Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

D STEL HAS OCT ISSUE

August, 1952

Volume 40 Number 8

Federal Telecommunication Labs

MICROWAVE TUBE

An all-metal traveling-wave amplifier tube, shown above, is intended as the output stage of a microwave transmitter. It provides more than 20-db gain over the frequency range of 5,900-7,100 mc and without tuning adjustment. Its output is 10 watts.

PROCEEDINGS OF THE I.R.E.

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

The IRE Standards on Radiation Counter Tubes, Definitions of Terms and Methods of Testing, appear in this issue.

The Institute of Radio Engineers

for *Stock* Hermetically Sealed Components

For over fifteen years UTC has been the largest supplier of transformer components for military applications, to customer specifications. Listed below are a number of types, to latest military specifications, which are now catalogued as UTC stock items.

RCOF CASE

Length 1 25/64
Width 61/64
Height 1 13/32
Mounting 1 1/8
Screws .... 4-40 FIL.
Cutout .... 7/8 Dia.
Unit Weight .... 1.5 oz.

MINIATURE AUDIO UNITS...RCOF CASE

Type No.	Application	MIL Type	Pri. Imp. Ohms	Sec. Imp. Ohms	DC in Pri., MA	Response ± 2db. (Cyc.)	Max. level dbm	List Price
H-1	Mike, pickup, line to grid	TF1A10YY	50,200 CT, 500 CT*	50,000	0	50-10,000	+ 5	$16.50
H-2	Mike to grid	TF1A11YY	82	135,000	50	250-8,000	+21	16.00
H-3	Single plate to single grid	TF1A15YY	15,000	60,000	0	50-10,000	+ 6	13.50
H-4	Single plate to single grid, DC in Pri.	TF1A15YY	15,000	60,000	4	200-10,000	+14	13.50
H-5	Single plate to P.P. grids	TF1A15YY	15,000	95,000 CT	0	50-10,000	+ 5	15.50
H-6	Single plate to P.P. grids, DC in Pri.	TF1A15YY	15,000	95,000 split	4	200-10,000	+11	16.00
H-7	Single or P.P. plates to line	TF1A13YY	20,000 CT	150/600	4	200-10,000	+21	16.50
H-8	Mixing and matching	TF1A16YY	150/600	600 CT	0	50-10,000	+ 8	15.50
H-9	82/41:1 input to grid	TF1A10YY	150/600	1 meg.	0	200-3,000 (4db.)	+10	16.50
H-10	10:1 single plate to single grid	TF1A15YY	10,000	1 meg.	0	200-3,000 (4db.)	+10	15.00
H-11	Reactor	TF1A20YY	300 Henries-0 DC, 50 Henries-3 Ma. DC, 6,000 Ohms.					12.00

RC-50 CASE

Length 1 5/8
Width 1 5/8
Height 2 5/16
Mounting 1 5/16
Screws .... #6-32
Cutout .... 1 1/2 Dia.
Unit Weight .... 8 oz.

COMPACT AUDIO UNITS...RC-50 CASE

Type No.	Application	MIL Type	Pri. Imp. Ohms	Sec. Imp. Ohms	DC in Pri., MA	Response ± 2db. (Cyc.)	Max. level dbm	List Price
H-20	Single plate to 2 grids, can also be used for P.P. plates	TF1A15YY	15,000 split	80,000 split	0	30-20,000	+12	$20.00
H-21	Single plate to P.P. grids, DC in Pri.	TF1A15YY	15,000	80,000 split	8	100-20,000	+23	23.00
H-22	Single plate to multiple line	TF1A13YY	15,000	50/200, 125/500**	8	50-20,000	+23	21.00
H-23	P.P. plates to multiple line	TF1A13YY	30,000 split	50/200, 125/500**	8 BAL.	30-20,000	+19	20.00
H-24	Reactor	TF1A20YY	450 Hys.-0 DC, 250 Hys.-5 Ma. DC, 6000 ohms . . . 65 Hys.-10 Ma. DC, 1500 ohms.					15.00

SM CASE

Length 11/16
Width 1/2
Height 29/32
Screw .... 4-40 FIL.
Unit Weight .... 8 oz.

SUBMINIATURE AUDIO UNITS...SM CASE

Type No.	Application	MIL Type	Pri. Imp. Ohms	Sec. Imp. Ohms	DC in Pri., MA	Response ± 2db. (Cyc.)	Max. level dbm	List Price
H-30	Input to grid	TF1A10YY	50***	62,500	0	150-10,000	+13	$13.00
H-31	Single plate to single grid, 3:1	TF1A15YY	10,000	90,000	0	300-10,000	+13	13.00
H-32	Single plate to line	TF1A13YY	10,000****	200	3	300-10,000	+13	13.00
H-33	Single plate to low impedance	TF1A13YY	30,000	50	1	300-10,000	+15	13.00
H-34	Single plate to low impedance	TF1A13YY	100,000	60	.5	300-10,000	+ 6	13.00
H-35	Reactor	TF1A20YY	100 Henries-0 DC, 50 Henries-1 Ma. DC, 4,400 ohms.					11.00

The impedance ratings are listed in standard manner. Obviously, a transformer with a 15,000 ohm primary impedance can operate from a tube representing a source impedance of 7700 ohms, etc. In addition, transformers can be used for applications differing considerably from those shown, keeping in mind that impedance ratio is constant. Lower source impedance will improve response and level ratings . . . higher source impedance will reduce frequency range and level rating.

* 200 ohm termination can be used for 150 ohms or 250 ohms, 500 ohm termination can be used for 600 ohms.
** 200 ohm termination can be used for 150 ohms or 250 ohms, 125/500 ohm termination can be used for 150/600 ohms.
*** can be used with higher source impedances, with corresponding reduction in frequency range. With 200 ohm source, secondary impedance becomes 250,000 ohms . . . loaded response is −4 db. at 300 cycles.
**** can be used for 500 ohm load . . . 25,000 ohm primary impedance . . . 1.5 Ma. DC.

United Transformer Corp.

150 VARICK STREET • NEW YORK 13, N. Y.

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

PROCEEDINGS OF THE I.R.E. August, 1952, Vol. 40, No. 8. Published monthly by the Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927. Printed in U.S.A.

Table of Contents will be found following page 64A

Meetings with Exhibits

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

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

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

September 19-20, 1952
Cedar Rapids IRE Technical Conference Roosevelt Hotel, Cedar Rapids, Iowa.
Exhibits: Lauren K. Findley, Collins Radio Co., Cedar Rapids, Iowa.

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

October 1-November 1
Audio Fair Hotel New Yorker, New York, N.Y.

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

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

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

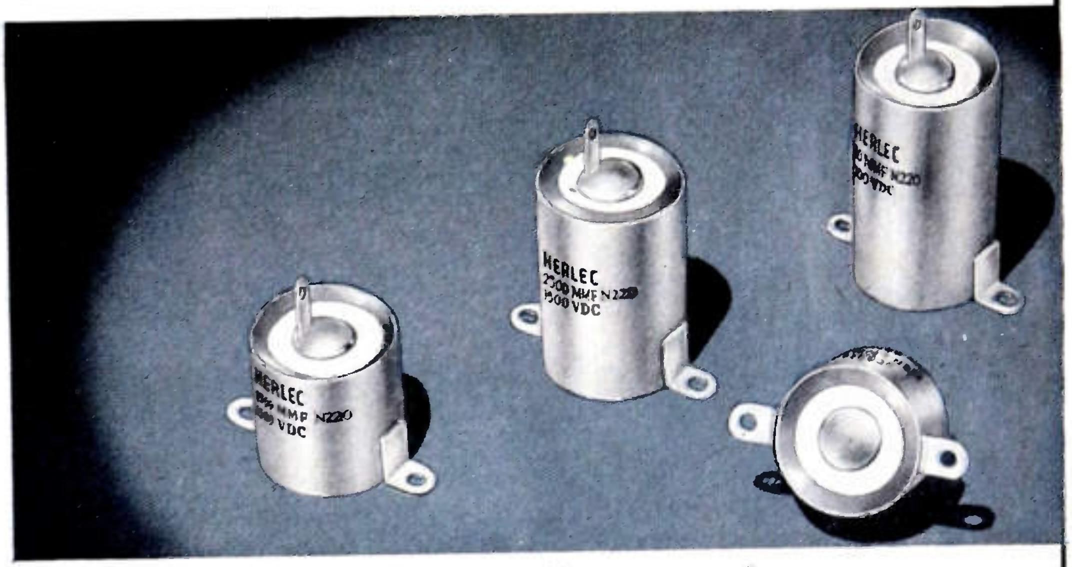
April 25, 1953
NEREM—New England Radio Engineering Meeting University of Connecticut, Storrs, Conn.
Accept Exhibits

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

9 standard terminal designs fit every mounting need

20,000-Volt Molded Ceramic Capacitors

Molded in moisture resistant, non-flammable thermosetting plastic, these new Sprague Type 700C Ceramic Capacitors offer exceptional reliability and economy as filters for TV receivers and C-R instrument high-voltage supplies. Standard capacitance is 500 mmf. and the units are conservatively rated for operation at 20,000 volts d-c. Write on letterhead for Engineering Bulletin 606.



Extended Capacitance Ranges for Precision Circuitry

These new Sprague-Herlec Precision Tubular Ceramic Capacitors make it possible to control the capacitance tolerance of exacting 500, 1000 and 1500 V. d-c precision circuits within ±1%. Temperature coefficient tolerances may be reduced to as little as ±10 parts in a million!

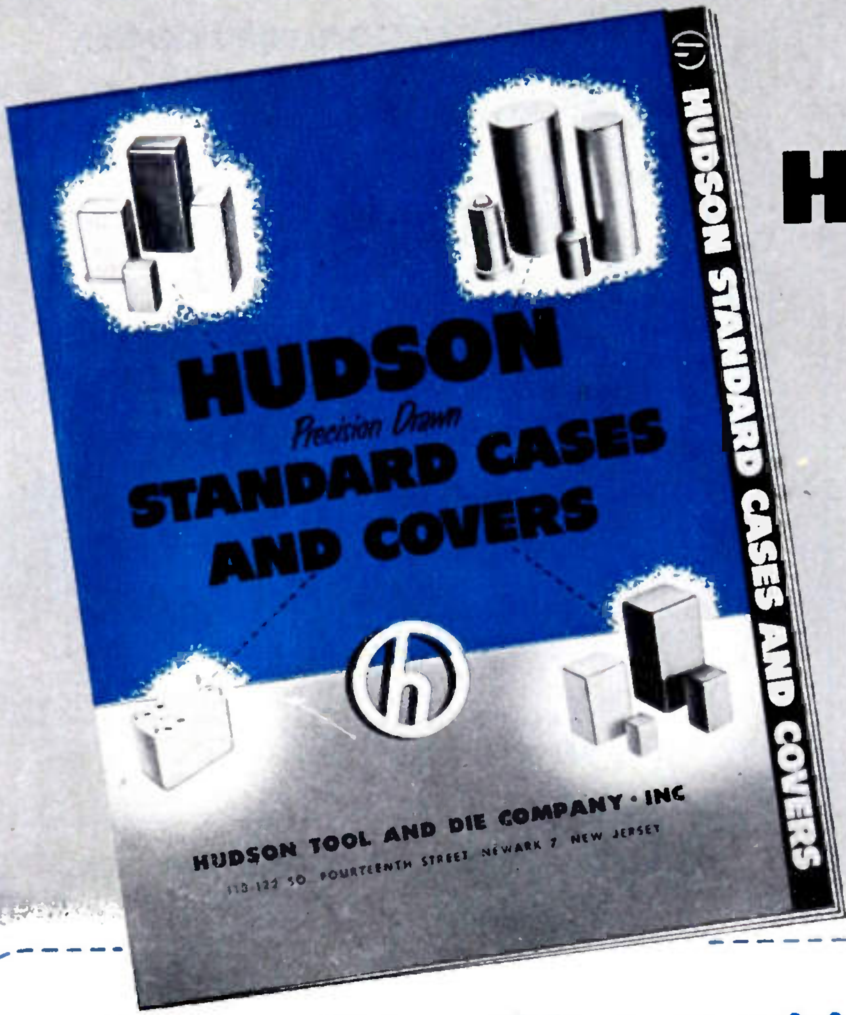
A logical development of the design first popularized in Sprague-Herlec cup ceramics, they greatly extend the capacitance range available to designers. "Q" and capacitance stability are high and the units have excellent retrace characteristics. Hermetically sealed in metal tubes, they operate over the range from −55°C. to +85°C. Bulletin 607 sent on letterhead request to Sprague Electric Company, 235 Marshall St., North Adams, Mass. or to the wholly owned Sprague subsidiary, The Herlec Corp., 422 N. 5th St., Milwaukee 3, Wis.

SPRAGUE®

WORLD'S LARGEST CAPACITOR MANUFACTURER

When SPECIFICATIONS *call for*

STANDARD CASES & COVERS



—consult the new

HUDSON DATA FILE!

—including

- STANDARD DRAWN CASES & COVERS
- SPECIAL METAL STAMPINGS
- SPECIAL COVER & CAN ASSEMBLIES

—the complete guide to precision metal parts

The new Hudson Handbook contains full engineering data and specifications on hundreds of standard cases and covers. All are craftsmen-made to provide a quick, economical solution to component problems. Keep a copy of the Handbook on file in your engineering and purchasing departments. Request yours now!

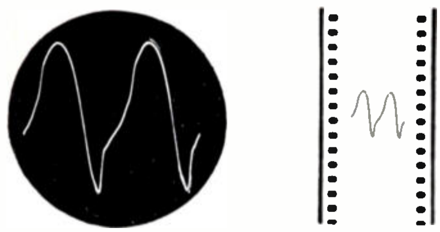
—plan your production around Hudson

All drawn cases and covers in the new Hudson Data File are standard and orders are filled promptly from regular factory stocks. Quality metal stampings are precision made to your exact specifications. For complete information on standard components, consult the Hudson Data File—for specification metal stampings, call or write Hudson for prompt service! Address inquiries to Desk 210.

HUDSON TOOL AND DIE COMPANY · INC

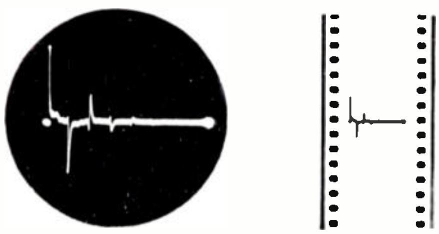
118-122 SO. FOURTEENTH STREET, NEWARK 7, NEW JERSEY

Want an oscilloscope camera NOW?



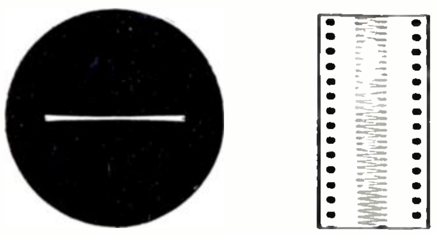
Scope Image — Film Recording

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



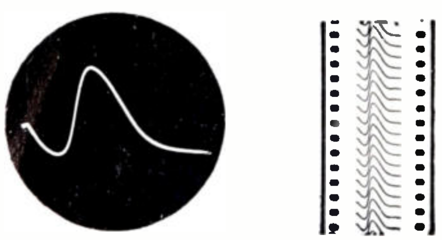
Scope Image — Film Recording

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



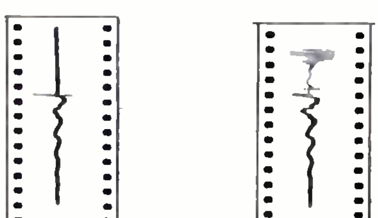
Scope Image — Film Recording

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



Scope Image — Film Recording

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



FILM MOTION TIME BASE — FILM MOTION AND SCOPE SWEEP

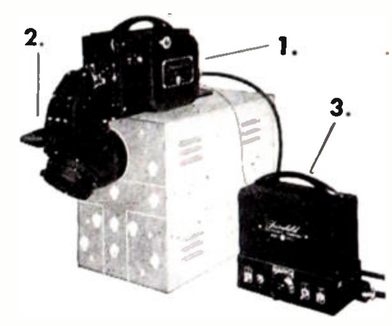
5. Continuous-motion photography employing combination of film motion and oscilloscope sweep as a time base.

Fairchild Oscillo-Record Cameras are now available *from stock* for immediate shipment. With these units you can make *permanent* photographic records of oscilloscope traces, thereby eliminating possible errors in making hand sketches from memory. In time-saving and convenience alone, these cameras will pay for themselves many times over.

FAIRCHILD OSCILLO-RECORD CAMERA IS UNUSUALLY VERSATILE

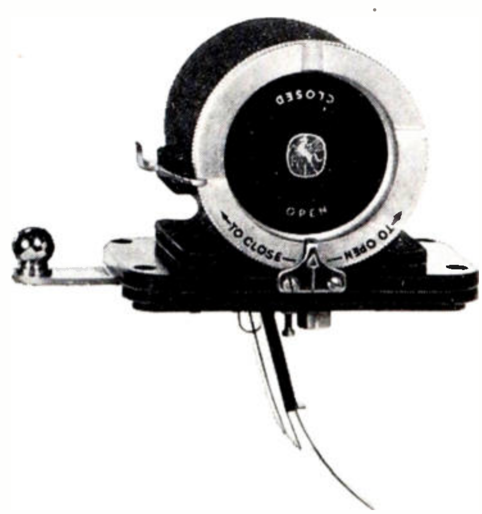
Users of the Fairchild Oscillo-Record Camera like its versatility. Designed for both still and continuous-motion photography on 35-mm film, it records non-recurring phenomena that are too rapid for visual study, others that are so slow that continuity is lost, and the occasions where very high-speed transients are combined with very slow-speed phenomena. For some idea of the types of jobs this instrument can do, study the examples at the left. Each solves a particular problem. Oscillo-Record camera users especially like its:

- **CONTINUOUSLY VARIABLE SPEED CONTROL** — 1 in/min. to 3600 in/min.
- **TOP OF SCOPE MOUNTING** that leaves controls easily accessible.
- **PROVISION FOR 3 FILM LENGTHS**—100, 400 or 1,000 feet.



1. Camera, 2. periscope, 3. electronic speed control. Accessories include 400- and 1,000-ft. film magazines, magazine adaptor and motor, universal mount for camera and periscope, binocular split-beam viewer.

FAIRCHILD TAKE-UP CASSETTE FOR SHORT RUNS



Where only a few pictures are required for quick development and study, a small Take-up Cassette is available as an accessory. The convenience afforded by this unit results in the saving of considerable time in handling short runs and reduces film wastage to a minimum. It is easily attached to the top of the camera by means of an adapter. A built-in knife permits short lengths of exposed film (up to 10 feet) to be cut off and removed with the cassette for developing.

Complete information about applications and operation of both the Fairchild Oscillo-Record Camera and the Fairchild-Polaroid Oscilloscope Camera is available. Write today to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, New York, Department 120-18C1.*

FAIRCHILD

OSCILLOSCOPE RECORDING CAMERAS

NEW EXPANDED LINE – OF E-I STANDARD terminals

48 STANDARD TYPES NOW AVAILABLE TO SIMPLIFY YOUR DESIGN PROBLEMS, SPEED DELIVERIES, REDUCE COSTS!

16 ADDITIONAL TYPES!

The increasingly popular E-I STANDARD LINE of SEALED TERMINALS now includes 16 additional types making a total of 48 items that can be ordered direct from stock with prompt delivery preassured. Our application engineers believe that this new expanded group of standard items could readily solve the majority of sealed terminal problems thereby eliminating much of the time and expense involved in custom design and production.

All 48 types are currently being specified in great numbers for an extremely wide range of applications, thus users of these types benefit by the additional economy of large scale production. For complete information covering all 48 types write today for Bulletin 949-A.

TYPE: RR-60W-PP

TYPE: CCS-80W-XP

TYPE: CCS-80W-HP

TYPE: C-80W-2/56P

TYPE: C-75T-SS

TYPE: B-60W-SS

TYPE: ABS-40W-XP

TYPE: AAA-30W-HS

TYPE: AAA-30W-HP

TYPE: AAA-30W-SX

TYPE: AA-40W-SP

TYPE: AB-60T-SX

TYPE: AB-60W-SS

TYPE: AB-60T-LX

TYPE: ABS-40W-HH

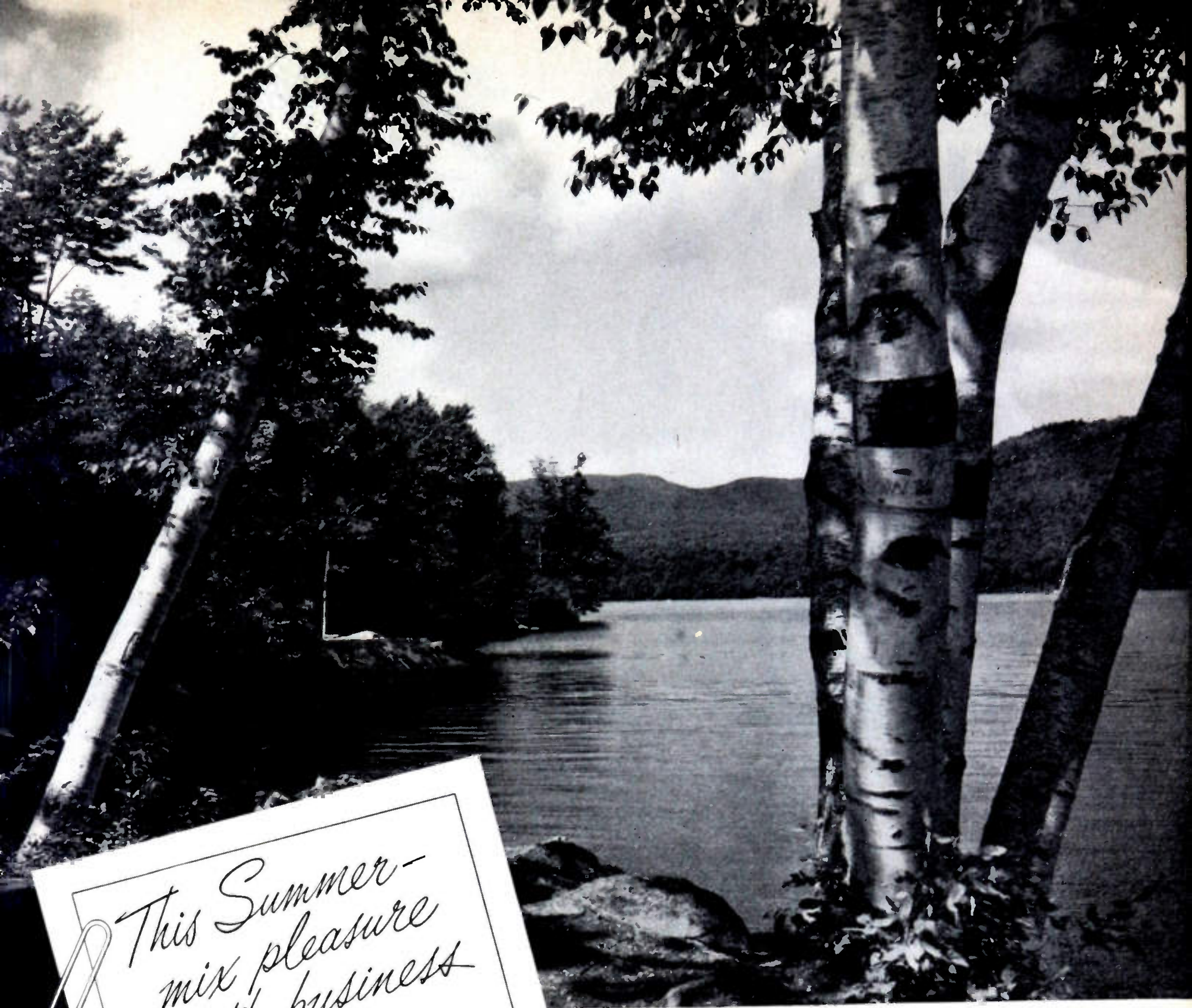
TYPE: ABS-40W-HP

E-I . . . YOUR HEADQUARTERS FOR HERMETICALLY-SEALED MULTIPLE HEADERS, OCTAL PLUG-INS, TERMINALS, COLOR CODED TERMINALS, END SEALS, etc. WRITE FOR CATALOGS.

