 of 1948. The wave equation for any lossless system of parallel conductors is derived in terms of the characteristic impedance or admittance coefficients obtained directly from the partial capacitances of pairs of the conductors. Illustrative examples are discussed.

621.392.26† 1283

**Theory of Symmetrical Waves in a Circular Waveguide with an Open End**—L. A. Weinstein. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1543–1564; December, 1948. In Russian.) A mathematical analysis of the propagation of symmetrical electric and magnetic waves in a circular waveguide toward the open end. An integral equation (7) determining the current density in the wall of the waveguide is derived and methods for its solution are proposed. Coefficients of reflection of the waves from the open end are calculated and also the coefficients of transformation of the waves into waves of different modes. Huyghens' principle for calculating the diffraction field is discussed; formulas and graphs for determining the radiation field of the waveguide are derived. Approximate formulas of adequate accuracy involving simple calculations are also given.

621.392.26† 1284

**A Note on Reflection from Dielectric Structures in Wave Guide**—J. Shuoys. (*Jour. Appl. Phys.*, vol. 19, p. 797; August, 1948.) A method of representing the effect of an obstacle in a waveguide by the equivalent capacitive loading on a transmission line, valid for obstacles of small cross section

621.392.26† 1285

**The Calculation of the Critical Frequency of Waveguides with Π or H Cross-Section.**—L. N. Deryugin. (*Radiotekhnika* (Moscow), vol. 3, pp. 49–61; November and December, 1948. In Russian.) General formulas are derived linking the critical frequency of the $H_{10}$ wave with the cross-sectional dimensions of the waveguide. From these formulas, graphs are obtained for use in design work. Some limiting cases are considered separately.

621.392.26† 1286

**The Theory of Disk-Loaded Wave Guides**—W. W. Hansen. (*Atti del Congresso internazionale della Radio* (Rome), pp. 111–132; September and October, 1947. In English.) The phase velocity in a circular waveguide can be reduced to any desired value by introducing equally spaced baffles or disks. The group velocity can also be controlled, and the structure may be regarded as a band-pass filter. The characteristics of such structures are evaluated by various methods and shown graphically.

621.392.26†:621.3.09 1287

**Theory of the Propagation of an E.M. Field along a Dielectric Waveguide of Circular Cross-Section**—M. Abele. (*Atti del Congresso internazionale della Radio* (Rome), pp. 3–13; September and October, 1947. In Italian.) If the waveguide is lossless, two distinct types of electromagnetic modes can be set up in which the longitudinal components $E_l$, $H_l$ of the electric and magnetic field are both present, except in the case of waves of order zero. For all other orders, there is a fundamental mode which can be propagated longitudinally whatever the frequency. All other modes possess a cut-off frequency, below which only transverse propagation is possible.

If the conductivity is small compared to the product of the frequency and the permittivity of the dielectric, the attenuation tends, for increasing frequency, to a limit independent of the mode and of the diameter of the waveguide.

621.392.26†:621.3.09 1288

**Rectangular Waveguides with Several Dielectrics**—M. Abele and C. M. Garelli. (*Atti del Congresso internazionale della Radio* (Rome), pp. 14–29; September and October, 1947. In Italian.) The case of two dielectrics is here considered. Results can easily be extended to the general case, provided there is symmetry about a central plane. In a metallic waveguide, dielectrics can be chosen to obtain any desired field configuration and in particular a plane wave within a given region. Transformers and filters can be constructed in this way. The radiation diagram of a waveguide open at one end can be much altered by the use of dielectrics, and directivity can be greatly increased.

621.392.26†:621.317.335.3† 1289
**Universal Curves for Dielectric-Filled Wave Guides and Microwave Dielectric Measurement Methods for Liquids**—Surber. (*See* 1433.)

621.392.26†:621.392.52 1290
**Maximally-Flat Filters in Waveguide**—Mumford. (*See* 1320.)

621.396.67 1291
**Microwave Beam-Shaping Antennas**—L. J. Chu. (*Atti del Congresso internazionale della Radio* (Rome), pp. 52–64; September and October, 1947. In English.) Optical ray principles are applied to line sources with cylindrical reflectors, line sources with lenses, and point sources with double-curvature reflectors to produce radiation diagrams of any desired shape. The cosecant diagram is especially considered.

621.396.67:538.56.029.8 1292
**Microwave Optics Between Parallel Conducting Sheets**—H. B. DeVore and H. Iams. (*RCA Rev*, vol. 9, pp. 721–732; December, 1948.) Discussion of various microwave systems which focus in one plane only. These range from simple lens-like elements to more elaborate systems involving dielectric elements. They produce a fan-shaped beam of radiation which may be converted to a pencil beam by using cylindrical reflectors or lenses to focus in the perpendicular direction. A method of scanning in this direction is described.

621.396.67:621.396.93 1293
**An Investigation of Resonances and Asymmetry in the Adcock Aerial Systems**—A. Z. Fradin and V. A. Khatskelevich. (*Radiotekhnika* (Moscow), vol. 3, pp. 6–28; November and December, 1948. In Russsian.) A rigorous analysis is given of the resonance phenomena taking place in the H-type Adcock system (Fig. 1), in order to determine the type and magnitude of errors introduced into the system by asymmetry in various elements. The usual practice of identifying the frequency characteristic of the system with that of its input impedance appears in general to be incorrect.

621.396.671 1294
**On the Radiation Resistance of a Radiator Surrounded by a Spherical Magneto-Dielectric Envelope**—A. R. Vol'pert. (*Radiotekhnika* (Moscow), vol. 3, pp. 29–48; November and December, 1948. In Russian.) It is often necessary to reduce the physical length of a linear vibrator without decreasing its radiation resistance. Methods for increasing the effective length of the vibrator have been considered by various authors. Pistol'kors examined the effect of a dielectric envelope in the form of an infinitely long cylinder. Here a spherical envelope is considered, because this shape allows a rigorous solution (34) to be obtained, based on Maxwell's equations and the corresponding boundary conditions. This solution involves only elementary functions. The effects of the permeability and permittivity of the sphere and of its radius on the radiation resistance of the vibrator are examined in detail. The physical interpretation of the results is discussed; it can be extended to other shapes of envelope.

621.396.671 1295
**Mutual Impedance of Two Centre-Driven Parallel Aerials**—P. Starnecki and E. Fitch. (*Wireless Eng.*, vol. 25, pp. 385–389; December, 1948.) A formula and curves are given for symmetrically placed antennas with a sinusoidal current distribution. Measurements at frequencies from 3 to 20 Mc, using vertical antennas above a horizontal earth mat 400 feet in diameter, are in fair agreement with theory.

621.396.671 1296
**Ring-Aerial Systems**—H. Page. (*Wireless Eng.*, vol. 25, p. 402; December, 1948.) Correction to 308 of March.

621.396.671 1297
**The Radiation Patterns of Dielectric Rods—Experiment and Theory**—R. B. Watson and C. W. Horton. (*Jour. Appl. Phys.*, vol. 19, pp. 661–670; July, 1948.) The radiation from rods used as terminations to rectangular waveguides is discussed theoretically. Radiation patterns are computed from the fields due to equivalent magnetic and electric currents on the surfaces of the rods. Specific calculations are made for rods of cross section about 0.9 inch×0.25 inch and 3λ to 10λ long, in the $TE_{0,1}$ mode. Theoretical and experimental patterns for polystyrene rods 3λ to 6λ long are in agreement. Discrepancies for longer rods are discussed.

621.396.679.4 1298
**A Method of Feeding Turnstile Antennas**—R. E. Taylor. (*Electronics*, vol. 22, pp. 164, 170; February, 1949.) To obtain correct phasing, the dipoles are fed by separate coaxial feeders whose physical lengths are equal, but whose electrical lengths differ by λ/4. This is achieved by using different dielectric materials in the feeders. The characteristics of the feeders for the antenna used in telemetering from V-2 rockets (2536 and 3242 of 1947) are discussed.

621.396.679.4:621.392.43 1299
**High-Frequency Balancing Devices**—Meinke. (*See* 1314.)

621.396.671 1300
**Radio Research Special Report No. 16: A Method of Determining the Polar Diagrams of Long Wire and Horizontal Rhombic Aerials ]Book Notice]**—W. R. Piggott. H.M. Stationery Office, London, 1948. 9d. (*Govt. Publ.* (London), p. 14; December, 1948.)

## CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.396.622 1301
**Transient Processes in detecting an E.M.F. with a Linearly Increasing Amplitude**—R. D. Leites. (*Radiotekhnika* (Moscow), vol. 3, pp. 62–75; November and December, 1948. In Russian.)

621.3.018.7:621.396.813 1302
**Trapezoidal Waveform with Minimal Distortion**—W. Bader. (*Frequenz*, vol. 2, p. 208; August, 1948.) Comment on 2470 of 1948 (Päsler), in which "distortion" should be substituted for "noise factor" in both title and text, and whose UDC number should be as above.

621.3.018.7:621.396.813 1303
**A Representation of the Distortion Formula in Closed Form**—M. Päsler. (*Frequenz*, vol. 2, pp. 208–210; August, 1948.) The so-called completeness condition (Parseval's equation), which expresses the sum of the squares of the coefficients in the series development of an arbitrary function $F(z)$ in the form of a definite integral, enables the usual expression for the distortion $\chi$ to be put into a closed form. The general formula is given and applied to the case where $F(z)$ is represented by a Fourier series. When $F(z)$ has certain symmetrical properties, relatively simple formulas for $\chi$ are obtained. See also 2470 of 1948 and 1302 above.

621.314.3† 1304
**Simplified Magnetic Amplifier**—M. Marinescu. (*Wireless Eng.*, vol. 25, p. 402; December, 1948.) Comment on 315 of March (Bell). An alternative type of amplifier, giving balance of output current for zero signal input, proportionality between output and input, and discrimination in the response to inputs of different polarity, is briefly discussed with illustrative diagram.

621.314.3†:623.82 1305
**Magnetic Amplifiers for Shipboard Applications**—L. W. Buechler. (*Elec. Eng.*, vol. 68, pp. 33–37; January, 1949.) Advantages include: ruggedness and reliability; no real limitation on power handled; no warming-up time; power can be taken direct from the 50-cps mains, although speed of response increases with increasing frequency; power gains of 1,000 to 10,000 per stage are possible with response times of 0.005 to 0.1 second; power gains up to $10^8$ can be obtained with response times of 1 to 5 seconds. Design principles and relevant properties of magnetic materials are discussed.

621.318.4 1306
**Calculation of $Q$ for Single-Layer Coils**—F. Benz. (*Elektrotech. und Maschinenb.*, vol. 66, pp. 7–12; January, 1949.) A formula is derived which gives results in good agreement with measured values. The effects of self-capacitance and dielectric losses and the reduction of $Q$ by magnetic screens are discussed. A simple empirical formula is also given for the inductance of single-layer coils.

621.319:679.5 1307
**Electrostatic "Magnets"**—(*Elec. Times*, vol. 114, p. 696; December 9, 1948.) Electrets, dielectric bodies which can retain an electric moment after an externally applied field has been removed, can be made from certain ceramic or plastic materials. The plastics require an emf of 4 to 12 kv for satisfactory electrification. Specimens tested after a year have not changed their electrical properties appreciably.

621.319.4.011.5:519.272 1308
**The Theory of Extreme Values and Its Implications in the Study of the Dielectric Strength of Paper Capacitors**—B. Epstein and H. Brooks. (*Jour. Appl. Phys.*, vol. 19, pp. 544–550; June, 1948.) Statistical theory of the distribution of extreme values is used to deduce the dependence of paper-capacitor breakdown on capacitor size.

621.362 1309
**Thermomagnetic Generator**—L. Brillouin and H. P. Iskenderian. (*Elec. Commun.* (London), vol. 25, pp. 300–311; September, 1948.) A mathematical analysis of an electric generator using a coil wound on a core of high permeability through which a magnetic field is maintained by an external magnet. The core is heated above and cooled below the Curie point, and the resulting changes in flux induce electromotive forces in the coil. Expressions are derived for the optimum power and efficiency of such a generator.

621.385.012 1310
**The Correspondence between the Static Characteristics and the Dynamic Parameters of Electrical Negative-Resistance Devices**—Cartianu. (*See* 1542.)

621.385.012.8 1311
**Equivalent Circuit and Input Admittance of Retarding-Field Valves**—A. Pinciroli and R. Ferrero. (*Alta Frequenza*, vol. 17, pp. 196–211; October, 1948. In Italian, with English, French, and German summaries.) Assuming transit-time effects and control-grid current can be neglected, an expression is derived for the equivalent input admittance of a retarding-field tube used as a negative-transconductance triode. For the case of a resistive output impedance, the equivalent input capacitance and conductance variations are plotted as functions of the ratio of output impedance to the differential resistance of the tube, with the differential resistance as a parameter. Discussion of the case of two distinct output impedances shows that when these consist of resistances of suitable value, the input admittance can be made zero and, within certain frequency limits, independent of frequency.

621.392 1312
**Relation between Amplitude and Phase in Electrical Networks**—T. Murakami and M. S. Corrington. (*RCA Rev.*, vol. 9, pp. 602–631; December, 1948.) A simple graphical method is

presented for computing the phase curve from a given amplitude characteristic, and the attenuation characteristic from a given phase curve in a minimum phase-shift network. A large number of universal curves are given, to simplify the application of the theory.

621.392:517.43 1313
**History of the Operational Calculus as Used in Electric Circuit Analysis**—T. J. Higgins. (*Elec. Eng.*, vol. 68, pp. 42–44; January, 1949.) Bibliography, pp. 44–45.)

621.392.43:621.396.679.4 1314
**High-Frequency Balancing Devices**—H. H. Meinke. (*Fernmeldetech. Z.*, vol. 1, pp. 193–199; November, 1948.) Description of various arrangements for converting from an unsymmetrical feeder, such as a coaxial cable, to the symmetrical type of feeder required for a dipole.

621.392.5 1315
**Equivalent Circuits of Linear Active Four-Terminal Networks**—L. C. Peterson. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 593–622; October, 1948.) The representation of a triode as a linear quadripole and the concept of "effective transconductance" are examined. Electron transit-time effects and circuit behavior were considered independently in earlier theories. The present paper develops a more general theory of active linear quadripoles, in which circuit and electron stream parameters are combined in a single equivalent-circuit representation. It is shown that the equations for current or voltage equilibrium may yield a number of alternative tee or pie networks, the passive elements of which represent faithfully any three of the four independent parameters. The method is applied to grounded-cathode, grounded-grid or grounded-anode triodes, and the complete sets of quadripole parameters are listed, together with approximate parameters and equivalent circuits for moderately low frequencies.

621.392.5 1316
**Generalized Quadripoles**—R. Malvano. (*Atti del Congresso internazionale della Radio* (Rome), pp. 141–159; September and October, 1947. In Italian.) A general theory is developed for the case of axial symmetry, and applied to the study of filters and impedance transformers.

621.392.52 1317
**New Methods of Filter Design by means of Frequency Transformations**—G. Neovius. (*Kungl. Tekn. Högsk. Handl.*, Stockholm, no. 3, 40 pp.; 1946. In English.) Frequency transformations, discussed by Laurent (1339 of 1938), replace the angular frequency by a function thereof, which has the dimensions of radians per second, and also multiply all impedances by a dimensionless function. Such transformations are further examined, and a tabular method of calculation is considered and applied to the following process of filter design: (a) the required attenuation curve of the filter in the real-frequency region is converted to the negative imaginary-frequency region and plotted on a logarithmic scale; (b) an approximation to the curve thus obtained is derived from the attenuation curves of all-pass filter sections; (c) attenuation peaks are retransformed to the real-frequency region; (d) the characteristic frequencies thus obtained are used for designing L-sections which can be connected in cascade to form the requisite filter. Numerical examples are given.

621.392.52:517.53 1318
**Splitting of an Analytical Function into a Linear and a Nonlinear Part, and also a Method of Determining, for a Known Linear Component, the Nonlinear Component and the Function as a Whole**—Pleijel. (*See* 1422.)

621.392.52:517.727 1319
**Application of sn Elliptic Functions to the Calculation of Filters**—K. Steffenhagen. (*Funk und Ton*, vol. 3, pp. 44–47; January, 1949.)

621.392.52:621.392.26† 1320
**Maximally-Flat Filters in Waveguide**—W. W. Mumford. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 684–713; October, 1948.) Good match within the pass band and efficient suppression outside it are achieved by giving the loss characteristic the greatest possible number of zero derivatives at midband. The required loaded $Q$s of the resonant elements are given, and it is shown that sections of waveguide between elements reduce mutual coupling and allow the use of shunt reactances only. Loaded $Q$s are given for resonant cavities formed by various reactive obstacles. Up to 75 obstacles have been used in practice, giving good agreement with theory.

621.392.6:512.831 1321
**Transmission Losses in $2n$-Terminal Networks**—V. Belevitch. (*Jour. Appl. Phys.*, vol. 19, pp. 636–638; July, 1948.) The efficiency matrix is defined for such networks, and its application to the solution of reactive and resistive, transformer-type, and general $2n$-terminal network problems is discussed.

621.394/.395].66+621.315.2 1322
**Construction of Cables and Loading Coils**—Belus. (*See* 1276.)

621.396.611.1:518.4 1323
**Graphical Solution of Oscillation Problems in Circuits with Negative Resistance**—A. Sabbatini. (*Atti del Congresso internazionale della Radio* (Rome), pp. 374–388; September and October, 1947. In Italian.) Discussion of various circuits of the dynatron and arc types. Current versus voltage curves are obtained by a geometrical construction; a method of deducing the wave form is given.

621.396.611.1:621.3.015.3 1324
**The Occurrence of Needle Impulses and their Effect on the Transient Behaviour of Forced Oscillations**—M. Päsler. (*Frequenz*, vol. 2, pp. 322–331; December, 1948.)

621.396.611.1:621.317.6 1325
**Erratum: Response of Linear Resonant Systems to Excitation of a Frequency Varying Linearly with Time**—G. Hok. (*Jour. Appl. Phys.*, vol. 19, p. 623; July, 1948.) Correction to 671 of April.

621.396.611.21:534.6 1326
**Low-Loss Crystal Systems**—W. J. Fry. (*Jour. Acous. Soc. Amer.*, vol. 21, pp. 29–34; January, 1949.) Discussion of the effect of damping or resistive mechanical loads on the characteristics of a piezoelectric crystal system vibrating in either a longitudinal or a thickness mode. Results are applied to the double crystal acoustic interferometer (1257 above).

621.396.611.3:517.942 1327
**On Linear Differential Equations with Slowly Variable Coefficients. Application to the Study of Nonlinear Coupling**—Blanc. (*See* 1423.)

621.396.611.3:621.396.813 1328
**On the Reduction of Phase Distortion in Stages with Coupled Circuits**—J. Laplume. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 1213–1215; December 8, 1948.) The case of two coupled resonant circuits tuned to the same frequency is analyzed in a manner analogous to that used for stagger-tuned circuits (681 of April). It is not found possible to annul the second-degree term in the formula obtained for the phase distortion, but the third-degree term can be annulled by suitable choice of the coupling coefficient $k$. When both circuits have the same $Q$, $k=1/Q\sqrt{3}$. The top of the output amplitude curve is rounded, in agreement with a general rule given previously (40 of February).

621.396.611.4 1329
**The Radiation of Oscillations from a Cavity Resonator through an Aperture as an Analogue of the Tunnel Effect**—P. E. Krasnushkin and E. R. Mustel'. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1378–1393; November, 1948. In Russian.)

621.396.611.4 1330
**On Avoiding Low Frequencies in a Rectangular Cavity Resonator used as Part of a Triode Generator**—K. F. Niessen. (*Atti del Congresso internazionale della Radio* (Rome), pp. 312–329; September and October, 1947. In English.) A mathematical analysis of optimum shapes and proportions for resonators built up from rectangular units to exclude resonances at frequencies below that desired. Other shapes where the cross section is made up of triangles are also discussed.

621.396.611.4.029.64 1331
**Design of a Resonant Cavity for Frequency Reference in the 3 cm Range**—R. R. Reed. (*Atti del Congresso internazionale della Radio* (Rome), pp. 364–373; September and October, 1947. In English.) Discussion of the design and performance of a temperature-compensated resonator, suitable for quantity production, having a frequency error of less than 1 Mc and a loaded $Q$ of 2,000.

621.396.615+621.396.645 1332
**Limits of Tube-Gain and Power Output at Extreme-High Frequencies**—M. J. O. Strutt. (*Atti del Congresso internazionale della Radio* (Rome), pp. 655–667; September and October, 1947. In English.) Amplifier and oscillator stages without feedback are discussed. A formula for the optimum gain is derived. The importance of the "border" frequency at which this gain drops to unity is stressed; the gain formula is transformed in various ways according to the type of reception tube to which it is applied. Grounded-cathode and grounded-grid stages are compared. The theory does not cover traveling-wave tubes, with which an effective bandwidth of 800 Mc at 3,300 Mc has been successfully amplified. Similitude rules for determining the frequency dependence of the power output of power tubes at uhf are tabulated and discussed.

621.396.615 1333
**High-Stability Pilot Generator for the 3.5-Mc/s and 7-Mc/s Bands**—L. Liot. (*Radio Franç.*, pp. 21–23; January, 1949.) An electron-coupled oscillator using a single EL39 tube. The tuned-plate output circuit is isolated from the oscillator by connecting the suppressor grid directly to earth. Both high- and low-impedance outputs are provided. Performance figures for different load resistances are tabulated.

621.396.615:621.396.645 1334
**The Theory of the Oscillator using a Cathode-Coupled Amplifier**—G. A. Khavkin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1416–1420; November, 1948. In Russian.)

621.396.615.17 1335
**Aperiodic Frequency Doubler for a Sinusoidal Electric Quantity**—M. Nuovo. (*Atti del Congresso internazionale della Radio* (Rome), pp. 330–335; September and October, 1947. In Italian.) Discussion of a doubler which uses a single tube with two control grids. The doubler can operate at all frequencies for which this tube can amplify. The output is practically free from unwanted harmonics.

621.396.615.18 1336
**Quasi-Aperiodic Divider for Low Acoustic Frequencies**—E. Gatti. (*Atti del Congresso internazionale della Radio* (Rome), pp. 297–301; September and October, 1947. In Italian.) A very stable divider has been used for over a year to reduce a standard frequency of 100 kc to 1,000 cps. Associated apparatus to extend the reduction to 50 cps is discussed. See also 3687 of 1939 (Miller) and 1598 of 1948.

621.396.615.18 1337
**Synchronous Frequency Division**—P. G. Bordoni. (*Atti del Congresso internazionale*

*della Radio* (Rome), pp. 252–272; September and October, 1947. In Italian.) By studying mechanical oscillatory systems whose parameters vary periodically, analogous electrical circuits can be realized for frequency halving in which instability is impossible. Practical experience with two different circuits confirms theoretical predictions and shows that considerable voltage amplification and good wave form can be obtained together. Only one triode is needed for each stage of division.

621.396.645 1338
**Note on the Sensitivity of an Amplifier Stage**—W. Kleen. (*Ann. Radioélec.*, vol. 3, pp. 299–301; October, 1948.) Calculation shows that the signal-to-noise ratio of a grounded-grid circuit is given by the same expression as for a grounded-cathode circuit. For a given tube, therefore, the optimum sensitivities of two such amplifier stages should be the same.

621.396.645 1339
**The Quasistationary Wave Amplifier**—W. Kleen. (*Elektrotechnik* (Berlin), vol. 2, pp. 341–342; December, 1948.) A short discussion of distributed amplification, based on the account given by Ginzton et al. (3375 of 1948.)

621.396.645 1340
**Study of Amplifiers for Square-Wave Signals**—H. Gilloux. (*Radio Franç.*, pp. 6–12; January, 1949.) Methods of producing square-wave signals are briefly considered and the response of an amplifier stage to such signals is discussed with particular reference to circuit time constants. Practical design details are given for *RC*-coupled and cathode-coupled circuits. Low-frequency and high-frequency compensation methods are outlined and transformer coupling and the use of feedback are discussed. These design principles have been used for amplifiers with outputs of the order of 20 and 50 watts respectively, both having only 0.5 per cent distortion at full power and passing square-wave signals of frequencies from 25 to 15,000 cps.

621.396.645.029.3 1341
**A Miniature Audio-Frequency Amplifier**—W. T. Duerdoth and J. Garlick. (*P. O. Elec. Eng. Jour.*, vol. 41, part 4, pp. 228–233; January, 1949.) Discussion of the design techniques required for a line amplifier occupying a space of $2\frac{1}{2}\times2\frac{1}{2}\times4\frac{1}{2}$ inches. More than one feedback path is used.

621.396.645.37 1342
**Group Transmission Time in Feedback Amplifiers**—G. Schaffstein. (*Frequenz*, vol. 2, pp. 291–295; November, 1948.) The variations of phase and group transmission time due to feedback are calculated. With pure positive or negative feedback, the phase difference between input and output voltages is constant. The phase variation is greatest when the feedback voltage is 90° out of phase with respect to the input voltage. Variations of the group transmission time occur for both positive and negative feedback and for 90° phase displacement between feedback and input voltages. For pure phase feedback, the transmission-time variation is proportional to $k$, the ratio of feedback and input voltages, but for 90° phase displacement between feedback and input voltages the variation is proportional to $k^2$. For a single-stage amplifier, the gain and group transmission time vary in the same way. For a multistage amplifier, the transmission-time variations for the same gain variation differ according to whether (a) each of the $n$ stages has its own feedback, or (b) feedback is applied between input and output. In the latter case, the variations are $n$ times larger. Experiments with a 2-stage amplifier confirmed the theory.