ELECTRICAL INDUSTRIES • INC

MANUFACTURERS OF SPECIALIZED ELECTRONIC EQUIPMENT

44 SUMMER AVENUE, NEWARK 4, NEW JERSEY



This Summer— mix pleasure with business

This is only one of the many beauty spots that make New Hampshire an outstanding vacation land in America. Here you can find the answer to vacation yearnings . . . seashore or mountains . . . lakes, rivers or streams . . . hunting, fishing, golf . . . or just plain "resting", New Hampshire offers them all, together with modern conveniences.

Also in New Hampshire is the Marion Electrical Instrument Company — located in one of the historic Amoskeag Mill buildings in Manchester. Not far from Boston (Mass.), on your way to the White Mountains, Manchester offers excellent stop-over facilities. To customers and friends alike we say — this summer mix business with pleasure . . . vacation in New Hampshire and be sure to visit us at Marion and see how fine instruments are made.

marion meters

Reg. U. S. Pat. Off.

MANUFACTURERS OF RUGGEDIZED, HERMETICALLY SEALED AND STANDARD PANEL INSTRUMENTS

SEABORNE SERVICE

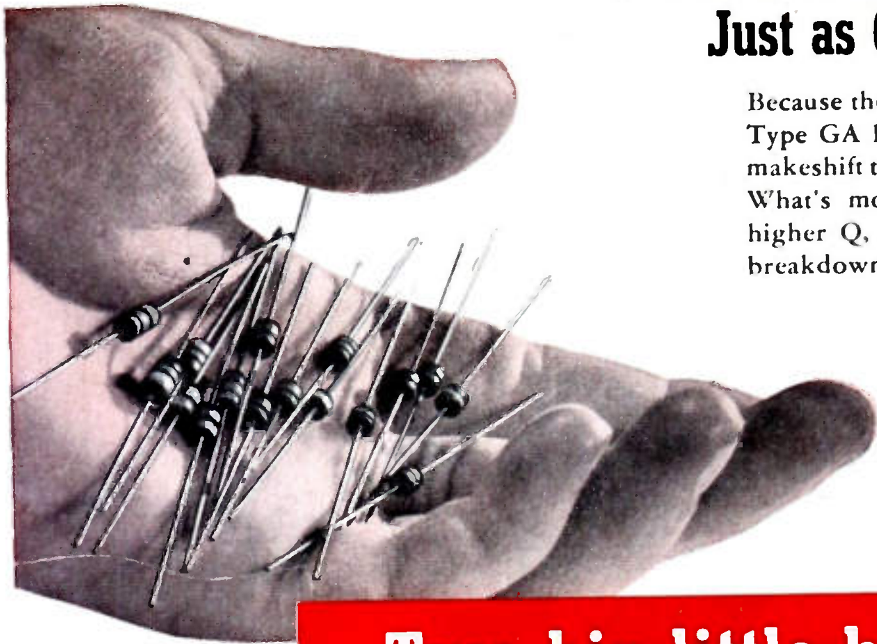
...Custom Designed for every Government and Military Application

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

COMMUNICATION PRODUCTS COMPANY • Inc

MARLBORO, NEW JERSEY
Telephone: FReehold 8-1880

DIPOLE ANTENNAS, SWITCHES, Q-MAX LACQUER AND CEMENT



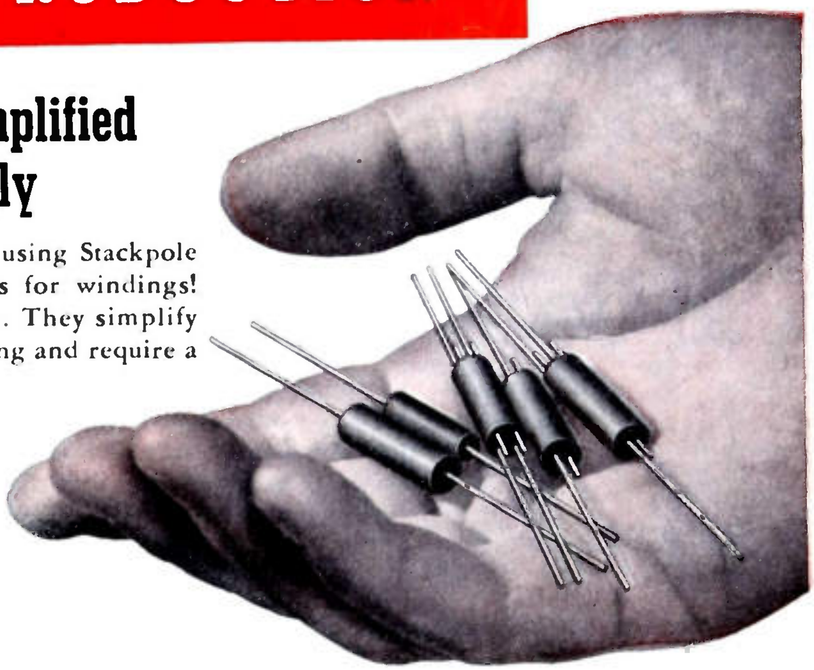
A lot better than "Gimmicks"... and Just as Cheap in the Long Run!

Because they're so much easier to install, Stackpole Type GA low-value capacitors cost no more than makeshift twisted-wire "gimmicks" in the long run. What's more, they offer much greater stability, higher Q, better insulation resistance and higher breakdown voltage. They are far superior mechanically and eliminate the inductive characteristic common to twisted wires.

Samples on letterhead request.

Two big little helps to BETTER DESIGN and PRODUCTION

For Smaller Coils... Simplified Equipment Assembly



Chances are you'll gain in several ways by using Stackpole Molded Coil Forms as mechanical supports for windings! They cost little. They permit smaller coils. They simplify equipment assembly with point-to-point wiring and require a lot fewer soldered connections. Forms are available with iron core sections that increase Q materially while decreasing the amount of wire needed for a given inductance. Stray magnetic fields are greatly reduced.

Electronic Components Division

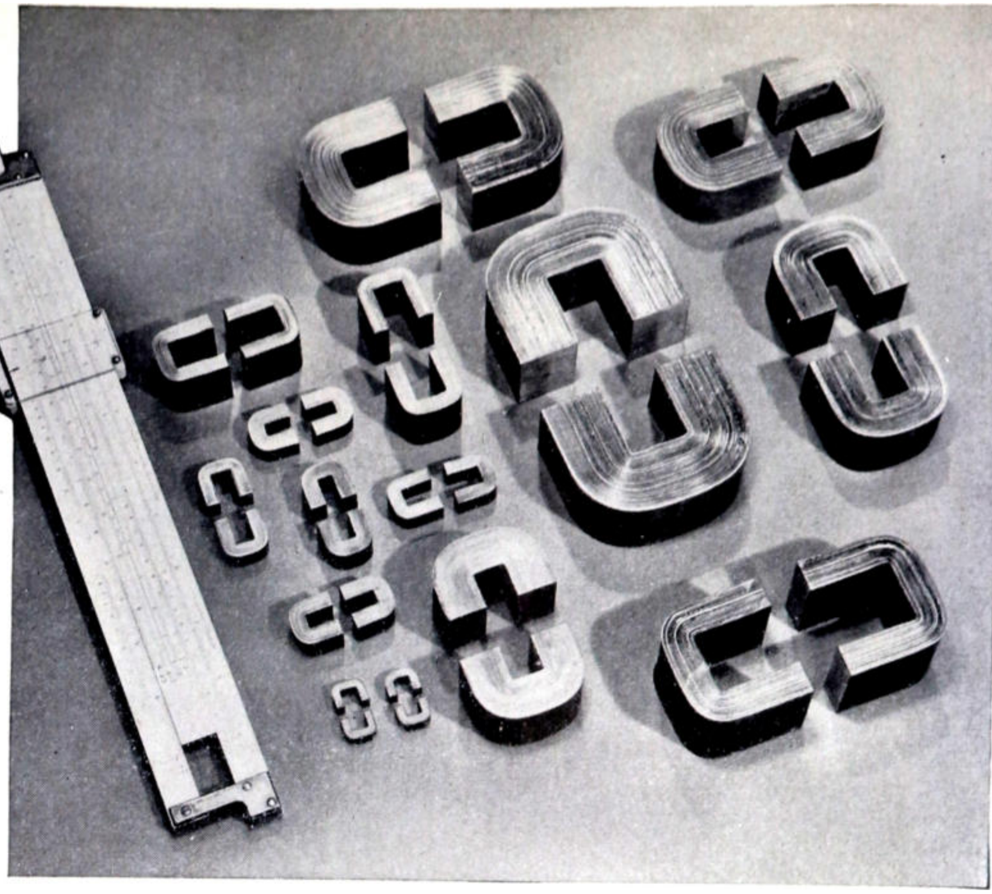
STACKPOLE CARBON COMPANY

St. Marys, Pa.

STACKPOLE

FIXED and VARIABLE RESISTORS • SPECIAL RESISTORS • LINE and SLIDE SWITCHES • IRON CORES • CERAMAG® CORES (FERRITES)

SILECTRON "C" CORES *for quick delivery* IN PRODUCTION QUANTITIES



. . . wound from strip as thin as 0.00025"

Quality-Tested and Proved

★ Arnold "C" Cores are made to highly exacting standards of quality and uniformity. Physical dimensions are held to close tolerances, and each core is tested as follows:

★ 29-gauge Silectron cut cores are tested for watt loss and excitation volt-amperes at 60 cycles, at a peak flux density of 15 kg.

★ 4-mil cores are tested for watt loss and excitation volt-amperes at 400 cycles, at a peak flux density of 15 kg.

★ 2-mil cores are tested for pulse permeability at 2 microseconds, 400 pulses per second, at a peak flux density of 10 kg.

★ 1-mil cores are tested for pulse permeability at 0.25 microseconds, 1000 pulses per second, at a peak flux density of 2500 gauss.

★ ½ and ¼-mil core tests by special arrangement with the customer.

Now available—"C" Cores made from Silectron (oriented silicon steel) thin-gauge strip to the highest standards of quality.

Arnold is now producing these cores in a full range of sizes wound from ¼, ½, 1, 2 and 4-mil strip, also 29-gauge strip, with the entire output scheduled for end use by the U. S. Government. The oriented silicon steel strip from which they are wound is made to a tolerance of plus nothing and minus mill tolerance, to assure designers and users of the lowest core losses and the highest quality in the respective gauges. Butt joints are accurately made to a high standard of precision, and careful processing of these joints eliminates short-circuiting of the laminations.

Cores with "RIBBED CONSTRUCTION"* can be supplied where desirable.

Ultra thin-gauge oriented silicon steel strip for Arnold "C" Cores is rolled in our own plant on our new micro-gauge 20-high Sendzimir cold-rolling mill. For the cores in current production, standard tests are conducted as noted in the box at left—and special electrical tests may be made to meet specific operating conditions.

● **We invite your inquiries.**

*Manufactured under license arrangements with Westinghouse Electric Corp.

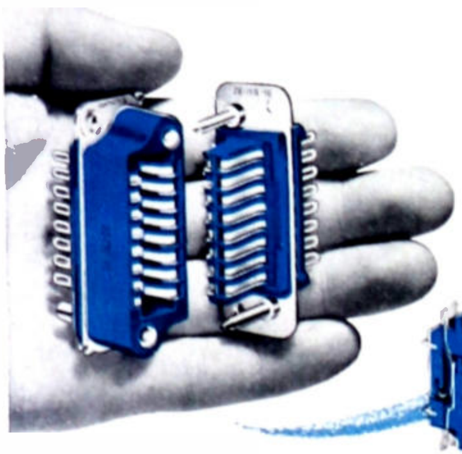
W&D 4211

THE ARNOLD ENGINEERING COMPANY

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

General Office & Plant: Marengo, Illinois

ANNOUNCING AMPHENOL BLUE RIBBON CONNECTORS



Low insertion and extraction force with high individual contact pressure

Unique spring contact construction maintains positive contact under vibration

Wiping action insures positive contact at all times

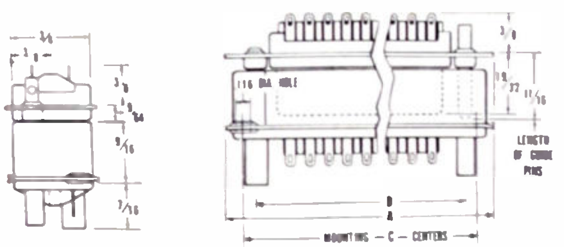
TYPICAL CROSS SECTION

VOLTAGE RATING 5 AMPS { 600 VOLTS D. C. AT SEA LEVEL
150 VOLTS D. C. AT 60,000 FEET ALTITUDE

RACK and PANEL TYPE

These new compact and lightweight connectors have been designed by Amphenol's Engineering Department to meet the demand for connectors that are easily mated even when out of sight. They provide quick disconnect, with low insertion and withdrawal requirements, for electronic sub-assemblies.

The rugged construction features high quality dielectric, silver base plated contacts with gold plated finish and stainless steel mounting plates. Plug contacts are supported their full length on the tough dielectric. The unique spring contact construction is self-cleaning and maintains full contact at *all* times! This same contact design makes it impossible to overstress or fatigue the spring members. The contact terminals are designed to accommodate up to No. 16 stranded conductors.



CONTACTS

26-159 / 26-190	16	24	32
"A"	2.437	3.118	3.798
"B"	1.842	2.522	3.202
"C"	2.024	2.704	3.384

These new Blue Ribbon Connectors are available in 8, 16, 24 and 32 contact sizes. Circuit switching or re-routing is easily done by proper wiring between contacts and plug-in member.

AMPHENOL 1-501 BLUE DIELECTRIC

This new dielectric, used in the Blue Ribbon Connectors, has been developed by Amphenol to meet the demand for a new and better dielectric. It easily meets the requirements of the Army-Navy Specifications and is far superior to melamine.

This diallyl phthalate resin-based compound combines nearly perfect dimensional stability with high insulation resistance, a lifetime shrinkage of less than 0.3% and an arc resistance exceeding 135 seconds on the standard ASTM test.

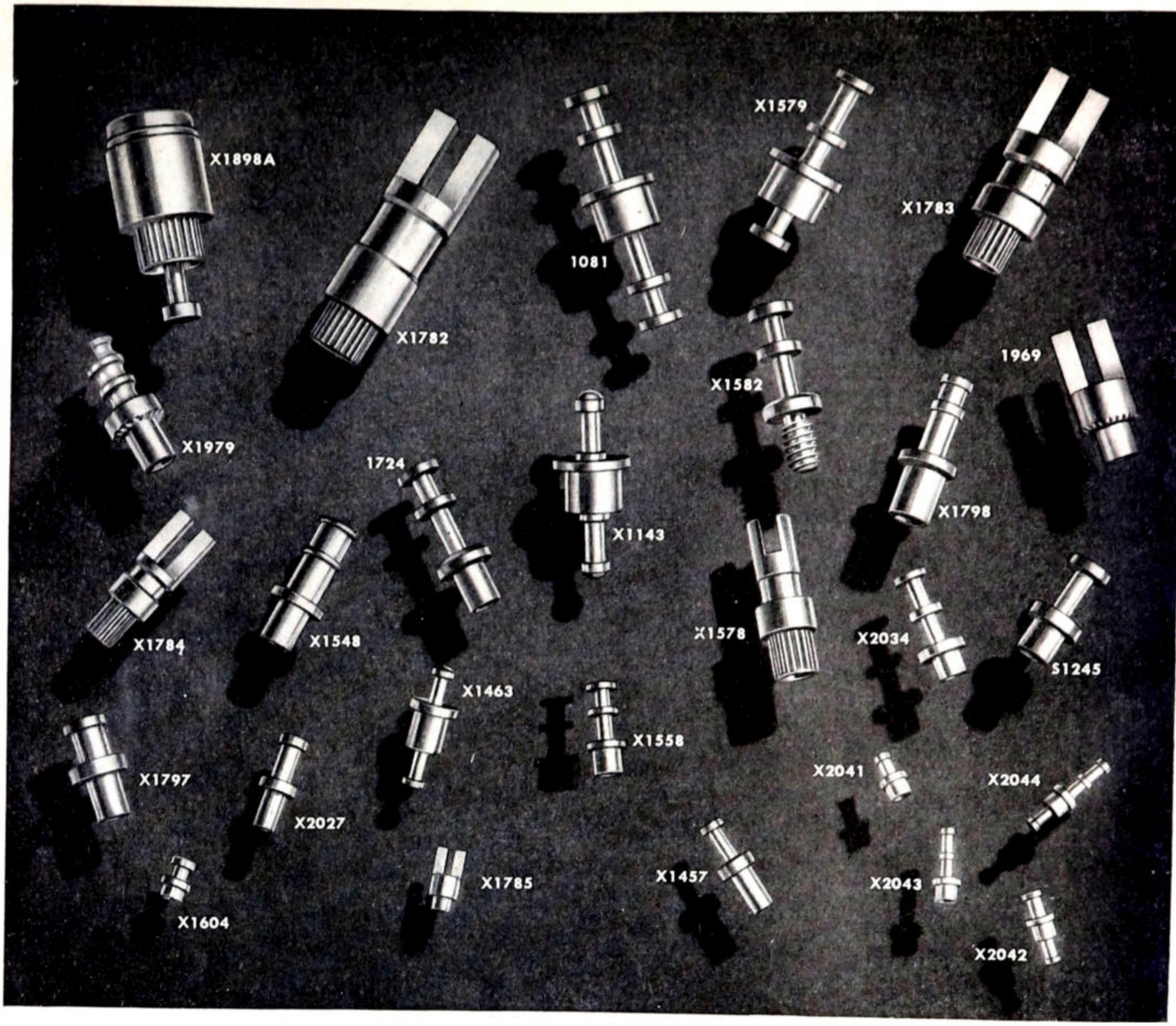
SPECIAL PURPOSE CONNECTORS

The Amphenol Blue Ribbon principle of low insertion and withdrawal force can be adapted to many special types and purposes. Pictured is a hermetically sealed plug with an adapted 16 contact receptacle. Special round configurations with and without keying shells are available. Mounting plates are available for special applications such as small complete circuit enclosures. The Amphenol Engineering Department offers consulting service in the designing of special purpose Blue Ribbon Connectors.

Write Dept. 13B for a copy of the four page Blue Ribbon Bulletin.

AMPHENOL

AMERICAN PHENOLIC CORPORATION
1830 SOUTH 54th AVENUE • CHICAGO 50, ILLINOIS



Shown Approximately Twice Size

Everything you need in standard terminal lugs
. . . or made to your own specifications!

C.T.C. has exactly the types and sizes of terminal lugs you want . . . or will quickly make them to your specifications in any production quantity. Very likely you'll find what you're looking for in the broad C.T.C. line of standard terminals. There are 28 different types, each available in varied shank lengths.

C.T.C. standard terminals are of silver plated brass, coated with water dip lacquer to keep them chemically clean for soldering.

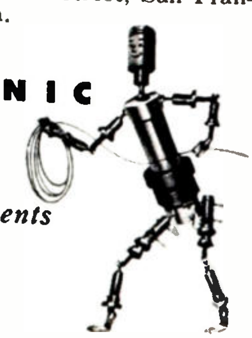
In addition, combination screw and solder terminals are available in 3 sizes, and a complete line of phenolic or ceramic terminals can be furnished.

All materials, processes and finishes meet applicable government specifications. Finishes include hot tinned, electro-tin, cadmium plate or gold plate on special order. In the event standard terminals don't meet your needs, C.T.C. offers a special consulting service to solve your solder terminal problems without extra cost or obligation.

For all specifications and prices, write to Cambridge Thermionic Corporation, 456 Concord Avenue, Cambridge 38, Mass. West Coast Manufacturers contact: E. V. Roberts, 5068 West Washington Blvd., Los Angeles 16 and 988 Market Street, San Francisco, California.

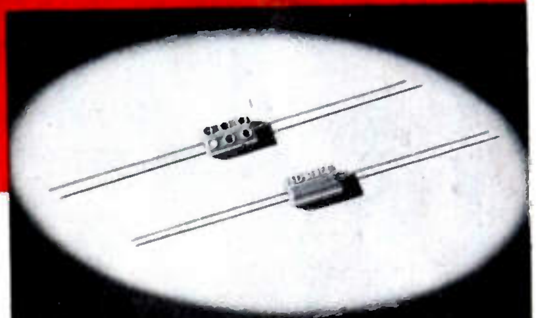
CAMBRIDGE THERMIONIC CORPORATION

custom or standard . . . the guaranteed components



Close DOESN'T COUNT

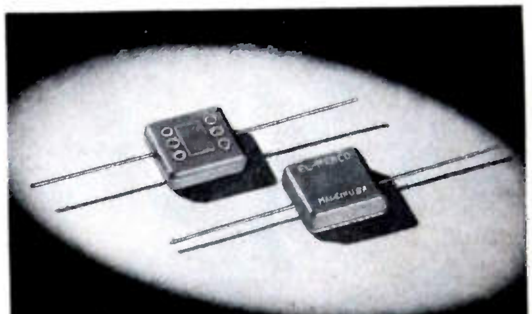
In the hushed white of the operating room, precision and dependability mean life to the quiet patient. **Almost** is the same as failure. In electronics the identical holds true . . . close just isn't good enough.



CM-15

This is why El-Menco Capacitators are designed for the ultimate in reliability and are built with razor-edge accuracy.

Lessons have been learned from surgery . . . today a doctor always allows a large margin of safety in standard operations. For long life and freedom from failure in your electronic applications every El-Menco Silvered-Mica Capacitator is factory-tested at more than **double** its working voltage.



CM-35

For peak performance in compact form . . . for higher capacity values, which require extreme temperature and time stabilization . . . there are no substitutes for El-Menco Capacitators. Available for every specified military capacity and voltage.

JOBBERS AND DISTRIBUTORS: For information write to Arco Electronics, Inc., 103 Lafayette St., New York, N. Y.—Sole Agent for Jobbers and Distributors in U. S. and Canada.