621.396.645.371 1343
**Linear Power Amplifiers in American Broadcasting**—W. H. Doherty. (*Atti del Congresso internazionale della Radio* (Rome), pp. 280–291; September and October, 1947. In English.) Discussion of: (a) operating principles and practical adjustment of the Doherty amplifier, (b) the application of negative feedback to increase linearity, and (c) the advantages of this system, particularly for high-power broadcasting transmitters. See also 378 of 1940 (Doherty and Towner).

621.397.645 1344
**Stagger-Peaked Video Amplifiers**—A. Easton. (*Electronics*, vol. 22, pp. 118–120; February, 1949.) The use of stagger tuning to obtain a desired amplitude versus frequency characteristic was discussed in 2491 of 1948 (Wallman). A similar general principle is here applied to video amplifiers.

621.3.012:621.38.001.8 1345
**Handbook of Industrial Electronic Circuits [Book Review]**—J. Markus and V. Zeluff. McGraw-Hill, London, 272 pp., 39s. (*Wireless Eng.*, vol. 25, p. 400; December, 1948.) The book contains circuit diagrams, usually with typical component values and short explanatory notes. The subjects covered include af capacitance control, cathode-ray control, counters, dc amplifiers, electronic switches, limiters, measurement, metal location, motor control, multivibrators, oscillators, photoelectric circuits, power supplies, stroboscopy, telemetry, temperature control, timing, ultrasonics, voltage regulation, and welding. Only American work is mentioned.

## GENERAL PHYSICS

53.081+621.3.081 1346
**The Rationalized Giorgi System in the Theory of Electricity**—P. Cornelius. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 1–9; January, 1949. In Dutch, with English summary.) Reasons are advanced for the adoption of the Giorgi system with the absolute volt and ampere. It enables the relations between electrical quantities and mechanical forces to be expressed in a particularly simple form. Maxwell's laws are presented in rationalized units.

535.215.3 1347
**The Capacitor Method for Investigations of Photo-E.M.F.**—S. M. Ryvkin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1521–1542; December, 1948. In Russian.)

535.343.4:535.61-15 1348
**The Infra-Red Spectra of Atmospheric Gases other than Water Vapour**—G. B. B. M. Sutherland and G. S. Callendar. (*Rep. Progr. Phys.*, vol. 9, pp. 18–28; 1942 and 1943.)

535.343.4:535.61-15:546.212 1349
**The Absorption of Water Vapour in the Far Infra-Red**—T. G. Cowling. (*Rep. Progr. Phys.*, vol. 9, pp. 29–41; 1942 and 1943.)

536.3 1350
**The Efficiency of Radiation Shields**—A. E. De Barr. (*Rev. Sci. Instr.*, vol. 19, p. 922; December, 1948. Comment on 1022 of May (Garrison and Lawson).

537.122:538.3 1351
**General Properties of the Paraxial Electron-Optic Image**—F. Borgnis. (*Helv. Phys. Acta*, vol. 21, pp. 461–479; December 20, 1948. In German.) Discussion, for fields of axial symmetry, based on the differential equation for an electron path near the axis.

537.2 1352
**Potential and Field of a Plane Electrode with an Elliptical Hole**—M. Cotte. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 377–378; January 31, 1949.) Exact formulas are derived.

537.311.62 1353
**Calculation of the Magnetic Skin Effect in Steel Sheets taking into account the Relationship between Magnetic Permeability and Magnetic Field Intensity**—S. D. Margolin. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1306–1316; October, 1948. In Russian.)

537.311.62 1354
**The Theory of the Anomalous Skin Effect in Metals**—G. E. H. Reuter and E. H. Sondheimer. (*Proc. Roy. Soc. A*, vol. 195, pp. 336–364; December 22, 1948.) The problem of high-frequency metallic conduction is reviewed and exact solutions are obtained which are valid at all frequencies and temperatures. For large values of the free path of the conduction electrons, the electric field is propagated through the metal as a "surface wave" which differs considerably from the classical exponential solution. For the temperature variation of the surface impedance in the microwave range, Pippard's simplified theory is qualitatively correct. The frequency variation of the surface impedance at low temperatures is also discussed; relaxation effects are negligible in the microwave range but become important in the infrared and eventually restore the validity of the classical theory. Theory predicts that the reflection coefficient of metals should pass through a minimum in the far infrared. See also 1014 of 1948 (Pippard).

537.312.62 1355
**Superconductivity**—K. Mendelssohn. (*Rep. Progr. Phys.*, vol. 10, pp. 358–375; 1944 and 1945. Bibliography, pp. 375–377.) A general review, with particular reference to Russian and German work.

537.533:530.145.6:621.385.833 1356
**Field Emission of Electrons**—R. O. Jenkins. (*Rep. Progr. Phys.*, vol. 9, pp. 177–197; 1942 and 1943.) A survey of experimental work begins with that of Lilienfeld and includes the development of the field-emission electron microscope. The application of wave mechanics is discussed.

538.22 1357
**Modern Theory of Magnetism: Part 2**—S. V. Vonsovski. (*Uspekhi Fiz. Nauk*, vol. 36, no.1 pp. 30–82; 1948. In Russian.) Discussion of diamagnetic and paramagnetic materials and magnetic cooling.

538.3 1358
**The Field of a Point Source of Current located on the Surface of the Earth above an Inclined Layer**—I. P. Skal'skaya. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1242–1254; October, 1948. In Russian.)

538.3 1359
**Classical Electrodynamics without Singularities**—H. McManus. (*Proc. Roy. Soc. A*, vol. 195, pp. 323–336; December 22, 1948.) It is possible to construct a theory of the electron with an extended charge distribution in a Lorenz invariant way by introducing a 4-dimensional form function. Electromotive field quantities reduce to those given by the ordinary theory at distances large compared with the electron radius, but remain finite on the world line. See also 698 of 1948 (Eliezer).

538.6:537.525 1360
**The Influence of a Transverse Magnetic Field on a Cylindrical Plasma**—L. Beckman. (*Proc. Phys. Soc.*, vol. 61, pp. 515–520; December 1, 1948.)

621.39.001.11 1361
**A Mathematical Theory of Communication**—C. E. Shannon. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 379–423 and 623–656; July and October, 1948.) The methods of statistical mechanics are applied to determine rates for the efficient transmission of information in generalized coded forms. The properties of discrete and continuous sources and the effect of noise are considered. Entropy of the source and capacity of the channel are defined, and it is shown that statistical matching of channel to source is required. By suitable choice of coding, errors in transmission may be made as small as desired, provided the rate does not exceed the channel capacity. For a continuous

source, the maximum rate is expressed in terms of a measure of required fidelity.

621.396.822 1362
**The Energy of Electrical Fluctuations in Conductors**—E. Ya. Pumper. (*Zh. Eksp. Teor. Fis.*, vol. 18, pp. 1112–1129; December, 1948. In Russian.) The limitations of Nyquist's formula (p. 1112) determining the thermal effect are discussed. Experiments were designed to determine the distribution of probabilities of electrical fluctuations in conductors at frequencies from 4 to 136 kc, and for the absolute measurement of fluctuational energy. The main conclusions reached are: (a) In many thin metallic and graphite conductors, the noise level exceeds the value calculated from the thermal effect theory; the discrepancy increases with the frequency. (b) Experiments with an electrolyte (a weak solution of KCl) and annealed resistances have shown that other variations, besides those due to thermal effect, occur more rarely and disturb the distribution of probabilities. (c) Experiments with heated resistances have shown that additional fluctuations can be eliminated by rapid annealing. (d) There are reasons to suppose that the excess of the fluctuation level over the theoretical value is due to emission of energy from the crystal lattice and is related to the process of crystallization.

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

522.56:523.7 1363
**Physical Basis of Solar Research and Technical Aids**—V. von Keussler. (*Z. Angew. Phys.*, vol. 1, pp. 232–242; November, 1948.) A review, for the last 20 years, of research methods and apparatus for optical wavelengths. 72 references are included.

523.53:551.510.535 1364
**The Exploratory Properties of Radio Waves and Their Application to the Detection of Meteor Trails**—R. Naismith. (*Atti del Congresso internazionale della Radio* (Rome), pp. 160–167; September and October, 1947. In English.) The transient echoes observed near the sporadic-*E* region are due to reflections from the ionization trails of meteors. There is a marked seasonal variation in the number of meteors recorded around noon. The response of the medium also has a seasonal variation. In temperate latitudes, sporadic-*E* ionization is mainly due to the fine dust of meteors. The atmosphere is continually being bombarded with meteoric dust particles, most of which are too small to produce visible meteor trails.

523.53:621.396.9 1365
**The Theory of the Radio Detection of Meteors**—Manning. (*See* 1392).

523.72:523.854 1366
**The Radio Spectrum of the Sun and of the Milky Way**—C. E. Krüger. (*Atti del Congresso internazionale della Radio* (Rome), pp. 133–140; September and October, 1947. In Italian.) Discussion and theoretical interpretation of results obtained by various authors. See also 1028 of 1945 (Reber) and 3270 of 1946 (Hey, Parsons, and Phillips).

523.72+523.854]:621.396.822 1367
**Radio Astronomy**—C. R. Burrows. (*Electronics*, vol. 22, pp. 75–79; February, 1949.) A general review of knowledge obtained from solar and galactic rf radiation measurements, and of possible future developments.

523.72.029.62 1368
**Some Characteristics of Solar Radio Emissions**—J. S. Hey, S. J. Parsons, and J. W. Phillips. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 5, pp. 354–371; 1948.) Continuation of experimental investigations noted in 1825 of 1946 (Hey) and 3508 of 1947 (Appleton and Hey). Discussion of the relation between solar radio emissions observed mainly at $\lambda = 4.1$ meters between March, 1946 and September 1947, and associated visual solar and geophysical phenomena. The most intense flares are the most likely to produce radio bursts; in general, the radio emission lags several minutes behind the visual or ultra-violet flare radiations. The heliographic distribution of coincidences between flares and bursts is considered and contrasted with the sharply beamed pattern of continuous radiation associated with sunspots.

523.745:550.385/.386 1369
**Geomagnetic Disturbances and Solar Activity**—A. I. Ol'. (*Priroda*, pp. 3–10; July, 1948. In Russian.) Discussion of corpuscular solar radiation and its connection with disturbances of the earth's magnetic field, with special reference to the solid angle of the radiation, the "M-regions" in the sun which emit the radiation, observations of the absorption lines of ionized Ca in the solar spectrum during magnetic storms, and the two types of corpuscular radiation thought to cause the two classes of magnetic storms.

523.746:550.38 1370
**Observational Aspects of the Sunspot-Geomagnetic Storm Relationships**—H. W. Newton. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 5, p. 423; 1948.) Summary only. Using international magnetic character figures, the statistical rise of geomagnetic activity is investigated for the solar disk passage of four area groupings of sunspots, between 1914 and 1944. For large sunspots, there is an increase of geomagnetic activity centered at about 2 days after the central meridian passage of the sunspots. For smaller sunspots, groupings based on solar flare incidence are better correlated with geomagnetic activity than those based on area. 27-day recurrence tendencies of geomagnetic activity are not well defined except in the case of the smaller storms, especially those without associated sudden commencements, sunspots, or solar flares. The statistical nonrecurrence of great magnetic storms is confirmed.

523.746 "1941" 1371
**Mean Areas and Heliographic Latitudes of Sunspots in the Year 1941**—(*Mon. Not. R. Astr. Soc.*, vol. 108, no. 5, pp. 420–422; 1948.)

523.746.5 1372
**Double Cycle of Solar Activity**—A. I. Ol'. (*Priroda*, pp. 39–41; August, 1948. In Russian.) Additional confirmation of the existence of the 22-year cycle of solar activity suggested by M. N. Gnevyshev and A. I. Ol'. (*Astronomicheski Zhurnal*, vol. 25, no. 1. 1948. In Russian.)

523.76 1373
**Solar Physics**—A. Hunter. (*Rep. Progr. Phys.*, vol. 9, pp. 101–112; 1942 and 1943.)

538.12:521.15 1374
**Magnetic Field of Massive Rotating Bodies**—Yu. G. Pliner. (*Priroda*, pp. 16–24; July, 1948. In Russian.) The implications of Blackett's theory (3112 of 1947) are discussed and the following three methods of its verification are considered in turn: (a) application of the theory to stars, (b) extension of the theory to rotating bodies on the earth, and (c) extension of the theory to elementary particles such as Bohr's magnetron. The possibility is suggested that the magnetic field of the moon could be measured by means of rockets which would overcome the force of gravity of the earth and be attracted by the moon while sending radio messages back to the earth. Experiments conducted by P. N. Lebedev in 1909 are mentioned in connection with (b).

551.5:621.396 1375
**Applications of Radio in Modern Meteorology**—R. Bureau. (*Atti del Congresso internazionale della Radio* (Rome), pp. 683–695; September and October, 1947. In French.) Various radio techniques have combined to increase enormously the quality and quantity of information available for the forecaster.

551.51:541.14 1376
**Photochemical Processes in an Oxygen-Nitrogen Atmosphere**—C. H. Bamford. (*Rep. Progr. Phys.*, vol. 9, pp. 75–89; 1942 and 1943. Bibliography, pp. 89–91.)

551.51:541.14:546.21 1377
**The Photochemistry of Atmospheric Oxygen**—S. Chapman. (*Rep. Progr. Phys.*, vol. 9, pp. 92–100; 1942 and 1943.) An outline of the attempts made to account for the existence of ozone and atomic oxygen. The diurnal and seasonal changes in the ozone content of the atmosphere are discussed.

551.510.5:537.56 1378
**Ionization Phenomena in the Earth's Atmosphere**—J. Sayers. (*Rep. Progr. Phys.*, vol. 9, pp. 52–61; 1942 and 1943.) For the first 20 km above sea level the ionization balance is simple and measurements of conductivity confirm the theoretical values. The balance equations for free electrons and for ions in the ionosphere are developed. The effective recombination coefficients for the *E* and *F* regions are discussed and compared with values deduced from radio observations

551.510.5:546.21 1379
**The Properties of Neutral and Ionized Atomic Oxygen and their Influence on the Upper Atmosphere**—H. S. W. Massey and D. R. Bates. (*Rep. Progr. Phys.*, vol. 9, pp. 62–74; 1942 and 1943.) The energy levels of positive and negative ions and neutral atoms of oxygen are briefly considered and the probabilities of various ionization, recombination, and attachment processes are discussed. In the ionosphere, the observed recombination coefficients are too large to be accounted for by direct radiative recombination of electrons and ions; the presence of negative ions must be assumed.

551.510.535 1380
**The Present State of Investigations of the Ionosphere: Part 2**—Ya. L. Al'pert. (*Uspekhi Fiz. Nauk*, vol. 36, no. 1, pp. 1–29; 1948. In Russian.) The formation of ionized layers and the processes taking place in them are discussed.

551.510.535 1381
**The Analysis of Ionospheric Reflections: Part 1**—A. H. de Voogt. (*Tijdschr. ned. Radiogenoot.*, vol. 13, pp. 83–195; November, 1948. In English. Corrections, *ibid.*, vol. 14, January, 1949; insert.) Formulas for the complex refractive index and polarization are based on the Appleton-Hartree formula; curves are given representing the results of careful calculations, by the mathematical bureau of the Dutch Post Office, of refractive index, group velocity, attenuation, and polarization for a frequency of 6 Mc, assuming conditions similar to those over the Netherlands. Experiments to verify these results will be described in part 2.

551.510.535:525.624 1382
**The Lunar Atmospheric Tide at Twenty-Seven Stations Widely Distributed over the Globe**—S. Chapman and K. K. Tschu. (*Proc. Roy. Soc. A*, vol. 195, pp. 310–323; December 22, 1948.) The tides were determined by the Chapman-Miller method from bi-hourly values of barometric pressure covering periods of 5 years or more at each station. Results are discussed.

551.510.535:621.317.79 1383
**Ionosphere Reflections Recorded Mechanically by Means of a Repetition Frequency Converter**—Stoffregen. (*See* 1438.)

551.510.535:621.396.11 1384
**Regular Ionosphere Observations over Mid-Germany**—Dieminger. (*See* 1475.)

551.510.535:621.396.11 1385
**Ionosphere Review: 1948**—T. W. Bennington. (*Wireless World*, vol. 55, pp. 56–60; February, 1949.) A survey of sunspot and short-wave propagation conditions for 1948, with forecast for 1949.

551.593.9 1386
**The Spectrum of the Night Sky**—R. W. B. Pearse. (*Rep. Progr. Phys.*, vol. 9, pp. 42–51; 1942 and 1943.)

551.593.9 1387
**Variations of the Colour of the Night Sky from March 1947 to May 1948**—R. Grandmontagne and C. Delestrade. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 415–416; January 31, 1949.) Results obtained with a photoelectric recorder at the Haute-Provence observatory are tabulated and discussed.

551.594 1388
**Electric Field Intensity Inside of Natural Clouds**—R. Gunn. (*Jour. Appl. Phys.*, vol. 19, pp. 481–484; May, 1948.) Maximum intensities measured by instruments carried on aircraft flying through the clouds were of the order of 3 kv per centimeter. This is much higher than previous values based on free-balloon soundings. See also 2425 of 1947.

551.594.21 1389
**Propagation of Atmospherics**—H. Norinder. (*Atti del Congresso internazionale della Radio* (Rome), pp. 168–195; September and October, 1947. In English.) Discussion of initial results of a projected long-term investigation. Two mobile stations and a fixed station near Uppsala (Sweden) were almost in line, with intervals of 27 and 37 km. Numerous oscillograms of the received electric field are shown; a striking feature is a change of polarity beyond a certain distance, which may partly be due to reflection from ionized layers at a great height in the thunderstorm. Simultaneous observations of atmospherics at two fixed stations 200 km apart are also discussed; the importance of direction finders at these stations is emphasized. See also 3907 of 1947 and back references.

551.594.21:621.396.11 1390
**Guided Propagation of Radar in Thunderstorm Conditions**—Coons. (*See* 1478.)

551.594.221 1391
**The Lightning Discharge**—J. M. Meek and F. R. Perry. (*Rep. Progr. Phys.*, vol. 10, pp. 314–354; 1944 and 1945. Bibliography, pp. 354–357.) The formation of thunderclouds and the mechanism of the lightinng discharge are discussed from the theoretical aspect and accounts of experiments using a Boys' camera are given. The technique of the measurement of the current and voltage of a discharge is considered generally. An electrometer to measure atmospheric voltage gradients is described and results obtained with it are given. Future work is discussed.

## LOCATION AND AIDS TO NAVIGATION

621.396.9:523.53 1392
**The Theory of the Radio Detection of Meteors**—L. A. Manning. (*Jour. Appl. Phys.*, vol. 19, pp. 689–699; August, 1948.) Methods of observation for both bursts of signal strength and whistles caused by reflection of radio waves from meteoric ionization are discussed. Meteor velocity and range can be accurately determined by combined burst and whistle observations. The possibility of determining the direction and position of a meteor path by observations at 3 receiving stations is examined. The percentage of meteors capable of producing bursts depends on the angle between the meteor path and the vertical and on the direction from which signals are received; various possible formulas for it are discussed.

621.396.93 1393
**Recent Progress in Radio Direction Finding**—R. L. Smith-Rose. (*Atti del Congresso internazionale della Radio* (Rome), pp. 877–901; September and October, 1947. In English.) A detailed survey of methods used in the frequency range 10 kc to 1,200 Mc. Antenna systems, recent advances in instrumental technique and calibration, accuracy limitations, and the effect of the ionosphere on the accuracy of bearings are discussed. Methods of direction-finding transmission include the rotating-loop beacon, the equi-signal beacon and the Consol system. Future trends in high-frequency direction finders include the use of highly directional receiving systems for the reduction of site errors, more accurate transfer of angular data from the direction finder to the plotting center, and methods to enable the operator to assess the approximate bearing accuracy.

621.396.93 1394
**Automatic Direction-Finder**—J. R. Steinhoff. (*Electronics*, vol. 22, pp. 97–99; February, 1949.) A light-weight and compact instrument having bearing accuracy within 1°. The externally mounted loop assembly is stationary and hermetically sealed; it consists of 4 coils arranged as two pairs at right angles. A special switching system enables each coil voltage to be sampled in succession 50 times a second and transmitted to the corresponding coil of the indicating meter, which is calibrated in degrees. Block and circuit diagrams are included.

621.396.93 1395
**The Interconnection of Dead-Reckoning and Radar Data for Precision Navigation and Prediction**—B. Chance. (*Jour. Frank. Inst.*, vol. 242, pp. 355–372; November, 1946.) The importance of accurate dead-reckoning data for both ship and aircraft navigation is stressed. Methods of providing an index on a radar display which continuously indicates the dead-reckoning position of an identifiable object are discussed. The solution of problems involving the relative motion of two ships, wind velocity, or ground range in antenna navigation can thus be greatly simplified. See also 1396 below.

621.396.93 1396
**G.P.I. [ground position indicator]—An Automatic Navigational Computer**—W. J. Tull and N. W. MacLean. (*Jour. Frank. Inst.*, vol. 242, pp. 373–398; November, 1946.) A device indicating continuously the present position on the earth's surface of the ship or aircraft in which it is carried. Wind velocity and velocity relative to the ground are determined from data provided by a true airspeed meter and the radar and compass. The total distance traveled is presented numerically and the heading required in order that a particular course may be followed is also indicated. The computing mechanism used is described in detail and its operation is explained. The main types of error due to the instruments which supply the GPI with data are discussed. Applications are considered. See also 1395 above.

621.396.93:621.396.67 1397
**An Investigation of Resonances and Asymmetry in the Adcock Aerial Systems**—Fradin and Khatskelevich. (*See* 1293.)

621.396.932 1398
**Radar for Tilbury-Gravesend Ferry**—(*Engineer* (London), vol. 186, p. 595; December 10, 1948.) Scanner unit 80 feet above ground; frequency 9,425 to 9,525 Mc; output power 30 kw; pulse width 0.2 microsecond and repetition rate 2,000 pulses per second; 2-way radio telephone communication between the radar operators and the masters of the ferries; PPI display with ranges of 0.8, 1.2, or 3 miles.

621.396.933 1399
**Review of Recent Trends in Radio Aids to Air Navigation**—H. Busignies, P. R. Adams, and R. I. Colin. (*Atti del Congresso internazionale della Radio* (Rome), pp. 696–716; September and October, 1947. In English.) A comprehensive account of basic principles and development work, with a brief description of various existing systems. International recommendations are summarized.

621.396.933 1400
**Aerial Radio Navigation in Wartime**—Guigonis. (*Onde Élec.*, vol. 29, pp. 21–25; January, 1949.)

621.396.933 1401
**Decca Navigator System of Hyperbolic Navigation**—P. Giroud and L. Couillard. (*Onde Élec.*, vol. 29, pp. 5–20; January, 1949.) A detailed discussion of basic principles, method of operation, practical equipment, and applications.

621.396.933.2 1402
**High-Stability Radio Distance-Measuring Equipment for Aerial Navigation**—H. Busignies. (*Elec. Commun.* (London), vol. 25, pp. 237–243; September, 1948.) A performance specification with brief technical description of a 51-channel interrogator-responder system. Challenging and responding frequencies, which are crystal controlled, are separated by 125.5 Mc; both are in the antenna-navigation band of 960 to 1,215 Mc. Adjacent channels may be used by different beacons, and one beacon can deal with up to 50 aircraft. Distances up to 100 nautical miles are indicated on a meter with an accuracy within ± 0.2 mile or 1 per cent whichever is the greater.

## MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 1403
**On High-Vacuum Technique**—G. W. Oetjen. (*Elektrotechnik* (Berlin), vol. 2, pp. 333–340; December, 1948.) A review of modern methods and equipment, with particular reference to pumping rates, the different oils used in oil diffusion pumps, high-vacuum tubes, gauges, etc.

535.37 1404
**The Relation between Efficiency and Exciting Intensity for Zinc-Sulfide Phosphors**—H. A. Klasens, W. Ramsden, and Chow Quantie. (*Jour. Opt. Soc. Amer.*, vol. 38, p. 649; July, 1948.) Corrections to 3422 of 1948.

535.37:621.385.832 1405
**Silicate Phosphors for Cathode Ray Tubes**—P. N. Campbell. (*Jour. Soc. Chem. Ind.* (London), vol. 66, pp. 191–194; June, 1947.) The object of the experimental work described was to produce a phosphor which would give a white screen suitable for television. A mixture of powders to produce the correct color has been found but the brilliance is not yet satisfactory. Possible future developments are indicated. "Settling" is thought to be the most effective and economical method of preparing screens on a large scale.

537.228.1 1406
**Recent Progress in Piezoelectric Technique**—A. Tournier. (*Atti del Congresso internazionale della Radio* (Rome), pp. 485–518; September and October, 1947. In French.) An illustrated account of the production of quartz crystals from silica at the Laboratoire Central de Télécommunications. Crystallization of other salts can be accelerated by rotation of the crystal during formation and by addition of crystal impurities. Piezoelectric characteristics of rochelle salt and other crystals are discussed in detail. Certain phosphates and arsenates of ammonium or potassium can be used to obtain directly a bandwidth of 4 kc at a carrier frequency of 80 kc, but they have a large characteristic impedance and

temperature coefficient of frequency. See also 2811 of 1948.

538.13+538.221 1407
**The Magnetic After-Effect of Different Types of Silicon-Iron**—H. Wilde and G. Bosse. (*Frequenz*, vol. 2, pp. 214–215; August, 1948.) The effect of temperature and of frequency on the complex permeability of three grades of material, containing respectively 4 per cent, 3.5 per cent, and 2.5 per cent Si, is shown graphically. Curves are also given for the permeability for various ac fields.

538.221 1408
**Some New Ferromagnetic Manganese Alloys**—F. A. Hames and D. S. Eppelsheimer. (*Nature* (London), vol. 162, p. 968; December 18, 1948.) The results of an investigation on three Cu-Mn-Ga alloys are given. Two specimens were strongly magnetic when quenched from 750° C. X-ray diffraction data are tabulated, but cannot as yet be satisfactorily interpreted.

538.221:538.213 1409
**Ferro-Magnetics at Ultra-High-Frequencies**—M. J. O. Strutt. (*Atti del Congresso internazionale della Radio* (Rome), pp. 448–459; September and October, 1947. Bibliography, pp. 459–462. In English.) It seems probable that in the range of very low frequencies (from zero to a few kc) the effective permeability declines from several thousand to several hundred for certain very pure ferromagnetic materials. Anomalies found by some experimenters near 3 Mc are believed to be spurious. At room temperatures, a marked decline from several hundred to about unity occurs in the frequency range 300 to 30,000 Mc; it is suggested that the increased conductivity at low temperatures would have the same effect as a corresponding increase of frequency. Even slight differences in treatment of a well-defined material may result in large differences in measured permeability values; hence the entire range 0 to 30,000 Mc should be investigated in one laboratory. A theoretical explanation of the fall in permeability is given in terms of eddy-current effects.

538.244 1410
**Magnetization in Perpendicularly Superposed Direct and Alternating Fields**—F. J. Beck and J. M. Kelly. (*Jour. Appl. Phys.*, vol. 19, pp. 551–562; June, 1948.) General investigation of the behavior of an iron specimen. Curves of the dynamic longitudinal magnetic induction versus transverse magnetic field intensity and transverse magnetic induction are included.

621.315.59 1411
**Properties of Poorly Conducting Layers between Metals and Semiconductors**—V. E. Lashkarev. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1347–1355; November, 1948. In Russian.)

621.315.59:621.314.63 1412
**Rectification at the Boundary of Two Semiconductors**—A. V. Ioffe. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1498–1510; December, 1948. In Russian.) A detailed report upon an experimental investigation. For rectification, the two semiconductors in contact must have different types of conductivity and there must be a difference of contact potentials. The hole contact potential should be higher than the electron potential. The following processes are also of importance in rectification: (a) the formation of barrier layers of low conductivity, (b) variations in their thickness during the passage of the current, (c) increase in the conductivity of semiconductors under the influence of the electric field, and (d) transfer of charges through the boundaries of the semiconductors.

621.315.612 1413
**New Ceramic Materials with Very High Dielectric Constant**—A. Pascucci and H. Wolf-Stawski. (*Atti del Congresso internazionale della Radio* (Rome), pp. 336–363; September and October, 1947. In Italian.) The work of various authors is reviewed and an experimental investigation into the relationship between dielectric constant and loss angle for materials containing titanates of Ba and Sr is described. Applications are discussed.

621.315.612.011.5 1414
**Variation of the Dielectric Properties of Ceramic Materials with Magnesium-Orthotitanate Base and Its Representation by the Logarithmic Law of Mixtures**—E. Albers-Schoenberg and W. Soyck. (*Ann. Radioélec.*, vol. 3, pp. 290–292; October 1948.) Results for materials containing titanates of Mg and Ca or Sr show the logarithmic law to hold for the dielectric constant, which increases from about 20 for 7 per cent of $CaTiO_3$ to 120 for 85 per cent of $CaTiO_3$, the corresponding temperature coefficients being $-10\times10^{-6}$ per °C and $-1,600\times10^{-6}$ per °C.

621.318.2 1415
**Permanent Magnets**—W. E. Burnand. *Elec. Rev.*, (London) vol. 144, pp. 63–67; January 14, 1949.) Discussion of economical methods of magnetization for magnets of awkward shapes.

621.318.22 1416
**Materials for Permanent Magnets**—S. Wintergerst. (*Funk und Ton*, vol. 3, pp. 48–50; January, 1949.) The composition and some magnetic and physical properties of 12 alloys are tabulated and discussed.

621.318.22 1417
**Magnetic Materials: a Review of Progress**—F. Brailsford: D. A. Oliver and D. Hadfield: G. R. Polgreen. *Jour. IEE* (London), part I, vol. 95, pp. 522–543; December, 1948.) Part 1: Electrical sheet steels. Part 2: Permanent magnet materials. Part 3: Magnetic powder cores.

621.775.7:538.221 1418
**Verification of Néel's Theory for the Coercive Field of Finely Powdered Ferronickel Alloys**—L. Weil. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 1347–1349; December 20, 1948.) Measurements on alloys containing 50 to 90 per cent Ni confirm that the mean coercive field for agglomerated powders with little or no magnetic anisotropy is directly proportional to the magnetization at saturation. See also 3151 and 3152 of 1947 (Néel).

666.1.037.5 1419
**The Theory of Stresses in Two-Component Glass to Metal Tube Seals**—H. Rawson. (*Jour. Sci. Instr.*, vol. 26, pp. 25–27; January, 1949.) Theory based on that of Lamé is given, with curves showing the dependence of the principal stresses at the seal interface on the seal dimensions.

666.192.037.5 1420
**New Quartz-Metal Seal**—E. H. Nelson. (*Elec. Rev.* (London), vol. 144, pp. 60–62; January 14, 1949.) The metal part of the seal consists of a Mo annulus of lenticular cross section soldered coaxially upon a Mo rod. The rod provides a large current-carrying capacity whilst the annulus gives a hermetic seal.

679.5:535.316/.317:534.321.9 1421
**Ultrasonic Lenses of Plastic Materials**—D. Sette. (*Jour. Acous. Soc. Amer.*, vol. 21, p. 61; January, 1949.) Summary of Acoustical Society of America paper. Propagation velocity and attenuation of elastic waves have been measured in various synthetic materials; plexiglas appears to be the most suitable material for such lenses.

## MATHEMATICS

517.53:621.392.52 1422
**Splitting of an Analytical Function into a Linear and a Nonlinear Part, and also a Method of Determining, for a Known Linear Component, the Nonlinear Component and the Function as a Whole**—H. Pleijel. (*Arch. Elek.* (Übertragung), vol. 2, pp. 307–320; August and November, 1948.) Mathematical discussion. In many cases, the problem has a simple solution. Applications to the theory of filter networks are considered.

517.942:621.396.611.3 1423
**On Linear Differential Equations with Slowly Variable Coefficients. Application to the Study of Nonlinear Coupling**—C. Blanc. (*Bull. Tech. Suisse Romande*, vol. 74, pp. 185–188 and 209–213; July 17, and August 14, 1948.) For an equation with slowly variable coefficients and such that the integrals of the equation with the right-hand side replaced by zero tend sufficiently rapidly toward zero when the independent variable increases indefinitely, a solution is obtained involving a function which plays the part of a variable admittance. This result is applied to discussion of the maintenance of the oscillations of a pendulum by ac, and also to the case of two circuits coupled by the variable capacitance between two plates linked elastically.

517.942.1:518.5 1424
**Electro-Integrator for the Solution of the General Linear Differential Equation with Constant Coefficients**—N. V. Korol'kov and G. K. Kuz'minok. (*Bull. Acad. Sci. URSS*, pp. 517–532; April, 1948. In Russian.) The device described solves systems of such differential equations with a speed determined only by the duration of the transient process in the circuit; any relevant dynamic conditions can be simulated. The integrator is based on a circuit proposed by L. I. Gutenmacher and consisting of a number of interconnected two-terminal and four-terminal networks (tube amplifiers). The solution obtained is shown on a cro or a loop (mechanical) oscillograph and can be photographed. The operation of the integrator is discussed in detail. A year's experience has proved its suitability for many kinds of research work.

518.5 1425
**A Discussion on [general-purpose, automatic, digital] Computing Machines**—(*Proc. Roy. Soc. A* vol. 195, pp. 265–287; December 22, 1948.) The following papers were read: A Historical Survey of Digital Computing Machines, by D. R. Hartree. General Principles of the Design of All-Purpose Computing Machines, by M. H. A. Newman. The Design of a Practical High-Speed Computing Machine, the EDSAC, by M. V. Wilkes. A Cathode-Ray Tube Digit Store, by F. C. Williams. The Automatic Computing Engine at the National Physical Laboratory, by J. H. Wilkinson. Recent Computer Projects, by A. D. Booth. See also 1157 and 1493 of 1947, 464 and 3448 of 1948, 759 of April and 1110 of May.

517.432.1 1426
**Modern Operational Calculus, with Applications in Technical Mathematics [Book Review]**—N. W. McLachlan. Macmillan, London, 1948, 218 pp., 21s. (*Nature* (London), vol. 162, p. 945; December 18, 1948.) Based on a transform simply related to that of Laplace. The book is intended for post-graduate engineers and technologists and is unusually rigorous.

## MEASUREMENTS AND TEST GEAR

531.761+621.317.361 1427
**Limits for the Comparison of Frequency and Time at Great Distances**—M. Boella. (*Atti del Congresso internazionale della Radio* (Rome), pp. 213–224; September and October, 1947. In Italian.) A series of measurements on the stability of standard signals received from the Bureau of Standards confirms that irregular variations up to ±2 parts in $10^7$ exist and are due to a propagation effect; their magnitude depends on the time of day. The afternoon is

the best time for precise comparisons at 15 Mc, the maximum errors then observed are 2 parts in $10^8$. See also 1163 of May (Rawer) and back reference.

531.761 1428
**High-Precision Comparison of Times**—C. Egidi. (*Atti del Congresso internazionale della Radio* (Rome), pp. 292–296; September and October, 1947. In Italian.) Discussion of apparatus specially designed for comparing WWV with other standard signals.

621.317.3 1429
**An Experimental Method for Determining the Fundamentals and Harmonics of Stationary Low-Frequency Electrical Quantities**—F. Koppelmann. (*Frequenz*, vol. 2, pp. 296–303; November, 1948.) A moving-coil instrument is used with a mechanical rectifier (synchronized with the supply voltage) and suitably connected resistors to obtain the integral curve of the quantity in question. The values of the fundamental and harmonics are then calculated from measurements of selected ordinates. The method is illustrated by results for a transformer core.

621.317.324† 1430
**Production of a Uniform H.F. Field for Measurement Purposes**—E. Roeschen. (*Funk und Ton*, vol. 3, pp. 18–32; January, 1949.) Sensitive methods of field-strength measurement are based on measurements of mutual inductance. Curves are given showing the field strength for single circular coils and for combinations of pairs of coils. For a particular 4-coil system, the deviations from uniformity of the field were $<1$ per cent.

621.317.335.3†+621.317.374 1431
**Determination of Dielectric Quantities by Balance Methods**—T. Gast and E. Alpers. (*Z. Angew. Phys.*, vol. 1, pp. 228–232; November, 1948.) The basic principles of the methods are outlined and the apparatus used is described. Typical results are given showing the dependence of the dielectric constant and loss factor of polyvinyl chloride on temperature and frequency.

621.317.335.3†:546-145 1432
**On the Measurement of the Dielectric Constant of Solutions at High Frequencies**—U. Tiberio. (*Atti del Congresso internazionale della Radio* (Rome), pp. 463–484; September and October, 1947. In Italian.)

621.317.335.3†:621.392.26† 1433
**Universal Curves for Dielectric-Filled Wave Guides and Microwave Dielectric Measurement Methods for Liquids**—W. H. Surber, Jr. (*Jour. Appl. Phys.*, vol. 19, pp. 514–523; June, 1948.) A method is given for constructing a set of curves for the variation of waveguide parameters in terms of $D$, a dissipation factor of the dielectric filling analogous to $\tan\delta$ for coaxial lines. An experimental method is described for measuring the dielectric constant $\epsilon$ of a high-loss liquid contained in a waveguide, from the variation of reflection coefficient with sample length. The constants of the medium can then be determined using $\epsilon$ and $D$ as primary independent parameters.

621.317.335.3†:621.396.611.4 1434
**A Method for Measuring the Complex Dielectric Constant of Gases at Microwave Frequencies by Using a Resonant Cavity**—C. K. Jen. (*Jour. Appl. Phys.*, vol. 19, pp. 649–653; July, 1948.) For low-pressure gases at X-band frequencies (8,500 to 9,750 Mc). The real part of the dielectric constant of a gas affects the natural frequency of a resonant cavity containing it, while the imaginary part affects the amplitude and breadth of the cavity response curve. By rapid variation of the frequency through resonance, the real and imaginary parts can be conveniently and accurately determined from cro measurements. Sample results are given. The method can also be used for the measurement of (a) the resonant dispersion and absorption of microwaves by gas molecules, and (b) the loaded and unloaded $Q$ of a cavity.

621.317.725:621.317.733 1435
**Analysis of Bridge-Type Valve Voltmeters**— P. Popper and G. White. (*Wireless Eng.*, vol. 25, pp. 377–384; December, 1948.) Two arms of the bridge are formed by the tube and its associated resistors. The other two can be formed either by two resistors or by a similar tube and a resistor. These two types of bridge can be further subdivided according as the tube resistor is placed in the plate or cathode circuit. The relative merits of these four types are compared from the points of view of sensitivity, stability, linearity, and the effect of choice of output meter upon performance. For a single-tube circuit (with positive biasing battery) the transfer of the resistor from the plate to the cathode lead improves stability and linearity, but with the two-tube circuit this is not the case. A graphical solution for nonlinear operation in the two-tube circuit is given. Optimum deflection sensitivity occurs when the meter resistance equals the output resistance of the circuit; this cannot be realized in practice when plate resistors are used in double-tube circuits.

621.317.727 1436
**An A.C. Potentiometer**—J. M. Vanderleck. (*Elec. Eng.*, vol. 67, pp. 173–181; February, 1948.) Description of an instrument of the Gall type for general measurements. Simplicity of operation rather than absolute accuracy was the primary consideration.

621.317.761:621.396.611.4 1437
**Electrical Measurements on Transmission Cavity Resonators at 3 cm Wavelength**—M. S. Wheeler. (*Atti del Congresso internazionale della Radio* (Rome), pp. 520–525; September and October, 1947. In English.) Energy from a 3-cm FM oscillator is applied, through a variable attenuator and the cavity under test, to a rectifier crystal which provides, after amplification, the voltage for the vertical deflection of a cro beam. A small part of the oscillator energy is transmitted through a high impedance to another crystal, to which a standard frequency is also applied. The difference-frequency voltage intensifies the trace on the cathode-ray screen whenever the difference frequency falls within the pass band of the Z-axis amplifier. The horizontal sweep is derived from the oscillator modulating voltage, so that a response versus frequency curve is produced with bright spots superimposed and spaced equally on either side of the reference frequency. Accuracy of measurement of the resonance frequency is estimated as within 1 part in 150,000.

621.317.79:551.510.535 1438
**Ionosphere Reflections Recorded Mechanically by Means of a Repetition Frequency Converter**—W. Stoffregen. (*Jour. Appl. Phys.*, vol. 19, pp. 487–490; May, 1948.) The high repetition frequency (50 cps) of received ionospheric echoes is converted to a much lower frequency (0.2 to 0.5 cps) by a stroboscopic technique. Mechanical recording of echoes having a delay of some milliseconds, using a pulse width of 20 to 100 microseconds is thus facilitated. See also 789 of April.

621.319.4(083.74) 1439
**Standards for Low Values of Direct Capacitance**—C. Moon and C. M. Sparks. (*Jour. Res. Nat. Bur. Stand.*, vol. 41, pp. 497–507; November, 1948.) Discussion of the design and measurement of capacitors from 0.001 pF to 5 pF. Capacitors above 0.1 pF are of normal guard-ring type, but below this value a guard-well construction is used.

621.38/.39.001.4 1440
**Transients in Mechanical Systems**—J. T. Muller. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 657–683; October, 1948.) High-impact shock tests for electronic apparatus are analyzed by the Laplace transform for various types of impact pulse. Curves for the transient displacement and the maximum amplitude of the ensuing harmonic motion are given.

621.397.62.001.4 1441
**Airline TV Receiver Installation Tests**—W. S. Smoot. (*Communications*, vol. 29, pp. 8–9, 34; January, 1949.) A 12″ direct-viewing receiver was used, with alternative antennas above and below the aircraft. Power was supplied by means of an inverter. Performance on a run from Washington to Norfolk, Virginia, is discussed.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.001.82 1442
**Generation of Sonic and Ultrasonic Waves in Liquids for Industrial Purposes**—W. Janovsky and R. Pohlman. (*Z. Angew. Phys.*, vol. 1, pp. 222–228; November, 1948.) A special type of high-frequency whistle is described which gives frequencies from 4 to 32 kc and is very effective for the production of emulsions. The operating cost of emulsification by means of the whistle is about 0.003 of that when magnetostriction oscillators are used.

539.16.08 1443
**Further Experiments with an Adjustable Geiger-Müller Counter**—A. G. Fenton and E. W. Fuller. (*Proc. Phys. Soc.*, vol. 62, pp. 32–40; January 1, 1949.) Discussion of the effect of various counter variables on the occurrence of multiple pulses, and of the performance of the counter as an ionization chamber at operating potentials below the Geiger region. For earlier work see 2578 of 1948 (Chaudhri and Fenton).

539.16.08 1444
**Recent Research on [Geiger-Müller] Counter Tubes**—J. D. Craggs. (*Rep. Progr. Phys.*, vol. 9, pp. 137–155; 1942 and 1943. Bibliography, pp. 155–157.) The work done on the mechanism of counters since about 1938 is reviewed. The main sections deal with proportional and nonproportional counters and their absolute efficiency. Their application to the counting of visible photons, X-rays, and neutrons is discussed. Directional, discriminating, and other special types are briefly described.

539.16.08 1445
**Two Methods of Measurement of Dead Time in Geiger-Müller Counters**—J. L. Putman and E. H. Cooke-Yarborough. (*Jour. Sci. Instr.*, vol. 25, pp. 409–411; December, 1948.)

539.16.08:621.383 1446
**Performance of 931-A Type Multiplier in a Scintillation Counter**—G. A. Morton and J. A. Mitchell. (*RCA Rev.*, vol. 9, pp. 632–642; December, 1948.) The scintillation type of nuclear radiation detector uses such a multiplier to convert into an electrical pulse the light flash produced by a suitable phosphor crystal when it absorbs a nuclear particle. The main properties investigated are the number and distribution of spurious pulses generated by the multiplier in darkness and the pulse performance when very few photoelectrons contribute to the pulse.

620.179.16 1447
**Supersonic Flaw Detectors**—D. C. Erdman. (*Elec. Eng.*, vol. 67, pp. 181–185; February, 1948.) Echo techniques and apparatus.

621.38.001.8:621.3.012 1448
**Handbook of Industrial Electronic Circuits [Book Review]**—Markus and Zeluff. (*See* 1345.)

621.38.001.8:621.9 1449
**The Application of Electronics to Machine Tools**—(*Machinery* (London), vol. 73, pp. 663–668; November 11, 1948.) Discussion of

devices shown at the recent Machine Tools Exhibition at Olympia. In most cases, the number of electronic devices used has been kept to a minimum,and the circuit arrangements are designed for long life and ease of maintenance. Service requirements have stimulated the production of rugged, shock-proof, and moisture-proof components. Devices for obtaining stepless speed control, especially for small machines, were a prominent feature.

621.38.001.8:623.95 1450
**Developments on Magnetic and Acoustic Mines at the Admiralty Mining Establishment**—A. J. Baggott and C. H. Fawcett. (*Jour. IEE* (London), part I, vol. 95, pp. 550–551; December, 1948.) Discussion on 790 of 1948.

621.384.62† 1451
**The Choice of Operating Mode for Standing Wave Type Linear Accelerators for Electrons**—E. J. Lawton. (*Jour. Appl. Phys.*, vol. 19, pp. 534–539; June, 1948.)

621.384.62† 1452
**Experimental Results on Standing Wave Type Linear Accelerators for Electrons**—E. J. Lawton and W. C. Hahn. (*Jour. Appl. Phys.*, vol. 19, pp. 642–648; July, 1948.)

621.385.833 1453
**Chromatic Aberration and Resolving Power in Electron Microscopy**—E. G. Ramberg and J. Hillier. (*Jour. Appl. Phys.*, vol. 19, pp. 678–682; July, 1948.)

621.385.833 1454
**Electron Microscopy**—L. Marton. (*Rep. Progr. Phys.*, vol. 10, pp. 204–248; 1944 and 1945. Bibliography, pp. 248–252.) Magnetic lenses are discussed in some detail and formulas for focal length, magnification, chromatic and spherical aberration, and optimum aperture are given together with graphs showing their variation. Some experimental justification is given, and the superiority of magnetic over electrostatic lenses is shown. The theory of resolving power is not so well understood but there is experimental evidence for assuming a limit to resolution of about 20 Å for bright-field observation. For dark-field observation, the limit is somewhat larger. General descriptions are given of the main types of electron microscopes and allied instruments, and, in particular, of a microanalyzer and a microtome; the operational techniques involved are discussed.

621.385.833 1455
**The Refractive Index in Electron Optics and the Principles of Dynamics**—W. Ehrenberg and R. E. Siday. (*Proc. Phys. Soc.*, vol. 62, pp. 8–21; January 1, 1949.) A relation between the ray direction and the wave normal is obtained, and an expression for the optical path difference is given in terms of the magnetic flux enclosed. The results are applied to the differential equations for trajectories, the focusing properties of an axially symmetrical field and the interference pattern due to two converging bundles of rays enclosing a magnetic flux.

621.385.833 1456
**Electron Lenses of Hyperbolic Field Structure: Part 2**—R. Rüdenberg. (*Jour. Frank. Inst.*, vol. 246, pp. 377–408; November, 1948.) If the lens voltage is increased above the incident electron voltage, the beam is at first blocked. A further increase converts the lens into a curved mirror, the focal properties of which are analyzed. Boundary conditions which may be rigorously or approximately satisfied in actual designs are discussed, and the effects of lateral apertures and windows are considered in detail. Expressions are derived for various defects which are small and depend on only a few parameters of the lens field. The case of electron cylinder lenses is also considered. Part 1: 813 of April.

621.385.833 1457
**Electron Lens Corrected for Spherical Aberration**—P. Hubert. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 233–235; January 17, 1949.) The method of correction proposed consists in sending the electron beam through a narrow metal cylinder in order to use the attraction of the charges induced on the walls. Numerical calculations show that, with such a system, the resolving power is greatly improved. The method may have important applications in the case of the proton microscope.

621.385.833 1458
**Electron Lens with Curvilinear Axis**—P. Hubert. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228 pp. 302–304; January 24, 1949.) A magnetic lens with circular axis has been used by Svartholm and Siegbahn in the construction of a $\beta$-ray spectrograph (1868 of 1947). It is possible to construct a lens whose magnetic field does not possess rotational symmetry. The theory of such a system is given and the practical advantages are enumerated.

621.385.833:537.533:530.145.6 1459
Field Emission of Electrons—Jenkins. (*See* 1356.)

621.385.833.032.29 1460
**On a New Electron Gun for High-Voltage Tubes**—M. Bricka and H. Bruck. (*Ann. Radioélec.*, vol. 3, pp. 339–343; October, 1948.) Detailed description of a gun used on C.S.F. electron microscopes and diffraction analyzers. At 45 kv it gives a current density ten times that of the ordinary type of gun and has an efficiency of about 20 per cent. The high-voltage current is about 40 microamperes.

534.321.9.001.8:620.179.16 1461
**Ultrasonic Material Testing and Other Applications [Book Notice]**—B.I.O.S. Final Report No. 1679, Item No. 9 H.M. Stationery Office, London, 68 pp. plus reprints in German from various German journals. Many different workers had obtained suggestive results indicating the possibility of industrial application of ultrasonics to emulsification, precipitation, preparation of fine grain alloys, etc., but there was little evidence of actual use on an industrial scale. A magnetostriction technique was successfully used for both vertical and horizontal echo soundings. An ultrasonic glass delay line was also used in the Rehbock radar test set. Success in techniques for testing materials is claimed, though insufficient information is available as to the amount of material so tested. See also B.I.O.S. Final Reports Nos. 724 and 1504.

## PROPAGATION OF WAVES

538.566 1462
**On Formulae for the Principle of Huyghens for Electromagnetic Waves**—F. Croze and É. Durand. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 236–239; January 17, 1949.) Formulas have been given by L. de Broglie (3083 of 1944) which are equivalent to those of Kottler but in which the elements of a surface wave have only superficial densities of current and electric and magnetic charges. Application of these formulas to the determination of the electric field in a secondary wave emanating from an element of a surface wave leads to formulas established by Love, Larmor, and Bromwich, so that de Broglie's formulas should be regarded as the correct expression of the principle of Huyghens, taken in its usual sense, for electromagnetic waves.

538.566 1463
**On the Expression of the Principle of Huyghens for Electromagnetic Waves**—F. Croze and P. Boillet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 305–307; January 24, 1949.) The formulas of Love, Larmor, and Bromwich for electromagnetic waves are derived directly from the classical formulas for the fields produced by electric and magnetic doublets. See also 1462 above (Croze and Durand).

538.566 1464
**Generalized Magneto-Ionic Theory**—N. C. Gerson and S. L. Seaton. (*Jour. Frank. Inst.*, vol. 246, pp. 483–494; December, 1948.) A vector equation relating the polarization of a homogeneous isotropic medium to the externally impressed magnetic field is derived for the case of a medium penetrated by an electromagnetic wave. The equation involves the fourth degree of the refractive index and is a generalization of that obtained by Booker (422 of 1939) for a 2-dimensional case.

538.566:551.510.535 1465
**On the Polarization in the Ionosphere**—J. Malsch. (*Arch. Elek.* (Übertragung), vol. 2, pp. 231–237; June and July, 1948.) Evidence from various sources is discussed which suggests that the so-called Lorentz term is not applicable in ionospheric propagation.