WRITE ON YOUR BUSINESS LETTERHEAD FOR CATALOG AND SAMPLES

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

CAPACITORS

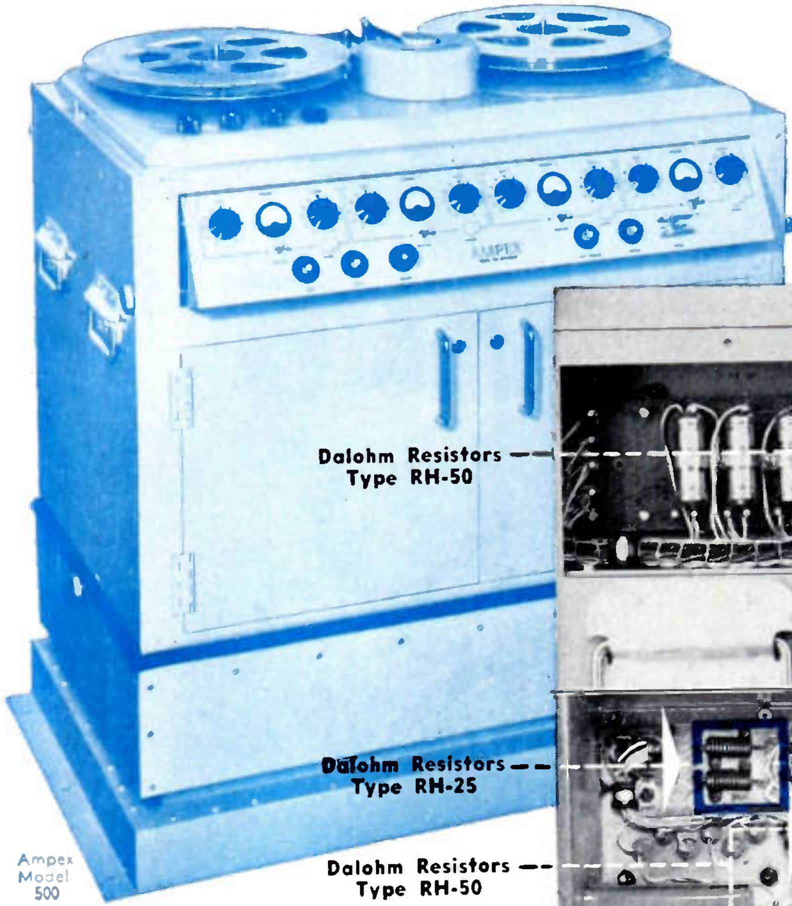
Radio and Television Manufacturers, Domestic and Foreign, Communicate Direct With Factory—

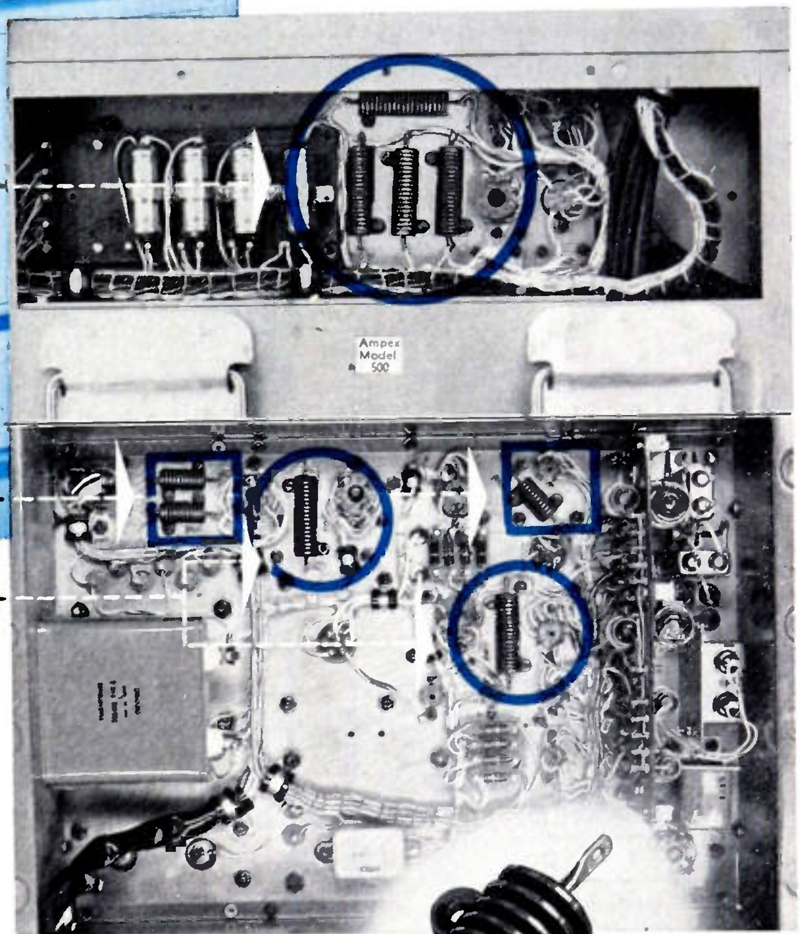
THE ELECTRO MOTIVE MFG. CO., INC. **WILLIMANTIC, CONNECTICUT**

DALOHM RESISTORS now standard

AMPEX

components!





Dalohm Resistors Type RH-50

Dalohm Resistors Type RH-25

Dalohm Resistors Type RH-50

AMPEX MODEL 375W

The Ampex Electric Corporation uses Dalohm Resistors in their Amplifiers to assure highest quality of reproduction and trouble-free performance. Ampex users find their equipment will operate 18 hours per day with but infrequent inspection emphasizing the superiority of the Ampex Magnetic Tape Recorder and the precise workmanship and dependable performance of the Dalohm 25 and 50 watt miniature power resistors. Manufacturers who seek the answer to that space problem find it in the Dalohm Resistor. It's real power in miniature!

Write today for full details and information on Dalohm 25 and 50 watt miniature power resistors.

"Also available in 2, 5, and 10 watt sizes."

"A complete line of the deposited carbon resistors is also available for prompt delivery."

25-Watt
Type RH-25
Resistance Range
.2 OHM to 15,500 OHMS
Tolerance
.05%, .1%, .25%,
.5%, 1%, 3%, & 5%
Temp. Coef. 0.00002/Deg. C

50-Watt
Type RH-50
Resistance Range
.5 OHM to
55,000 OHMS
Tolerance
.05%, .1%, .25%, .5%, 1%, 3%, & 5%
Temp. Coef. 0.00002/Deg. C

DALE

For those tight specifications

DALE PRODUCTS, Inc.

COLUMBUS, NEBRASKA U. S. A.

Bradleyunits are available in all standard R. T. M. A. values.

This HONEYCOMB RESISTOR CARTON is a Time- and Laborsaver in the Production Line

Differentially tempered leads

The leads of all Bradleyunits are differentially tempered. This graduated softness of leads near the body of the resistor prevents sharp bends and avoids damage to resistor.

SIZES OF UNITS

Rating	L	D
$\frac{1}{2}$-w	3/8″	9/64″
1-w	9/16″	7/32″
2-w	11/16″	5/16″

Give your assemblers the laborsaving advantage of Allen-Bradley honeycomb packaging. This unique container keeps resistor leads straight and free from tangling. It makes it easy to pick up a Bradleyunit from the patented Allen-Bradley carton, which holds the resistors in perfectly spaced rows. The removal of one or even fifty resistors does not affect the alignment of the remaining units.

Bradleyunits have permanent characteristics because they are rated to operate continuously at 70 C ambient temperature and not at 40 C. Therefore, they can withstand extremes of temperature and humidity. Bradleyunits need no wax impregnation to pass salt water immersion tests.

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

IMPORTANT NOTICE

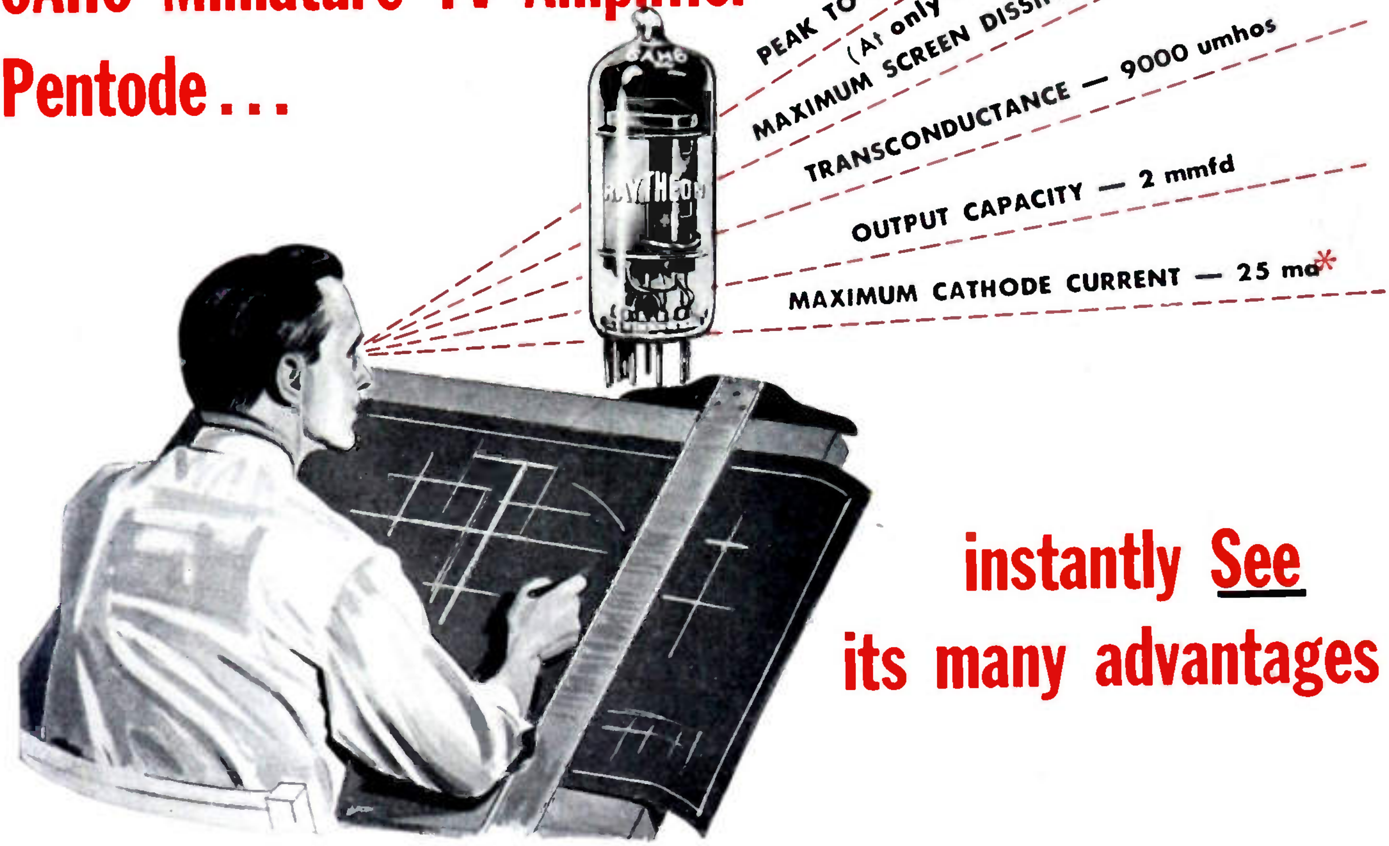
The tremendous demand for Bradleyunits has, in the past, resulted in disappointments due to extended deliveries. Our production facilities have been substantially increased, and your demands for Bradleyunits can now be satisfied quite promptly.

ALLEN-BRADLEY

FIXED & ADJUSTABLE RADIO RESISTORS

Sold exclusively to manufacturers QUALITY of radio and electronic equipment

Video Designers who **Look** into the RAYTHEON 6AH6 Miniature TV Amplifier Pentode...



instantly **See** its many advantages

As a result of extensive life tests and continued excellent field performance of the *6AH6*, cathode current and screen dissipation ratings are now increased. These new ratings are in line with the increased picture tube drive conditions required by trends to a larger and more brilliant picture.

What's more, despite these increased ratings the inherent low grid current level of the *6AH6*, achieved by carefully controlled manufacture, still permits the use of 1 megohm grid resistor in AC coupled video amplifiers.

* New higher rating

Input Coupling and Sync. Polarity	Output Volts P/P	Voltage Gain	Max. Watts Dissipation Screen	Max. Watts Dissipation Plate	Cathode Resistor Ohms	Cathode Current No Sig. (ma.)	Cathode Current With Sig. (ma.)	Grid Resistor Ohms
DC –	66	22	0.6	3.2	39	20	13	5000
DC +	100	25	0.4	3.2	270	8	15	5000
AC –	100	25	0.6	3.2	39	20	21	1 meg.
AC +	100	25	0.6	3.2	39	20	18	1 meg.

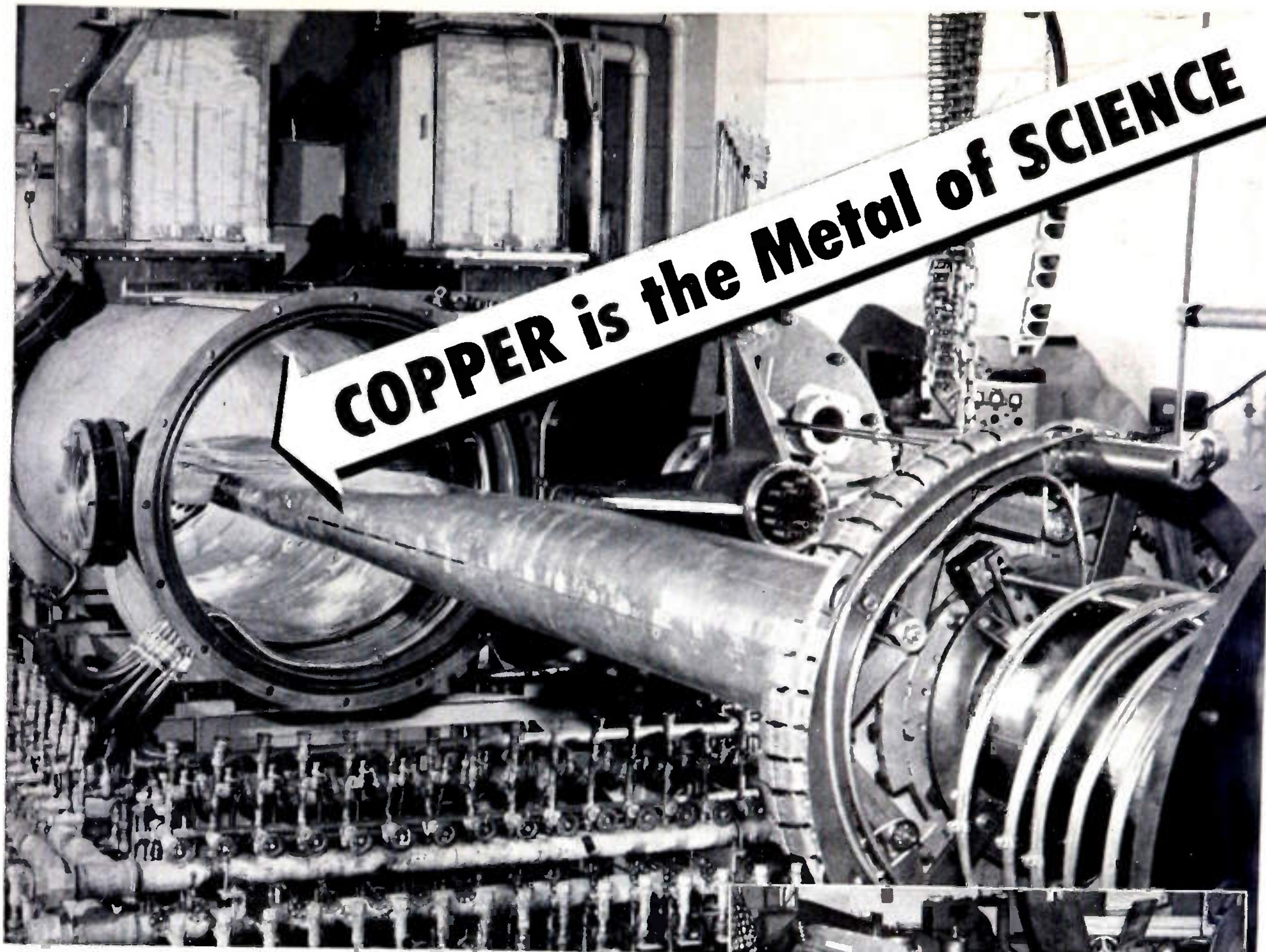
All data taken with Screen voltage of 150 and Plate load of 4000 ohms with typical on-the-air television signals and average production tubes.

RAYTHEON

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

Excellence in Electronics

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



D and D Stem being inserted in Cyclotron; made of Revere Electrolytic high-conductivity copper, hot rolled and annealed, 1/8" thick. Note also large number of bronze valves to control flow of cooling water through brass pipe.

● For many years Revere has been saying that "Copper is the metal of invention." It has high electrical and heat conductivity, excellent resistance to corrosion, is easily fabricated and formed, so that it is attractive to designers and inventors, as well as to manufacturers. Now we say it is also "The metal of science," because it is so essential to the operation of most scientific devices.

The pictures on this page illustrate some of its uses in a cyclotron, built by and for the Nuclear Physics Laboratory of the University of Washington in Seattle. The instrument was designed and constructed so far as possible by University personnel, who were completely successful in working copper into the most complicated shapes.

Revere collaborated on the project in various ways, and furnished copper bar, sheet, rod and tube to the University's high specifications. Remember that Revere will be glad to consult with you on your problems concerning copper and copper alloys, and aluminum alloys.

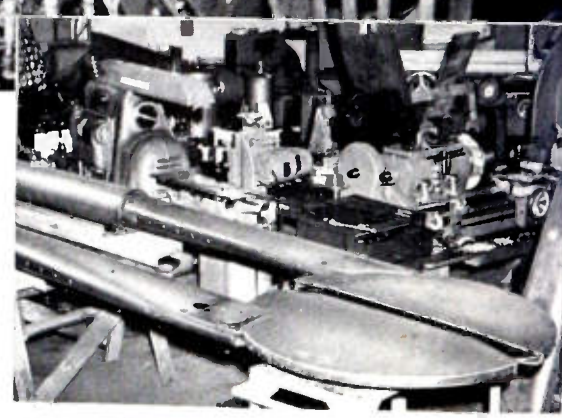


Photo taken in the University of Washington shop during fabrication of the two Ds and D Stems.

Seven miles of Revere copper bus bar were wound into great coils for the cyclotron electromagnet. The University built the winding machine itself, and wound the coils in its own shop. The special Revere bar is soft temper, free from scale, with rounded edges.

REVERE

COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

230 Park Avenue, New York 17, N. Y.

• • •

Mills: Baltimore, Md.; Chicago and Clinton, Ill.; Detroit, Mich.; Los Angeles and Riverside, Calif.; New Bedford, Mass.; Rome, N. Y.—Sales Offices in Principal Cities. Distributors Everywhere

SEE "MEET THE PRESS" ON NBC TELEVISION EVERY SUNDAY

News—New Products

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

Synchoscope

Browning Laboratories, Inc., 750 Main St., Winchester, Mass., announces Model P4-EX, a new Synchroscope. This instrument is designed for those applications requiring a triggered sweep. An internal trigger generator with continuous adjustment from 50–5,000 pps enables the scope to be used as a timing source. The triggered sweep is continuously variable and calibrated from 1.0 microsecond per inch to 25,000 microseconds per inch. Output triggers may be phased from 500 microseconds before the sweep to 500 microseconds following the start of the sweep.

A vertical amplifier with a flat response from 5 cps to 5 mc makes this unit useful for examination of various pulse waveforms. Used with a suitable crystal detector, it is possible to observe the modulation envelope of pulsed rf sources such as radar modulators. A direct connection to a vertical deflection plate is also provided. Both deflection arrangements make use of a panel-controlled calibration source of 0.3, 1, 3, 10, 30, and 100 volts.

The P4-EX uses a 5-inch cathode-ray tube and is housed in a compact steel cabinet 14½ inches high, 10 inches wide and 16¾ inches deep. Weight is 50 pounds.

RCA Training Film Available

A color motion picture entitled, "Your Gateway to New Opportunities," containing many inside shots of the radio-television industry, has recently been produced by RCA for use in their training program for new engineers. Because of its educational value and general appeal, this 20-minute film, together with projection equipment, is being made available to any interested group on request.

Requests should be made to Robert E. McQuiston, Manager, Specialized Employment Division, **Radio Corp. of America**, 30 Rockefeller Plaza, New York 30, N. Y.

(Continued on page 40A)

3 UNIVERTERS

. . . for extending the coverage of B.R.C. Signal Generators

UNIVERTER—Type 207-A

A frequency converter for use with FM-AM Signal Generators, 202-B and 202-C. Output frequency range of Signal Generators is 54 to 216 mc. Additional output when using 207-A:

Frequency Range: 0.1 to 55 mc.
Output: 0.1 to 100,000 microvolts at X1 jack, approximately 7.5 times these values at high output jack.
Frequency Increment Dial: ±300 kc in 5 kc increments.
Modulation: FM and AM controlled by Signal Generator.
Price: $345.00 fob Factory.

UNIVERTER—Type 207-B

A frequency converter for use with FM-AM Signal Generator 202-D. The 202-D is applicable to telemetering problems over frequency range of 175-250 mc. Additional output when using 207-B:

Frequency Range: 0.1 to 55 mc.
Output: 0.1 to 100,000 microvolts at X1 jack and approximately 7.5 times these values at high output jack.
Frequency Increment Dial: ±300 kc in 5 kc increments.
Modulation: FM and AM controlled by Signal Generator.
Price: $345.00 fob Factory.

UNIVERTER—Type 207-C

A frequency converter for use with FM Signal Generator 206-A. The 206-A is applicable to mobile communications problems over a frequency range of 146 to 176 mc. Additional output when using 207-C:

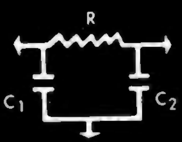
Frequency Range: 0.1 to 50 mc.
Output: 0.1 to 100,000 microvolts at X1 jack and approximately 7.5 times these values at high output jack.
Modulation: FM controlled by Signal Generator.
Price: $345.00 fob Factory.

Write for complete information

BOONTON RADIO Corporation
BOONTON, N. J.

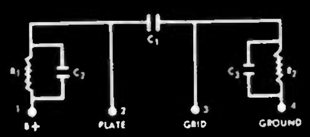
ERIE PRINTED CIRCUITS

DIODE FILTER



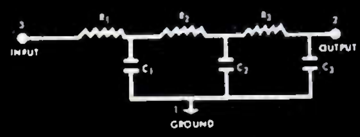
1403-01 1403-02
1403-03

TRIODE PLATE COUPLER



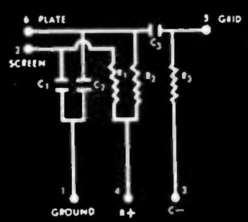
1404-01 1406-01
1404-02 1406-02

VERTICAL INTEGRATOR



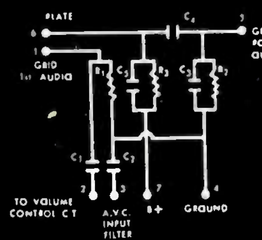
1405-01

PENTODE PLATE COUPLER



1407-01 1407-02
1407-03

AUDIO OUTPUT CIRCUIT



1408-01 1408-02

ERIE PRINTED CIRCUITS

offer these advantages:

- Fewer soldered connections mean less installation time.
- One installation unit replaces several.
- Fewer connections mean fewer wiring errors.
- Circuit stability is improved through simplification.
- Lower costs for procurement and stock maintenance.
- Other material costs are decreased by smaller size, lighter weight.

Save Space . . . Time . . . Cost and Improve Stability

Erie Resistor began the development of Printed Circuits in 1940. Since then the advantages of Printed Circuits have been amply demonstrated and Erie has made important contributions in the field.

By bonding the complete or partial circuit to a ceramic base plate, the work of several capacitors may be combined in one installation unit. Erie Printed Circuits have simplified design and production problems for manufacturers of radio and television receivers, hearing aids, and other electronic products, including various military equipment requiring sub-miniaturization. Such products may be reduced in size, weight, and cost, at the same time that they are made more reliable in service.