538.566.2 1466
**Propagation of Electromagnetic Perturbations in an Atmospheric Waveguide**—T. Kahan and G. Eckart. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 228, pp. 235–236; January 17, 1949.) An outline of the physical phenomena in an ideal plane atmospheric duct bounded by a perfectly conducting earth and a parallel plane at a height $h$ where the index of refraction has a discontinuity. Mathematical theory will be published elsewhere. For low values of $h$, the field $F$ of the dipole source at distance $r$ varies as $1/r^3$, but above a limiting value of $h/\lambda$, $F \propto 1/\sqrt{r}$ and the energy is propagated in cylindrical waves of the *TE* type.

538.566.2 1467
**Representation of the Radiation in an Atmospheric Duct by a Series of Virtual Sources**—T. Kahan and G. Eckart. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 304–305; January 24, 1949.) Near the origin, the field in the duct can be derived by superposing, on the field of a dipole of double strength, the fields due to successive images of this dipole with respect to (a) the plane forming the upper boundary of the duct, and (b) the earth. This representation is valid up to a distance such that the ray from the first virtual image makes an angle with the horizontal at least equal to the angle of total reflection at the boundary layer. At greater distances, the field strength follows either the $1/r^3$ or the $1/\sqrt{r}$ law (see 1466 above).

621.396.1 1468
**Diversity Reception in U.S.W. Radio Links**—G. Barzilai and G. Latmiral. (*Wireless Eng.*, vol. 25, pp. 390–395; December, 1948.) The proper disposition of the receiving antennas is considered theoretically, taking account of the way in which they are connected to the receiver. Results are applied to a practical case.

621.396.11 1469
**Coastal Refraction**—T. L. Eckersley. (*Atti del Congresso internazionale della Radio* (Rome), pp. 97–110; September and October, 1947. In English.) The fact that the velocity of electromagnetic waves over land is less than their velocity over sea is inconsistent with Zenneck's surface-wave theory, but this theory is not considered relevant. The result can be obtained from pure diffraction theory. The difference in velocity (and, therefore, coastal refraction) is only significant at wavelengths between 15 and 1,500 meters.

621.396.11 1470
**The Future of Wave Propagation Research**—S. B. Smith and K. W. Tremellen. (*Atti del Congresso internazionale della Radio* (Rome), pp. 196–210; September and October, 1947. In English.) A survey of present knowledge and possible future developments for (a) atmospherics and extraterrestrial radiations, (b) vhf propagation (above 30 Mc) and superrefraction, (c) high-frequency propagation (3 to 30 Mc), (d) medium-frequency ground-wave propagation and ionospheric transmission

paths (0.15 to 3 Mc), (e) low-frequency propagation (10 to 150 kc).

621.396.11 1471
**General Remarks on the Theory of Wave Propagation**—H. Bremmer. (*Atti del Congresso internazionale della Radio* (Rome), pp. 30–42; September and October, 1947. In English.) Discussion of the effects of earth curvature, refraction and diffraction, and the relevance of both ray theory and mode theory.

621.396.11 1472
**Mechanisms of Propagation**—C. R. Burrows. (*Atti del Congresso internazionale della Radio* (Rome), pp. 43–51; September and October, 1947. In English.) Discussion of refraction, diffraction, and ionospheric and guided tropospheric propagation.

621.396.11 1473
**Studies in Propagation**—T. L. Eckersley. (*Atti del Congresso internazionale della Radio* (Rome), pp. 78–96; September and October, 1947. In English.) A critical survey. The properties of the ionosphere are discussed with reference to the problems of scatter clouds in the *E* layer, abnormal-*E*, and abnormal-*F*. Propagation over land and sea paths is discussed in some detail for $\lambda \leqq 10$ meters. The importance of the phase-integral method is stressed.

621.395.11:551.510.535 1474
**A Note on the Phase Difference between Two Waves Reflected from the Ionosphere**—J. M. Kelso. (*Jour. Appl. Phys.*, vol. 19, pp. 590–591; June, 1948.) Two relations given without proof by Appleton and Beynon (2895 of 1947) are discussed and shown to have wide validity.

621.396.11:551.510.535 1475
**Regular Ionosphere Observations over Mid-Germany**—W. Dieminger. (*Fernmeldetech. Z.*, vol. 1, pp. 222–224; November, 1948.) Diagrams show the diurnal variations of the monthly mean values of limiting frequency for the normal and abnormal *E* layers and for the $F_2$-layer ordinary component from January to October, 1948. Disturbances which occurred during October are briefly discussed.

621.396.11:551.510.535 1476
**The Verification of Magneto-Ionic Theory from the Gyrointeraction of Radio Waves in the Ionosphere**—M. Cutolo. (*Atti del Congresso internazionale della Radio* (Rome), pp. 65–77; September and October, 1947. In Italian.) Experimental results confirm V. A. Bailey's theory of gyrointeraction. The fact of gyrointeraction shows that there is an ionospheric resonance frequency in the sense indicated by Nicholls and Shelleng. A nocturnal variation was observed in the gyromagnetic frequency. The experimental methods discussed can be used to study the distribution of the earth's magnetic field in the ionosphere. See also 513 of 1947 (Cutolo, Carlevaro, and Ghergi) and 2055 of 1948.

621.396.11:551.510.535 1477
**Measurement of the Parasitic Modulation in Gyrointeraction Phenomena**—M. Cutolo and R. Ferrero. (*Alta Frequenza*, vol. 17, pp. 212–216; October, 1948. In Italian, with English, French, and German summaries.) An account of measurements on transmission paths between Turin and Taranto, Augusta and Palermo, with the Vatican transmitter as the interfering element. Curves show the variation of the parasitic modulation as a function of (a) the modulation frequency of the interfering station, and (b) time.

621.396.11:551.594.21 1478
**Guided Propagation of Radar in Thunderstorm Conditions**—R. D. Coons. (*Bull. Amer. Met. Soc.*, vol. 28, pp. 324–329; September, 1947.) Discussion of the behavior of radar echoes during thunderstorm conditions. A series of photographs of a PPI presentation shows the passage of storm clouds, and the abnormal range of ground echoes resulting from guided propagation caused by stratification of water vapor in the rain areas under the cloud.

621.396.11:621.396.619.13:621.396.65 1479
**The Possibility of Transatlantic Transmission by Means of Frequency Modulation**—Arguimbau and Granlund. (*See* 1506.)

621.396.81:621.397.5 1480
**Field Test of Ultra-High-Frequency Television in the Washington Area**—G. H. Brown (*RCA Rev.*, vol. 9, pp. 565–584; December, 1948.) A survey of the field strength of the signals from a broadcast transmitter operating in the 500-Mc band is described. The measurements were made at points along eight radial lines and also in about 50 homes with typical home receivers. From the results, the power required for satisfactory coverage was determined. Measurements in the homes were also made of the signal voltage for the existing 67.25-Mc television service and for transmissions on 505.25 Mc. Performances of various types of receiving antennas are compared. See also 3224 of 1948 (Brown, Epstein, and Peterson).

621.396.812.029.64 1481
**A Method of Determining the Angle of Arrival**—A. W. Straiton, W. E. Gordon, and A. H. LaGrone. (*Jour. Appl. Phys.*, vol. 19, pp. 524–533; June, 1948.) The angles of arrival of the direct and reflected rays are determined by measuring the phase difference for two points at different heights and the field strength at each point. Values obtained for 3-cm radiation over a 27-mile path are consistent with those calculated by ray theory, taking account of the refractive-index distribution deduced from meteorological measurements. See also 1182 of 1947 (Sharpless) and 1183 of 1947 (Crawford and Sharpless).

## RECEPTION

621.396.1 1482
**Diversity Reception in U.S.W. Radio Links**—Barzilai and Latmiral. (*See* 1468.)

621.396.621 1483
**On the Use of a Superregenerator in the Nonlinear Mode for the Reception of F.M. Signals**—G. B. Ol'derogge. (*Radiotekhnika* (Moscow), vol. 3, pp. 76–87; November and December, 1948. In Russian.)

621.396.621:534.861 1484
**The Acoustics of Broadcasting Receivers**—Govyadinov. (*See* 1267.)

621.396.621:621.3.015.3 1485
**Transient Response of an F.M. Receiver**—M. K. Zinn. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 714–731; October, 1948.) The frequency detector is represented by two resonant impedances with low-frequency outputs in opposition. Mathematical treatment permits the calculation of the output voltage for rapid variation of the instantaneous frequency. Very large impulsive variations, resulting in abrupt changes of phase, such as may be produced by ignition interference, are analyzed and curves of output voltage are obtained. The method is also used to analyze the reception of a simple sinusoidal modulation; the signal-to-noise ratio is deduced for severe impulsive interference.

621.396.82:621.397.5 1486
**The Effect of Aircraft on the Reception of Transmissions in the 45-Mc/s Band**—R. A. Rowden and G. I. Ross. (*BBC Quart.*, vol. 3, pp. 251–256; January, 1949.) Reflection of radio waves from aircraft can have a disturbing effect upon vhf reception. Tests with random aircraft indicate that interference is serious only when the aircraft is within 6 miles of the receiver or within 3 miles of the line joining the transmitter and receiver. The fluctuations in field strength of horizontally polarized transmissions are three or four times as great as those with vertically polarized transmissions. Experiments with co-operating aircraft gave similar results.

621.396.822 1487
**Report on the Present Limits of V.H.F., U.H.F. and S.H.F. Reception**—M. J. O. Strutt. (*Atti del Congresso internazionale della Radio* (Rome), pp. 412–442; September and October, 1947. Bibliography, pp. 442–447. In English.) A review of various causes of unwanted rf noise, for frequencies between 30 and 30,000 Mc with discussion of (a) the admissible signal-to-noise ratio for various types of transmission and the present definition of noise figure, (b) thermal external noise, the antenna being regarded as enclosed in a space of dimensions large compared to λ, and in thermal equilibrium at the temperature of that space, (c) atmospheric and man-made noise, including the propagation of atmospherics, (d) ionospheric and extraterrestrial noise, with special reference to the temperature, location, and frequency bands of the various sources, (e) thermal fluctuations in resistive elements of a network, (f) tube and crystal noise, (g) low-noise circuits and methods of noise reduction, and (h) actual noise levels obtainable at the various frequencies.

621.396.822 1488
**Noise in Linear Networks**—E. J. Schremp. (*Atti del Congresso internazionale della Radio* (Rome), pp. 402–411; September and October, 1947. In English.) A mathematical analysis of the possible alternative representations of a passive or active linear network containing bilateral or nonbilateral elements. The results are applied mainly to thermal noise and various forms of shot noise in multi-electrode tubes.

621.396.822 1489
**Thermal Noise Output in A.M. Receivers**—M. V. Callendar. (*Wireless Eng.*, vol. 25, pp. 395–399; December, 1948.) Formulas given by Burgess for noise and signal outputs from rectifiers are examined, and criteria for the variation of signal-to-noise ratio with predetector bandwidth and signal input to the detector are shown graphically. Practical conclusions are given for television sound, telephony, and telegraphy receivers. See also 534 of 1948 (Thomas and Burgess).

## STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 1490
**A Mathematical Theory of Communication**—Shannon. (*See* 1361.)

621.391.63:534.321.9 1491
**On the Modulation of Light at Radio Frequencies by means of Ultrasonic Waves**—Giacomini. (*See* 1255.)

621.394/.395+621.315.2 1492
**French Telecommunication Networks**—J. Mailley; J. Gastebois. (*Câbles and Trans.*, (Paris), vol. 3, pp. 8–30; January, 1949. With English summary.) Developments from 1924 to 1949 of carrier-current, overhead, underground, and underwater cables are discussed and present-day local and long-distance telegraphy and telephony systems are described.

621.394/.395].724 1493
**Repeater-Station Equipment**—J. Malézieux and R. Sueur. (*Câbles and Trans.* (Paris), vol. 3, pp. 66–82; January, 1949. With English summary.) Review of developments since 1924, with special reference to standardization and carrier-current systems.

621.394/.395].74 1494
**Maintenance of the Network of Long-Distance Underground Lines**—R. Croze; A. Chavigner. (*Câbles and Trans.* (Paris), vol. 3, pp. 106–133; January, 1949. With English summary.)

621.394/.395].74 1495
**Electro-Mechanical Installations of the Amplification Centres of the [French] Long-Distance Cable Network**—A. Romanet. (*Câbles*

*and Trans.* (Paris), vol. 3, pp. 83–95; January, 1949. With English summary.) A general description of the power supplies normally required at an amplification center, the operating conditions, and the evolution of suitable equipment, including batteries, rectifiers, voltage regulators, and generators.

621.394/.395].74 1496

**The Future of the French Telecommunication Network**—P. Marzin. (*Câbles and Trans.* (Paris), vol. 3, pp. 134–136; January, 1949. With English summary.)

621.395.44 1497

**Carrier-Current Systems on Overhead Lines**—R. Sueur. (*Câbles and Trans.* (Paris), vol. 3, pp. 96–105; January, 1949. With English summary.) A short account of various single- and multi-channel systems developed in France.

621.396 1498

**Recent Developments in Radio Communication in Great Britain**—M. Faulkner. (*Atti del Congresso internazionale della Radio* (Rome), pp. 752–787; September and October, 1947. In English.) A general account of some developments in the radio and allied services of the British Post Office. Details are given of point-to-point radio systems, alterations at pre-war stations, and post-war developments. A proposed radio-telephone terminal in N.W. London is to have initially at least 48 terminal equipments. The extension of single-sideband operation to long-distance point-to-point radiotelephone links is planned; the main features of the equipment are described. Results of laboratory tests with a fading machine (2299 of 1948) on the effect of propagation conditions are discussed and illustrated graphically. The relaying of television signals by balanced, coaxial, and unloaded telephone cables is considered and details are given of a special universal equalizer for quick reliable testing of video circuits. Plans are outlined for the extension of (a) outside broadcast networks in London, and (b) long coaxial-cable routes for transmission from the program source to distant service areas. The proposed method of relaying television signals by radio from London to Birmingham is described.

621.396.1:061.3 1499

**International Telecommunication Convention, Atlantic City, 1947**—P. E. Erikson. (*Elec. Commun.* (London), vol. 25, pp. 232–236; September, 1948.) Discussion of important changes or additions to the Madrid Convention of 1932. For other accounts see 1748, 1749, and 2357 of 1948.

621.396.3.029.56.58 1500

**Multi-Channel Radio-Telegraph System for High-Frequency Circuits**—T. E. Jacobi. (*RCA Rev.*, vol. 9, pp. 704–720; December, 1948.) Sub-carrier channels using FM in conjunction with single-sideband operation and space-diversity reception appeared from field tests to be the best of several systems tried; error rates of 0.02 to 0.14 per cent were obtained on a transcontinental circuit. Double-sideband circuits had high error rates during periods of selective fading; these could be reduced to the above values by using exalted-carrier receivers. Single-sideband systems make better use of the frequency band available and use the transmitter power more efficiently. Under present conditions of high-frequency congestion, reduction of bandwidth and increase of message capacity must be regarded as of paramount importance.

621.396.324 1501

**Teleprinting over Radio Circuits**—H. C. A. van Duuren. (*Atti del Congresso internazionale della Radio* (Rome), pp. 902–914; September and October, 1947. In English.) Description of a reliable telegraph circuit between Amsterdam and Batavia. Each letter is transmitted on a 7-unit code. When a fault is detected, the printer stops, a signal is given on the return path, and the transmitter is stepped back to the beginning of the faulty letter.

621.396.41:621.396.619.16 1502

**Multiplex Radiotelephony Link between the Mainland and Corsica**—P. Rivère. (*Ann. Radioélec.*, vol. 3, p. 338; October, 1948.) Corrections to 3508 of 1948.

621.396.5 1503

**Radio for Telephone Service in America**—H. I. Romnes. (*Atti del Congresso internazionale della Radio* (Rome), pp. 851–876; September and October, 1947. In English.) An account of methods used by the Bell System. A description of overseas radio telephone includes a world map of the direct and principal interconnecting services, existing and planned. Service to ships, a general mobile service, a domestic point-to-point system using the 30 to 300-Mc band, and a microwave relay system between New York and Boston are also discussed.

621.396.619.11/.13 1504

**A.M. and Narrow-Band F.M. in U.H.F. Communications: Parts 1 and 2**—E. Toth. (*Electronics*, vol. 22, pp. 84–91 and 102–108; February and March, 1949.) A comparison based on a long-term study undertaken at the U. S. Naval Research Laboratory, dealing with typical conditions for mobile Navy communications. Advantages of AM include: (a) better weak-signal performance, (b) relative freedom from co-channel and adjacent-channel interference, (c) lower susceptibility to multipath propagation difficulties, (d) AM occupies a narrower frequency range, especially at low carrier frequencies, (e) ease of equipment alignment, (f) tolerance of severe detuning, and (g) circuit simplicity. FM has advantages for geographically fixed communication and broadcast systems, particularly for high-fidelity reproduction of speech and music. In general, FM permits the use of a smaller transmitter and requires less power for a given carrier-output rating. Transmitter modulation is difficult for FM crystal-controlled systems.

621.396.619.11/.13 1505

**Reduction by Limiters of Amplitude Modulation in an Amplitude- and Frequency-Modulated Wave**—A. G. Clavier, P. F. Panter, and W. Dite. (*Elec. Commun.*, vol. 25, pp. 291–299; September, 1948.) The limiting action of ideal and imperfect limiters is analyzed mathematically by the use of the Fourier transform. Curves are given, for various applied carrier voltages, showing the reduction of AM by theoretical limiters having static characteristics represented by known mathematical functions. See also 3110 of 1944 (Bennett).

621.396.619.13:621.396.65:621.396.11 1506

**The Possibility of Transatlantic Transmission by Means of Frequency Modulation**—L. B. Arguimbau and J. Granlund. (*Atti del Congresso internazionale della Radio* (Rome), pp. 671–679; September and October, 1947. In English.) An analysis of the problem of multipath interference to FM telephonic communication. An experimental circuit is described in which a source using FM is fed to the receiver through two paths, one a simple attenuator and the other a ½-ms supersonic delay line. Preliminary tests appear promising.

621.396.65:621.396.41 1507

**Multiplex U.H.F. Radiotelephone Links**—H. Chireix. (*Atti del Congresso internazionale della Radio* (Rome), pp. 717–745; September and October, 1947. In French.) Methods of calculating signal-to-noise ratio, fully discussed in 2902 of 1947, are summarized. For systems using simple carrier-current FM, high signal-to-noise ratio can be obtained for a large number of channels. Overmodulation effects are much less serious than for AM. The pass band is relatively narrow. These advantages are offset by high cost, which is greatly reduced if two coaxial cables equipped for carrier current are available. Double carrier-current FM is a system particularly suitable at cm $\lambda$; reflex klystrons are used whose frequency can, at present, be varied by as much as $\pm$ 7 Mc at 3,000 Mc; the disadvantage of low power output is likely to be rectified soon. Pulse-time modulation systems have the advantage of requiring simple apparatus which can be easily installed. The main disadvantages are that signal-to-noise considerations tend to limit their use to multiplex links with only a small number of channels, crosstalk may be very objectionable and the pass band required is appreciably wider than for the FM systems considered. See also 2307 of 1946.

621.396.65(45) 1508

**Experimental U.S.W. Multiplex Radiotelephone Link between Milan and Rome**—F. Vecchiacchi. (*Atti del Congresso internazionale della Radio* (Rome), pp. 915–927; September and October, 1947. In Italian.) The total direct distance is 475 km; two intermediate mountain relay stations are used, so that this is split up into optical paths of 160, 260, and 70 km. Frequencies are between 150 and 220 Mc, with horizontal polarization. There are 7 channels. Performance has been continuously satisfactory; crosstalk is completely absent. See also 3750 of 1946.

621.396.931 1509

**Railroad Radio**—W. D. Hailes. (*Elec. Eng.*, vol. 68, pp. 1–7; January, 1949.) Description of two systems and the associated equipment: (a) a narrowband FM system inductively coupled to telegraph wires near the track; frequency is between 70 and 200 kc, input power 270 watts, output 30 to 40 watts, range about 30 miles train-to-station or 15 to 20 miles between trains; (b) a 4-channel FM space radiation system; input power is 425 watts, radiated power 50 watts, bandwidth 1.2 Mc; range depends upon terrain but is usually less than for system (a). The advantages and uses of both systems are discussed.

621.396.932.029.63 1510

**An Early Application of Decimetre Waves to Communication between Ships**—E. C. S. Megaw and W. E. Willshaw (*Atti del Congresso internazionale della Radio* (Rome), pp. 790–828; September and October, 1947. In English.) Experimental work in the development, for the Admiralty, of equipment to provide reliable all-round communication at distances up to 3 miles. The stabilized transmitter used a 4-segment magnetron and provided frequencies of 515, 530, 635, and 650 Mc by means of plug-in resonators. Modulation was effected by af interruption of the carrier, with marking and spacing frequencies of 1,000 cps and 3,200 cps respectively. The oscillatory circuit of the superregenerative receiver consisted of a short length of parallel-strip line; this was tuned by a symmetrical capacitor, on the stator plates of which an acorn triode was mounted. The frequency range 415 to 705 Mc was covered in 8 steps, selected by changing the position of a preset capacitor bridge. The antenna, of constant impedance 70 $\Omega$, comprised a $\lambda/4$ radiator with a $\lambda/4$ compensating reactance transformer. Details are given of equipment performance during trials in 1939 and of measurements made to determine diffraction losses produced by masts and superstructure. A much simplified model was built in 1940 to provide a single communication channel.

621.396.933 1511

**Aircraft Radio Communication Set A.R.I. 5272**—E. C. Fielding. (*Elec. Commun.*, vol. 25, pp. 244–255; September, 1948.) Description of an AM transmitter-receiver originally designed for naval aircraft. Any of 4 predetermined crystal-controlled spot frequencies in the range 115 to 140 Mc can be selected by a switch; channel-changing only takes 3 seconds. The transmitter output is 3.5 watts and the range is 100 miles to ground or 200 miles to another aircraft. The set can be used under extreme tropical or arctic conditions.

621.396.97(73) 1512

**F.M. Broadcasting in the United States**—

E. M. Ostlund and G. S. Wright. (*Atti del Congresso internazionale della Radio* (Rome), pp. 829–850; September and October, 1947. In English.) Results of tests in the broadcast band 88 to 105 Mc on FM and AM receivers show that the former have better signal-to-noise ratio and freedom from interference. FM transmitters, methods of modulation, and and transmitting antenna systems are discussed. A brief account of networks in operation in the United States includes a contour radiation pattern for a typical station. The advantages of using FM at vhf are summarized.

## SUBSIDIARY APPARATUS

621.316.7.078:016 1513

**List of Russian Articles on Questions of Automatic Regulation and Following Systems for the Period 1917–1947**—(*Avtomatika i Telemekhanika*, vol. 9, pp. 397–411; September and October, 1948. In Russian.)

778.53 1514

**Motion Picture Photography at Ten Million Frames per Second**—B. O'Brien and G. Milne. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 30–40; January, 1949.) The area to be photographed is transformed by a stationary multiple-lens and slit optical system into a number of narrow rectangles placed end to end across the moving film. The negative so formed is afterwards projected back through a similar optical system to reform a set of normal images for printing. The equipment is described in detail and illustrated, and examples of records are shown.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.26 1515

**The [British] Post Office Phototelegraph Service to Europe**—A. Wilcock. (*P.O. Elec. Eng. Jour.*, vol. 41, part 4, pp. 189–192; January, 1949.) The picture to be transmitted is scanned by light which is reflected on to a photo cell whose output is used to provide a single-sideband AM transmission at 1,300 cps. Synchronization and phasing signals are also transmitted. Both transmitter and receiver are driven by phonic motor from the output of oscillators with tuning-fork control. At the receiver, the fork frequency is synchronized manually with that at the transmitter; the receiver drum is automatically started in phase with the transmitter drum. The received picture signal is demodulated in a metal bridge rectifier and converted into a light trace by a Duddell oscillograph recording on photographic paper or film.

621.397.331.2:621.397.5 1516

**Electro-Optical Characteristics of Television Systems: Part 4—Correlation and Evaluation of Electro-Optical Characteristics of Imaging Systems**—O. H. Schade (*RCA Rev.*, vol. 9 pp. 653–686; December, 1948.) Continuation of 540 of March.

621.397.5 1517

**Recent Progress in Television**—V. K. Zworykin. (*Atti del Congresso internazionale della Radio* (Rome), pp. 928–946; September and October, 1947. In English.)

621.315.212:621.397.5 1518

**The London-Birmingham Television Cable: Part 1—General System and Electrical Requirements**—Stanesby and Weston. (*See* 1279.)

621.397.5:621.396.81 1519

**B.B.C. Television Map**—(*Wireless World*, vol. 55, pp. 55, 74; February, 1949.) Field-strength contours for the Alexandra Palace 45-Mc video transmitter.

621.397.5:621.396.81 1520

**Field Test of Ultra-High-Frequency Television in the Washington Area**—Brown. (*See* 1480.)

621.397.5:621.396.813 1521

**Delay Distortion in Television Transmission and its Measurement**—S. H. Padel (*BBC Quart.*, vol. 3, pp. 235–244; January, 1949.) Delay distortion can be estimated from the phase versus frequency characteristic or from modulation phase shift. The methods are described in detail. Block diagrams of measuring apparatus are given, and the degree of improvement made possible by delay equalizers is discussed.

621.397.5:621.396.82 1522

**The Effect of Aircraft on the Reception of Transmissions in the 45-Mc/s Band**—Rowden and Ross. (*See* 1486.)

621.397.5:791.45 1523

**Television in the Cinema**—A. G. D. West. (*Wireless World*, vol. 55, pp. 42–44; February, 1949.) A discussion of distribution methods, based on a paper read at the International Television Convention at Zürich (867 of April). A long-term proposal involves a very-high-definition system (900 to 1,200 lines) which will link up cinemas all over Britain and provide a variety of television programs. An experimental large-screen television service is also being developed with the existing 405-line definition; various programs can be relayed over 480-Mc radio links to selected London cinemas. See also 240 of February.

621.397.5(083.74) 1524

**Comments on Standardisation of European Television Services**—B. J. Edwards. (*Atti del Congresso internazionale della Radio* (Rome), pp. 746–751; September and October, 1947. In English.) A frame scan frequency of 50 cps, a standard interlacing index, and 405 lines per frame are advocated; detailed reasons are given. Various technical improvements are also suggested.

621.397.5(44) 1525

**French Work on Television**—R. Barthélemy. (*Atti del Congresso internazionale della Radio* (Rome), pp. 680–682; September and October, 1947. In French.)

621.397.62 1526

**Motorola Television Receiver—Model VT. 71**—(*Radio Franç.*, pp. 10–13; February, 1949.) Discussion of general characteristics and special features, together with a detailed circuit diagram.

621.397.62 1527

**Superheterodyne Television Unit: Part 1**—(*Wireless World*, vol. 55, pp. 61–65; February, 1949.) Circuit and construction of a long-range sound and vision receiver. Single-sideband operation and rejector circuits are used to secure adequate sound-channel rejection. It is expected that only minor modifications will be necessary for reception of the future BBC Birmingham transmissions. The unit can be used with an external audio amplifier and may replace the straight receiver described previously (1186 of 1948 and back references). To be continued.

621.397.645 1528

**Stagger-Peaked Video Amplifiers**—Easton. (*See* 1344.)

## TRANSMISSION

621.396.61 1529

**Description of a 1-kW Medium-Wave Broadcasting Transmitter**—P. Paris and J. Polonsky. (*Ann. Radioélec.*, vol. 3, pp. 293–298; October, 1948.) The feedback principles previously described (3377 of 1948) are applied in a transmitter for the range 272.5 to 750 kc; it can be used with or without crystal control and can form the driving unit of a high-power transmitter. Performance curves are given.

621.396.61.029.62 1530

**The "Little Slugger"**—P. S. Rand. (*QST*, vol. 33, pp. 11–17, 122; February, 1949.) Constructional and operating details of a low-power crystal-controlled 10-meter transmitter using narrow-band FM. Adequate filtering prevents television interference. The unit may be used as a narrow-band FM exciter for an AM transmitter of higher power.

621.397.61.029.63 1531

**Developmental Television Transmitter for 500–900 Mc/s**—R. R. Law, W. B. Whalley, and R. P. Stone. (*RCA Rev.*, vol. 9, pp. 643–652; December, 1948.) The development of tubes and circuits for a wide-band transmitter giving a peak output power of 1 kw is described. For video modulation, a pair of uhf pulse triodes are used for cathode modulation of the push-pull rf amplifier. A feature of the output tubes is the tungsten-wire grid which has negligible emission because of its excellent cooling properties.

621.396.615.141.2 1532

**The Turbator, a Single-Cavity Magnetron**—Lüdi. (*See* 1556.)

621.396.619:621.396.5 1533

**Systems of Modulation for Radio-Telephone Transmitters**—R. Vaudetti. (*Atti del Congresso internazionale della Radio* (Rome), p. 519; September and October, 1947. In Italian.) Summary only. In the first system, the modulation transformer is replaced by two inductances, each with two windings, acting as autotransformers. In the second, applicable where the final stage has tubes in parallel, the tubes of the modulated stage are arranged in series with those of the modulator stage; load matching is thus improved.

## VACUUM TUBES AND THERMIONICS

621.383 1534

**The Physical Characteristics of Silver Sulphide Photocells**—V. E. Kosenko and E. G. Miselyuk. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1369–1377; November, 1948. In Russian.)

621.383 1535

**The Structure and the Peculiarities in the Operation of Silver Sulphide Photocells**—I. R. Potapenko. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1356–1368; November, 1948. In Russian.)

621.383 1536

**The Frequency Characteristics of Lead Sulphide Photocells**—B. T. Kolomiets. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1456–1457; November, 1948. In Russian.)

621.383 1537

**Comparison of Lead-Sulfide Photoconductive Cells with Photoemissive Tubes**—N. Anderson and S. Pakswer. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 41–48; January, 1949.) A comparison of the electrical characteristics and a discussion of the modifications required in present practice in the motion-picture industry to obtain optimum performance.

621.385 1538

**Electron Tube Development and its Place in the Progress of Radio Art**—I. E. Mouromtseff. (*Atti del Congresso internazionale della Radio* (Rome), pp. 544–577; September and October, 1947. In English.) A general review of the main developments during the author's lifetime, with discussion of (a) water-cooled tubes, (b) sealing, (c) high-frequency and uhf tubes, (d) the split-anode magnetron, (e) normal and reflex klystrons, (f) tubes for pulse operation, (g) multi-cavity magnetrons, (h) the resnatron, and (i) the traveling-wave tube.

621.385 1539

**Use of Optical Polish in Valve Construction**—A. Danzin and E. Despois. (*Ann. Radioélec.*, vol. 3, pp. 280–289; October, 1948.) The conditions which must be satisfied for an airtight joint between two surfaces are discussed, experimental work is described and applications to the construction of both demountable and ordinary tubes are illustrated.

621.385 1540

**Analysis of a Simple Model of a Two-Beam Growing-Wave Tube**—L. S. Nergaard. (*RCA Rev.*, vol. 9, pp. 585–601; December, 1948.) The gain and bandwidth of a mathematical model of a tube in which a growing wave is produced by the interaction of two electron beams is investigated. The model consists of two admixed beams, infinite in extent, and uniform except in the common direction of their velocities. The gain per unit length of the model is found to depend on the operating frequency, the current densities and the velocities of the two beams. With a tube having an interaction space 30 cm long, a gain of 120 db at 3,000 Mc, with a bandwidth of 860 Mc, should be practicable.

621.385:621.316.726.078 1541

**The Transitrol, an Experimental Automatic-Frequency-Control Tube**—J. Kurshan. (*RCA Rev.*, vol. 9, pp. 687–703; December, 1948.) The transitrol combines the function of local oscillator and reactance device into a single unit particularly suitable for the vhf band. Experiments with commercial tubes which led up to its design are discussed; its operation is analyzed. Its uses in FM receivers and as a FM transmitter are considered. See also 1235 of May.

621.385.012 1542

**The Correspondence between the Static Characteristics and the Dynamic Parameters of Electrical Negative-Resistance Devices**—G. Cartianu. (*Onde Élec.*, vol. 29, pp. 44–50; January, 1949.) Certain types of static characteristic can only correspond to certain dynamic parameters and hence to certain conditions of stability. Thus a device with a dynatron type of characteristic behaves as an inductance, while one with a characteristic of the arc type behaves as a capacitance. General rules are given for determining, from the features of the static characteristic, whether a particular device is of the inductance or the capacitance type.

621.385.029.63/.64 1543

**Transverse Fields in Traveling-Wave Tubes**—J. R. Pierce. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 732–746; October, 1948.) A mathematical treatment of the problem of weak focusing fields. Traveling-wave tubes will have gain even if the rf field at the mean position of the electron stream is purely transverse. The addition of a longitudinal magnetic focusing field reduces the gain due to transverse fields and increases the electron velocity for optimum gain. See also 2284 of 1947.

621.396.615.141.2+621.385.029.63/.64 1544

**On the Properties of Valves using a Constant Magnetic Field: Part 3**—J. Brossart and O. Doehler. (*Ann. Radioélec.*, vol. 3, pp. 328–338; October, 1948.) A new type of tube is described, the traveling-wave tube with magnetic field. Two possible designs are considered, in each of which the cathode is a short cylinder with a gap, the anode a delay line of zigazg type surrounding the cathode, and a uniform magnetic field is applied in the direction of the axis. A high-frequency signal is applied to one end of the anode line and is picked up near the other end by a collector plate which also serves as a screen between input and output. The gain of such a tube is calculated, neglecting space-charge effects. The new tube has considerable advantages compared with the normal traveling-wave tube. The gain is much greater and electron currents of the order of 1 ampere can be used, compared with 10 to 20 milliamperes for the normal type of tube. The efficiency is also greater and there are only two systems of forced oscillations, while the normal tube has three. Parts 1 and 2, 261 of February. To be continued.

621.385.029.64:621.397.6 1545

**On the Help which some Recent Ideas concerning U.H.F. Valves can give in Television**—R. Warnecke and P. Guénard. (*Ann. Radioélec.*, vol. 3, pp. 259–280; October, 1948. Erratum, vol. 4, p. 92; January, 1949.) Full paper. Summary noted in 904 of April.

621.385.032.213 1546

**Spectral Emissivity and Electron Emission Constants of Thoria Cathodes**—T. E. Hanley. (*Jour. Appl. Phys.*, vol. 19, pp. 583–589; June, 1948.) The relation between true and brightness temperatures for cataphoretically deposited thoria cathodes is obtained and the constants of the Richardson equations are found from electron emission studies.

621.385.032.213:621.314.57 1547

**Hot-Cathode Gas-Filled Rectifying Valves**—R. Suart. (*Radio Franç.*, no. 1, pp. 17–20; January, 1949.) The characteristics of such tubes are discussed and two tubes with xenon filling (Types XV.550 and VX.7400) are described. These have mean rectified anode currents of 0.35 and 1.25 amperes respectively and a maximum inverse voltage of 10,000 volts.

621.385.032.216 1548

**Poisoning Effects in Oxide-Cathode Valves**—G. H. Metson. (*P.O. Elec. Eng. Jour.*, vol. 41, part 4, pp. 204–205; January, 1949.) Poisoning effects due to electron bombardment of the control grid of a tube occur at energies of the bombarding electrons corresponding to the heats of formation of compounds (oxide, chloride, and sulphate) contained in normal, cathode paste and apt to be evaporated from the cathode and deposited on the grids of a tube during manufacture. It is suggested that this poisoning occurs throughout the normal life of a tube and is reversible, the cathode reactivating by expulsion of gas when the supply of primary poisoning compounds on the grids is exhausted. In addition, there appears to be an allied, but irreversible, effect due to progressive failure of minute areas of the cathode to reactivate.

621.385.032.216 1549

**Note on the Ionic Conductivity of Oxide-Coated Cathodes**—S. Wagener. (*Proc. Phys. Soc.*, vol. 61, pp. 521–525; December 1, 1948.)

621.385.83.032.29 1550

**Design of an Electron Gun, taking Account of the Space Charge of the Beam**—H. Huber. (*Ann. Radioélec.*, vol. 4, pp. 26–32; January, 1949.) Relations between gun dimensions and electrical parameters are given for an electron beam with high current density and minimum dispersion. Experiments with the gun supplying the beam for a helix-type traveling-wave tube confirm the theory. The gun gives a current of 14 ma at 1,400 volts and after traversing the 28-cm length of the helix, the beam current reduction is only 3 per cent.

621.385.832 1551

**Cathode-Ray Tube with Cylindrical Screen**—A. Pinciroli. (*Atti del Congresso internazionale della Radio* (Rome), pp. 605–629; September and October, 1947. In Italian.) The essential parts are (a) an electron gun, (b) an arrangement of electrodes for providing a rotating electric field, (c) two trumpet-shaped electrodes between which the beam is deflected at right angles to the tube axis, (d) a cylindrical focusing lens, and (e) two deflecting electrodes to which a voltage proportional to the quantity under observation is applied. The advantages over a cathode-ray tube using normal cartesian or polar co-ordinates are: (a) extension of the axis of time, which is determined by means of two sinusoidal voltages of lower frequency than that of the quantity under observation, (b) two distinct electron beams can be used, and (c) there is no interaction between deflection electrodes and those which are used to determine the time axis.

621.396.615.14 1552

**The Resnatron**—G. E. Sheppard. (*Atti del Congresso internazionale della Radio* (Rome), pp. 643–654; September and October, 1947. In English.) A detailed description of the design and operating principles, with diagrams. The advantages over conventional tubes for high-power use at uhf are stressed.

621.396.615.141.2 1553

**Rising-Sun Magnetrons with Large Numbers of Cavities**—A. V. Hollenberg, N. Kroll, and S. Millman. (*Jour. Appl. Phys.*, vol. 19, pp. 624–635; July, 1948.) Disturbance from unwanted modes and the critical nature of anode and cathode dimensions prevented satisfactory operation with more than 26 open-end cavity-type vane resonators. Stable pie mode operation of pulsed magnetrons having up to 38 cavities was achieved for $\lambda \approx 1.25$ cm by the use of closed-end resonators. Increasing the number of resonators should enable existing magnetron techniques to be extended to millimeter wavelengths, and the rf power obtainable for $\lambda \geqq 1$ cm to be increased.

621.396.615.141.2 1554

**The Cavity Magnetron**—J. T. Randall. (*Atti del Congresso internazionale della Radio* (Rome), pp. 630–642; September and October, 1947. In English.) A review of the experimental development, carried out during the war, at Birmingham University; it includes design and method of construction, design of the high-frequency output circuit, and an outline of the underlying theory.

621.396.615.141.2 1555

**A Magnetron Resonator System**—E. C. Okress. (*Atti del Congresso internazionale della Radio* (Rome), pp. 578–600; September and October, 1948. In English.) A mathematical investigation of the normal modes of the symmetrical multi-sectional or vane-type magnetron. Graphs and equations enable the modes to be determined from anode-block dimensions. The relationship between the wavelengths for strapped and unstrapped resonator systems is investigated. Tables are given for designing reduced-scale prototypes for use on other wavelengths and voltages.

621.396.615.141.2 1556

**The Turbator, a Single-Cavity Magnetron**—F. Lüdi. (*Atti del Congresso internazionale della Radio* (Rome), pp. 529–543; September and October, 1947. In German.) See also 1660 of 1943.

621.396.615.141.2:[537.291+538.691 1557

**Motion of an Electron in a Cavity Magnetron**—M. Panetti. (*Atti del Congresso internazionale della Radio* (Rome), pp. 601–605; September and October, 1947. In Italian.) The resonance wavelengths for the different modes are first calculated by Abele's method (3823 of 1947). The electron trajectory is regarded as differing by a small amount from the cycloidal orbit due to uniform electromagnetic and electrostatic fields; equations of motion are given applicable when certain simplifying conditions are satisfied.

621.396.615.142 1558

**On the Efficiency of Velocity-Modulation Valves**—P. Guénard, R. Warnecke, and C. Fauve. (*Ann. Radioélec.*, vol. 3, pp. 302–327; October, 1948. Erratum, *ibid.*, vol. 4, p. 92; January, 1949.) The factors which tend to reduce efficiency are discussed. Methods of correcting imperfections of the form of the electron packets in the 2-cavity type of tube are considered; efficiency can thus be considerably improved. The 3-cavity tube appears to be the most suitable for obtaining high-power output with good efficiency. Calculation shows that an efficiency of the order of 45 per cent is to be expected for wavelengths of 10 to 20 cm.

621.396.615.142.2:[537.291+538.691 1559

**Electron Optics of H.F. Valves**—D. Charles. (*Ann. Radioélec.*, vol. 4, pp. 33–47; January, 1949.) A description of the electrolyte-tank method of determining electron trajectories is illustrated by results obtained for a reflex klystron. Details are also given of a simple method which makes use of a resistance grid. For systems of cylindrical symmetry, the resistances are all equal; for revolution symmetry, the resistances are graded. Results obtained are sufficiently accurate for practical design. An analytical method is described which enables corrections for space-charge effects to be applied.

621.396.822 1560

**Noise in Linear Networks**—Schremp. (See 1488.)

## MISCELLANEOUS

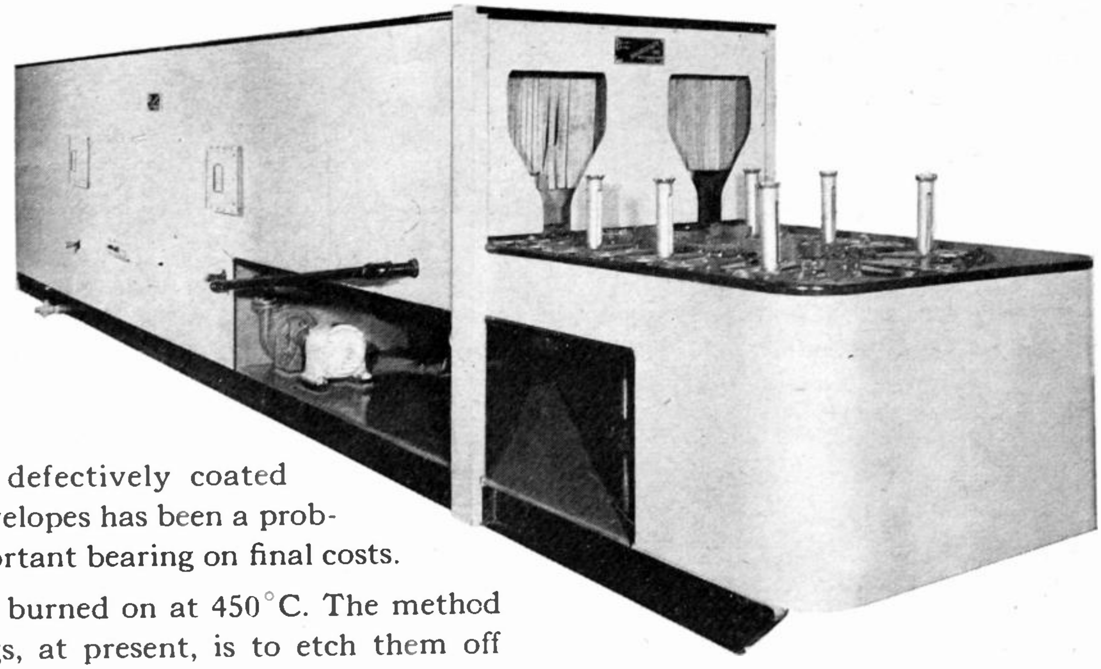
022.3:621.38.001.8 1561

**The Patent Office Library and Electronics**—R. Neumann. (*Electronic Eng.* (London), vol. 21, pp. 52–57; February, 1949.) A general description, with special reference to electronic subjects and the indexing system.

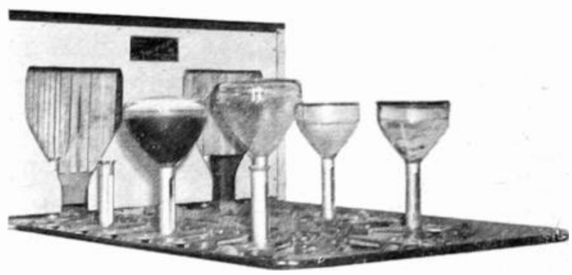
621.396 1562

**Elektrische Wellen [Book Review]**—W. O. Schumann. C. Hanser Verlag, München, 340 pp. (*Wireless Eng.*, vol. 25, p. 370; November, 1948.) Based on a course of lectures at Munich University. The treatment is detailed and well illustrated.









## SECTION MEETINGS

### BALTIMORE

"Development of a Large Metal Kinescope for Television," by H. P. Steier, Radio Corporation of America; March 22, 1949.

### BUFFALO-NIAGARA

"Electrical Problems in Stratovision," by A. A. Nimms, Westinghouse Electric Company; February 17, 1949.

"Some Recent Developments and Applications of Germanium Crystals," by S. T. Martin, Sylvania Electric Products Company; March 16, 1949.

### CEDAR RAPIDS

"Some Aspects of Annular Circuits," by D. Priest, Eitel-McCullough, Inc.; March 15, 1949.

"Current Problems in Frequency Allocation," by G. F. Leydorf, Radio Station WJR; April 13, 1949.

### CINCINNATI

"New Seeds for the Fields of Engineering," by G. Wenot, Science Illustrated; March 23, 1949.

### COLUMBUS

"Recent Developments in Sound Recording," by T. Lynch, Electronic Engineering Brush Development Company; February 9, 1949.

"Television Demonstration," by J. A. Hatchwell, Engineering RCA Service Organization; March 23, 1949.

### CONNECTICUT VALLEY

"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; March 24, 1949.

### DALLAS-FORT WORTH

"High Frequency—Applications in Aircraft, by D. T. Doherty, Student, Southern Methodist University; March 21, 1949.

"Electrical Techniques in Geophysics," by J. W. Roby, Student, Southern Methodist University; March 21, 1949.

"Amplifying the Brain Waves," by H. J. Spurlock, Student, Southern Methodist University; March 21, 1949.

"A Photoelectric Brightness Meter," by J. Fleck, Student, Southern Methodist University; March 21, 1949.

### DAYTON

"Related Frontier Problems in Electronics and Aerodynamics," by S. Ramo, Hughes Aircraft Company; April 14, 1949.

### DENVER

"Geiger Counters and Uranium Exploration," by P. M. Lahue, Consulting Engineer; April 8, 1949.

### DES MOINES-AMES

"Television Networks—Coaxial and Microwave," by F. E. Baird, Northwestern Bell Telephone Company; March 16, 1949.

"A Report on the Annual Convention," by C. F. Quentin, Cowles Broadcasting Company; March 16, 1949.

Election of Officers; March 16, 1949

### DETROIT

"Engineering Aspects of the Transistor," by J. A. Morton, Bell Telephone Laboratories; March 22, 1949.

### FORT WAYNE

"The Development and Application of Railroad Communication," by A. A. Curry, Farnsworth Television and Radio Company; March 28, 1949.

*(Continued on page 36A)*

## 10 CENTIMETER

WAVEGUIDE DIRECTIONAL COUPLER, 27 db. Navy type CABV-47AAN, with 4 in. slotted section as shown ........$42.50
SQ. FLANGE to rd choke adapter, 18 in. long OA 1½ in. x 3 in. guide, type "N" output and sampling probe ........$32.00
"S" BAND CRYSTAL MOUNT, gold plated, with 2 type "N" connectors ........$12.50
PICKUP LOOP, Type "N" Output ........$2.75
TR BOX Pick-up Loop ........$1.25
POWER SPLITTER: 726 Klystron input, dual "N" output ........$5.00
MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated ........$27.50
10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or dc arranged switching motor. Mfg. Raytheon. CRP 24AAS. New and complete ........$150.00
10 CM END-FIRE ARRAY POLYRODS ....$1.75 ea.
"S" BAND Mixer Assembly, with crystal mount, pick up loop, tunable output ........$3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers ........$12.50
10 CM. McNALLY CAVITY Type SG ........$3.50
WAVEGUIDE SECTION, MC 445A, rt. angle bend, 5½" ft. OA. 8" slotted section ........$21.00
10 CM OSC. PICKUP LOOP, with male Homedell output ........$2.00
10 CM DIPOLE WITH REFLECTOR in lucite ball, with type "N" or Sperry fitting ........$4.50
10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola ⅞" Rigid Coax Input $8.00

### ⅞" RIGID COAX—⅜" I.C.

RIGHT ANGLE BEND, with flexible coax output pick-up loop ........$8.00
SHORT RIGHT ANGLE bend, with pressurizing nipple ........$3.00
RIGID COAX to flex coax connector ........$3.50
STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length ........$5.00
RT. ANGLES for above ........$2.50
RT. ANGLE BEND 15" L. OA ........$3.50
FLEXIBLE SECTION, 15" L. Male to female $4.25
MAGNETRON COUPLING to ⅞" rigid coax, with TR pickup loop, gold plated ........$7.50
ROTATING JOINT—P O SCR584 ........$35.00
DIPOLE ASSY—P/O SCR584 ........$25.00
FLEX COAX SECT. Approx 30 ft. ........$16.50

### ⅞" RIGID COAX—¼" I.C.

⅞ RIGID COAX. Bead Supported ........$1.20
SHORT RIGHT ANGLE BEND ........$2.50
Rotating joint, with deck mounting ........$15.00
RIGID COAX slotted section CU-60/AP ........$5.00

## MICROWAVE ANTENNAS

APS-15 Antennas. New ........$99.50
AN MPG-1 Antenna. Rotory feed type high speed scanner antenna assembly, including horn parabolic reflector. Less internal mechanisms. 10 deg. sector scan. Approx. 12'L x 4'W x 3'H. Unused. (Gov't Cost—$4500.00) ........$250.00
APS-4 3 cm. antenna. Complete. 14½ dish. Cutler feed dipole directional coupler, all standard 1" x ½" waveguide. Drive motor and gear mechanisms for horizontal and vetical scan. New, complete ..$65.00
AN/TPS3. Parabolic dish type reflector approx. 10' diam. Extremely lightweight construction. New. In 3 carrying cases ........$89.50
RELAY SYSTEM PARABOLIC REFLECTORS: approx. range: 2000 to 6000 mc. Dimensions: 4¼' x 3' rectangle, new ........$85.00
TDY "JAM" RADAR ROTATING ANTENNA, 10 cm. 30 deg. beam, 115 v.a.c. drive. New ........$100.00
SO-13 ANTENNA, 24" dish with feedback dipole 360 deg. rotation, complete with drive motor and selsyn. New ........$128.00
Used ........$45.00
DBM ANTENNA, Dual, back-to-back parabolas with dipoles. Freq. coverage 1,000-4,500 mc. No drive mechanism ........$65.00
AS125/APR Cone type receiving antenna, 1080 to 3208 megacycles. Now ........$4.50
140-600 MC. CONE type antenna, complete with 25' sectional steel mast, guys, cables, carrying case, etc. New ........$49.50
ASD 3 cm. antenna, used, ex. cond ........$49.50