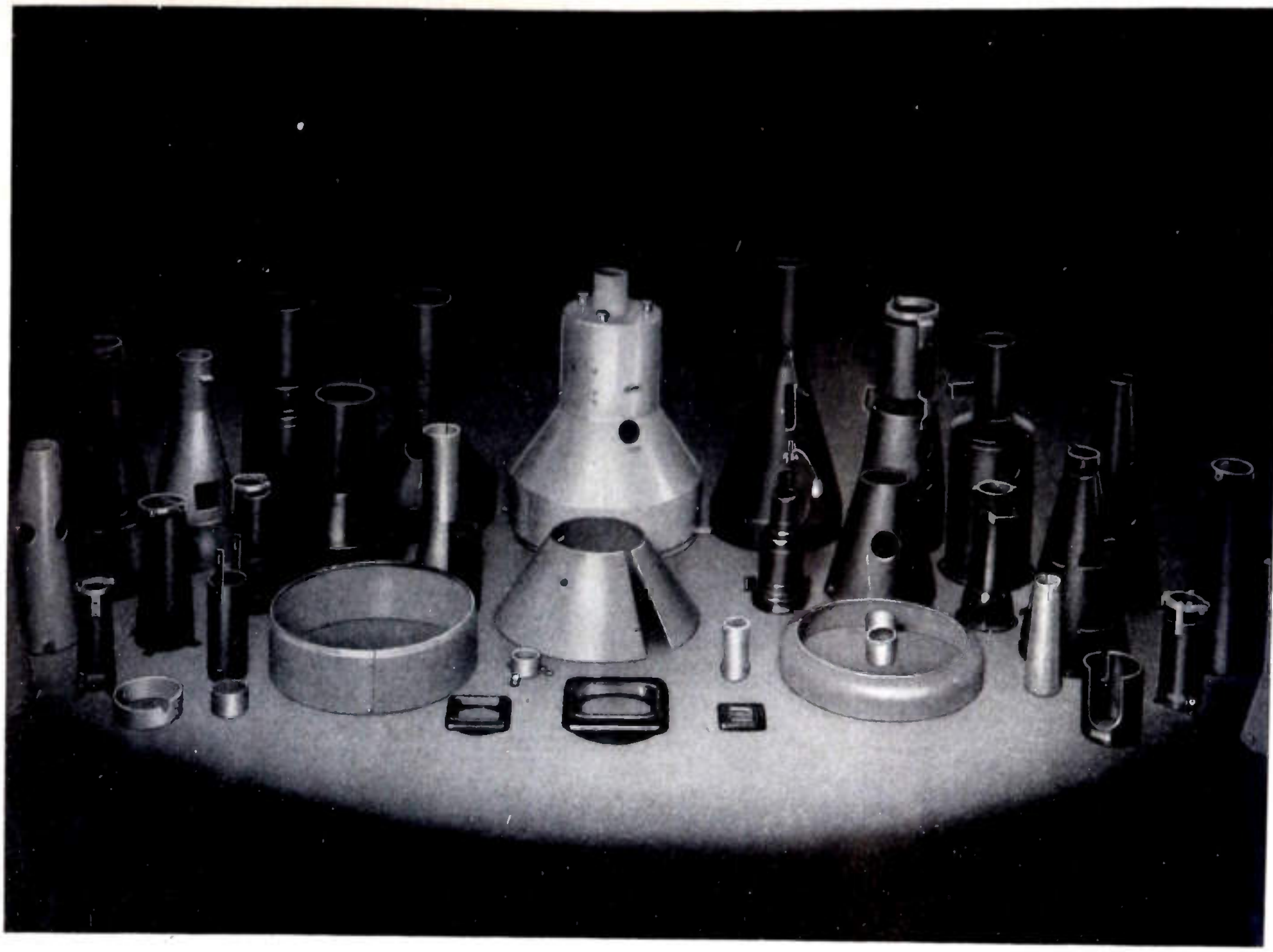
TRADE **ERIE** MARK
RESISTOR CORP.

Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND • • • • TORONTO, CANADA

Cliffside, N. J. • Philadelphia, Pa. • Buffalo, N. Y. • Chicago, Ill.
Detroit, Mich. • Cincinnati, Ohio • Los Angeles, Calif.



Designed for Application

Mu Metal Shields

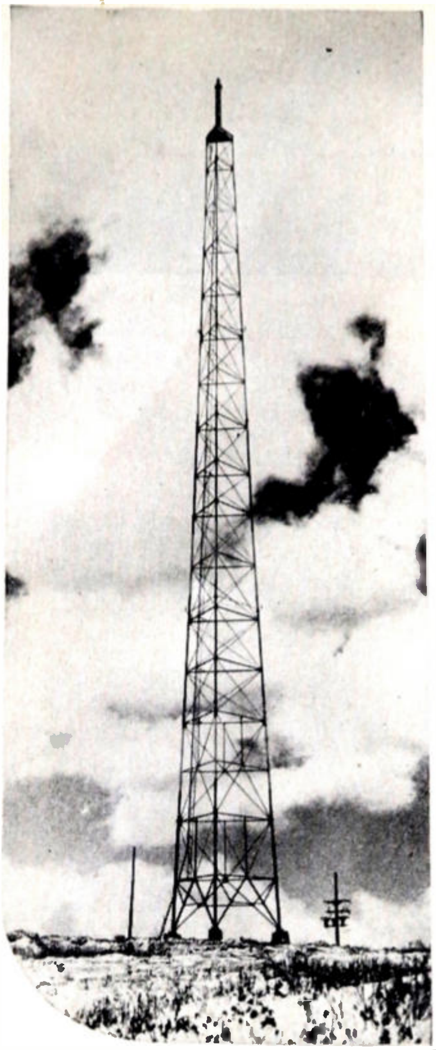
The James Millen Mfg. Co. Inc. has for many years specialized in the production of magnetic metal cathode ray tube shields for the entire electronics industry, supplying magnetic metal shields to manufacturing companies, laboratories and research organizations. Stock shields are immediately available for all of the more popular sizes and types of cathode ray tubes as well as bezels for 2″, 3″ and 5″ size tubes.

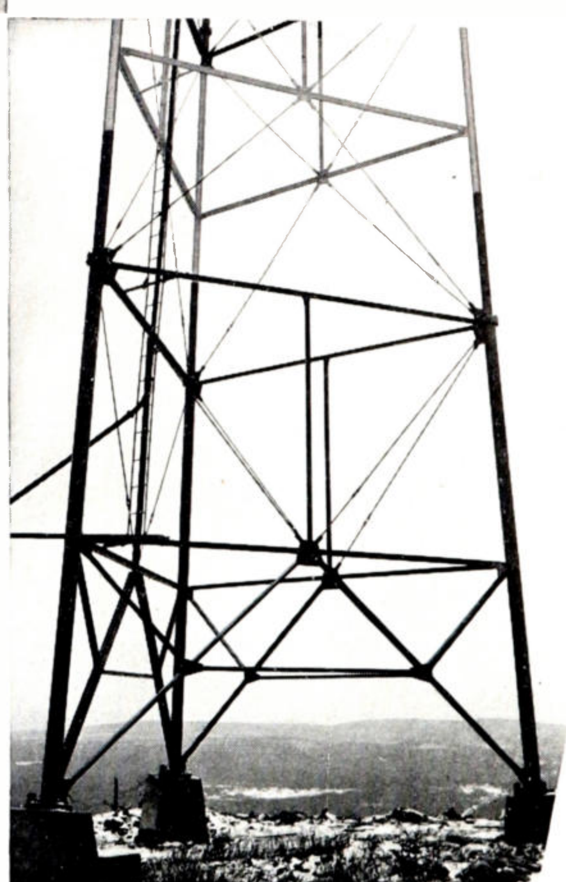
Many production problems, however, make desirable special shields designed in conjunction with the specialized requirement of the basic apparatus. Herewith, are illustrated a number of such custom built shields. Our custom design and fabrication department is at the service of our customers for the development and manufacture of magnetic metal shields of either nicoloi or mumetal for such specialized applications.

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U. S. A.

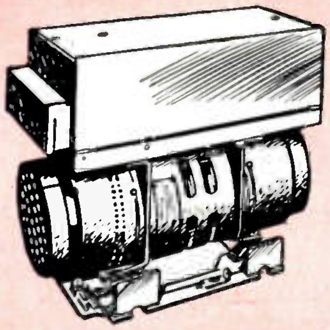


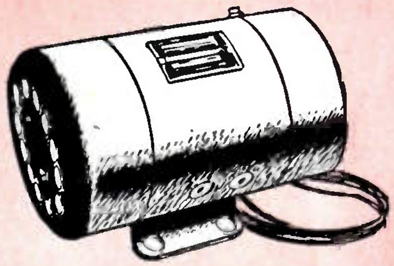


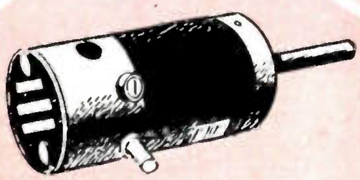
Truscon Self-Supporting Radio Tower operated by Station WILK, Wilkes-Barre, Pa.

Bendix ROTARY POWER SUPPLIES

DYNAMOTORS, INVERTERS, MOTOR GENERATORS
Designed to meet the EXACT Requirements of Each Application!







● Whenever DC is available, Bendix will tailor a complete power supply or motor from standard, mechanical parts to provide the exact voltage—either AC or DC—called for by your equipment.

DYNAMOTORS—essentially DC transformers—will supply one, two, or three DC outputs for direct application to electronic circuits. Radio filtering and voltage regulation are available. Compact, efficient units can be provided with outputs of 10 to 500 watts.

INVERTERS—will produce an AC output for supplying transformer-type power supplies or operating power for servos, synchros, etc. Standard models to work from 28 volts and deliver 115 volts, 400 cycles, single or three phase are available in ratings up to 2500 VA. Frequency and voltage are closely regulated in all models.

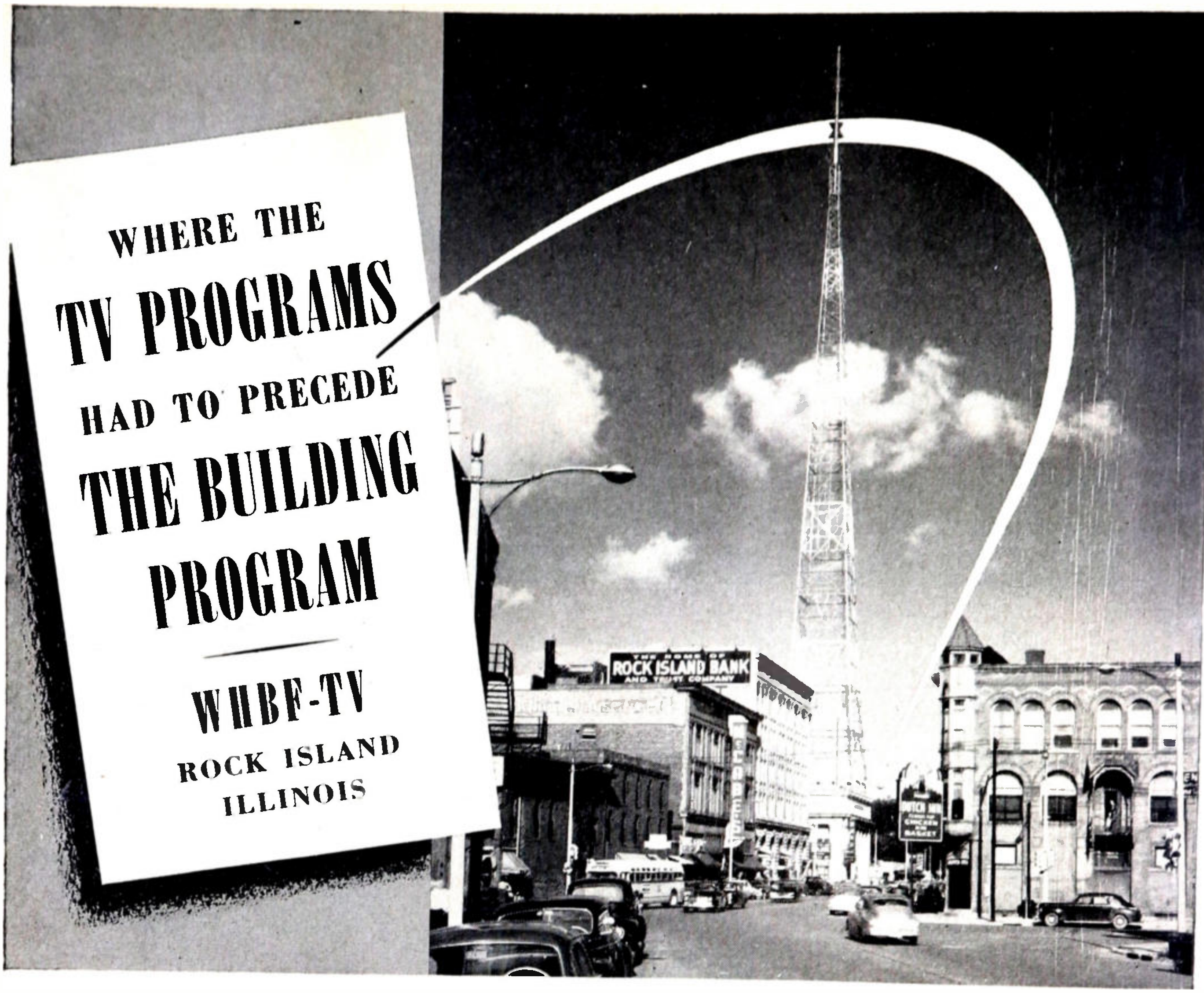
MOTOR GENERATORS—are available for furnishing combinations of DC and AC and for various special requirements.

MOTORS—for performing mechanical functions are designed by Bendix engineers for the most efficient utilization of space and power.

YOUR POWER SUPPLY PROBLEM will receive prompt engineering attention at Bendix. Please send a complete description of the performance required and the condition under which the supply must work. You will be answered with detailed information and specific recommendations for the most practical solution to your problem.

BENDIX AVIATION CORPORATION
RED BANK DIVISION
EATONTOWN, NEW JERSEY
Export Sales: Bendix International Division, 72 Fifth Avenue, New York 11, N. Y.

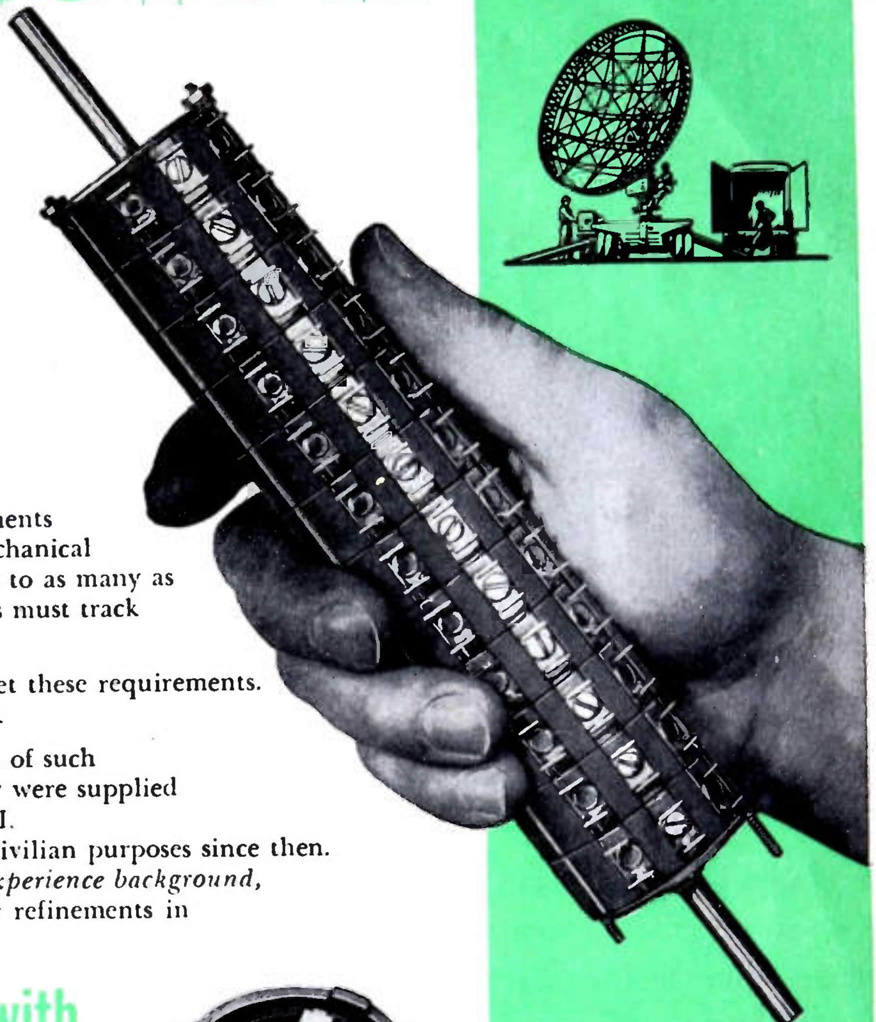
Bendix
AVIATION CORPORATION



WHERE THE
TV PROGRAMS
HAD TO PRECEDE
THE BUILDING
PROGRAM
WHBF-TV
ROCK ISLAND
ILLINOIS
ROCK ISLAND BANK

PRECISION

Controls



At the *very heart* of highly critical equipment such as electronic computers, electronic gunsights and radar assemblies, the control requirements call for outstanding electrical and mechanical precision. Indeed, from single section to as many as twenty sections, the precision controls must track with mathematical accuracy.

Clarostat Series 42 Controls fully meet these requirements. Thus *the climax in precision controls.*

Clarostat has made the major portion of such precision controls in use today. Many were supplied to the armed forces in World War II. Many more have been supplied for civilian purposes since then. And now, *based on an unparalleled experience background,* Clarostat engineers offer you further refinements in their latest Series 42 design.

You can stand pat with CLAROSTAT

Engineering Bulletin No. 142 sent on request. And remember, when your control or resistor requirements call for *quality, quantity* and *economy,* you can meet them with Clarostat's engineering and production facilities. *Submit that problem!*

New Clarostat Series 42 potentiometer. Available in single and multiple assemblies up to 20 sections. Precision windings to plus/minus 0.5% and better. Positive contact rotor, smooth rotation, minimum wear. Perfect tracking of all units in assembly. No backlash or play. Rotor of each potentiometer mounted on centerless-ground shaft passing through all sections.

Controls and Resistors

CLAROSTAT

CLAROSTAT MFG. CO., INC., DOVER, NEW HAMPSHIRE
In Canada: Canadian Marconi Co., Ltd., Toronto, Ontario


JK
PRODUCTS
keeping communications ON THE BEAM
JK STABILIZED
H-17 CRYSTAL
CRYSTALS FOR THE CRITICAL
The small, compact H-17 is designated as a military type crystal for its use in mobile units common to the military. Frequency range: 200 kc to 100 mc. Hermetically sealed holders; wire-mounted, silver-plated crystals.
the JK
FD-12
FREQUENCY AND MONITOR MODULATION
Monitors any four frequencies anywhere between 25 mc and 175 mc, checking both frequency deviation and amount of modulation. Keeps the "beam" on allocation, guarantees more solid coverage, too!
"High Gear" Response to High Power Maintenance!
Dawn or dusk, it doesn't matter. These heroes of the high wires arrive to stop power trouble before it starts. Their "nose for disaster" is in the service truck, in the mobile radio unit which often relies on JK crystals and monitors to keep their assigned radio frequency on the beam!
THE JAMES KNIGHTS COMPANY
SANDWICH 1, ILLINOIS

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 27A)

IF Swept Signal Generator

Avion Instrument Corp., 299 State Highway #17, Paramus, N. J., announces a signal generator which combines all functions normally required in IF amplifier work. The generator incorporates four instruments in one: (1) A standard signal generator covering 4.5 mc to 95 mc at 0.25 per cent accuracy, with calibrated output to ½ volt RMS. (2) A wide range swept frequency generator covering center frequencies from 7 to 70 mc, sweep width continuously adjustable up to ±30 per cent of any center frequency. Calibrated output adjustable up to ½ volt RMS, and constant to 0.25 db while sweeping. (3) A precision frequency marker generator covering the full swept range, accurate to ±100 kc. (4) A pulsed rf signal generator covering the 4.5 to 95 mc range with internal pulse amplifier for introduction of external pulse modulation. Output to ½ volt peak to peak. Rise time 0.1 microsecond.

Additional features include an internal crystal calibrator, sweep voltage output for oscilloscope, variable, and fixed-step push button attenuator calibrated over a 100 db range, and a complete accessory cable complement.

Scaler

A restyled model of the Autoscaler, incorporating the latest techniques in scaler construction, is announced by **Tracerlab, Inc.,** 130 High St., Boston 10, Mass.

The new Autoscaler has a completely electronic scaling circuit of 12 scales of-two and is so designed that any predetermined count corresponding to the powers-of-two from 4 to 4096 may be selected. Panel-mounted neon lights indicate the progress of the count and the elapsed time is shown on the odometer-type timer which reads to 999.99 minutes with an

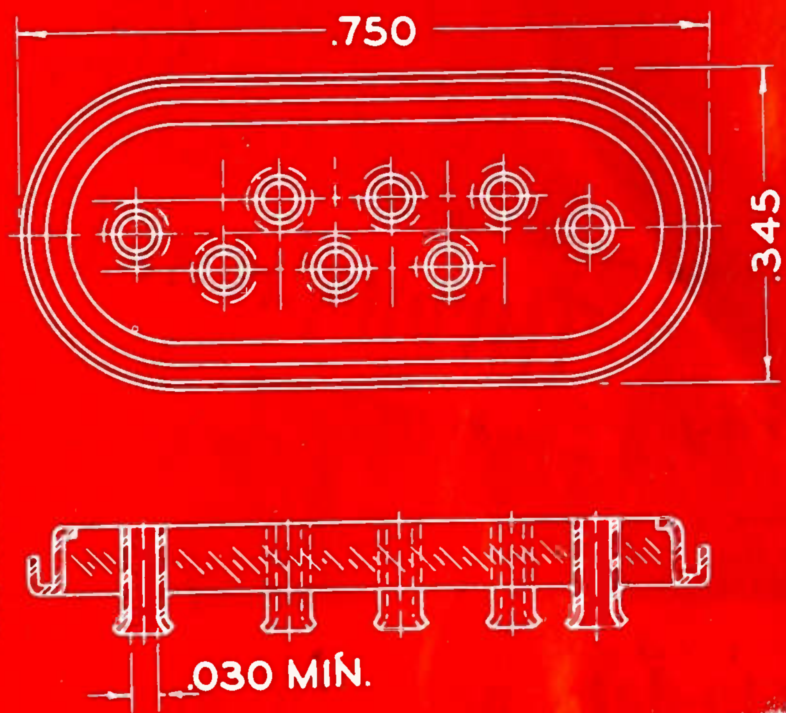
(Continued on page 44A)

SUB
SUB
SUB

Part 1011 — .570
Part 1012 — .765
Part 1228 — 1.516

Part 1531

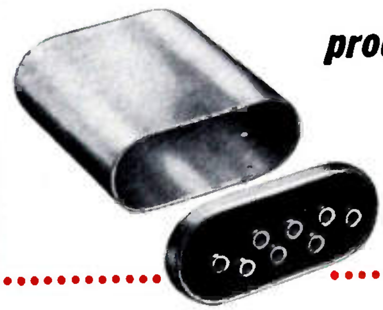
Ceramic-metal part shown with tubulations is also available with pins.



SUB
SUB

SUB-miniature HERMETICALLY SEALED RELAY ENCLOSURE

produced in 10 DAYS from modified EXISTING TOOLS.



Hermetic was called upon to develop a sub-miniature relay enclosure in a matter of weeks and came up with a solution in exactly 10 working days, by utilizing existing tooling and modifying one of its standard parts. This points up the unusually wide range of Hermetic's production and its ability to adapt it to particular problems, no matter how difficult. Naturally, the customer was saved a considerable amount of time and money and was able to meet a deadline with ease.

Because the solution of this problem is characteristic of Hermetic's ability to serve **you,** contact the one and only **dependable** source of supply, and be **sure** that **your** problems will be solved, too.

Write for your FREE copy of Hermetic's colorful, informative, new brochure, the most complete presentation ever offered on hermetic seals.

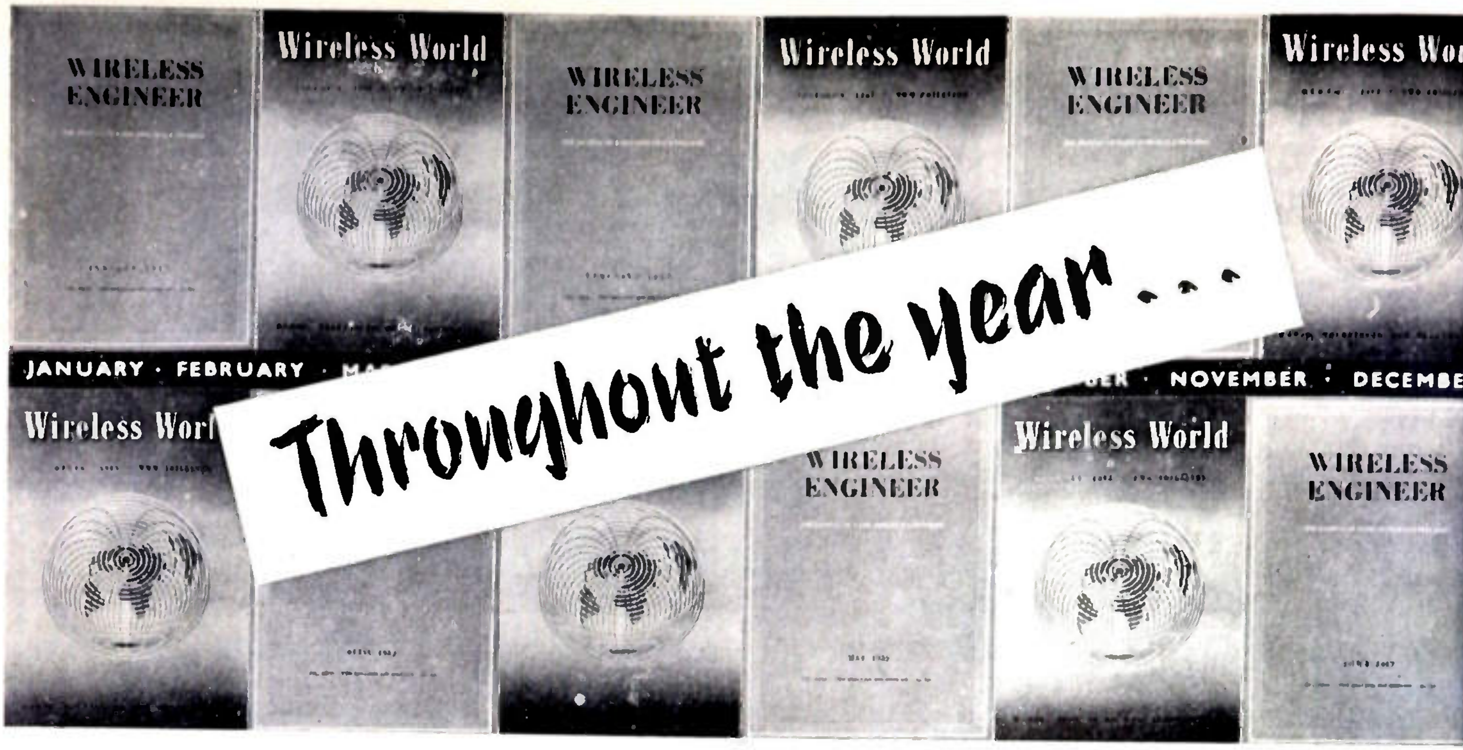
Visit Hermetic's booth #418 at the Western Electronic Show and Convention, Long Beach, Cal., August 27, 28 and 29.

SEALED IN by HERMETIC HEADERS

Hermetic Seal Products Co.

FIRST AND FOREMOST IN MINIATURIZATION

29 SOUTH SIXTH STREET NEWARK 7, NEW JERSEY



up-to-date news of every British development

WIRELESS ENGINEER—the magazine of radio research and progress—is produced for research engineers, designers and students in the fields of radio, television and electronics. Its editorial policy is to publish only original work, and its highly specialized content is accepted as the authoritative source of information for advanced workers everywhere. The magazine's Editorial Advisory Board contains representatives of the National Physical Laboratory, the British Broadcasting Corporation, and the British Post Office. Regular features include an Abstracts and References Section compiled by the Radio Research Organization of the Department of Scientific and Industrial Research. *Published monthly, $7.00 a year.*

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

Published monthly, $4.50 a year.

Recent Editorial Features :

Visibility of Radar Echoes. Shunt-Regulated Amplifiers. Dielectric Lens Aerial. Directional-Coupler Errors. Impedance Changes in Image Iconoscopes. Precision Calibrator for Low Frequency Phase-Meters. Television Camera Tubes.

ASSOCIATED ILIFFE PUBLICATIONS

Recent Editorial Features :

Speech Reinforcement in Reverberant Auditoria : Use of Time Delays and Line-source Loudspeakers. Magnetic Recording: Mechanism of Asymmetrical Hysteresis. Valve Voltmeter without Calibration Drift : "Infinite-input, Zero-output-resistance" Adaptor for D.C. Voltmeters.

MAIL THIS ORDER TODAY

To ILIFFE & SONS LIMITED, DORSET HOUSE, STAMFORD STREET, LONDON, S.E.1, ENGLAND

Please forward .... for 12 months. Payment is being made* ....

NAME ....

ADDRESS ....

CITY .... ZONE .... STATE ....

**Payment can be made by Banker's Draft or International Money Order*

Simplify your production procedure with High-precision

Stupakoff

CERAMIC to Metal ASSEMBLIES

for Electrical and Electronic Applications

To combine ceramic and metal parts into one permanent unit, Stupakoff draws upon extensive experience with both materials. Methods of assembly employed by Stupakoff include: metallizing, soldering, pressing, spinning and others. Among the metals assembled to ceramics are silver, copper, brass, stainless steel and monel.

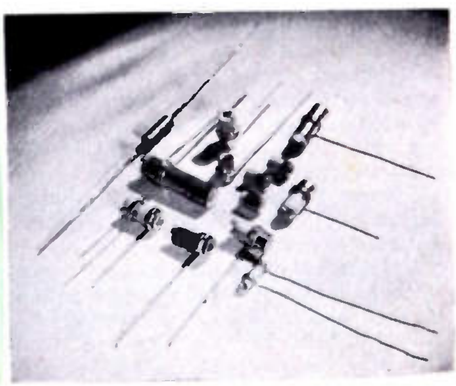
The rotor shafts shown above consist of metal bands attached securely to ceramic rods, and exemplify Stupakoff precision manufacture. On a mass production basis, concentricity of components, for example, are held to less than ± 0.001 in. Likewise, the strains and spreaders, stand-offs and trimmers shown below meet the exacting requirements of the service for which they are made.

Stupakoff high-precision ceramic assemblies offer many opportunities to reduce costs, increase production and improve electrical and electronic equipment.

We will be glad to discuss your requirements with you and to submit samples for your inspection.

STUPAKOFF CERAMIC & MANUFACTURING CO.

LATROBE, PENNSYLVANIA



(Continued from page 40A)

accuracy of 0.01 minutes. The timer stops automatically when the predetermined count is completed.

The end of a counting run is signaled by an aural monitor which emits a 120 cps tone. This monitor also indicates the arrival of each pulse by means of a sharp click.

Other features include automatic resetting of the scales when the count stops, a coincidence loss of less than 0.8 per cent for up to 100,000 random pulses per minute, a resolving time of 5 microseconds, and a stop button to interrupt the count at will.

A shielded high-voltage electronically-regulated power supply is provided so that 1 per cent change in the line voltage will cause only 0.01 per cent change in high voltage. The high voltage is continuously adjustable from 300 to 2,500 volts by means of a multiturn potentiometer.

Noise Level Meter

General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass., announces the GR Type 1555-A sound-survey meter.

Among the industrial uses of the instrument are the measurement of machinery noise, deafness-risk surveys, and noise studies on industrial equipment and household appliances. Audio engineers will find the sound-survey meter useful for adjusting relative level of speakers, checking the dynamic range, and frequency response.

Although shaped to fit the hand, the sound-survey meter can be set on a table or mounted on a tripod. Only two controls are used and both are mounted on the front together with the meter. The instrument can be carried in a pocket but a leather carrying case, with room for spare batteries, is available as an accessory.

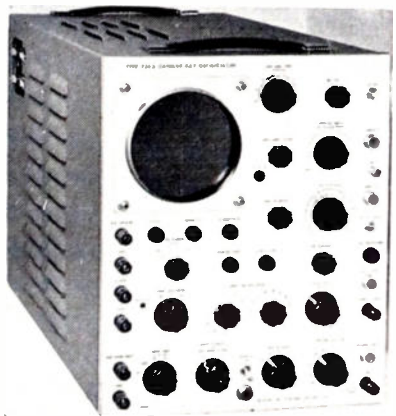
The total sound-pressure-level range of the meter is from 40 to 136 decibels and three frequency weighting networks are provided. Although low cost and small size were design objectives, high-quality components have been used throughout.

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

The Type 1555-A sound-survey meter is priced at $125.00 net including batteries.

Television Oscilloscope

The Type 524-D Oscilloscope designed to meet the needs of television broadcasters in adjusting and maintaining television transmitters and studio equipment, is available from **Tektonix, Inc.,** P.O. Box 831, Portland 7, Ore.



A variable sweep delay circuit provides a zero to 25 millisecond delay. Delayed sweeps, triggered by any line sync pulse throughout the picture, are available through the entire sweep range of 0.01 sec/cm to 0.1 μsec/cm. Field selector permits switching from one field of the frame to the other at will. An internal sync separator permits triggering from sync pulses of the composite video signal. In the new sweep magnifier, the center of the sweep remains fixed and the sweep is expanded to right and left of center. Magnifications of three times and ten times are provided, permitting detailed examination of sync and equalizing pulses. A 60 cps sine wave sweep has front panel phasing control and amplitude control.

An internal time mark generator modulates the trace brightness. Pips spaced 1 μsec, 0.1 μsec, 0.05 μsec, and 200 pips per television line are available. The new amplitude calibrator produces a variable duty cycle square wave, accurate within 3 per cent. The calibrator is continuously variable through 7 ranges, from 0.05 v to 50 v. Duty cycle is adjustable from 1 to 99 per cent.

More than 6 cm undistorted deflection is available on a flat faced c-r tube. Accelerating potential is 4 kv. Vertical sensitivity dc to 10 mc—0.15 v/cm, 2 cps to 10 mc—0.015 v/cm. Risetime is 0.04 μsec, and a signal delay of 0.25 μsec is provided. All dc voltages are electronically regulated.

Mechanical Development Apparatus Brochure

New 16-page illustrated brochure, MDA-200 which describes the complete line of precision components for rapid and economical assembly of control systems instruments and analog computers for breadboard and semi-permanent assembly is announced by **Servomechanisms, Inc.,** Post & Stewart Aves., Westbury, L. I., N. Y.

In addition to the previously standard articles the brochure features a greatly expanded line of new mechanical development apparatus components. The new components include: lead screw unit, clutch, bevel gears, limit stop, dials, cams, bellows and Oldham Couplings, spring-loaded split gears, shaft adapters, block and switch assembly, and larger mounting boards.

Brochure is available on request on standard company letterhead, contact L. C. Briggs.

Power Supply

A continuously variable 0–325 volt electronic-regulated dc power supply has been developed by the **Perkin Engineering Corp.,** 345 Kansas St., El Segundo, Calif. This unit (Model M30) has a dc current rating of 150 ma, and also is equipped with a low-voltage ac output of 6.3 volts at 6 amperes.

Voltage regulation is within ½ per cent for voltages between 30–325 volts from no load to full load, and the ripple is less than 2 mv. This unit has 2–3 inch rectangular meters and is finished in a blue-gray wrinkle color. Further information can be obtained by writing the company at the above address.

Capacity Bridge

Simpson Electric Co., 5200 W. Kinzie St., Chicago 44, Ill., has announced engineering changes on their Model 381 Capacity Bridge.

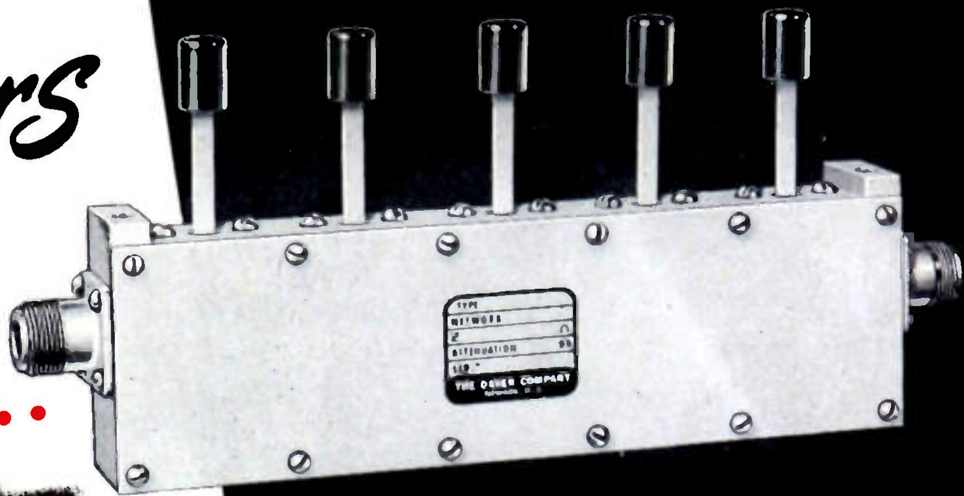
Model 381 circuit enables the inexperienced to make capacity measurement with ease. Merely press a button for the desired range, adjust the bridge arm for maximum meter deflection, and read the capacity on the scale.

The small size of this tester, together with its wide range of capacity measurement and low price makes it a helpful instrument for Radio and Television service, broadcast engineers, electric repair shops, X-ray servicing, industrial maintennace

(Continued on page 56A)

IN *Attenuators*

WHY DOES ONE NAME.. DAVEN STAND OUT?



Series 550-RF Attenuator

In addition to Daven being the leader in audio attenuators, they have achieved equal prominence in the production of RF units. A partial listing of some types is given below.

DAVEN Radio Frequency Attenuators, by combining proper units in series, are available with losses up to 120 DB in two DB Steps or 100 DB in one DB Steps. They have a zero insertion loss and a frequency range from DC to 225 MC.

Standard impedances are 50 and 73 ohms, with special impedances available on request. Resistor accuracy is within ± 2% at DC. An unbalanced circuit is used which provides constant input and output impedance. The units are supplied with either UG-58/U* or UG-185/U** receptacles.

TYPE	LOSS	TOTAL DB	STANDARD IMPEDANCES
RFA* & RFB 540**	1, 2, 3, 4 DB	10	50/50Ω and 73/73Ω
RFA & RFB 541	10, 20, 20, 20 DB	70	50/50Ω and 73/73Ω
RFA & RFB 542	2, 4, 6, 8 DB	20	50/50Ω and 73/73Ω
RFA & RFB 543	20, 20, 20, 20 DB	80	50/50Ω and 73/73Ω
RFA & RFB 550	1, 2, 3, 4, 10 DB	20	50/50Ω and 73/73Ω
RFA & RFB 551	10, 10, 20, 20, 20 DB	80	50/50Ω and 73/73Ω
RFA & RFB 552	2, 4, 6, 8, 20 DB	40	50/50Ω and 73/73Ω

GREATLY EXPANDED PRODUCTION FACILITIES ENABLE DAVEN TO MAKE DELIVERY FROM STOCK ON A LARGE NUMBER OF STANDARD ATTENUATOR TYPES.

Because DAVEN makes the most complete, the most accurate line of ATTENUATORS in the world!

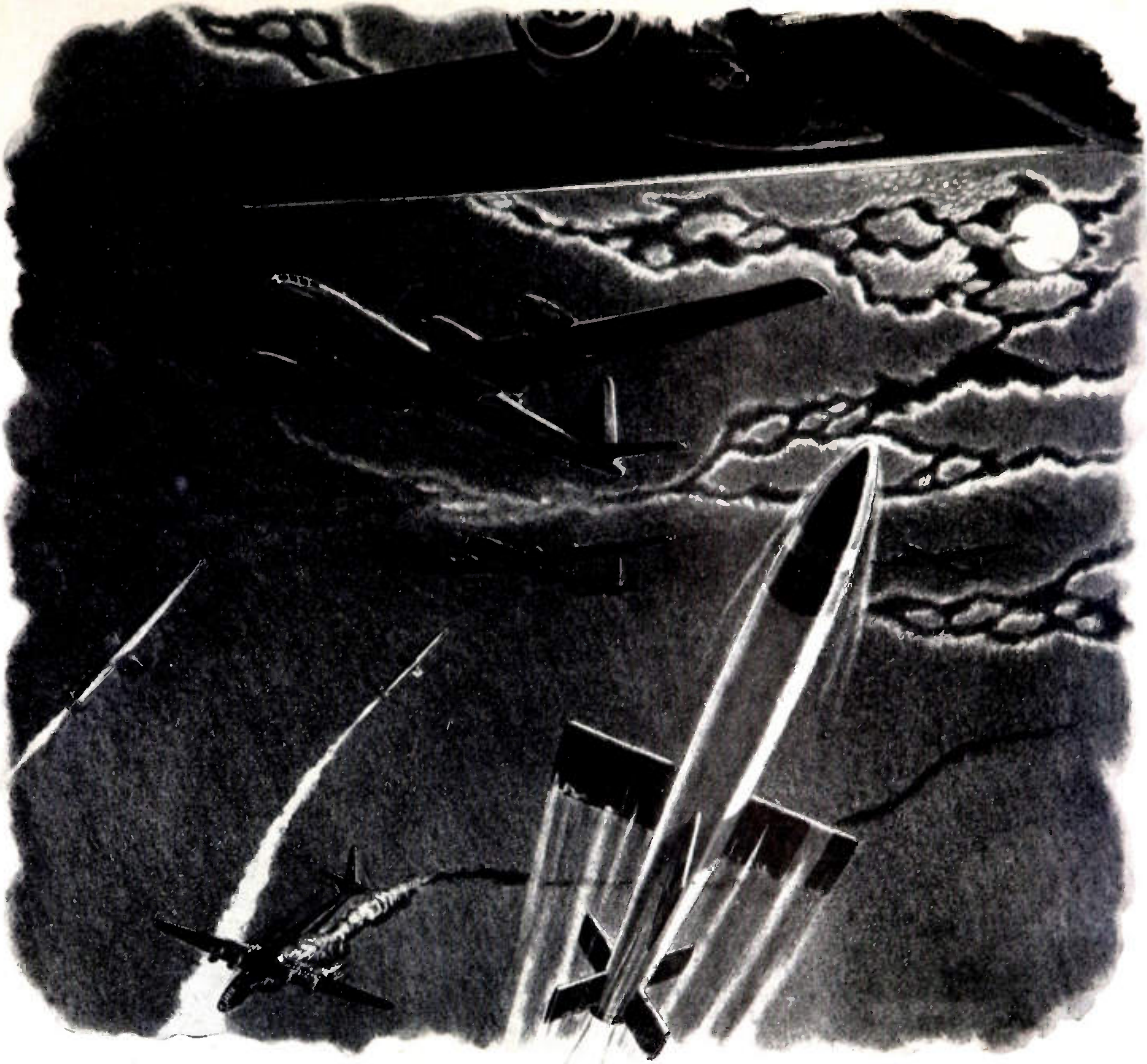
Series 640-RF Attenuation Network

These units are now being used in equipment manufactured for the Army, Navy and Air Force.

Write for Catalog Data.

THE DAVEN CO.

195 CENTRAL AVENUE
NEWARK 4, NEW JERSEY



Rauland Tubes give you a prettier profit picture

When you rely on Rauland picture tubes you get the benefit of acknowledged leadership in picture tube engineering . . . *which usually means that you'll be first to know of the latest picture tube improvements.* Rauland research has developed more picture tube improvements in the past 5 years than any other picture tube source. And naturally, Rauland customers get the break in announcing these firsts in their sets.

You get quality you can count on, too. Rauland production employs machines unique in the industry—many of them designed by Rauland engineers and built in Rauland's own plant.

And finally, you get assurance of customer satisfaction beyond what any other line can give you. Installation and adjustment of sets in the field is faster and better with Rauland's patented Indicator Ion Trap. It gives the surest known protection against ion burn and shortened tube life.

• • •

Specify Rauland—deliver Rauland—and assure yourself of pleased dealers and consumers.

THE RAULAND CORPORATION

Perfection Through Research

4245 N. KNOX AVENUE • CHICAGO 41, ILLINOIS

A page from the note-book of Sylvania Research

Sylvania Research Improves Semiconductor Performance

Development of improved processes for producing germanium crystals and improved methods of measuring their properties is a continuous program of the Sylvania Research Laboratories. The result is an ever increasing proficiency of semiconductor devices such as Sylvania germanium diodes and varistors.

The range of useful work performed by the many forms of Sylvania germanium crystals in communications and other electronic applications seems almost without limit. It fully justifies the large investment in research which Sylvania puts behind it. And it is another reason for Sylvania leadership in the world of electronics.

In the illustration, the fixture holds a sample of germanium to which electrical connections are made by means of tungsten wire whiskers. The pressure of contact is controlled by the knurled micrometer knob at the right. When the proper voltages are applied, a pattern is produced on the oscilloscope which permits the technician to measure electrical properties of the crystal and to study variations due to changes in methods of preparation.

SYLVANIA

Sylvania Electric Products Inc., 1740 Broadway, New York 19, N. Y.

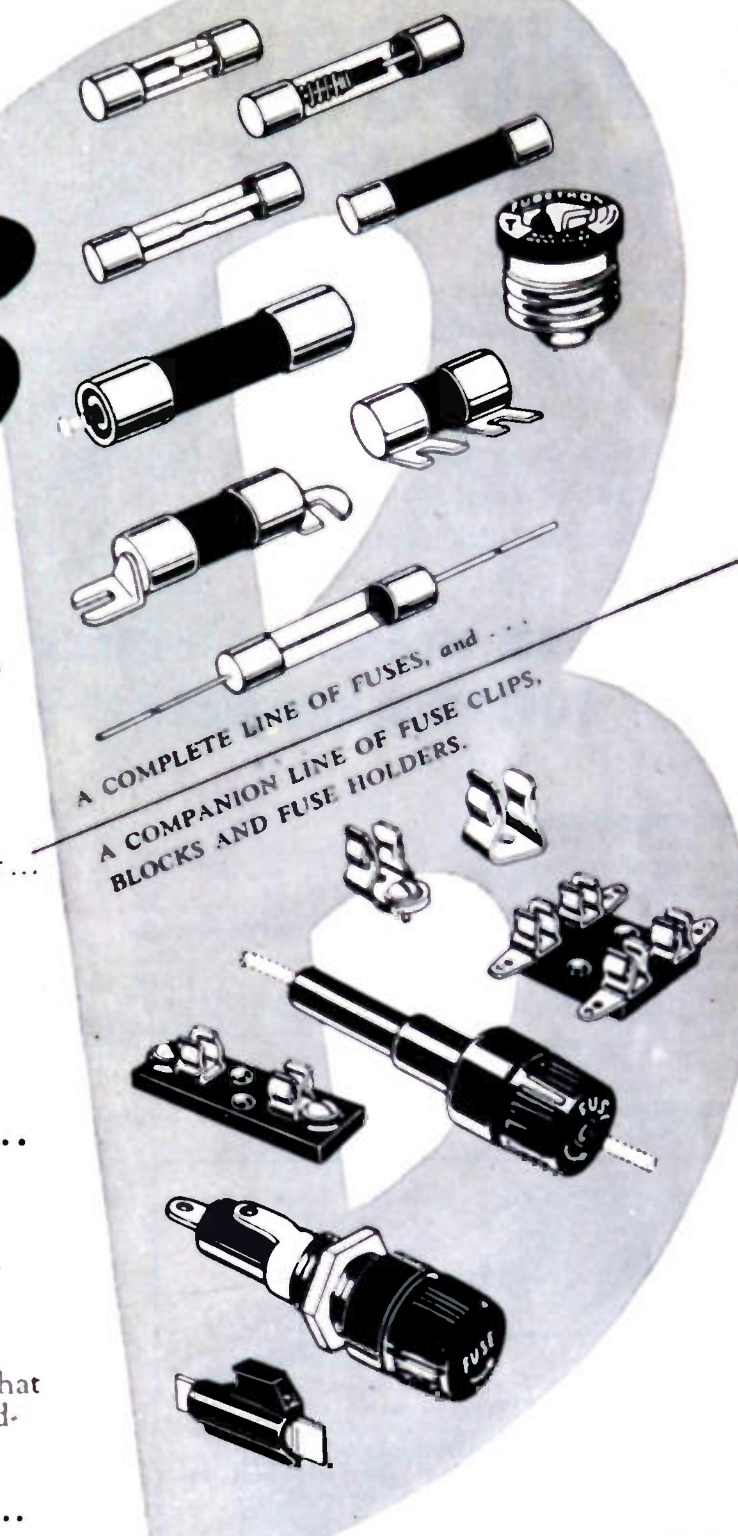
TELEVISION PICTURE TUBES; ELECTRONIC TEST EQUIPMENT; ELECTRONIC DEVICES; RADIO TUBES; FLUORESCENT TUBES, FIXTURES, SIGN TUBING, WIRING DEVICES; LIGHT BULBS; PHOTOLAMPS; TELEVISION SETS

turn to BUSS

THE ONE SOURCE FOR ANY

FUSE..