## RADAR SETS

RC 145 IFF SET. Consists of BC 1267 xmtr-revr, remote antenna controller and indicator 1-221, power supply RA 105-A, 1 kw. pulse oscillator operates on 154-186 mc. Operates from 117 v., 60 cy. New $190.00
SN RADAR-GE, low power, 5 and 25 miles ranges, Uses GL446 as pulsed oscillator, 5" "A" scope, "S" band. Extremely compact, ideal for demonstration and laboratory work. 115V 60 C operation. Used. Excel. cond. ........$600.00
SE 10 CM SURFACE SEARCH RADAR. W.E.. 20,000 to 80,000 yds. range. 250 KW. pk. power input to 706 magnetron. Thyratron modulator, variable pulse rate. Complete set including spare parts, tubes, waveguide and fittings. Send for price and additional information.
R85TPL-1 RADAR RCVR, Sperry ........$85.00

## *Buy with Confidence*

## MICROWAVE PLUMBING

### 1.25 CENTIMETER

"K" BAND FEEDBACK TO PARABOLA HORN, with pressurized window ........$30.00
MITRED ELBOW cover to cover ........$4.00
TR/ATR SECTION choke to cover ........$4.00
FLEXIBLE SECTION 1" choke to choke ........$5.00
ADAPTER, rd. cover to sq. cover ........$5.00
MITRED ELBOW and S sections choke to cover..$4.50
WAVE GUIDE ½ x ¼ per ft. ........$1.00
K BAND CIRCULAR FLANGES ........50¢

*************************************************

## PULSE EQUIPMENT

### MODULATOR UNIT BC 1203-B

Provides 200-4,000 PPS, Sweeptime: 100 to 2,500 microsec. in 4 steps, fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse sweep voltages, calibration voltages, fil. voltages. Operates 115 vac. 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio 804B or 804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New as shown ........$125.00
MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 144 KW (12 KV at 12 amp). Duty Ratio: .001 max. Pulse duration: .5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-72's, 1-73. New ........$110.00
APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro Sec. Rep. rate 624 to 1348 Pps. Pk. pwr. out 35 KW. Energy 0.018 Joules ........$49.00
TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes ........$49.50
APS-10 MODULATOR DECK. Complete, less tubes ........$78.00
APS-10 Low voltage power supply, less tubes ..$18.50

### PULSE NETWORKS

15A—1-400-50: 15 KV, "A" CKT, 1 microsec., 400 PPS, 50 ohms imp ........$42.50
G.E. $6E3-5-2000-50P2T, 6KV, "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance ..$6.50
G.E. $3E (3-.84-810; 8-2.24-405) 50P4T: 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec, 810 PPS, 50 ohms imp.; Unit 2, 8 Sections, 2.24 Microsec, 405 PPS, 50 ohms imp. ........$6.50
7.5E3-1-200-67P, 7.5 KV, "E" Circuit, 1 microsec, 200 PPS, 67 ohms impedance, 3 sections ........$7.50
7.5E4-16-60-67P, 7.5 KV, "E" circuit, 4 sections, 16 microsec, 60 PPS, 67 ohms impedance ........$15.00
7.5E3-3-200-6PT, 7.5 KV, "E" Circuit, 3 microsec, 200 PPS, 67 ohms imp., 3 sections ........$12.50

### DELAY LINES

D-168184: .5 microsec. up to 2000 PPS, 1800 ohm term ........$4.00
D-170499: .25/.50/.75/. microsec. 8 KV, 50 ohms imp. ........$16.50
D-165997: 1¼ microsec. ........$7.50

### PULSE TRANSFORMERS

G.E.K.-2745 ........$39.50
G.E.K.-2744-A, 11.5 KV High Voltage, 3.2 KV Low Voltage @ 200 KW oper. (270 KW max.) 1 microsec. or ¼ microse. @ 660 PPS ........$39.50
W.E. $D166173 HI-Volt input transformer W.E. Impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc to 2 mc. 2 sections parallel connected, potted in oil ........$36.00
W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1. Frequency range: 380-520 c.p.s. Permalloy core ........$6.00
G.E. $K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms, Sec. Imp: 450 Ohms, Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK, Sec. Output: 28 KV PK. Peak Output: 800 KW, Bifilar 2.75 Amp. ....$64.50
W.E. $D169271 Hi Volt input pulse Transformer $27.50
G.E. K2450A Will receive 13KV, 4 micro-second pulse on pri., secondary delivers 14KV. Peak power out 100KW GE ........$45.00
G.E. $K2748A Pulse Input, line to magnetron ..$36.00
$9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1 1/8 x 19/32" ........$1.50

## MISCELLANEOUS

DEHYDRATING UNIT, 60 lb. capacity, 115 v. 60 cy. operation, 2' x 22" x 15". New and complete $425.00
F129 SPP-2. COAX WAVETRAP ........$12.50

### 3 CENTIMETER PLUMBING

(STD. 1" x ½" GUIDE, UNLESS OTHERWISE SPECIFIED)

TRANSITION: 1 x ½ to ¼ x ⅝, 14 in. L $8.00
"X" BAND PREAMPLIFIER, consisting of 2-723 A/B local oscillator-beacon feeding waveguide and TR/ATR Duplexer sect. incl. 60 mc IF amp $67.50
Random Lengths waveguide 6" to 18" Lg. ...$1.10 Ft.
WAVEGUIDE RUN, 1¼" x ½" guide, consisting of 4 ft. section with Rt. angle bend on one end 2" 45 deg. bend other end ........$8.00
WAVEGUIDE RUN, 1¼" x ½" guide, consisting of 4 ft. long ........$10.00
DUMMY LOAD, Ts 332/UP ........$22.50
"X" BAND PRESSURIZING gauge section w/15-lbs gauge & Pressurizing Nipple ........$18.50
45 DEG. TWIST 6" Long ........$10.00
12 SECTION 45 deg twist 90 deg. bend ........$6.00
11 STRAIGHT WAVEGUIDE section choke to cover Special heavy Construction, silver plated ........$4.50
15 DEG BEND 10" choke to cover ........$4.50

5 FT. SECTIONS choke to cover. Silver Plated ........$14.50

18" FLEXIBLE SECTION ........$17.50
"E" and "H" PLANE BENDS ........$12.50
BULKHEAD FEED THRU ........$15.00
"X" BAND WAVEGUIDE 1¼" x ⅝" OD 1/16" wall Aluminum ........Per Foot $ .75
WAVEGUIDE 1" x ½" I.D. ........Per Foot $1.50
TR CAVITY For 724-A TR Tube ........$3.50
3" FLEX SECT. sq. flange to Circ Flang Adapt. $7.50
724 TR TUBE (41-TR-1) ........$2.50
SWR MEAS. SECTION, 4" L with 2 type "N" output probes MTD full wave apart. Bell size guide, Silver plated ........$10.00
ROTARY JOINT with slotted section and type "N" output pickup ........$17.50
WAVEGUIDE SECTION, 12" long choke to cover 45 deg. twist & 2½" radius, 90 deg. bend ........$4.50
SLUG TUNER/ATTENUATOR, W.E. guide, gold plated ........$6.50
TR/ATR DUPLEXER Sect. w/iris flange ........$8.00
TWIST 90 deg. 5" choke to Cover w/press nipple $6.50
WAVEGUIDE SECTIONS 2½ ft. long silver plated with choke flange ........$5.75
WAVEGUIDE 90 deg. bend "E" Plane, 18" Lg $4.00
ROTARY JOINT choke to choke ........$17.50
ROTARY JOINT choke to choke with deck mounting ........$17.50
S. CURVE WAVEGUIDE 8" Lg. cover to choke $3.50
DUPLEXER SECTIONS FOR 1B24 ........$10.00
CIRCULAR CHOKE FLANGES solid brass ........$.55
SQ. FLANGES FLAT BRASS ........ea. $.85
APS-10 TR/ATR DUPLEXER section with additional IRIS Flange ........$10.00
FLEXIBLE WAVEGUIDE ........$4.00/FT.
723 A B Klystrow Mount shielded on X Band Guide with Rough Attenuator Output ........$75.00 ea.

## MICROWAVE GENERATORS

AN/APS-15A "X" Band compl. RF head and modulator, incl. 725-A magnetron and magnet, two 723A/B klystrons (local osc. & beacon), 1B24 TR, rcvr-ampl, duplexer, HV supply, blower, pulse xtmr. Peak Pwr. Out: 45 KW apx. Input: 115, 400 cy. Modulator pulse duration .5 to 2 micro-sec. apx. 13 KV Pk Pulse. Compl. with all tubes incl. 715-B, 829B, RKR 73, two 72's. Compl. pkg., new ........$210.00
APS-15B. Complete pkg. as above, less modulator ........$150.00
"S" BAND AN/APS-2. Complete RF head and modulator, including magnetron and magnet, 417-A mixer, TR, receiver, duplexer, blower, etc., and complete pulser. With tubes, used, fair condition ....$75.00
10 CM. RF Package. Consists of SO Xmtr.-receiver using 2I27 magnetron oscillator, 250 KW peak input, 707 receiver-mixer, rotating yolk PPI for above $150.00
Modulator-motor-alternator unit for above....$ 75.00
Receiver-rectifier power unit for above........$ 25.00
Rotating Ant. with parabolic reflector for above. New ........$ 75.00

| THERMISTORS | |
|---|---|
| D-167332 (tube) | $.95 |
| D-170396 (bead) | $.95 |
| D-167613 (button) | $.95 |
| D-166228 (button) | $.95 |
| D-164699 for MTG. in "X" band Guide | $2.50 |
| D-167018 (tube) | $.95 |

| COAX PLUGS | |
|---|---|
| 831SP | $.35 |
| 831AP | $.35 |
| 831HP | $.15 |
| UG 21/U | $.85 |
| UG 86 U | $.95 |
| UG 254/U | $.75 |

| VARISTORS | |
|---|---|
| D170225 | $1.25 |
| D-167176 | $.95 |
| D-168687 | $.95 |
| D-171812 | $.95 |
| D-171528 | $.95 |
| D-168549 | $.95 |
| D-162482 | $3.00 |
| D163298 | $1.25 |
| D-99428 | $2.00 |
| D161871A | $2.85 |
| D171121 | $.95 |
| 3A(12-43) | $1.50 |
| D167020 | $3.00 |

## MAGNETRON MAGNETS

| GAUSS | POLE DIAM. | SPACING | PRICE |
|---|---|---|---|
| 5200 | 21/32 in. | ⅝ in. | $17.50 |
| 4850 | ¾ in. | ⅝ in. | $12.50 |
| 1860 | 1⅝ in. | 1½ in. | $14.50 |
| D161392* | 1⅝ in. | 1 5/16 in. | $12.50 |

* Mfr's Number.

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RADAR TELEVISION ELECTRONICS
FM MICROWAVE RADIO SONAR

## SECTION MEETINGS

*(Continued from page 34A)*

INYOKERN

"Radio the Hard Way," by G. R. Sams, United States Navy; March 30, 1949.

"Summaries of IRE National Convention," by F. M. Ashbrook and T. Parkin, United States Navy; March 30, 1949.

KANSAS CITY

"The Baldwin Electronic Organ," by J. F. Jordan, Baldwin Company; January 12, 1949.

"Stratovision," by A. Nims, Westinghouse Electric Corporation; January 25, 1949.

"Oscillographs and their Application to Industry and Research," by A. Rich, Dumont Laboratories; February 15, 1949.

"Technical and Operational Aspects of the VHF Omni-Range," by F. L. Moseley, Collins Radio Company; March 15, 1949.

LOS ANGELES

"Electronic Applications in Seismic Prospecting," by M. L. Swan, United Geophysical Company; March 15, 1949.

LOUISVILLE

"The Operation of a Modern Television Station," by W. E. Hudson, Radio Station WAVE; April 11, 1949.

MONTREAL

"Creative Technology," by M. J. Kelly, Bell Telephone Laboratories; February 24, 1949.

"An Experimental Design of a Simple FM Receiver," by T. L. Taylor, Student, McGill University; March 9, 1949.

"Link-Coupled Tank Circuits," by R. Tennet, Student Laval University; March 9, 1949.

"Varistors in Negative Feedback Amplifiers," by G. W. Holbrook, Queens University; March 9, 1949.

"An Adjustable Frequency Stroboscope Suitable for Taking Multiple Exposure Photographs," by H. M. Sullivan, Carleton College; March 9, 1949.

"A Radiating Pattern Computor and Phase-Modulating Tube," by L. Pigeon, Student, University of Montreal; March 9, 1949.

"Acoustic Measurements of Broadcasting Studios," by R. E. Santo, Canadian Broadcasting Company; March 23, 1949.

"Television Site Survey Measurements," by J. E. Hayes, Canadian Broadcasting Corporation; March 23, 1949.

"The Selection of Magnetic Core Materials," by H. F. Porter, Magnetic Metals Company; April 6, 1949.

NEW MEXICO

"Telemetering on Power Transmission Lines," by V. B. Wilfey, Westinghouse Electric Company; March 25, 1949.

OTTAWA

"Acoustic Measurement of Broadcast Studios," by R. E. Santo, Canadian Broadcasting Corporation; March 17, 1949.

"Television Site Survey Measurements," by J. E. Hayes, Canadian Broadcasting Corporation; March 17, 1949.

"Some Developments in Direction Finding and the Speed of Radio Waves," by R. L. Smith-Rose, Department of Scientific and Industrial Research; March 31, 1949.

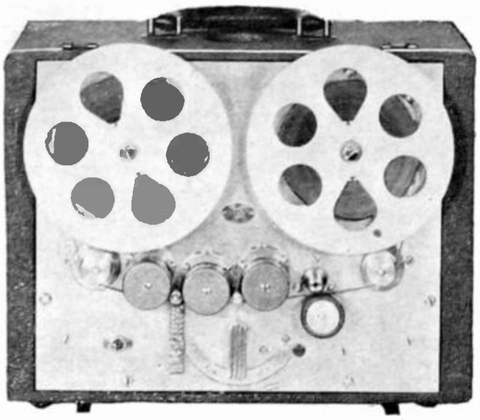
PHILADELPHIA

"Magnetic Amplifiers," by E. Weber, Brooklyn Polytechnic Institute; April 7, 1949.

PORTLAND

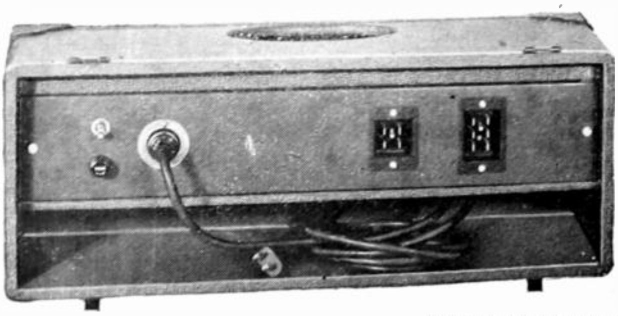
"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; February 15, 1949.

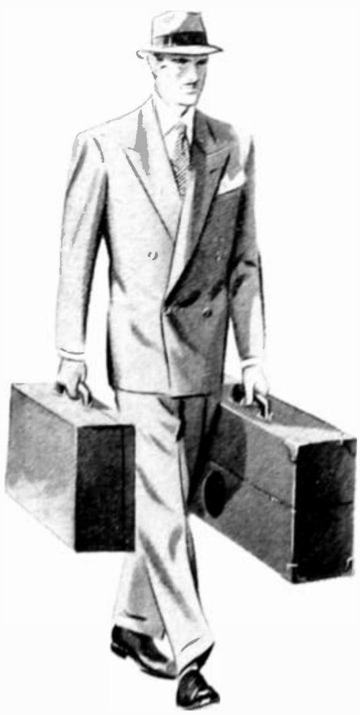
*(Continued on page 38A)*



# comparisons

**indicate this is the world's finest recorder of its type**





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(Continued from page 36A)

PRINCETON

"Television Receiver with 16-Inch Metal Kinescope," by E. L. Clark, Radio Corporation of America; April 14, 1949.

ROCHESTER

"The Block Signal System for Airways Traffic Control," by H. C. Kendall, General Railway Signal Company, March 24, 1949.

SALT LAKE

"Measurements at Ultra-High Frequencies, by W. R. Hewlett, Hewlett-Packard Company; March 21, 1949.

SAN ANTONIO

"Impedance Matching with Transmission Lines and Waveguides," by F. E. Brooks, Jr., University of Texas; March 21, 1949.

SAN FRANCISCO

"Power Systems Communication," by R. H. Miller, Pacific Gas and Electric Company; March 14, 1949.

TORONTO

"Acoustic Measurements of Broadcasting Studios," by R. E. Santo, Canadian Broadcasting Corporation; March 21, 1949.

"Television Site Survey Measurements," by J. E. Hayes, Canadian Broadcasting Corporation; March 21, 1949.

"CBL—CJBC Transmitters," by J. C. Punchard, Northern Electric Company, Ltd.; April 11, 1949.

TWIN CITIES

"The Mechanism of Electronic Digital Computers," by A. A. Cohen, Engineering Research Associates, Inc.; March 1, 1949.

WASHINGTON

"Electronic Arithmetic," by C. N. Hoyler, RCA Laboratories; April 11, 1949.

*SUBSECTIONS*

LANCASTER

"A Record Changer and Record of Complimentary Design," by A. D. Burt and H. I. Reiskind, RCA Victor Division; April 6, 1949.

Election of Officers; April 6, 1949.

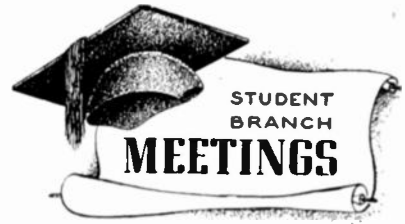
LONG ISLAND

"Intercarrier Sound Television Receiver Circuits," by S. W. Seeley, Radio Corporation of America; February 16, 1949.

NORTHERN NEW JERSEY

"Nuclear Energy," by K. K. Darrow, Bell Telephone Laboratories; March 23, 1949.

"The Ampex Magnetic Tape Recorder," by W. O. Summerlin, Audio and Video Products Corporation; March 23, 1949.




UNIVERSITY OF ALBERTA, IRE BRANCH

"The Value of Education in Engineering," by R. M. Hardy, Dean of Faculty of Engineering, University of Alberta; March 15, 1949.

(Continued on page 40A)

## STUDENT BRANCH MEETINGS

*(Continued from page 38A)*

UNIVERSITY OF CALIFORNIA SOCIETY OF ELECTRICAL ENGINEERS, IRE-AIEE BRANCH

"Silicones and their Application to Transformers," by M. L. Manning, Kuhlman Electric Company; March 2, 1949.

CASE INSTITUTE OF TECHNOLOGY, IRE BRANCH

"Television," by E. M. George, WXEL (TV); March 1, 1949.

"Synthetic Piezoelectric Crystals," by Mr. Arnst, Brush Development Company; March 15, 1949.

COLUMBIA UNIVERSITY, IRE-AIEE BRANCH

"Selenium Rectifiers," by Mr. Lobenstein, Selenium Rectifier Division of Radio Receptor Corporation; March 25, 1949.

UNIVERSITY OF DAYTON, IRE BRANCH

"Television Receivers," by W. Horst, Student, March 1, 1949.

"Colored Television," by D. Moore, Student; March 8, 1949.

"Telephony," by B. O'Brien, Student; March 15, 1949.

"Strowger System," by J. Quinlisk, Student; March 22, 1949.

"Telephone Trouble Shooting," by E. Johnson, Student; March 29, 1949.

UNIVERSITY OF DETROIT, IRE-AIEE BRANCH

"FM versus AM," by E. Clark, Radio Station WJLB, R. MacDonald, Radio Station WJR, C. Wesser, Radio Station WWJ, C. Kocher, Radio Station WXYZ, L. West, Radio Station WDET, J. F. Steadley, Radio Station WLDM and C. Dook, Radio Station WTDR; January 6, 1949.

"Technical Problems in the Broadcasting Studio," by A. Friedenthal, Radio Station WJR; March 9, 1949.

"Industrial Applications of Television," by J. Good, Diamond Power Specialties Company; March 29, 1949.

UNIVERSITY OF FLORIDA, IRE-AIEE BRANCH

"The Development of the Scientific Method in Physics," by S. Dushman, General Electric Company; March 23, 1949.

GEORGE WASHINGTON UNIVERSITY, IRE BRANCH

"Telemetering," by W. J. Mayo-Wells, John Hopkins University; March 2, 1949.

STATE UNIVERSITY OF IOWA, IRE BRANCH

"Symmetrical Components," by C. Lodge, Graduate Assistant; March 16, 1949.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH

"Difficulties of Long-Range Television Reception," by D. Berk, Student; March 10, 1949.

"Series Computers," by D. Seib, Student; March 10, 1949.

"Wire Recorder," by J. Wright, Student; March 10, 1949.

UNIVERSITY OF MAINE, IRE BRANCH

"New Type of FM Detector," by P. Howells, Faculty staff; March 17, 1949.

"Nockel-Cadmium Batteries," by N. B. Hamlin, Student; March 17, 1949.

"Illumination—Development of Lamps and their Applications," by R. G. Slauer, Sylvania Electric Products Company; April 14, 1949.

Election of Officers; April 14, 1949.

MASSACHUSETTS INSTITUTE OF TECHNOLOGY, IRE-AIEE BRANCH

"Electric Power Systems," by G. Orrok, Boston Edison Company; March 23, 1949.

*(Continued on page 41A)*

## STUDENT BRANCH MEETINGS

(Continued from page 40A)

"Industrial Research," by M. J. Kelley, Bell Telephone Laboratories; April 6, 1949.

University of Michigan, IRE-AIEE Branch

"Engineering Opportunities," by J. R. North, Commonwealth and Southern Corporation; March 23, 1949.

Election of Officers; March 23, 1949.

Missouri School of Mines, IRE-AIEE Branch

"Investigation of a Uniformly-Inductively Loaded Antenna," by E. L. Hughes, Student; March 23, 1949.

"Application of Tensor Analysis to Synchronous Machines," by P. F. Schaefer, Student; March 23, 1949.

University of Nebraska, IRE-AIEE Branch

"Power Production from Nuclear Chain Reaction," by T. Jorgensen, University of Nebraska; April 6, 1949.

Newark College of Engineering, IRE Branch

"Psycho-Acoustics," by D. C. Spitz, Student; March 25, 1949.

University of New Mexico, IRE Branch.

"Antennas at Optimum Frequencies," by T. C. Church and B. J. Bittner, Sandia Base Laboraton; February 17, 1949.

City College of New York, IRE Branch

"Television Receiver Problems," by B. M. Cole, North American Philips Company, Inc.; March 29, 1949.

New York University, IRE Branch

"Radar Systems," by R. Shetzline, Bell Telephone Laboratories; April 6, 1949.

University of North Dakota, IRE-AIEE Branch

Election of Officers; February 16, 1949.

"Personnel Work," by C. E. Scott, University of North Dakota; March 16, 1949.

University of Notre Dame, IRE-AIEE Branch

"Electronics in the Electric Utility Field," by E. F. Kenefake, General Electric Company; February 23, 1949.

"Patent Law," by E. Knoblock, Patent Attorney; March 25, 1949.

Pratt Institute, IRE Branch

"Infinite Q Filters," by W. Heacock, Student; February 24, 1949.

"Wave Theory of Matter," by O. Shuart, Student; March 10, 1949.

St. Louis University, IRE Branch

"What is Expected of the Graduate Engineer in Industry," by W. Bennettsen, Emerson Electric Manufacturing Company; March 10, 1949.

Election of Officers; March 24, 1949.

"Illinois State Police Communications," by W. Friedrich, Student; March 31, 1949.

Syracuse University, IRE-AIEE Branch

"Expansion in the Power Field," by Mr. Pratt, New York Central Power Company; March 9, 1949.

"Electronics in the Utility Field," by E. W. Kenefake, General Electric Company; March 30, 1949.

Election of Officers; March 30, 1949.

University of Tennessee, IRE Branch

"Recent Developments in Electrical Engineering," by P. C. Cromwell, University of Tennessee, February 22, 1949.

(Continued on page 42A)

## Student Branch Meetings

*(Continued from page 41A)*

THE UNIVERSITY OF TEXAS,
IRE-AIEE BRANCH

"Installation of a 138 Kv Cable," by J. Foley, City Public Service Board of San Antonio; February 7, 1949.

"Mobile Telephone System in Austin," by M. Bassford, W. Graves, and R. Via, Southwestern Bell Telephone Company; February 21, 1949.

"Presentation of Student Papers—AIEE Technical Paper Contest," by C. C. Young, L. F. Bell, K. H. Powers, D. W. Spence, G. B. Wright, E. D. Lovelady, and Herman Willi, Students; March 7, 1949.

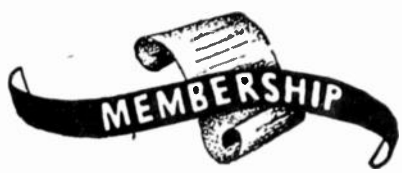
"Impedance Matching in Wave Guides," by F. E. Brooks, The University of Texas; March 21, 1949.

UTAH STATE AGRICULTURAL COLLEGE,
IRE BRANCH

"Teletype," by F. Gunnell, Bell Telephone System; March 30, 1949.

WORCESTER POLYTECHNIC INSTITUTE,
IRE-AIEE BRANCH

Election of Officers; March 22, 1949.