From 1/500 Amp. Up



BUSS offers the most complete line of fuses for Television...Radio...Radar... Instruments...Controls...Avionics... in standard types and dual-element (slow blowing) types.

By using BUSS as your one source for fuses you simplify your buying, stock handling and records.

Every BUSS Fuse is Electronically Tested... in a sensitive device that rejects any fuse that is not correctly calibrated, properly constructed and right in all physical dimensions. This means every BUSS Fuse is a good one.

Manufacturers and servicemen throughout the country have learned that they can rely on BUSS fuses for dependable protection.

If You Have a Special Protection Problem... let a BUSS engineer help you. He has at his command the world's largest fuse research laboratory and the world's largest fuse production capacity.

Send the Coupon

FOR MORE FACTS...

BUSSMANN MFG. CO. *Division McGraw Electric Company.*
University at Jefferson, St. Louis 7, Mo.

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

Please send me bulletin SFB containing complete facts on BUSS small dimension fuses and fuse holders.

Name ______________________________

Title ______________________________

Company ______________________________

Address ______________________________

City & Zone __________________ *State* __________ IRE-852

TUNG-SOL®

DAMPER DIODE

- doubles heater-to-cathode insulation rating
- eliminates external damper tube transformer
- no top cap—simplified wiring
- conserves critical materials
- lowers manufacturing costs

FOR TRANSFORMERLESS RECEIVERS
FOR "DIRECT DRIVE" DEFLECTION CIRCUITS

Mechanical Data

Coated unipotential cathode

Outline drawing RMA #9-11

Base RMA #B6-48

Bulb T-9

Short intermediate shell octal 6-pin

Maximum diameter 1-9/32"

Maximum overall length 3-5/16"

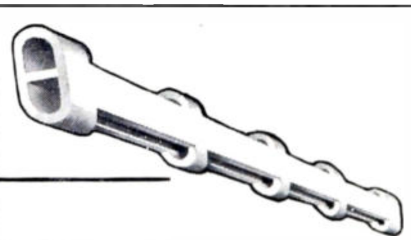
Maximum seated height 2-¾"

Pin connections

RMA basing..............#4CG

- Pin 1—no connection
- Pin 2—no connection
- Pin 3—cathode
- Pin 5—plate
- Pin 7—heater
- Pin 8—heater

Mounting position Any



Unique ceramic sleeve of aluminum oxide, fired to extreme hardness, completely isolates the cathode from the heater wire. Maximum heat transfer is obtained with fullest insulation protection.

Electrical Data

*(Interpreted according to RMA Standard M8-210)**

Ratings

Heater voltage (ac or dc).........................	6.3 VOLTS
Heater current....................................	1.2 AMPS.
Maximum heater-cathode voltage (heater negative)......	900 VOLTS
Maximum peak heater-cathode voltage (heater negative)..	4000 VOLTS **
Maximum heater-cathode voltage (heater positive)......	100 VOLTS
Maximum peak inverse plate voltage................	4000 VOLTS **
Maximum steady state peak plate current............	600 MA.
Maximum transient peak plate current...............	3.0 AMPS. ***
Tube voltage drop (measured with tube conducting 250 ma.)	32 VOLTS
Maximum dc plate current..........................	125 MA.

Interelectrode Capacitance

Heater to cathode..............................	7.5 $\mu\mu$f.

* These are design center ratings. Because of the nature of the service for which this tube is intended, it is important that these values not be exceeded by more than 10% under the most unfavorable operating conditions.

** This rating is applicable where the duty cycle of the voltage pulse does not exceed 15% of one scanning cycle, and its duration is limited to 10 micro-seconds.

*** This rating applies to hot switching where transient duration does not exceed 0.2 seconds.

This type is also available with 12.6 Volts, 600 MA. heater and is designated 12AX4GT.

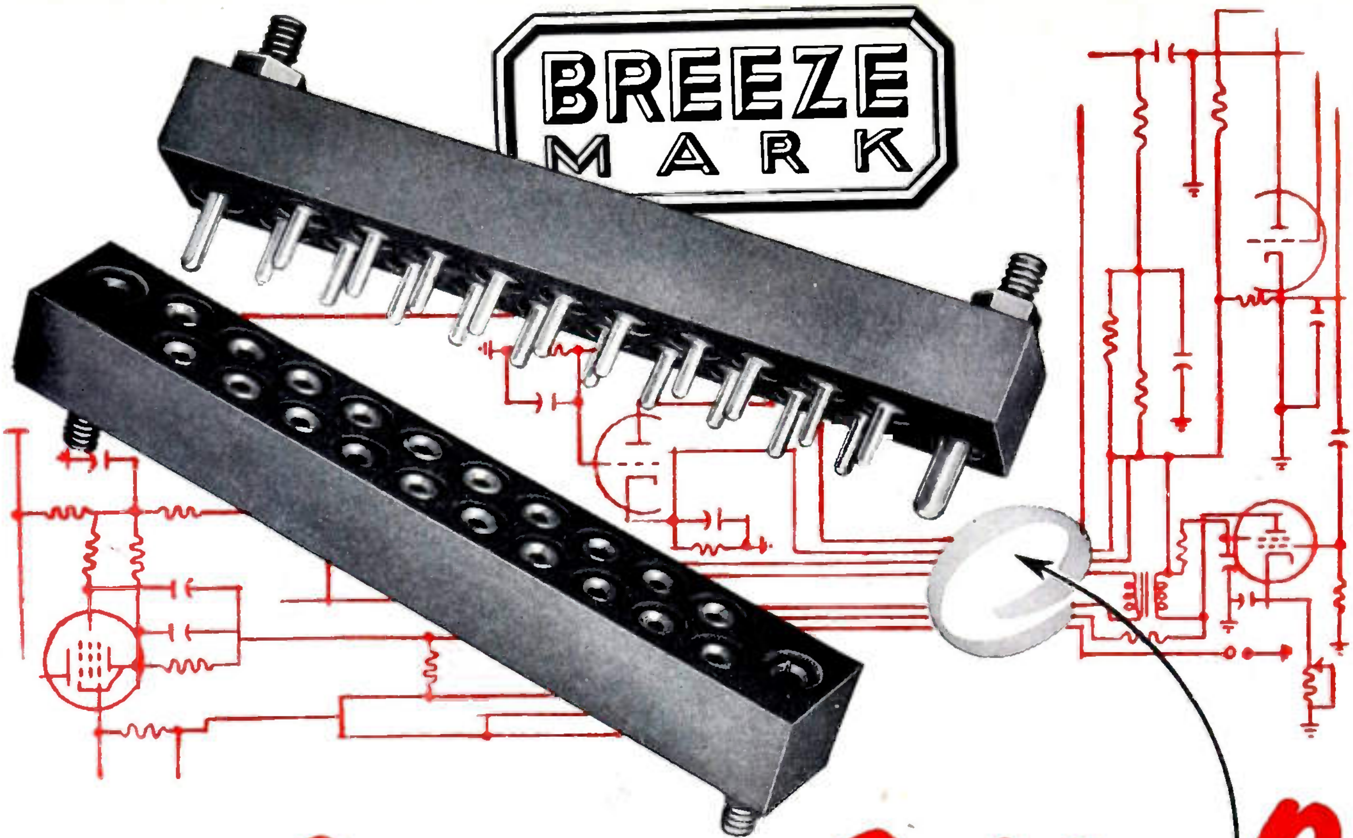
The TUNG-SOL engineering which has produced the 6AX4GT and the 12AX4GT is constantly at work on a multitude of special electron tube developments for industry. Many exceptionally efficient general and special purpose tubes have resulted. Information about these and other types is available on request to TUNG-SOL Commercial Engineering Department.

TUNG-SOL
ELECTRON TUBES

TUNG-SOL ELECTRIC INC., Newark 4, New Jersey

Sales Offices: Atlanta • Chicago • Culver City • Dallas • Denver • Detroit • Newark

TUNG-SOL MAKES ALL-GLASS SEALED BEAM LAMPS, MINIATURE LAMPS, SIGNAL FLASHERS, PICTURE TUBES, RADIO, TV AND SPECIAL PURPOSE ELECTRON TUBES



Connector Problem?

...We'll take it from HERE

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

Lightweight actuators for any requirement.

Job engineered, welded diaphragm bellows.

Flexible conduit and ignition assemblies.

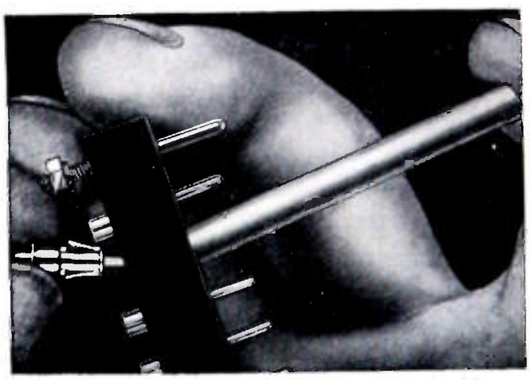
Aero-Seal vibration-proof hose clamps.

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

BREEZE
Special CONNECTORS

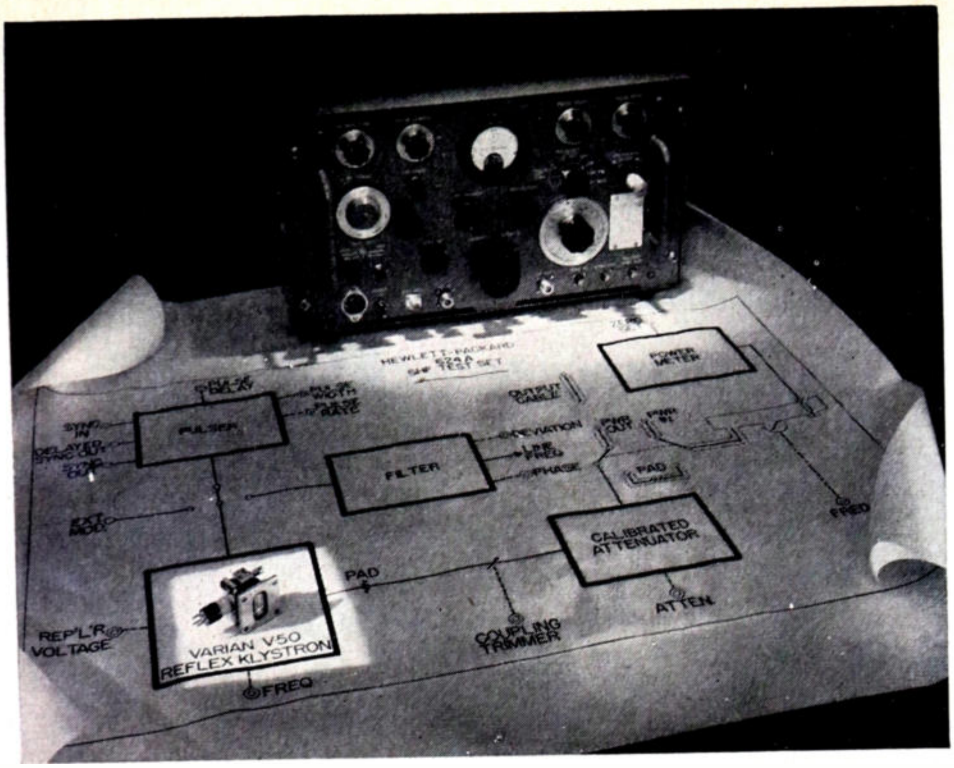
BREEZE CORPORATIONS, INC.

41 South Sixth Street Newark, New Jersey



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

VARIAN associates



V-50 REFLEX KLYSTRON

8.5-10.0 kmc

...smooth-tuning
—no backlash

...dependably uniform

...temperature and
shock-resistant

...simple to install

WHEN HEWLETT-PACKARD engineers designed the new -hp- Model 624A SHF Test Set they sought a signal source of dependable uniformity, high stability under shock and temperature changes, and smooth, chatter-free tuning. To meet these needs, they selected the Varian V-50 reflex klystron.

WHEREVER these characteristics are required in an x-band oscillator, the V-50 merits your consideration. For applications involving still greater shock and vibration, where single shaft tuning is not required, the extremely rugged V-51 may be more suitable.

BOTH THESE VARIAN klystrons are notable for integral-resonator construction; the exclusive Varian wideband mica-seal output window; extremely small space requirement; weight of only six ounces; power output, without special matching transformers, of 25 to 65 milliwatts for the V-50, 75 to 260 mw for the V-51. Both bolt directly, without adapters, to standard inch-by-half-inch x-band waveguide.

YOUR MICROWAVE PROBLEMS may be solved by one of these x-band oscillators. Or, your requirements may be different. Many Varian klystrons, for many different types of services, are in production or development but cannot be publicized. Correspondence is invited concerning klystrons for your specific needs.

VARIAN associates
990 VARIAN STREET · SAN CARLOS 2, CALIFORNIA

FIELD ENGINEERING REPRESENTATIVES IN PRINCIPAL CITIES

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 44A)

departments, or any other service where capacitors are tested.

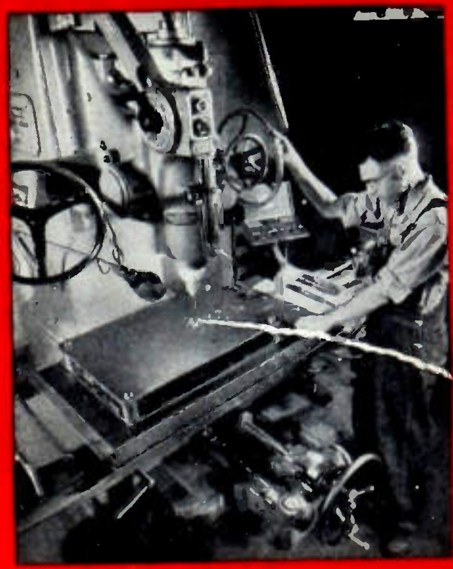
Model 381 comes in a bakelite case with an etched aluminum panel. Dealer's net price is $28.50.

24 Hour Recorder Reproducer

Press Wireless Mfg. Corp., Hicksville, L. I., N. Y., announces a single-channel magnetic recorder-reproducer (Model RRP) which will continuously record or transcribe voice-frequency intelligence for an entire 24 hours, or by utilization of its voice-actuated relay will record intermittant operations for a number of days. Separate recording and playback amplifiers permit simultaneous reproduction, or playback of earlier intelligence, while the machine is operating in its recording condition.

The RRP-24 uses an iron-oxide-coated cellulose-acetate magnetic medium that is 8½ inches wide and 206 feet in length. The medium winds itself on a drum and then, with a reversal time of only 1/40th second, proceeds in the opposite direction to rewind on another drum. At 4 inches per second the medium proceeds in each direction for a period of 10 minutes, thereby developing 144 recorded tracks in 24 hours. Each track is 0.30 inch wide with 0.20 spacing between adjacent tracks.

(Continued on page 58A)

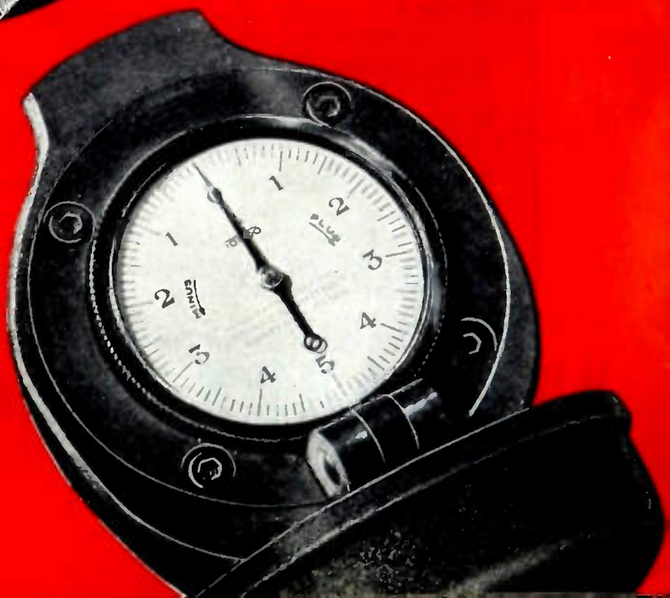


± $\frac{1}{10{,}000}$

Lives and property in the air are suspended on a beam of communications signals. Ground and air communications equipment must be right . . . it must have *unfailing dependability.*

Wilcox insures this dependability with *precision* design and engineering . . . *precision* tooling and fabrication of each component of Wilcox communication equipment.

This Pratt & Whitney Jig Borer used in the manufacture of Wilcox communication equipment has a resettable tolerance of ± 1/10,000 inch.



wilcox

ELECTRIC COMPANY, INC.

Fourteenth and Chestnut
Kansas City 1, Missouri

means dependable communications

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)

There is a fine index which consists of a helix drawn indelibly upon the recording medium, and as this line progresses laterally, being a spiral, it indexes to a 10 minutes scale divided into increments of 10 seconds. Accuracy of location of an earlier recording is better than 10 seconds in 24 hours.

To preclude the necessity of close supervision while recording, a compressor amplifier is incorporated in the audio input system. Its threshold level is preset to prevent material increase in gain in the total absence of a signal.

Scintillation Convertor Unit

A scintillation convertor unit, Model A-200, is now commercially available from **Nuclear Research & Development, Inc.**, 1094 Sutter Ave., St. Louis 5, Mo.

This unit is designed to allow users to employ scintillation counters with an ordinary Geiger scaler. Model A-200 provides the high voltage supply for the counter; the pulses from the counter are fed into the unit for amplification and pulse discrimination, and then into the scaler. The high gain amplifier has a rise time of 0.25 microseconds and a variable gain up to 2,000, providing sensitivity down to 0.75 millivolt pulses.

The discriminator is a modified Schmitt trigger circuit and is adjustable on the front panel for 0 to −100 volts.

Good high voltage regulation is obtained by adjusting the B+ and using that as a reference voltage. This results in better than 0.005 per cent change in output voltage for a 1 volt change in input voltage between 95 and 130 volts.

With this unit, one can be assured of dependable and quantitative scintillation counting even with a Geiger scaler.

Impedance Bridge

Brown Electro-Measurement Corp., 4635 S.E. Hawthorne Blvd., Portland 15, Ore., announces a Universal Impedance Bridge (Model 250C) and companion Amplifier-Oscillator (Model 855A). Measurements can be made with this new Impedance Bridge over the following ranges: Resistance, 1 milliohm to 11 megohms; Capacitance, 1 $\mu\mu f$ to 1100 μf; Inductance 1 micro-henry to 1100 henrys; Storage

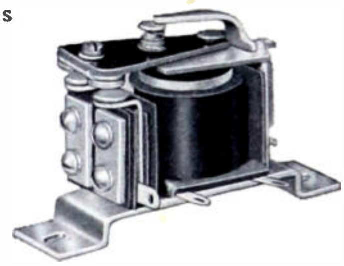
(Continued on page 76A)

Relays BY GUARDIAN

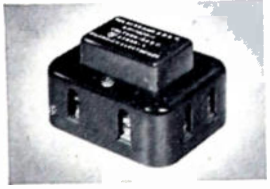
GUARD AGAINST WARTIME DESTRUCTION . . . GUARD AGAINST THEFT!

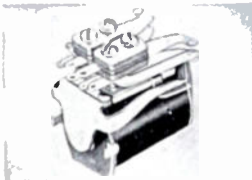
To guard our country against long-range bombing . . . to guard homes from prowling criminals . . . to guard property from fire . . . to guard against accidents in plants . . . in such alarm and safety systems Guardian Relays respond dependably at the crucial moment. The superior performance of "air-raid" and fire alarm sirens controlled by the Guardian Series 220 A.C. Relay, is known the world over. The Series 220 A.C. Relay provides contact combinations up to S.P.S.T. and is furnished *hermetically sealed* or with open-type mounting. The Guardian Series 5 D.C. Relay (with cover) is especially adapted to electronic tube plate circuits for photocell operation in bell-ringing, flood-lighting and coded ticker-tape fire and burglar alarm central systems. It has successfully followed 20 contacts per second for high speed counting, telephone dialing, timing and lighting control. We invite you—write for full details and cost-free advice on the use of Guardian Relays for your specific application.

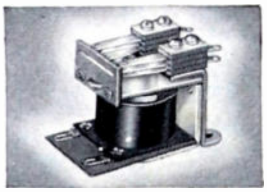

OCTAL PLUG HERMETICALLY SEALED CONTAINER


SERIES 220 A.C. RELAY


AN-3312-1 D.C.


AN-3324-1 D.C.


Series 595 D.C.


Series 610 A.C.—615 D.C.


Series 695 D.C.

Get Guardian's New HERMETICALLY SEALED RELAY CATALOG Now!

GUARDIAN G ELECTRIC

1628-J W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

MYCALEX the ideal insulation

INJECTION MOLDED GRADES:

MYCALEX 410

MYCALEX 410 IS APPROVED FULLY AS GRADE L-4B UNDER NATIONAL MILITARY ESTABLISHMENT SPECIFICATION JAN-I-10 "INSULATING MATERIALS, CERAMIC, RADIO, CLASS L." CAN BE INJECTION MOLDED, WITH OR WITHOUT METAL INSERTS, TO EXTREMELY CLOSE TOLERANCES.

CHARACTERISTICS

Power Factor, 1 megacycle	0.0015
Dielectric Constant, 1 megacycle	9.2
Loss factor, 1 megacycle	0.014
Dielectric Strength, volts/mil	400
Volume Resistivity, ohm-cm	1×10^{15}
Arc Resistance, seconds	250
Impact Strength, Izod ft.-lb./in. of notch	0.7
Max. Safe Operating Temp. °C	350
°F	660
Water Absorption, % in 24 hours	nil
Coefficient of Linear Expansion, °C	11×10^{-6}
Tensile Strength, psi	6000
Compression Strength, psi	25000

★ ★ ★

MYCALEX 410X

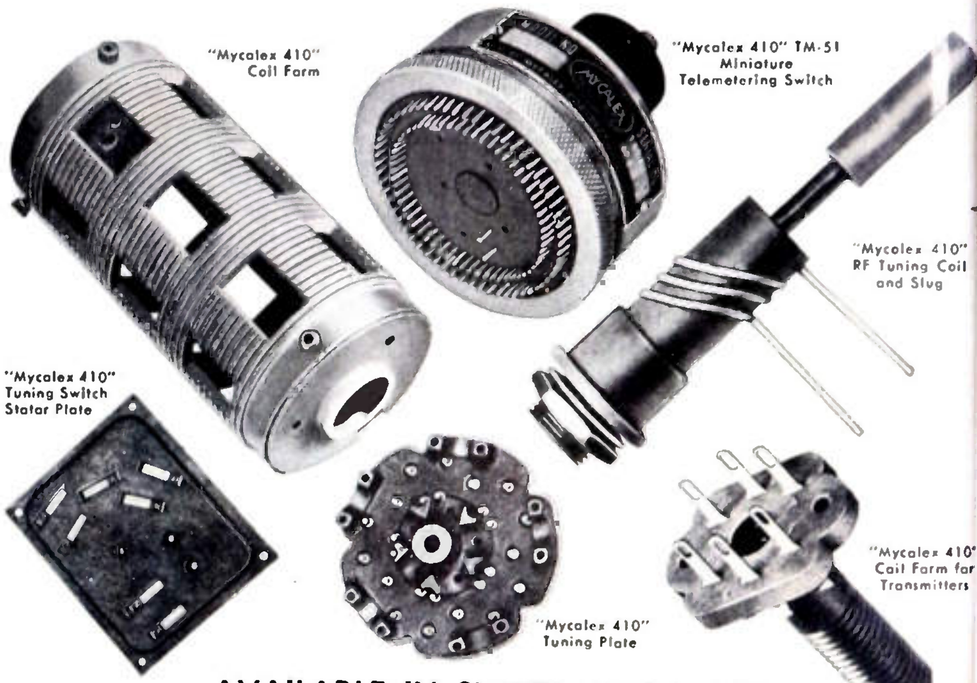
MYCALEX 410X CAN BE INJECTION MOLDED, WITH OR WITHOUT METAL INSERTS, TO EXTREMELY CLOSE TOLERANCES.

CHARACTERISTICS

Power Factor, 1 megacycle	0.012
Dielectric Constant, 1 megacycle	6.9
Loss factor, 1 megacycle	0.083
Dielectric Strength, volts/mil	400
Volume Resistivity, ohm-cm	5×10^{14}
Arc Resistance, seconds	250
Impact Strength, Izod, ft.-lb./in. of notch	0.6
Max. Safe Operating Temp., °C	350
°F	660
Water Absorption, % in 24 hours	nil
Coefficient of Linear Expansion, °C	12×10^{-6}
Tensile Strength, psi	6000
Compressive Strength, psi	28000

—a superior inorganic glass-bonded mica insulation featuring—

- LOW LOSS — HIGH EFFICIENCY
- HIGH DIELECTRIC STRENGTH
- IMPERVIOUS TO OIL OR WATER
- FREEDOM FROM CARBONIZATION
- PERMANENT DIMENSIONAL STABILITY
- WITHSTANDS HIGH TEMPERATURE
- IMMUNE TO HUMIDITY
- EXCELLENT MECHANICAL CHARACTERISTICS



AVAILABLE IN SHEETS, RODS, AND FABRICATED TO SPECIAL SHAPES

MYCALEX

LOW-LOSS MINIATURE, SUB-MINIATURE AND UHF TUNER

TUBE SOCKETS

These sockets cost no more than ordinary phenolic types yet electrical characteristics are far superior. Dimensional accuracy and uniformity are unexcelled.

PRODUCED IN TWO GRADES

MYCALEX 410 is approved fully as Grade L-4B under N.M.E.S. JAN-I-10 "Insulating Materials, Ceramic, Radio, Class L."

MYCALEX 410X is lower in cost but its insulating properties greatly exceed those of general purpose phenolics.

Both types available in 7- and 9-pin miniature, sub-miniature and UHF. All are injected molded to high precision.

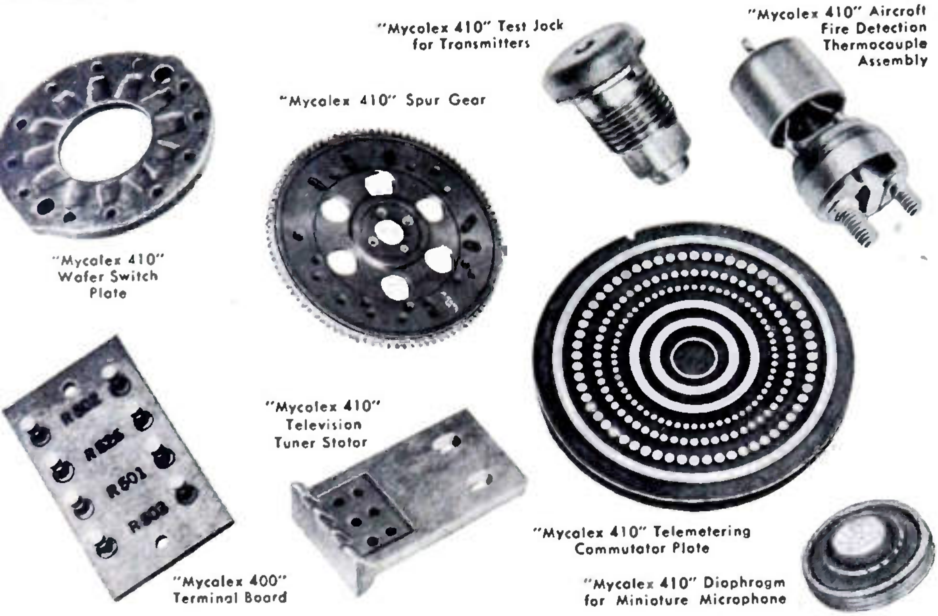
SINCE 1919 MYCALEX THE INSULATOR TRADE MARK REG. U.S. PAT. OFF.

MYCALEX CORPORATION OF AMERICA

or ALL frequencies MYCALEX

—economical, widely applicable machines and molds to close tolerances

Iycalex is a highly developed glass-bonded mica insulation backed by over a uarter-century of continued research and successful performance. In its present igh state of development, MYCALEX combines every important charactertic demanded of a modern, permanent, efficient electrical insulating material. IYCALEX is supplied in sheets or rods, can be injection or compression molded o close tolerance, is readily machineable, and can be fabricated in practically ny size or shape. Dielectric losses are very low from 60 cycles to 24 kilomegaycles per second, arc resistance, volume resistivity and other electrical characterstics meet the most rigorous requirements. Physical properties are equally dvantageous.



CAN BE TAPPED, DRILLED, THREADED GROUND, OR MOLDED WITH INSERTS

---MYCALEX---

GENERAL CATALOG and ENGINEERS' HANDBOOK

Your copy of the new up-to-the-minute MYCALEX CATALOG and ENGINEERS' HANDBOOK will be sent promptly on request. Tells the entire MYCALEX insulation story — Contains 20 pages of factual information including design, engineering, and practical manufacturing data — types of MYCALEX, characteristics, properties of dielectrics, injection molded grades, design of molded insulators, compression molded grades, design of machined insulators, etc.

Available on request, these data sheets cover the entire line of MYCALEX tube sockets—miniature, subminiature and UHF — available in all types of mountings: regular saddle for top or bottom mounting, snap saddle, four ground lug saddle and JAN types. MYCALEX tube sockets offer outstanding advantages over practically every other type, yet are priced competitively with the cheapest sockets available anywhere. Get the facts today!

FOR LITERATURE, INFORMATION OR QUOTATIONS— CALL, WRITE OR WIRE 190 CLIFTON BLVD., CLIFTON, N. J.

MACHINEABLE GRADES:

MYCALEX 400

MYCALEX 400 IS APPROVED FULLY AS GRADE L-4A UNDER NATIONAL MILITARY ESTABLISHMENT SPECIFICATION JAN-I-10 "INSULATING MATERIALS, CERAMIC, RADIO, CLASS L."

CHARACTERISTICS

Power Factor, 1 megacycle	0.0018
Dielectric Constant, 1 megacycle	7.4
Loss Factor, 1 megacycle	0.013
Dielectric Strength, volts/mil	500
Volume Resistivity, ohm-cm	2×10^{15}
Arc Resistance, seconds	300
Impact Strength, Izod, ft.-lb./in. of notch	1.85
Max. Safe Operating Temp., °C	370
°F	700
Water Absorption, % in 24 hours	nil
Coefficient of Linear Expansion, °C	10.2×10^{-6}
Tensile Strength, psi	6000
Compressive Strength, psi	35000

★ ★ ★

MYCALEX K-10

CHARACTERISTICS

Dielectric Constant, 1 megacycle	10.6
Q Factor, 1 megacycle	310
Loss factor, 1 megacycle	0.034
Volume Resistivity, ohm-cm	3.0×10^{14}
Dielectric Strength, volts/mil (0.10 in. thickness)	270
Fractional Decrease of Capacitance with Temperature Change	0.0056
Fractional Increase of Capacitance with Temperature Change	0.0076
Max. Safe Operation Temperature, °C	400

MYCALEX K EMBRACES AN ENTIRE SERIES OF CAPACITOR DIELECTRICS, EACH WITH SPECIFIC CHARACTERISTICS. THESE CAN BE SUPPLIED ON SPECIAL ORDER IN SHEETS 14" x 18" IN AREA AND FROM 1/8" TO 1" IN THICKNESS, ALSO AVAILABLE IN RODS. MYCALEX K CAN BE MACHINED, AND MYCALEX KM MOLDED TO CLOSE TOLERANCES.

SINCE 1919 MYCALEX THE INSULATOR
TRADE MARK REG. U.S. PAT. OFF.

Executive Offices: 30 Rockefeller Plaza, New York 20
Plant and General Offices: Clifton, New Jersey

VERSATILITY

ACTUAL SIZE

APPLICATIONS

in audio and radio fields up to 175 Mc

- Marine Transmitters
- Amateur Transmitters
- Mobile Transmitters
- Aircraft Transmitters
- Audio & Video Modulators
- Servo Amplifiers

The most widely accepted new beam power tubes in the field

Few tubes have won such wide popularity in so short a time as the RCA-developed 6146 and 6159 beam power tubes. Both types are extremely versatile as to application, and offer considerable latitude in circuit design because of their ability to operate efficiently over a wide range of plate voltages and frequencies. Low in cost, compact, and unusually rugged, these types are particularly suited to VHF transmitter designs for fixed and mobile services.

The RCA-6146, with a 6.3-volt heater, is designed for general service. The RCA-6159, with a 26.5-volt heater, is intended for use in VHF aircraft transmitters. In all other respects, the two tubes are identical.

Because of their high power sensitivity and low plate-voltage requirements, the RCA-6146 and 6159 permit compact, economical designs without a sacrifice in efficiency. A single tube as unmodulated Class C rf power amplifier will deliver an output of 69 watts up to 60 Mc with a plate supply of 600 volts . . . and 35 watts output at 175 Mc with a plate supply of 400 volts. A pair operated Class AB_1 will provide an audio output of 120 watts . . . in Class AB_2, 130 watts. A pair of either type triode connected will deliver an output of 19 watts with a total harmonic distortion of 4.6%.

For complete technical data on the RCA-6146 and RCA-6159, write RCA, Commercial Engineering, Section 47HR, Harrison, N. J., or your nearest Field Office.

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

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

RADIO CORPORATION of AMERICA
ELECTRON TUBES HARRISON, N. J.

TMK. ®

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

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

PROCEEDINGS OF THE I.R.E.®

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 40 — *August, 1952* — NUMBER 8

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

Reg. U. S. Pat. Off.

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

J. D. Ryder

Director, 1952-1954

J. D. Ryder was born in 1907 in Columbia, Ohio, and received the B.E.E. and M.S. degrees from the Ohio State University in 1928 and 1929, respectively. In 1944 he was awarded the Ph.D. degree in electrical engineering at Iowa State College.

Dr. Ryder was associated with the General Electric Company in Cleveland, Ohio, in 1929–1931, and then joined the research laboratory of the Bailey Meter Company, in Cleveland, as supervisor of the electrical and electronic section. As a result of his work, he was granted United States and foreign patents for such inventions as a form of electronic self-balanced ac resistance bridge, a high-speed photoelectrically-balanced potentiometer, ac and dc motor-control circuits, and smoke recording instruments.

Turning to the teaching field in 1941, Dr. Ryder was appointed assistant professor of electrical engineering at Iowa State College, Ames, Iowa, professor in 1944, and in 1946, he was in acting charge of the electrical engineering department. He became the assistant director of the Iowa Engineering Experiment Station in 1947, and in 1949, he was named head of the department of electrical engineering at the University of Illinois, Urbana, Ill.

Dr. Ryder is the author of 3 textbooks on electronics and networks, and numerous papers on technical and educational subjects. At present, he is directing research in high-frequency network analyzers and forms of computers.

As a member of the Board of Directors of the National Electronics Conference, Dr. Ryder was the program committee chairman of the 1951 Conference, and presently, is the executive vice-president. He is a member of the Illinois Engineering Council, a fellow of the American Institute of Electrical Engineers, and a member of the American Association for the Advancement of Science, and the American Society for Engineering Education, Eta Kappa Nu, and Sigma Xi.

Dr. Ryder joined the Institute of Radio Engineers in 1929 as an Associate, became a Senior Member in 1945, and in 1952, he was made an IRE Fellow. He has been the Vice Chairman of the IRE Electron Tube Conference Committee, is a member of the IRE Papers Review Committee, and Chairman of the Education Committee. In 1948, Dr. Ryder was the Vice Chairman of the IRE Des Moines-Ames Section. He is a member of the Institute's Policy Advisory Committee.

Let Us Re-Define Electronics

W. L. Everitt

The passing years modify the nature and scope of all human activities. This has been notably the case for the fields of activity of The Institute of Radio Engineers. There is presented below a keenly analytic consideration of this evolutionary trend in the communications and electronics field. It is a stimulating and thought-provoking treatment of a difficult subject, and includes as well an interesting and constructive semantic proposal.

The author of the following guest editorial is a Fellow and Past President of the IRE, and Dean of the College of Engineering of the University of Illinois. He is widely recognized as one of the leading communications and electronics engineers of this age, and is the author of certain outstanding technical texts.—*The Editor.*

All of us are continually asked, "What is electronics?" In defining a word, it must be remembered that language, like science and industry, is dynamic and constantly changing. As an example, consider the word "football." It started out denoting a game in which the manipulation of the ball was done largely with the feet. As any fan knows, the word now denotes a complicated sport involving strategy, mass movements in attack and defense, and the foot is in contact with the ball only occasionally. Similarly, the word "chivalry" is derived from the word "cheval," or horse. However, it brings to mind a whole code of conduct stemming from the day when knighthood was in flower and the knight was on a horse while the common man went afoot.

Understandably, in defining "electronics" the lexicographer thus far has been too greatly influenced by the etymology—by the meaning of the word "electron." But current definitions are not immutable. The lexicographer follows rather than determines common usage. In defining a word for a dictionary he records how the word is most commonly used by the expert. Therefore it is appropriate that the Institute of Radio Engineers should, from time to time, write its own definition of this word "electronics."

One definition has been, "Electronics is the science and technology of systems using devices in which electrons flow in a gas, no matter how dense or tenuous." This definition is too narrow in 1952. The field of electronics should be considered in terms of its broad concepts; it should not be limited by the devices which it employs. The electron tube has given its name to, but should not determine the boundaries of, electronics. DeForest with his grid introduced a method of continuous control of the flow of current in a circuit, made the amplifier possible, and opened the way to a whole new group of ideas. But control is only one of the many important contributions of electronics.

Fundamentally, electronics is interested primarily in extending man's senses in space, as by the radio, television, and radar; in acuity, as by the electron microscope; in visual or audible range, as by the infrared sniperscope and "ultrasound" detectors; and in speed, as by computers. It is also interested in supplementing man's brain, both by acting as a switchboard and by making comparisons and judgements, for example, servomechanisms and photoelectric inspection systems, and by solving mathematical problems.

Modern industry is founded on the use of mechanical and electrical power to replace man's muscles; electronics more and more provides the routine brain power and nervous system for its control. Man, too, has found need for the extension of his own brain power and sensory organs for his enjoyment and studies.

I would suggest therefore the definition:

> *"Electronics is the science and technology which deals primarily with the supplementing of man's senses and his brain power by devices which collect and process information, transmit it to the point needed, and there either control machines or present the processed information to human beings for their direct use."*

Such a definition will recognize the dynamic character of electronics and its broad scope. When a new device such as the transistor appears, it fits logically into the pattern, even though no electrons flow in an evacuated envelope. When the feedback principle is utilized in a hydraulic servomechanism, we find it also is part of the family.

At the same time, because electronics is so intimately interwoven with other sciences and techniques, we will always expect many byproducts such as the fluorescent light, the high power rectifier, and the atom smasher.

It is probable that the boundaries of this science can only be defined temporarily. Like other dynamic words, "electronics" will need to be re-defined as its expansion continues.

Scientific Manpower*

L. A. DUBRIDGE†

Summary—The most conservative figures on the shortage of engineers suggests that this shortage is now around 95,000 and will reach 156,000 by 1955. A less conservative view of the figures available suggests that this situation might be much worse. The number of new engineers now being produced each year may be actually less than the number lost to engineering activities through death, military service, and transfer to nonengineering duties. We might be 300,000 engineers short by 1955. Since it takes four years to train an engineer, all we can do during the next four years is to make better use of the engineering manpower which will be available. But high-school students are being discouraged from entering the fields of science and engineering by misleading statements of prominent people that science and engineering are the cause of the world's troubles. Engineers and scientists can do much to remove this misapprehension by pointing out that scientists and engineers also work for human welfare and that science and engineering are helping to solve the world's troubles rather than causing them. This must be done and additional scholarship funds be made available before the downward trend in engineering and science enrollments is reversed.

THERE IS no engineer in the country who has not read dozens of articles and speeches full of statistics on the engineering manpower problem. I apologize in advance for inflicting another one on you.

However, only a few of these articles and speeches go behind the statistics they present and attack the question of what these figures *mean*. I believe there is a deep meaning hidden behind them—one fraught with grave consequences for the future of this nation and the free world. It may be already too late to avoid some of them. But it is vitally important that we recognize them, that we identify the basic causes of the difficulty and try to tackle the stupendous task of remedying these causes.

First we must look at the figures themselves in order to appreciate the dimensions of the problem. And right at this point we run into a jungle of confusion. You can pick practically any figure you wish to represent the annual shortage of engineers, and you can find some authority to confirm your estimate. You may even take a *negative* figure and quote in support of it an article in *Life* magazine last year stating that "technicians are two bits a dozen in America." As I shall suggest later, thoughtless comments of this sort—often taken seriously by high-school students and their teachers—may be one of the causes of our present difficulty.

If we consult more authoritative sources, what do we find? Even here you can take your choice. For there is really no accurate way of adding up the present shortage and still less of projecting it into the future.

But if some rather conservative estimates are taken from such authorities as K. T. Compton, S. C. Hollister, and the Engineering Manpower Commission of the Engineers Joint Council, the following picture can be put together:

1. The present engineering population in this country is about 400,000, of which 300,000 are in industry, 90,000 in government agencies, and 10,000 in education.
2. The present *shortage* is about 95,000, i.e., there are 95,000 military and civilian jobs now vacant.
3. Between now and 1955 the country will need about 33,000 new engineers each year. Hence the accumulated need by 1955 will be for an engineering population of nearly 630,000.
4. The numbers of engineers we may actually have in 1955 can be more accurately predicted. It is the number we now have (400,000) plus the number now in engineering schools who will graduate by 1955. Making no allowance for losses in the meantime, this adds up to 474,000 against the predicted need of 630,000—156,000 short! A shortage increasing at the rate of some 16,000 per year.

These are the most generally accepted figures and are admittedly conservative.

The point I wish to make is this. The above figures of an annual shortage of 16,000 and an accumulated shortage by 1955 of 156,000 may be so conservative they distort the actual situation we face.

Taking a slightly less conservative view, one may arrive at the following picture:

1. If we allow for death, retirement, losses to nontechnical military service, and calling up of reserves and other diversions to nontechnical work, the present rate of supply of new engineers is actually 5,000 *less* than the expected annual losses.
2. The anticipated needs may have also been grossly underestimated. The technical requirements of the new 1.5 billion dollar a year program of military development (three times larger than 1950) have only begun to be felt. The military production program is rapidly climbing. The Atomic Energy Commission has been instructed to initiate a vast 5 billion dollar program of expansion. These national security programs alone could easily demand 30,000 more engineers a year for the next four years. Thus it could easily be true that by 1955 the number of engineers actually needed will be nearly 700,000. And we will actually have less than 400,000.

I do not claim, of course, that these figures are any more accurate than the previous ones. But we should not be blind to the possibility of a potential shortage of 300,000 or to the fact that we may actually be losing rather than gaining ground each year; otherwise, we may fail to understand the true dimensions of our problem. When we face a situation in which there are needs for nearly twice as many engineers (and I might add scientists too) as are available, we are meeting not merely a grave problem but something which more nearly approaches a national catastrophe For here we are, as a nation, spending 50 billion dollars or more a year to maintain world leadership in military, industrial, and other technology, yet we just do not have the basic wherewithal in trained manpower to do the job we are setting out to do. For the next four years there is practically nothing we can do about it. Furthermore, the causes for the situation lie so deep in American thought and practice that to do anything about it at all presents the gravest difficulties.

As to the first point—that there is nothing we can do about it—the reasons are fairly obvious. It takes four years to train an engineer. Even if the freshman classes next fall in all our engineering colleges were suddenly choked to overflowing, this would not make a dent on the situation until 1956.

There are, of course, a few things we might do to help. We could stop drafting engineers and engineering students. But the difficulties in national policy along that line are only too obvious. (The Armed Forces could, however, stop assigning engineers in uniform to nontechnical jobs.) We can recall to engineering work many fine engineers who have gone into executive, sales, or other nonengineering work. We could hurriedly give subprofessional training to a host of untrained youngsters to relieve the trained engineers of some of the drudgery of drafting and computation. We ought to pay really top grade engineers much higher salaries than they now receive. We ought to do all these things, and we no doubt will do some of them to a certain extent. But all together they will hardly make up a fifth of the projected shortage of 300,000.

So for the next four years we can adopt only certain palliative measures. And the results will be high wages offered to second grade men and—worst of all—second rate engineering jobs done on essential national products in places where only first-rate engineering should be accepted. We shall thereby be purchasing a colossal quantity of second-grade product at a high cost in maintenance and obsolescence—and at an incalculable cost to national security.

What about the long-term future?

This is a field in which one may speculate without restraint. There are those who say we should not worry. They say that high salaries will attract more men into engineering in the long run and things will be automatically adjusted. We can agree that higher salaries are desirable and will help. Also it may be that our high level of national expenditure will be cut back, business conditions may decline, and a reduced demand for engineers will match the then increasing supply. However, I hold a different view.

I believe that—except for occasional recessions—the need and the demand for scientists and engineers is going to continue to rise at an increasing rate for the indefinite future. I see no early easing of world tension; I see no decline in the opportunity for America to continue its world leadership in technology; I see *rising* and not falling opportunities for technological advance all over the world. The age of science and technology has only just begun. And while I am the last one to insist that all the world's problems can be solved by technology alone, I also believe technology can be an important weapon

* Decimal classification: R070. Original manuscript received by the Institute, May 15, 1952. Speech presented before the Los Angeles Section of the IRE.
† California Institute of Technology, Pasadena, Calif.

in advancing human welfare, i.e., in enabling human beings through social, economic, political and psychological advances to achieve the moral goals of human liberty, decency, and dignity to which all men aspire.

But I see serious obstacles in achieving the scientific and technological advances which are so essential to the progress of human welfare. The major obstacle is right here in this country where the intellectual atmosphere is such as to discourage rather than encourage men and women in entering the field.

This may sound like an extreme statement, but there are many facts to support it. Notice, for example, that in spite of increased population and in spite of increased demand we only had 3 per cent more freshman engineers enrolled in the nation's engineering colleges in the fall of 1951 than in 1940. It is true there was a big bulge in 1946–1949, but this hardly made up for the enormous war-time decrease.

Why are engineering enrollments decreasing?

Obviously, the number of college freshman engineers enrolled depends greatly on what is happening in the high schools. And what has been happening there in the past 50 years is not encouraging. In 1900 about 19 per cent of the nation's high-school students took a course in physics. Today only 5.5 per cent do. Many engineering schools very properly require high-school physics for entrance. Thus the number of available applicants is showing a relative decline. The situation is no better in mathematics, especially in advanced courses. Evidently high-school students, apparently encouraged by teachers, administrators, and counselors, now regard such courses as too tough or too "technical." Counselors advise students to take a "broad" course. This sounds fine. But what it usually means is taking literature or history or politics or "science survey" instead of mathematics and physics and chemistry. Just how poetry is more "broadening" than Newton's laws of motion however somehow escapes me.

High schools are not wholly to blame. There has been for many years a growing feeling among all Americans that the world has too much science and technology already. What we need, they say, is not better engineers but better citizens. Of course we need more and better citizens. But I deny the implication that a bank clerk or lingerie salesman or even a social psychologist is necessarily a better citizen than a scientist or engineer.

What after all is the prime prerequisite of a good citizen?

The first duty of a citizen is to perform a useful function for society. No matter how glibly a man is able to recite in proper order the names of all the presidents of the United States, if that man is not using his talents to the fullest possible extent in a constructive occupation I claim he is *not* being a good citizen. Whether his talents are for designing airplanes or writing good poetry, I don't much care, as long as those talents are being developed and being constructively used. I am not informed as to how many courses in history or political science or social problems were taken by Alexander Graham Bell, the Wright Brothers, or Lee De Forest. Whether they took any or not, they were great citizens because they made useful contributions to society. And when we let our promising high-school students believe you can't be both good engineer and good citizen, we are doing our nation and the world a disservice.