**The following transfers and admissions were approved and will be effective as of June 1, 1949:**

### Transfer to Senior Member

Barclay, J. N., 2915 Red River St., Austin 22, Tex.
Clewell, D. H., Magnolia Petroleum Co., Box 900, Dallas, Tex.
Dobosy, J. F., 31748 Lake Rd., Avon Lake, Ohio
Gambel, E. H., Curtiss-Wright Corp., 4300 E. Fifth Ave., Columbus 16, Ohio
Grimwood, F. O., 1325 N. Sixth St., Quincy, Ill.
Hershfield, S., 5 Glenwood Rd., Essex, Baltimore 21 Md.
Kean, W. F., 114 Northgate Rd., Riverside, Ill.
Muchmore, R. B., 571½ Oxford Ave., Hawthorne, Calif.
Parker, B. E., 338 S. 12 St., Quincy, Ill
Pasek, D. M., 255-04 West End Dr., Great Neck, L. I., N. Y.
Peake, H. J., 715 S. Washington St., Alexandria, Va.
Watts, H. M., Radiation Laboratory, The Johns Hopkins University, Baltimore 2, Md. (Effective December 1, 1948)

### Admission to Senior Member

Bischoff, A. F. H., 44 Hyde Blvd., Ballston Spa, N. Y.
Corrington, M. S., 12 E. Cottage Ave., Haddonfield, N. J.
Murphy, O. J., 410 Central Park West, New York 25, N. Y.
Townsend, F. H., 21 Park Lane, Histon, Cambs., England
Young, F. C., 100 Carlson Rd., Rochester 3 N. Y.

### Transfer to Member

Buehler, F. A., 665 W. Warren Ave., Detroit 1, Mich.
Carlstrom, T. H., Sylvania Electric Products, Inc., Emporium, Pa.
Davidson, H. H. A., 1376 Esquimalt Rd., Esquimalt, B. C., Canada
Ghose, A., Government Engineering College, Jubbulpore, India
Hillyer, C., 54-60 Lafayette St., New York 13, N. Y.
O'Donnell, J. A., 63 Wyneva St., Philadelphia 44, Pa.

*(Continued on page 43A)*

## MEMBERSHIP

(Continued from page 42A)

Oszy, A. J., 149-56 Elm Ave., Flushing, L. I., N. Y.
Robertson, A. S., 10 W. 48 St., New York 19, N. Y.
Schwab, C. B., 508—19 St., N.W., Massillon, Ohio
Shrock, J. B., Sandia Base Branch, Albuquerque, N. M.
Striplin, J. E., 2713 W. 154 St., Gardena, Calif.
Tilly, V. F., 31195 Merriman Pl., Inkster, R.F.D. 3, Mich.
Vane, A. B., 128 Blackburn Ave., Palo Alto, Calif.
Vought, N. F., 7455—64 Lane, Glendale, L. I., N. Y.
Walchli, H. E., 144 Waddell Circle, Oak Ridge, Tenn.
Webster, N. D., 515 Blackwood, N. Sacramento, Calif.
Weedman, W. F., 2712 Second St., S. E., Washington 20, D. C.
Wheeler, W. R., 8330 Kew Gardens Rd., Kew Gardens, N. Y.

### Admission to Member

Bibbero, R. J., 773 Potomac Ave., Buffalo 9, N. Y.
Biberman, L. M., 103 Mitscher Rd., China Lake, Calif.
Bright, F. W., 630 Randolph St., Dayton 8, Ohio
Carman, W. H., Station KOAT, Box 1419, Albuquerque, N. M.
Cotton, A. F. R., 2145 California St., N.W., Washington 8, D. C.
Dibos, R. A., 732 Charles St., Glenside, Pa.
Doane, H. J., 1851 AACS Liaison Unit, R.C.A.F., Station, Edmonton, Ala., Canada
Ewing, D. L., 309B Fowler, China Lake, Calif.
McCluskey, V., 5 W. 63 St., New York 23, N. Y.
Nelson, C. V., Ipswich Rd., East Boxford, Mass.
Purinton, R. M., 127 St. & Archer Ave., Lemont, Ill.
Rohrer, R. E., 1339 Wisconsin Ave., N.W., Washington 7, D. C.
Scott, H. G., Box 35, Glebe, N.S.W., Australia
Yewell, P. G., Box 102, West Medway, Mass.

### The following admissions to Associate were approved and were effective as of May 1, 1949:

Altmann, S., 264 W. 22 St., New York 11, N. Y.
Amick, R. H., 20070 Shiawassee, Detroit 19, Mich.
Angus, D. G., 117 E. Colorado St., Pasadena, Calif.
Argall, G. J., 1527 W. Rosemont Ave., Chicago 26, Ill.
Benson, G. G., 3915 N. State St., Jackson, Miss.
Berry, R. E., General Electric Co., Owensboro, Ky.
Bhargava, A. N., 30 Civil Lines, Roorkee, U.P., India
Borgelt, E. H., 742 Greenlawn Ave., Dayton 3, Ohio
Brooks, J., 1222 D St., N.W., Washington, D. C.
Campbell, D. H., 2216½ Ewing St., Los Angeles 26, Calif.
Chambers, A., 2 Main St., Passaic, N. J.
Close, E. F., 2153 W. LeMoyne, Chicago 22, Ill.
Codey, M. T., 119 E. 42 St., Covington, Ky.
Creighton, C. C., Sixth Radar Calibration Detachment, APO 942, c/o Postmaster, Seattle, Wash.
Cunetta, J. M., 314 Linden Blvd., Brooklyn 26, N. Y.
Dawson, A. W., 234 Charles St., Painted Post, N. Y.
Dolde, L. A., 9036 Cedros Ave., Van Nuys, Calif.
Elfving, A. L., Research Institute of National Defense, Stockholm 61, Sweden
Fairchild, F. E., 446 W. 38 St., New York 18, N. Y.
Fishman, S., 315 Chicago Ave., Dayton, Ohio
Fogg, G. E., Hotel Warner, Emporium, Pa.
Fonte, G., The Clough-Brengle Co., 6014 Broadway, Chicago 40, Ill.
Fujimoto, K. K., 4147 Drexel Blvd., Chicago 15, Ill.
Greathouse, D. E., 710 Poplar St., Waukegan, Ill.
Hagman, H. B., Commercial Engineering Department, Sylvania Electric Products, Inc., Emporium, Pa.
Herrman, R. F., 41 W. 82 St., New York 24, N. Y.
Holpuch, D. J., 3625 Arthur Ave., Brookfield, Ill.

(Continued on page 44A)

## MEMBERSHIP

*(Continued from page 43A)*

Horrigan, J. S., 245 Audubon Pk., Dayton 7, Ohio
Ierardi, G. S., John Wiley & Sons, 440 Fourth Ave., New York, N. Y.
Jackson, T. T., Department of Electrical Engineering, Pennsylvania State College, State College, Pa.
Johnson, H. B., 4426 Park Ave., Kansas City, Mo.
Kafka, N., 610 S. 23 Ave., Bellwood, Ill.
Kubitz, E. H., 5141 S. Mayfield Ave., Chicago 38, Ill.
LaGasse, S. G., 1068 Newby St., Glendale 1, Calif.
Larsson, N. H., Research Institute of National Defense, Stockholm, Sweden
LaRue, R. S., Physics Department, The University of Connecticut, Storrs, Conn.
Laudenslager, R. L., University of Connecticut, Storrs, Conn.
Lee, J. P., 1013 Bergenline Ave., Union City, N. J.
Lovick, R. C., 303 Oaklawn Dr., Rochester 12, N. Y.
Maginniss, F. J., Central Station Engineering Division, General Electric Co., Schenectady, N. Y.
Mahren, A. A., 5602—187 St., Flushing, N. Y.
Marshall, K. C., 151 Strand Lane, Radcliffe, Lancashire, England
Martin, G. E., 4655 Felton, San Diego 4, Calif.
McCoy, O. Z., 515 Aberdeen Ave., Dayton 9, Ohio
McKee, D. M., 218 Wabash Ave., Lansdowne, Pa.
McKinley, J. G., West Pennsylvania Power Co., Box 1736, Pittsburgh 30, Pa.
Morrison, G. B., 3749 N. Nottingham Ave., Chicago 34, Ill.
Murphy, E. J., 960 E. Tioga St., Philadelphia 34, Pa.
Nolen, J. F., 1909—37 St., N.W., Washington 7, D. C.
Pate, W. R., Cottage 658, Crooked Lake, Angola, Ind.
Peterson, A. E., Jr., 30-M Garden Terrace, North Arlington, N. J.
Regan, C. R., c/o Navigational Aids, Electronics Test, Patuxent River, Md.
Schlehlein, G. J., 3005 W. Capitol Dr., Milwaukee 10, Wis.
Schwalbe, M. L., Nichols Veterans Hospital, Louisville 2, Ky.
Sinnott, J., 4313 Ninth Ave., Brooklyn 32, N. Y.
Snyder, H., Box 753, Marietta College, Marietta, Ohio
Soltesz, E. S., 350 Gavin St., San Diego 2, Calif.
Spicer, W., 136 Hastings St., Brookville, Pa.
Stimmel, R. G., 725 Patterson Rd., Dayton 9, Ohio
Thatcher, R., 1728 S. Mansfield Ave., Los Angeles 35, Calif.
Tomberg, S., 97 Patterson Village Dr., Dayton 9, Ohio
Trutko, J. M., 833 Leonard St., Akron 2, Ohio
Votocek, J., 2511 S. Harding Ave., Chicago 23, Ill.
Wamboldt, H., 601 Meyer Ave., Dayton 3, Ohio
Wang, M. T., 2620 Garfield St., N.W., Washington 8, D. C.
Wellinghoff, J. C., 3725 N. Wilton Ave., Chicago 40, Ill.
Weiger, H. O., 4814 N. Lowell Ave., Chicago 30, Ill.
Wikland, T., Research Institute of National Defense, Stockholm, Sweden
Williams, E., 10 Vivian St., Bexley, Sydney, N.S.W., Australia
Wilson, W. D., 54 Wendell Ave., Pittsfield, Mass.
Zito, A. J., 2220 Fish Ave., New York 67, N. Y.

**The following transfers to the Associate grade were approved to be effective as of March 1, 1949:**

Bayley, R. M., 6046 Roy St., Los Angeles 42, Calif.
Berger, M., 3733 Laurel Ave., Brooklyn 24, N. Y.
De Agazio, E., Jr., 31 Belmont Pk., Everett 49, Mass.
Geppert, D. V., E. B. 207, University of Arkansas, Fayetteville, Ark.

*(Continued on page 45A)*

## MEMBERSHIP

(Continued from page 44A)

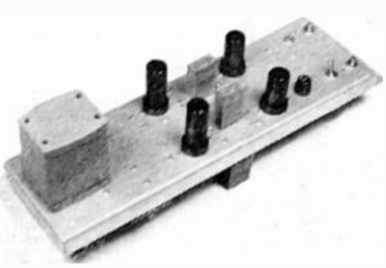
Giles, J. W., 40 Richlee Dr., R.F.D. 1, Camillus, N. Y.
Janislawski, A. G., c/o Tung-Sol Lamp Works, Inc., Dept. 170, Weatherly, Pa.
Knickel, E. R., Electronics Office, Naval Shipyard Boston, Mass.
Krueger, R. E., Box 1663, Los Alamos, N. M.
Laskin, H. J., 395 Belmont Ave., Brooklyn 7, N. Y.
Lawler, J. A., Jr., 268 Bowman Ave., Merion Station, Pa.
Miller, N. D., Box 95, Epworth, Iowa
Miller, P. G., 322 W. Cervantes St., Pensacola, Fla
Mitchell, A. V., Rm. 2100, 32 Sixth Ave., New York 13, N. Y.
Pomon, J. D., 4536 N. Magnolia, Chicago 40, Ill.
Popenoe, P., Jr., 2503 N. Marengo Ave., Altadena Calif.
Profumo, V. J., 336 Bidwell Rd., Mineola, L. I., N. Y.
Robinson, M. P., Jr., 30 Hawthorne St., Watertown 72, Mass.
Rosett, E. S., 35 Brookdale Ave., New Rochelle N. Y.
Russell, J. A., III, Winoka Village, Apt. 22D, Huntington Station, L. I., N. Y.
Shea, E. T., 1516 Darby Rd., Havertown, Pa.
Smit, J., Rua Capt. Salo Mao, S. Paula, Brazil
Stribling, J. L., Jr., 915 N. Jefferson, San Angelo Tex.
Walker, R. C., Box 414, Bucknell University, Lewisburg, Pa.

## News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**
(Continued from page 26A)

### Utility TV Amplifier

A new utility video amplifier, model 4TV17A1, for two-channel general-purpose use as a line or monitoring amplifier and as a line amplifier and supersync mixer, has been announced by **General Electric Co., Transmitter Div.,** Electronics Park, Syracuse, N. Y.



The 4TV17A1 is useful in raising remote programs as low as 1.2 volts up to the standard 2.0 volts for transmission. Either of the two channels can be used with a 75-ohm matched input, or with high impedance bridging the input for monitoring a line.

Low-frequency response on the unit will pass 25% negative 60-cps square wave with over-all tilt of 2% or less on a single channel, and 1% or less with both channels used in parallel. High-frequency response is flat within ±0.6 db to 7 megacycles.

Power input is 115 volts, 50 to 60 cps. The amplifier is designed for vertical rack mounting and is $5\frac{1}{4}$" high by 19" wide by 7" deep.

(Continued on page 46A)

# 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 45A)*

AS EXHIBITED AT THE
SPOTLIGHT THE NEW...
RADIO ENGINEERING SHOW

Beginning with this issue, the **News and New Products column** will describe as many as possible of the components and completed units which were labeled with "SPOTLIGHT THE NEW" posters at the 1949 IRE National Convention in New York City, on March 7–10.

One idea, which was used by two well-known manufacturers in this industry, was the immersion of equipment in large water-filled containers with gold fish swimming in them to demonstrate the quality and performance of these products in the most adverse conditions of humidity.

The **Commercial Sales Div., DeJur-Amsco Corp.,** 45-01 Northern Blvd., Long Island City 1, L. I., N. Y., employed this idea to exhibit their sealed meters, models 112 and 120. Six meters, each housing a photoelectric cell, were submerged in the bowl of water. They were so arranged that, when a beam of light directed on them was broken, the arrow in each would fluctuate according to the change in light intensity. The fish would break these beams as they swam through the water, causing a constant oscillation of the meter arrows. This protracted submersion and continued performance indicated that these meters were sealed and water tight.

For further information on these meters and the complete line of **DeJur** precision instruments, meters, potentiometers, and rheostats in a range from 10 to 1,000,000 ohms, contact the Commercial Sales Div.

The other firm to make use of this novel type of demonstration was the **General Electric Co., Electronics Dept.,** Electronics Park, Syracuse, N. Y.

To check maximum humidity effect on a standard model record player, pick-up cartridge, and loudspeaker, GE engineers placed them in a tank of water with no harmful effects.

Later is was decided that this test could be used as a demonstration at the IRE Convention. Gold fish, and a small loudspeaker which acted as a microphone, were added. The small speaker picked up the sound of music being played and released it through an external speaker, showing that the complete link went through the water, and two speakers, as well as the cartridge, actually functioned while emersed.

The fish displayed no ill effects from this perpetual serenade, but avoided the immediate area of the speaker, probably due to the increase in vibration.

*(Continued on page 47A)*

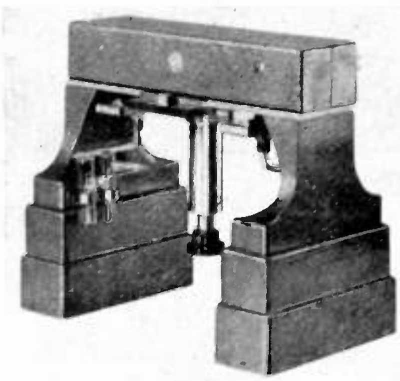
# News—New Products

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

*(Continued from page 46A)*

## A New Alpha Counter

A new methane-flow proportional counter, Model D45, has been designed to count alpha particles in the presence of high beta activity with 50% geometry and low coincidence losses, by **Nuclear Instrument & Chemical Corp.,** 223 West Erie St., Chicago 10, Ill.



When the D45 is used with a high-gain linear amplifier and a "fast" scaler, it is possible to count alpha particles in the presence of a beta activity of $5\times10^9$ disintegrations per minute.

The counter chamber has a polished cylindrical cathode. The anode is 0.002" tungsten wire mounted between glass insulators. The sample holder is designed for standard 1" diameter samples, with provisions for raising the sample up into the counting volume for improved geometry.

## Recent Catalogs

• • • A new catalog, No. 49, describing the complete line of resistors, controls, and resistance devices manufactured by **Clarostat Mfg. Co., Inc.,** Washington St., Dover, N. H., is ready for distribution.

• • • A new 4-page folder in color illustrates and describes various tube-testor models by **Hickok Electrical Instrument Co.,** 10541 Dupont, Cleveland 8, Ohio.

• • • A 4-page circular, entitled "Switches by Daven," with general information on switches used in broadcast, communications, and industrial fields, and in laboratory tests, is issued by the **Daven Co.,** 191 Central Ave., Newark, N. J.

• • • A new 4-page condensed catalog, which may be used as a brief guide for high-speed electronic counters, scalers, counter chronographs, and special electronic frequency-measuring computing equipment is offered by the **Potter Instrument Co., Inc.,** 136-56 Roosevelt Ave., Flushing, L. I., N. Y.

*(Continued on page 48A)*

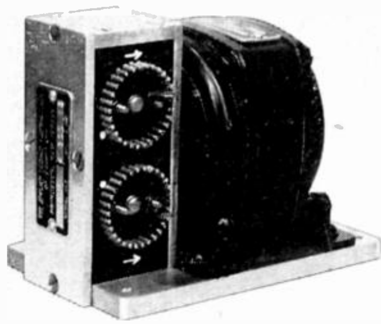
# News—New Products

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

*(Continued from page 47A)*

## Multichannel Switch

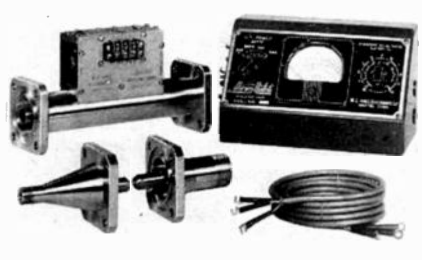
The development of a new multichannel switch, driven by a 60-cps synchronous motor, has been announced by the **Applied Science Corp. of Princeton,** Princeton, N. J. The use of a synchronous motor assures constant sampling rates and allows the convenient use of an ac line voltage source whose frequency is automatically synchronized with the sampling operation. The 4-pole model shown is 2″×3″×4″ in physical size and has 30 contacts per pole. The wiper is one of the shorting type for flexibility of use. Standard models have sampling rates of 720, 360, 240, 180, or 90 rpm.



These switches facilitate investigation of a large number of separate quantities or of a single quantity under a number of different conditions. In addition to telemetering applications, they may be used for the display of characteristic curves and multichannel voltage comparison. Electrical operations may be synchronized with mechanical motion by appropriate attachment of switch rotor.

## Rf Power and SWR Meter

An rf power and SWR meter, with a range of 50 to 500 Mc., has been designed and developed by **M. C. Jones Electronics Co.,** 96 N. Main St., Bristol, Conn.



The manufacturer claims that this is the first frequency-insensitive directional coupler to be produced by the industry.

This model, MM400, is one of a series of wide-frequency-range instruments for measuring rf power and standing-wave ratio on 51.5-ohm coaxial transmission line.

*(Continued on page 49A)*

# News—New Products

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

(Continued from page 48A)

These instruments are designed for laboratory measurements and for monitoring both transmitter and antenna performance.

A single meter reads incident power, reflected power, net power to load, and SWR of the load. A selector switch connects the meter to read any of these quantities.

The MM400 series "MicroMatch" may be used with 1⅝" or ⅞" air line, and with RG-17/U and RG-8/U coaxial cable. Full-scale power ranges of 400, 1,200, and 4,000 watts are available.

## Electrical Mismatch in TV Receiver Installations Diminished

An aid in the elimination of electrical mismatch existing between antenna and receiver in nearly all television receiver installations is made possible by a device described as "Telematch," according to its manufacturer, Standard **Transformer Corp.**, 3580 Elston Ave., Chicago 18, Ill.


Correction of line mismatch enables the full broadcast signal picked up by the antenna to be delivered, with minimum loss, to the receiver for improved reception. This often permits satisfactory reception with an inexpensive indoor antenna.

The "Telematch" has no tubes, and is installed by attaching two cable lugs to the receiver antenna input terminals.

Literature will be mailed on request.

## Recent Calalogs

• • • Engineering Bulletin 4, describing the complete line of audio attenuators, fixed pads, special attenuator networks, and other communications equipment components, will be sent by **Shallcross Mfg. Co.**, 520 Pusey Ave., Collingdale, Pa. to interested manufacturers on request.

(Continued on page 56A)

## POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

**The Institute reserves the right to refuse any announcement without giving a reason for the refusal.**

**PROCEEDINGS of the I.R.E.**
**1 East 79th St., New York 21, N.Y.**

### ENGINEERS

Location, Phoenix, Arizona. Excellent working conditions. Housing available. Motorola, Inc. announces a Research Laboratory devoted to armed service contract and company research in microwave, mobile communications, supervisory control, telemetering, miniaturization, and aviation electronics. Only fully qualified experienced inventors, engineers and scientists should apply. Send detailed statement of education and experience to Motorola Inc., D. E. Noble, 4545 Augusta Blvd., Chicago 51, Ill.

### SENIOR AND JUNIOR ENGINEERS

Senior and junior engineers needed with experience on SCR-548 radar or similar equipment. Location about 50 miles from Los Angeles. Electronic Engineering Co., 2008 W. 7th Street, Los Angeles 5, Calif.

### ELECTRONICS ENGINEER

Radio and industrial electronics in-instructor for two year technical college-Extension Division of Georgia Institute of Technology, Atlanta, Georgia. Write The Technical Institute, Chamblee, Georgia.

### ENGINEERS

Men needed immediately for permanent positions on an experimental, development, and production program of complex electronic and electro-mechanical equipment. Work covers computers, servos, amplifiers, instrumentation, small mechanisms in aircraft simulation for complex training equipment. Applicants must have college degree, equivalent experience or both. Apply Personnel Mgr. Link Aviation, Binghampton, N.Y.

### ELECTRO-ACOUSTIC ENGINEER

An attractive opportunity for an experienced engineer with thorough electro-acoustic training (transducers) in a small, sound rapidly growing concern. Permanent position with future opportunity as a research director. Write giving full details to Box 563.

### RESEARCH SCIENTISTS

Experienced research scientists with advanced degrees and experience in physics, aerodynamics, electronics, optics, mathematics, chemistry, metallurgy, or meteorology to perform supervisory research and act as permanent consulting group to engineering laboratories. Excellent opportunity for men with right qualifications. Salary $8,000-$14,000 bracket. Write Box 564.

### DEVELOPMENT ENGINEER

Newly formed and rapidly expanding aeronautical instrument company located in New York Metropolitan area is seeking an instrument development engineer

*(Continued on page 51A)*

## POSITIONS OPEN

*(Continued from page 50A)*

who is an electronics and servomechanism specialist to become Chief Engineer. Background must be well balanced in all phases of aeronautical instrument business, including laboratory and shop practice from first hand experience. Interesting salary and bonus. Box 565.

### LABORATORY TECHNICIAN

Excellent opportunity for experienced electronics laboratory technician having experience in aircraft instruments and allied fields. Location Metropolitan New York. Experience and ability in all phases of laboratory testing and shop work desired. Salary commensurate with ability. Box 566.

### COMMUNICATIONS ENGINEERS

Graduate communication engineers, Canadians citizens, interested in audio, radio and video frequency systems engineering. Salaries up to $350.00 per month depending on qualifications. Location, Montreal, Canada. Box 567.

### TELEVISION ENGINEERS

Due to wide expansion in television program of long established radio and television manufacturing company, positions are now available for engineers with 3 or more years television experience. Excellent working conditions in modern, well-equipped laboratory. Company located in northwest portion of New York State. Housing no serious problem. Salaries commensurate with ability. Box 568.

### SCIENTISTS AND ENGINEERS

Wanted for interesting and professionally challenging research and advanced development in the fields of microwaves, radar, gyroscopes, servomechanisms, instrumentation, computers and general electronics. Scientific or engineering degrees or extensive technical experience required. Salary commensurate with experience and ability. Direct inquiries to Mgr., Engineering Personnel, Bell Aircraft Corp. P.O. Box 1, Buffalo 5, N.Y.

### ENGINEERS

An unusual opportunity exists for qualified engineers to participate in the growth of the industrial television field. Experience in video, pulse, electronic display and circuit techniques with good educational background required. Openings range from Junior to Project Engineers. Write to: Leonard Mautner, Vice-President, Television Equipment Corp., 238 William St., New York 7, N.Y.

### RADIO ENGINEERS

Outstanding opportunity with progressive, rapidly growing company for radio engineers experienced in high frequency work. An interesting, attractive future assured, but requirements of jobs must be paralleled by necessary skill. If you have a background of high frequency work, tell us about yourself in résumé giving education, experience and salary requirements. Plant located in New Jersey within commuting distance of New York City. Box 569.

*(Continued on page 52A)*

## POSITIONS OPEN

*(Continued from page 51A)*

### DEVELOPMENT ENGINEER DESIGN ENGINEER

*Development Engineer* with at least 5 years experience in developing electronic equipment in the 100 to 1,000 megacycle range.

*Design Engineer* with at least 5 years experience with electro-mechanical design problems where space and difficult service conditions are important. Ideal laboratory conditions. 5 day week, paid vacation, paid medical and surgical policy, paid hospitalization policy and paid life insurance. Located in New Jersey. Box 570.

### DESIGN ENGINEER

Design Engineer on R.F. Transmission lines and fittings. Some experience also on microwave equipment. Box 571.

## Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

### RADIO PHYSICIST

B.Sc. May 1949, radio physics option of honor mathematics and physics, University of Western Ontario. 3 years radar in Canadian Navy. Age 27, married. Desires development, research or interesting position with good future. Box 243W.

### ELECTRONIC ENGINEER

B.S.E.E. June 1949, University of Cincinnati, age 30. Experience 2½ years co-op in development and test components, video and I.F. amplifiers, servo-mechanisms, instrument landing systems. Working knowledge of machine tools and machine shop practice. Desires position in research or development. Member of Eta Kappa Nu. Box 244W.

### ENGINEER

B.A.,Sc., E.E., April 1949, U.B.C.; age 27, married; appreciable experience as technician in F.M. and microwave radio relay, audio, mobile and harbour control F.M. with some design and development. Desires development or research engineering. Preference microwave. D. Nuttall, 2870 W. 13, Vancouver, B.C., Canada.

*(Continued on page 53A)*

# Positions Wanted

(Continued from page 52A)

### ENGINEER

Interested in television and its industrial application. Columbia University 1949, M.S. in I.E. City College 1948, B.S. in E.E. Engineer in training, State of New York. Prefers position in New York area. Box 245W.

### ELECTRONIC ENGINEER

B.E.E., February 1949, Cornell University. Age 24. Air Corps electronics officer, training at M.I.T., Harvard, Yale. Experience in electronics and power work. Interested in industrial electronics field. Box 246W.

### JUNIOR ENGINEER

Graduate R.C.A. Institutes; age 30; married. 3 years electrical engineering major in electronics, continuing studies for B.E.E. at night. Desires work as Junior Engineer in design and development of communication equipment under Senior Engineer. 4 years radio servicing experience preceding 3½ years U. S. Army radio servicing. 1st class Radiotelephone, 2nd class Radiotelegraph, and Class B amateur F.C.C. Licenses. Box 247W.

### ELECTRONIC ENGINEER

B.S. in E.E. 1948 University of Illinois. Experience with reputable firm in circuit design. Desires position anywhere in the U.S. Box 248 W.

(Continued on page 54A)

# Positions Wanted

*(Continued from page 53A)*

### ELECTRONIC ENGINEER

B.S.E.E. June 1949, Polytechnic Institute of Brooklyn. Single. 2 years experience as Navy electronic technician. Member Tau Beta Pi, Eta Kappa Nu. Desires position in electronic field; metropolitan area preferred. Box 265 W.

### FIELD SALES ENGINEER

B.S.E.E. communications major. 5 years experience in microwave, radar, and general communication equipment. Also some aircraft and electronic control experience. Finest references. Box 266 W.

### COMMUNICATIONS ENGINEER

B.S.E.E. June 1949, University of Minnesota, communications major. Age 25. Married, no children. 2 years experience GCA and other radar, 2 years airline radio and telephone work. Desires position with future in TV or electronics. Box 267 W.

### ELECTRONIC ENGINEER

Development or research in electronic control, instruments, or power preferred by ambitious veteran. Age 23. Married. M.S.E.E. June 1949. Tau Beta Pi, Sigma Pi Sigma member. 2 years varied industrial experience. Private pilot's license. Box 268 W.

### TELEVISION ENGINEER

Graduated American Television Institute of Technology with B.S.T.E. Age 22. Single. Desires position in design and development work in vicinity of Boston, Mass. Box 269 W.

### ENGINEER

June 1949 honor graduate of leading midwest engineering college B.S.E.E. Harvard-M.I.T. radar course. 3 years experience as research and development officer. MOS 7050. Experience with electronic bombing computers and telephone subs. equipment. Tau Beta Pi, Eta Kappa Nu—President, Student AIEE. Prefer connection with quality conscious firm in southwest U.S.A. Box 270 W.

### ELECTRONIC ENGINEER

Ex-flying officer. Industrial engineering and electronic engineering background. Both research and development and manufacturing experience on airborne radar, television and guidance systems. Schools—NCE, University of Dayton, Harvard, Columbia and M.I.T. Box 271 W.

### TELEVISION ENGINEER

Graduated American Television Institute of Technology January 1949 with B.S.T.E. Age 23. 1st Class FCC license. 2 years Army radio operator mechanic. Desires position as development or TV station engineer. Thorough understanding of RCA and Dumont TV equipment. Box 272 W.

### ADMINISTRATIVE OR EXECUTIVE

Retired Naval officer. M.S. Radio Engineering; Harvard 1928. Varied experience in charge reorganizing and operating

*(Continued on page 55A)*

# Positions Wanted

(Continued from page 54A)

major units Naval Communication System, radio procurement, electronics planning, radio stations maintenance, etc. Desires position as administrator or executive in communications, radio, or electronics, anywhere in U. S. Available immediately. Box 273W.

### JUNIOR ENGINEER

B.S. Radio Engineer, Chicago Technical College; Member Sigma Phi Delta, March 1949. Age 23; single. Navy experience Electrical Maintenance Technician. Desires position with a good opportunity. Box 274W.

### ENGINEER

M.S.E.E., June 1949, California Institute of Technology; age 23; married. Two years Navy radio technician. Summer work in small motor design. Member Tau Beta Pi. Desires work in control or telemetering. Box 275W.

### ELECTRONIC ENGINEER

B.E.E., New York University 1948, electronics major. One year GE test experience. GE advanced course in engineering. Interested electronic control equipment, servo-mechanisms; computers. New York Metropolitan area. Box 276W.

### ENGINEER

B.E.E., June 1949. Cooper Union. Age 28. Married. Three years electrical testing and inspection experience including supervision. Three years Army Signal Corps as radio serviceman. Desires position with future. Box 277W.

### EXECUTIVE ASSISTANT

Business administration plus engineering training. M.B.A., University of Chicago; B.S.E.E., B.S. (math.), University of Michigan. Eta Kappa Nu. Single. Age 26. Signal Corps technician; 1 year research, production quality control experience. Desires job in medium size concern. Résumé upon request. Box 278W.

### ENGINEER

B.S.E.E., September 1949, University of Florida. Age 30. Married. Two years Navy radio-radar technician and school. One year Chief Engineer 250 watt radio station. Motor and power transformer experience. Interested sales or construction. Will travel anywhere. Box 279W.

### SALES ENGINEER OR EXECUTIVE ASSISTANT

Former assistant to Vice President of a world-wide export organization invites inquiries from firms having need for addition to sales engineering staff or assistant to top executive. Young, energetic, widely traveled with a solid background of research design and supervision in all phases of electrical and electronic equipment. Capable of handling sales, engineering, production and personnel. Box 281W.

(Continued on page 56A)

# Positions Wanted

*(Continued from page 55A)*

### JUNIOR ENGINEER—PHYSICIST

Five and one-half years of physics, electrical engineering and electronics in Fordham, University of Rochester, Harvard, M.I.T. Three years Naval electronics officer; B.S. in physics, June 1949. Age 25; married; one child. Desires employment in electronic engineering, sales or development, anywhere, preferably west coast. Box 280W.

### ELECTRONICS ENGINEER

Guided missile electronics engineer. Currently engaged in production engineering of guided missiles. Three years experience in development of guidance and control equipment. Four years radar development experience for Army. Graduate engineer with postgraduate school. Box 282W.

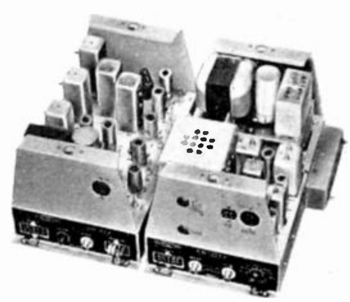
# News—New Products

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

*(Continued from page 49A)*

## Radiotelephone Equipment Operates on Minimum Power

Completion of the first production run of their new mobile radiotelephone equipment for the 152- to 162-Mc communication band, has been reported by **Kaar Engineering Co.**, Middlefield Rd., Palo Alto, Calif.



The manufacturer claims that this new equipment features lower standby battery drain than previously has been possible with equipment of this type. The transmitter consumes no power from the battery during standby periods; the receiver uses but 4 amps. In combination, the equipment consumes a total of just 4 amps during standby operation, yet permits full performance.

This feature means that radiotelephone systems can now be used in many installations from which they have heretofore been barred because of the bulky power supply required.

The Kaar equipment used is the FM-47X receiver and FM-177X 15-watt, or FM 179X 50-watt transmitter.

Literature describing the equipment in detail may be obtained by writing to the manufacturer.

*(Continued on page 57A)*

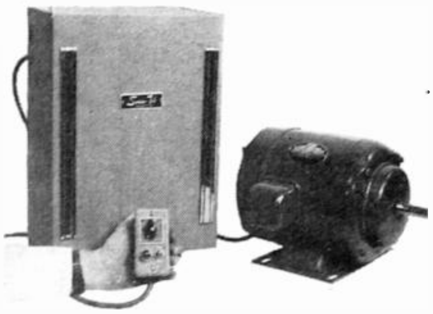
# News—New Products

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)

## Electronic Variable Speed Motor Control

A system for the stepless speed control of fractional hp motors has been developed by **Servo-Tek Products Co., Inc.** 4 Godwin Ave., Paterson 1, N. J.



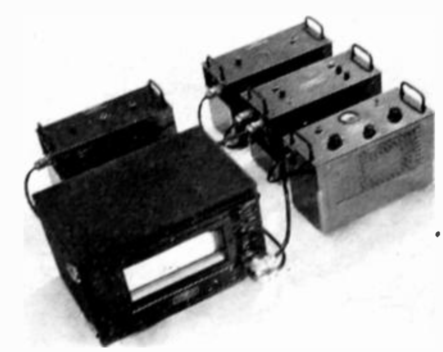
A simplified circuit is used which makes possible an economical electronic control system without sacrificing desirable operating features. Operating from a 110- or 220-volt 50–60 cps ac power source, the power unit utilizes full-wave gaseous rectification to supply the armature and the field of the ball bearing dc motor. The amount of voltage supplied to the armature is controlled by an electrical "brain" which reacts to change the applied voltage to correct for the slightest change in motor load, or by the operator.

The control unit, which may be remote, permits the use of any speed from 20 to 1,800 rpm standard.

Reversing control units are available, which incorporate features to achieve full speed reversals within 1 second.

## Rectangular Co-ordinate Recorder

A new instrument, the AIL Type 373 Rectangular Co-ordinate Recording System, which provides, in Cartesian co-ordinates, an inked plot voltage, or of the logarithm of voltage, as a function of the



displacement angle of a measure element, has been developed by **Airborne Instruments Laboratory, Inc.,** 160 Old Country Road, Mineola, L. I., N. Y.

(Continued on page 58A)

## News—New Products

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

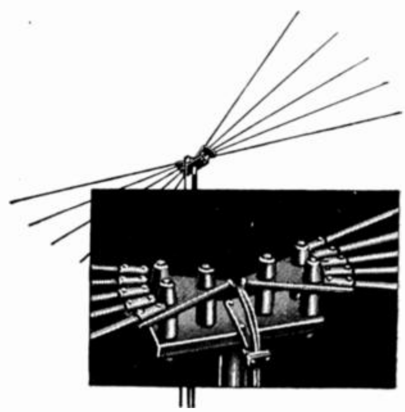
*(Continued from page 57A)*

Usable chart width is 10", corresponding to a voltage range of 10,000 to 1, or 80 db. Both pen and paper feed are servo-controlled, with chronograph paper feed optional. Maximum pen speed is 40" per second, equivalent to 320 db per second. At full scale expansion, the maximum paper feed rate is 10" per second.

The system finds application whenever it is desired to record voltage as a function of time or an angle. It can be used for recording light intensities, sound pressures, and heat levels at writing rates higher than formerly available. As originally designed to plot logarithmically the field-strength pattern of narrow-beam radar antennas, the system consists of a selective amplifier, pen and paper servo amplifiers, a power supply, and the recorder. The system operates from 115 volts, 60 cps power supply, and can be supplied in either portable carrying cases, or in standard rack and panel construction.

### Di-Fan Antenna for TV and FM

The new type 710 Di-Fan is a broadband receiving antenna tuned to receive all television and FM channels, according to the manufacturer, **Andrew Corp.**, 421 Seventh Ave., New York 1, N. Y.



The horizontal directivity pattern of the Di-Fan in TV channels 2 through 6 and in the FM band is a figure eight, broadside to the major axis of the antenna. In the high-frequency TV channels 7 through 13, the forward gain is decreased somewhat while the angle of acceptance is enlarged.

It is claimed that the antenna maintains the standard 300-ohm impedance over the TV spectrum; provides ghost-free reception from every TV station within range; and shows little mismatch loss over a wide range of frequencies. The fan elements are constructed of high-strength aluminum alloy with supports of plated steel.

For complete information contact J. F. White at Andrew.

*(Continued on page 59A)*

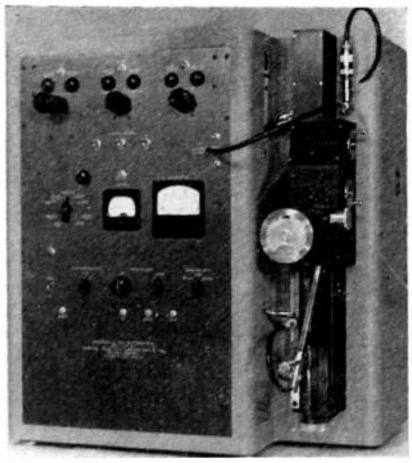
## News—New Products

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

*(Continued from page 58A)*

### A Microwave Dielectrometer

A microwave dielectrometer is now available for measuring the dielectric constant and loss of a wide variety of materials at nominal frequencies of 1,000, 3,000 and 9,000 Mc from **Central Research Laboratories, Inc.,** Red Wing, Minn.



The instrument consists of a slotted waveguide, precision traveling probe, modulated klystron oscillators, probe output amplifier, and associated power supplies and equipment.

The sample to be measured is inserted ahead of a short-circuiting plug and the effect of this arrangement on the standing-wave pattern in the guide provides data for calculating the dielectric constant and loss of the material.

At 1,000 and 3,000 Mc, the wave guide is used as a coaxial line operating in the *TEM* mode, and at 9,000 Mc either as a circular pipe operating in the $TE_{11}$ mode or as a coaxial line operating in the $TE_{10}$ mode. Solids are measured directly in the wave guide in the form of cylindrical samples 1" in diameter and about 1" to 2" long with a ⅜" hole coaxial with the outer surface except when the guide is operated as a circular pipe at 9000 Mc.

### Recent Catalogs

• • • Publication of the new Stancor catalog has been announced by **Standard Transformer Corp.,** 3580 Elston Ave., Chicago 18, Ill. This information will cover electrical and physical specifications, including list prices, of more than 400 items. Listings of transmitting and rectifier tubes, driver-modulator combinations, matched power supplies, output transformer-tube combinations, and detailed dimensional drawings of all Stancor transformer mounting styles are included. Requests for copies should be directed to G. C. Knoblock, Advertising Manager.

*(Continued on page 60A)*

# News—New Products

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

*(Continued from page 59A)*

## Above-Chassis-Mounted Vertical Power Resistor

Described as the Standee, a new type vertical power resistor for above-chassis mounting, is announced by **Clarostat Mfg. Co., Inc.,** Washington St. Dover, N. H.


Basically, the Standee comprises a wire-winding on fibre-glass core, bent in hairpin form with mica separator between the legs, placed in a ceramic tube filled with cold-setting inorganic cement and provided with bottom terminals and mounting bracket. The lugs are locked into the tube wall, in addition to being sealed in cement. The Standee can be mounted above the chassis with a large hole to clear the terminals, and a small hole to take a self-tapping screw or rivet for the mounting bracket. By having this power resistor mounted above the chassis, the problem of heat dissipation is avoided.

Standees are available in the standard 19/32" diameter, but in heights of 1½", 2", 2½", and 3", with respective power ratings of 10, 15, 20, and 25 watts. Maximum resistance values are 6,000, 9,000, 12,000 and 15,000 ohms, respectively. Standees are finished in the characteristic Greenohm green, and stamped with ohmage, brand name, and type number.

## NOTICE

Information for our News and New Products section is warmly welcomed. News releases should be addressed to **Industry Research Division, Proceedings of I.R.E., Room 707, 303 West 42nd St., New York 18, N.Y.** Photographs, and electrotypes if not over 2" wide, are helpful. Stories should pertain to products of interest specifically to radio engineers.

## INDEX AND DISPLAY ADVERTISERS



### DISPLAY ADVERTISERS

# PROFESSIONAL CARDS

## a power house nothing in the field can touch

## the CORNELL-DUBILIER

# BLACK CUB

### BAKELITE MOULDED PAPER TUBULAR

A midget-sized gargantua . . . a powerhouse of a capacitor that's armored to take all the stress and strain, vibration, humidity and extremes of temperature that comes its way. If the paper tubulars you're now using lack any of these 8 strong points, you're being short-changed. Switch to C-D BLACK CUBS as other leading manufacturers have done!


1. **"LEADWELD" CONNECTIONS!** Sturdy welded joints between wire leads and foil of capacitor section. Permanent connections! No intermittents! No open circuit defects!

2. **BAKELITE CASE!** Capacitor element molded in high quality Bakelite provides maximum protection under most severe service conditions.

3. **"POLYKANE" IMPREGNATION!** Polymerized resin impregnant provides excellent electrical properties and cannot leak through case at any temperature.

4. **HIGH TEMPERATURE OPERATION!** Excellent electrical properties maintained after long service at temperatures up to 100° C.

5. **EXCELLENT MOISTURE SEAL!** Will withstand long storage and service under extremes of humidity with minimum effect on electrical performance or appearance of tubular.

6. **STURDY CONSTRUCTION!** Withstands extremes of handling, soldering temperature, vibration and shock without damage to case material, moisture seal, circuit connections or electrical performance.

7. **HIGH INSULATION RESISTANCE!** Resistance exceeds 10,000 megohms per unit at 25° C.

8. **ALUMINUM FOIL ELECTRODES!** Provides best electrical properties and maximum service life at high temperature.

**C-D CAPACITORS — BEST BY FIELD TEST**

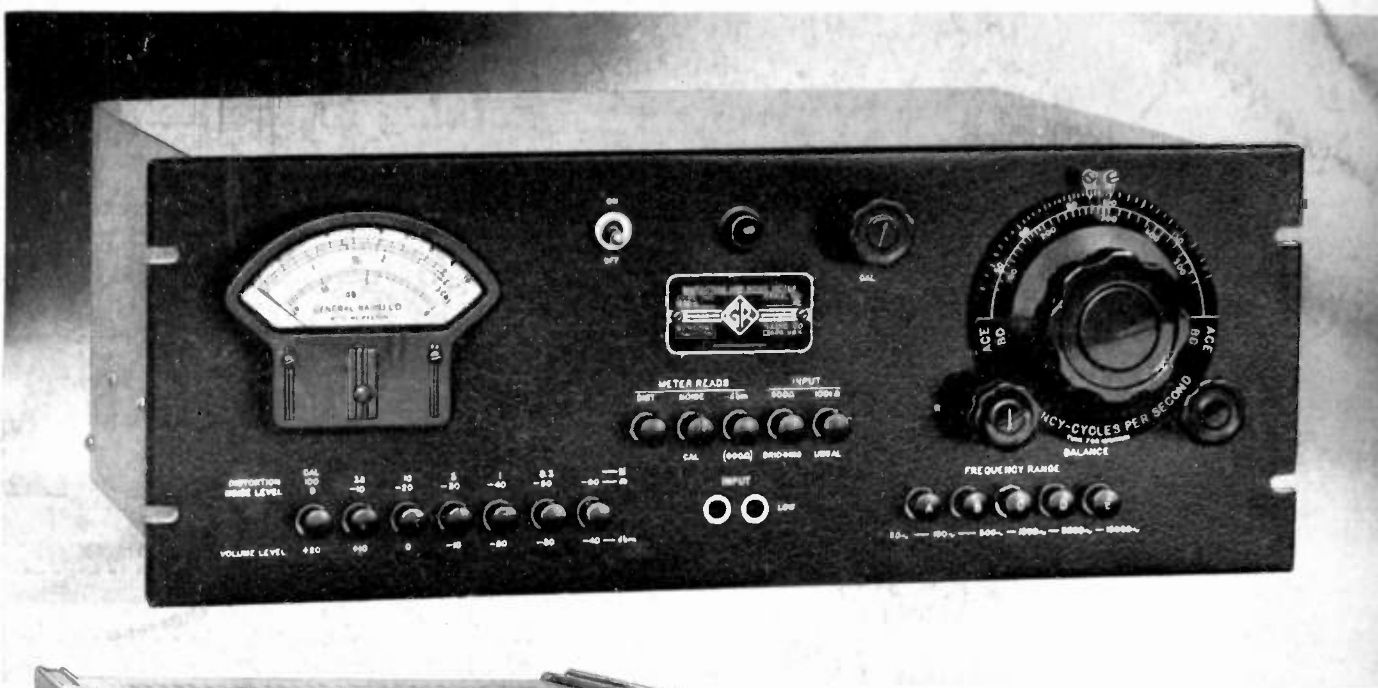
Write, wire or phone for samples and further information on the type MBT Black Cub Tubulars. Catalog on request. Cornell-Dubilier Electric Corporation, Dept. M69, South Plainfield, N. J. Other plants in New Bedford, Brookline and Worcester, Mass.; Providence, R. I.; Indianapolis, Ind.; and subsidiary, The Radiart Corporation, Cleveland, Ohio.

**CORNELL-DUBILIER**
**Consistently Dependable**

capacitors • vibrators
antennas • converters

1910 CORNELL-DUBILIER CAPACITORS 1949

Reg. U.S. Pat. Off.




Type 1301-A Low-Distortion Oscillator

## Here's Your "PROOF-OF-PERFORMANCE"

AS ANNOUNCED by the Federal Communications Commission,* effective August 1, 1949 all a-m and f-m broadcast stations will be required to make proof-of-performance checks of over-all noise and distortion of the complete station at least once a year.

Many stations already make these measurements at frequent intervals as routine operating maintenance to insure the continuous high-quality service the modern transmitter system is capable of supplying.

General Radio instruments for these measurements have been available for some time, and are in regular use by the leading stations where this equipment has given accurate, convenient-to-use and trouble-free service.

The G-R Type 1932-A Distortion and Noise Meter meets all of the F.C.C.'s requirements for measurements of this type for both a-m and f-m services; the Type 1301-A Low-Distortion Oscillator is the ideal companion unit for use with the Type 1932-A. Both of these instruments are relay-rack mounted and can be supplied in panel finishes to match most existing installations.

#### TYPE 1932-A DISTORTION & NOISE METER

For measurements of sine-wave voltages, distortion and noise throughout the audio range. Over-all pass-band of the voltmeter circuit extends to 45,000 cycles, thus including all noise and distortion products contained in this range; particularly the 3rd harmonic of a 15,000-cycle test is included.

This instrument is continuously adjustable and can be set to any frequency quickly since it has only one main tuning control plus a small trimmer. With it measurements can be made on a-f distortion in radio transmitters, line amplifiers, speech amplifiers, speech input equipment to lines; noise and hum levels of a-f amplifiers, wire lines to the transmitter, remote pick-up lines and other station equipment.

Full-scale deflections on the large meter read distortions of 0.3, 1, 3, 10 or 30 per cent; range for carrier noise measurements extends to 80 db below 100% modulation, or 80 db below an a-f signal of zero dbm level. The a-f range is 50 to 15,000 cycles, fundamental, for distortion measurements and 30 to 45,000 cycles for noise and hum.

Type 1932-A Distortion and Noise Meter: **$575.00**

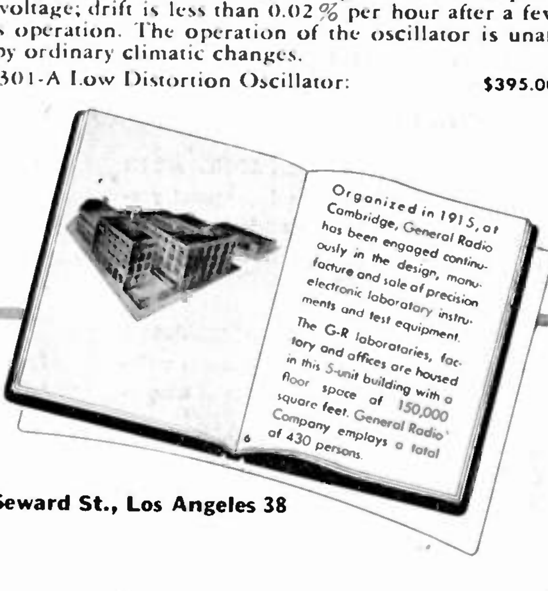
#### TYPE 1301-A LOW-DISTORTION OSCILLATOR

Especially designed for rapid measurements, this highly-stable oscillator has exceptionally low distortion. By means of push buttons, 27 fixed frequencies between 20 and 15,000 cycles may be selected in logarithmic steps. Any frequency between steps can be obtained by plugging in external resistors. The distortion over the entire range will not exceed the following percentages: with 5,000-ohm output, 0.1% from 40 to 7,500 cycles; 0.15% at other frequencies. With 600-ohm output 0.1% from 40 to 7,500 cycles; 0.25% from 20 to 40 cycles and 0.15% above 7,500 cycles.

The oscillator is calibrated to within ±(1½% + 0.1 cycle); the calibration is not affected by changes in load or plate supply voltage; drift is less than 0.02% per hour after a few minutes operation. The operation of the oscillator is unaffected by ordinary climatic changes.

Type 1301-A Low Distortion Oscillator: **$395.00**

*F.C.C. Rules and Regulations, Sections 3.254 and 3.46, as amended

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