Now, I don't want to be misunderstood. I also believe that the engineer or scientist who does not take an interest in the welfare of his community or his country is failing to fulfill his full obligations as a citizen. And I believe that somewhere in his schooling this point ought to be made clear to him. Furthermore, I believe that an engineer is more likely to be a more useful citizen—and also a more useful engineer—if he devotes an appreciable part of his efforts in high school and college to the study of subjects outside of his specialty and also to constructive activities outside the classroom. But this is not the same as saying he should do these things to the exclusion of accomplishing his primary task—as a citizen and as a man—of developing his primary talents.

One of the reasons for this recent drift away from science and technology is exemplified by the following quotation from a prominent scholar whose name I will not mention: " . . . the root cause of our current world crisis stems from the fact that our technical scientists in their perfection of the techniques of destruction are so far ahead of our social scientists whose principal job it is to teach men how to live together peaceably and constructively."

I can think of only one respectable word to describe such a statement—"eyewash." Because scientists helped prevent the world from being conquered by two groups of power-mad dictators in Berlin and in Tokyo, *therefore* they must be responsible for the current world crisis! Because they helped crush two dictators, it is *their* fault that a third one has appeared! And who says that it is only the social scientist whose aim it is to teach men to live together peaceably and constructively? That is a job for all of us. The ills of the world are not caused by the intelligence of the scientists but by general human cussedness. Not even social scientists are going to find a quick cure for *that*.

I won't bore you with further expostulations about such nonsensical statements—for you can expose them as nonsense as well as I. My point is that more and more people are believing that because science and technology is not *everything*, therefore it is *nothing;* that the way to advance social progress is to *stop* scientific progress; that because man can not live by bread alone, therefore bread is unnecessary and undesirable. When more and more people try to represent scientists and engineers as the villains of the world drama, it is not hard to understand why fewer and fewer youngsters are attracted into these fields.

I think then that you can see why I regard the present manpower situation in science and engineering as something more than a troublesome but unimportant phenomenon—why indeed I assert that it represents something more nearly in the nature of a national crisis.

The numerical shortage itself is serious enough. Everyone would recognize the absurdity of placing orders for 20,000 airplanes to be delivered this year if the proved total supply of aluminum was only enough to build 10,000. But we apparently think nothing of embarking on a program which would require 700,000 engineers when only 400,000 are in sight. The result is plain. We are going to fall flat on our faces. We simply can't deliver what we have ordered.

But the numerical shortage is only a symptom of a deeper ailment. We as a nation have grown dependent on scientists and engineers, and we don't know it and refuse to admit it. And so with one hand we appropriate billions of dollars for work that only scientists and engineers can do—and with the other hand we slap them in the face and accuse them of causing all the world's ills which we then call on them to help cure. As a symptom of all this, the House of Representatives has slashed the budget of National Science Foundation by 77 per cent, one agency of government set up to produce more scientists and engineers and to produce rather than consume basic knowledge.

Are we then wholly helpless to do anything about this crisis? I have already said that there is very little we can do for the short term. But for the long term we can do something. I suggest three things:

1. We can expose this nonsense about technology being the cause of the world's ills, about scientists being unconcerned about human welfare. We can let it be known that human welfare is the major goal for all of us and that we as scientists and engineers stand ready to join hands with all men of good will everywhere to advance that goal. And we have been *doing* it!
2. We can carry this same message to high-school students. You—the members of IRE—could initiate a high-school information campaign to tell high-school students and teachers that engineers are not villains; that the field of science and engineering offers great and exciting challenges for the future; that scientists and engineers can be—and for the most part *are*—good citizens, too. You can tell them that the best citizen is the useful citizen—the one who is using his talents to their fullest. You can invite students to visit your plants, factories, and laboratories and show them how exciting science and technology can be.
3. Finally, I would like to suggest something very definite which you—the members of IRE—can do right here in Southern California. Let us say that Southern California industry is going to need 100 engineers more each year than are now in sight. (I'll choose a modest number to avoid scaring you!) Why shouldn't IRE and the other engineering societies get together and raise, by industrial contributions, a scholarship fund so that each year 100 boys who need financial incentive could be sent to engineering school. For $200,000 a year you could offer 100 four-year scholarships, averaging $2,000 each—$500 a year—to the 100 most promising applicants. And my guess is that for each winner about 3 to 5 others would have their interest sufficiently aroused by the contest so that they would find other sources of funds and go to college anyway. If we in Southern California started such an enterprise, it might be copied in other areas. Properly promoted, such scholarship funds might help reverse the tide of declining interest in science and engineering, would make the voice of scientist and engineer heard again, and eventually help avert a real national calamity.

Experimental Verification of the Theory of Laminated Conductors*

H. S. BLACK†, FELLOW, IRE, C. O. MALLINCKRODT‡, SENIOR MEMBER, IRE, AND S. P. MORGAN, JR.†

Summary—Clogston[1] has discovered that if a conductor is properly laminated, there exists a particular phase velocity along the conductor for maximum penetration of the fields and minimum loss due to skin effect. An experimental coaxial line was constructed whose center conductor was laminated and whose phase velocity could be varied by changing the dielectric constant of the main dielectric. As predicted by theory, the measured attenuation was critically dependent upon phase velocity. With optimum phase velocity the attenuation, though greater than predicted by theory, was less than that of a conventional coaxial cable of the same dimensions and same main dielectric. A theoretical analysis of the experimental laminated conductor is described in an appendix.

I. Theory

In a recent theoretical paper[1] Clogston presents a discovery which offers a solution to one of the oldest problems of electrical engineering, namely, the problem of causing electromagnetic fields to penetrate deeply into a conductor rather than to travel only in a thin "skin" at the surface. This desirable result is achieved by insuring: first, that the conductor is so laminated that its many insulated layers of conducting material are parallel to the direction of current flow; second, that the thickness of each conducting layer is sufficiently small compared to a "classical skin" at the frequency under consideration; and, third that the phase velocity of the electromagnetic wave along the conductor is close to a certain critical value. Under these conditions, the fields will penetrate so far into the laminated conductor as to include a thickness of conducting material many classical skins deep. Laminated transmission lines which satisfy these conditions imply broad-band transmission with lower losses, or a wider frequency band, or flatter transmission or combinations of these.

II. Verification of the Theory

Verification of the theory has recently been accomplished by an experiment which, though not involving a low-loss line of optimum design, has brought out strikingly the dependence of field penetration on phase velocity. A short length of laminated conductor was made by hand, using metallic and dielectric layers of commercially available thinness. This laminated conductor was utilized as the center conductor of a coaxial quarter-wave resonator. The outer conductor was a large copper tube so that most of the loss was due to the laminated inner conductor. By adding cylindrical tubes and disks of titanium dioxide, the average relative dielectric constant of the main dielectric could be varied from unity to 4.55. At the first resonance frequency, the attenuation was measured as a function of the measured dielectric constant of the main dielectric.

To carry out these measurements, an oscillator and a detector were loosely coupled by small probes to the resonator, which was 94.1 inches long. Except for unimportant corrections, the relative dielectric constant of the main dielectric, ϵ_{1r}, is related to the observed resonance frequency, f, in megacycles by $\epsilon_{1r} = [31.36/f]^2$. The attenuation of the resonator is related to its sharpness of resonance, which is indicated by the change in detector input signal from s to $(s-\Delta s)$ as the frequency is changed from f to $(f \pm \Delta f)$. This relationship is closely approximated by equating the cable attenuation in decibels per 94.1 inches to

$$13.64 \frac{\Delta f}{f}\left[\left(\frac{s}{s-\Delta s}\right)^2 - 1\right]^{-1/2}.$$

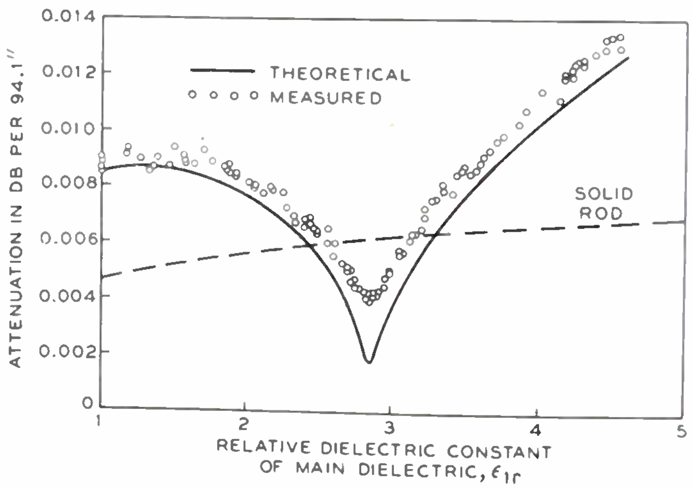


Fig. 1—Attenuation at resonance for a quarter-wavelength coaxial resonator 94.1 inches long with laminated center conductor, as a function of the relative dielectric constant of the main dielectric.

Fig. 1 displays a comparison between measured and calculated values. For purposes of calculation we considered an idealized coaxial line with homogeneous main dielectric, the idealization being regarded as well justified since the titanium dioxide tubes and disks in the experimental structure were arranged to simulate a homogeneous dielectric as closely as possible. The mathematical analysis underlying the calculations was an

* Decimal classification: R282.1. Original manuscript received October 10, 1951; revised manuscript received January 16, 1952.
† Bell Telephone Laboratories, Inc., Murray Hill, N. J.
‡ Formerly Bell Telephone Laboratories, Inc., Murray Hill, N. J.; now, Hughes Aircraft Co., Culver City, Calif.
[1] A. M. Clogston, "Reduction of skin-effect losses by the use of laminated conductors," Proc. I.R.E., vol. 39, pp. 767–782; July, 1951; also *Bell Sys. Tech. Jour.*, vol. 30, pp. 491–529; July, 1951.

extension to coaxial cables of Clogston's original work on parallel-plane transmission lines. Since the analysis is of practical engineering interest in itself, the major details are given in Appendix II, although it is expected that later papers will treat the theory of laminated coaxial cables much more fully.

In accordance with theoretical predictions, attenuations (Fig. 1) were measured which are lower than would be obtained with a solid rod of the same size surrounded by the same main dielectric and same outside conductor. The attenuation displays the expected critical dependence upon the dielectric constant of the main dielectric. The attenuation conforms to the theoretical pattern, exhibiting both the expected minimum at the Clogston value, where the resonance frequency is 18.58 mc, and the broad maximum predicted with light dielectric loading. One notes that if the dielectric constant of the main dielectric equals that of the insulating layers of the laminated conductor ($\epsilon_{1r}=2.4$ in Fig. 1), the calculated attenuation differs little from that of a solid conductor, that is, differs little from the attenuation of a conventional coaxial cable in which electrically thick conductors of the same material bound the same dielectric. That these two attenuations should be equal provided the laminated conductor is several "equivalent skins" deep has been observed by Clogston.[2]

Differences between measured and calculated attenuations (Fig. 1) can be partially accounted for. Errors of measurement are believed not to exceed ± 5 per cent. The effect of a small gap (Appendix I) in the longitudinal seam of each conductor was not included in the calculations. This effect is estimated to increase all calculated attenuations by about 5 per cent. Thus, except in the neighborhood of the critical value of ϵ_{1r}, agreement between calculation and experiment is satisfactory. In the region of least loss, it is conjectured that the unexplained extra loss is, for the most part, due to lack of uniformity in the hand-made laminated conductor and that the primary causes of this nonuniformity are irregularities in the surfaces of the conducting and insulating layers, variations in their thicknesses, and small variations in the length of the conducting layers at the high-impedance end of the resonator which, in turn, may lead to a slight amount of mode conversion.

The reduction in attenuation obtained with this particular conductor is not of course indicative of the potential improvement offered by laminations of optimum proportions. In practical applications of laminated conductors to transmission lines, it would be advantageous to laminate the outer as well as the inner conductor. Moreover, accurate machine-made laminations consisting of many more and much thinner layers of conducting material and also comparably thin insulating layers will be required in order to obtain the large factors of improvement which are theoretically possible.

[2] *Ibid.*, PROC. I.R.E., eq. (95); also *Bell Sys. Tech. Jour.* (III-57).

APPENDIX I

DESCRIPTION OF EXPERIMENTAL CABLE

An inner conductor 94.1 inches long was built up on a 0.146-inch diameter lucite core under tension, and consisted of 50 layers of copper foil 1.97×10^{-4} inches thick whose resistivity was 1.962×10^{-8} ohm meters at 25°C, alternated with 50 layers of insulation 1.063×10^{-3} inches thick whose relative dielectric constant was 2.4 and loss tangent 10^{-3}. Foils were longitudinally wrapped, with an average gap of 0.03 inches, successive gaps being staggered. The outside radius of the laminated conductor was 0.136 inches. The outer conductor was a thick copper tube 100.8 inches long of internal radius 2.0 inches, resistivity of the copper being 1.776×10^{-8} ohm meters at 25°C. To provide access to its interior, the outer conductor was cut in half longitudinally and hinged.

Cylindrical tubes and disks of titanium dioxide, so split as to permit easy addition and removal, afforded coarse and fine control of ϵ_{1r}. The relative dielectric constant of titanium dioxide is 99 and its loss tangent 6×10^{-4}. However, the effective loss tangent of the main dielectric was much smaller because most of the energy storage was in an air gap surrounding the inner conductor. The fact that this loss was negligible was established by testing the identical titanium dioxide parts in conjunction with a solid copper inner conductor in a like resonator whose losses were exactly predictable.

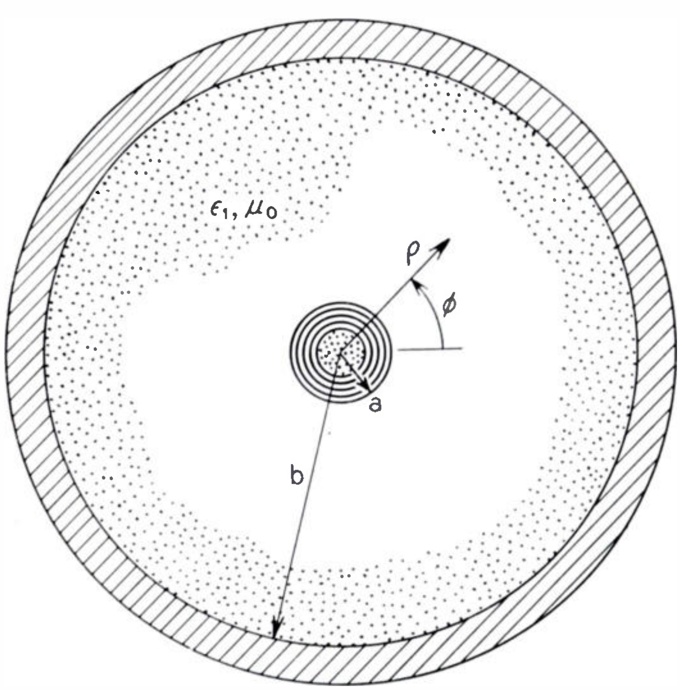


Fig. 2—Schematic cross section of idealized coaxial line with laminated center conductor.

APPENDIX II

MATHEMATICAL ANALYSIS

The theoretical attenuation plotted in Fig. 1 is the attenuation calculated for an idealized coaxial line shown schematically in Fig. 2 and equivalent to the experimental line described in Appendix I. In the idealized line a homogeneous main dielectric of dielectric constant ϵ_1 is bounded internally by a laminated center conductor of radius a, and externally by an electrically thick solid copper conductor of inner radius b. The permeabilities of all materials are taken equal to the permeability μ_0 of free space. MKS units are employed throughout. A cylindrical co-ordinate system (ρ, ϕ, z) has been introduced as shown, the z-axis coinciding with the axis of the cable. In this system, the only nonvanishing field components are E_ρ, E_z, and H_ϕ.

The attenuation constant γ of the line is given to a very good approximation by[3]

$$\gamma = \mathrm{Re}\,\frac{Z_a/a + Z_b/b}{2\sqrt{\mu_0/\epsilon_1}\,\log_e b/a}, \tag{1}$$

where Z_a and Z_b are the surface impedances of the inner and outer conductors respectively, defined by the field ratios

$$(E_z/H_\phi)_{\rho=a} = Z_a, \qquad -(E_z/H_\phi)_{\rho=b} = Z_b. \tag{2}$$

The surface impedance of an electrically thick solid metal tube of conductivity σ, whose radius is large compared to the classical skin thickness δ, is just

$$Z = (1+i)/\sigma\delta = (1+i)\sqrt{\omega\mu_0/2\sigma} \tag{3}$$

at the angular frequency ω; the surface impedance of the laminated center conductor must, however, be calculated at greater length.

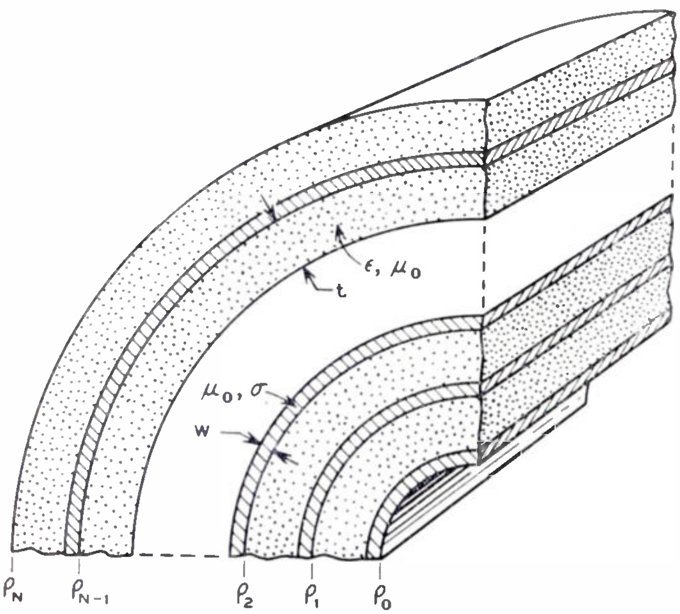


Fig. 3—Portion of laminated center conductor.

An enlarged view of part of the laminated stack is shown in Fig. 3. It consists of alternate conducting layers of thickness W and insulating layers of thickness t laid on an insulating core; a conducting layer has been drawn nearest the core, though this arrangement is not critical. The conductivity of the conducting layers is called σ (though it does not have to be equal to the conductivity of the outer conductor), and the dielectric constant of the insulation is ϵ. The radius of the insulating core is ρ_0, and the outer radius of the nth layer of insulation ($n = 1, 2, 3, \cdots$) is ρ_n.

If we assume that the fields in the coaxial line vary according to the propagation factor $e^{-ikz+i\omega t}$ and suppress this factor, the relevant Maxwell equations[4] can be reduced, in the conducting layers, to

$$\begin{aligned} E_\rho &= \frac{ik}{\sigma} H_\phi, \\ \frac{\partial}{\partial\rho}(\rho H_\phi) &= \sigma\rho E_z, \\ \frac{\partial E_z}{\partial\rho} &= \left(\frac{k^2}{\sigma} + i\omega\mu_0\right) H_\phi \approx i\omega\mu_0 H_\phi, \end{aligned} \tag{4}$$

if we neglect displacement currents and observe that in the present case $\omega\mu_0\sigma \gg k^2$. Similarly, in the insulating layers,

$$\begin{aligned} E_\rho &= \frac{k}{\omega\epsilon} H_\phi, \\ \frac{\partial}{\partial\rho}(\rho H_\phi) &= i\omega\epsilon\rho H_z, \\ \frac{\partial E_z}{\partial\rho} &= \left(\frac{k^2}{i\omega\epsilon} + i\omega\mu_0\right) H_\phi. \end{aligned} \tag{5}$$

If we now make the identifications

$$E_z \to \hat{V}, \qquad -\rho H_\phi \to \hat{I}, \tag{6}$$

and introduce Clogston's symbols,

$$\eta = \sqrt{i\omega\mu_0\sigma}, \qquad \xi = \sqrt{k^2 - \omega^2\mu_0\epsilon}, \tag{7}$$

we have in the conducting layers, from (4),

$$\begin{aligned} \partial\hat{I}/\partial\rho &= -\sigma\rho\hat{V}, \\ \partial\hat{V}/\partial\rho &= -(\eta^2/\sigma\rho)\hat{I}, \end{aligned} \tag{8}$$

which are formally identical to the equations of a nonuniform radial transmission line having series impedance $\eta^2/\sigma\rho$ and shunt admittance $\sigma\rho$ per unit length. Similarly, in the insulating layers we have, from (5),

$$\begin{aligned} \partial\hat{I}/\partial\rho &= -i\omega\epsilon\rho\hat{V}, \\ \partial\hat{V}/\partial\rho &= -(\xi^2/i\omega\epsilon\rho)\hat{I}, \end{aligned} \tag{9}$$

corresponding to a series impedance $\xi^2/i\omega\epsilon\rho$ and a shunt admittance $i\omega\epsilon\rho$ per unit length. If each layer be regarded as a four-terminal network consisting of a section of radial transmission line, the problem reduces to the calculation of the ratio $\hat{V}/\hat{I}$ at one end of a chain of such networks when its value at the other end is known.

[3] *Ibid.*, Proc. I.R.E., eq. (84); also, *Bell Sys. Tech. Jour.* (III-46). In this equation $C = 2\pi a$, $Z = \sqrt{\mu_0/E_1}(2\pi)^{-1}\log_e b/a$, and only the losses in the inner conductor are considered.

[4] See, for example, S. A. Schelkunoff, "Electromagnetic Waves," D. van Nostrand Co., Inc., New York, N. Y., p. 95, eqs. (12-9); 1943.

The radial line equations can be solved exactly in terms of Bessel functions; but when the thickness of each layer is very small compared to its radius, it is much simpler to replace the layer by a section of *uniform* line whose constants are determined by setting ρ equal to its mean value $\bar{\rho}$ for the given layer. Thus if we define

$$\bar{\rho}_{1n} = \rho_0 + (n-1)(t+W) + \tfrac{1}{2}W, \qquad \bar{\rho}_{2n} = \rho_0 + n(t+W) - \tfrac{1}{2}t, \tag{10}$$

we find that to a very good approximation the matrix relation between $\hat{V}$ and $\hat{I}$ at ρ_{n-1} and ρ_n is[5]

$$\begin{pmatrix} \hat{V}_n \\ \hat{I}_n \end{pmatrix} = \begin{pmatrix} \mathrm{ch}\,\xi t & -\dfrac{\xi}{i\omega\epsilon\bar{\rho}_{2n}}\,\mathrm{sh}\,\xi t \\ -\dfrac{i\omega\epsilon\bar{\rho}_{2n}}{\xi}\,\mathrm{sh}\,\xi t & \mathrm{ch}\,\xi t \end{pmatrix} \begin{pmatrix} \mathrm{ch}\,\eta W & -\dfrac{\eta}{\sigma\bar{\rho}_{1n}}\,\mathrm{sh}\,\eta W \\ -\dfrac{\sigma\bar{\rho}_{1n}}{\eta}\,\mathrm{sh}\,\eta W & \mathrm{ch}\,\eta W \end{pmatrix} \begin{pmatrix} \hat{V}_{n-1} \\ \hat{I}_{n-1} \end{pmatrix}, \tag{11}$$

where sh and ch represent the hyperbolic sine and cosine, respectively.

In the calculations we suppose that the propagation constant k is essentially determined by the properties of the main dielectric; that is,

$$k \approx \omega\sqrt{\mu_0\epsilon_1}, \tag{12}$$

and so from (7),

$$\xi \approx \omega\sqrt{\mu_0(\epsilon_1 - \epsilon)}. \tag{13}$$

If we put $\hat{I}_0 = 0$, $\hat{V}_0 = \text{constant}$, corresponding to an infinite-impedance dielectric core, then knowing the frequency and the electrical constants and the geometry of the stack we can calculate the ratio $\hat{V}_n/\hat{I}_n$ for successive values of n. If the total number of double layers in the stack is N, then the surface impedance Z_a defined by (2) is

$$Z_a = -a\hat{V}_N/\hat{I}_N. \tag{14}$$

The method just outlined, though accurate, requires high-speed computing machinery if there are many layers on the conductor. A much simpler procedure may be used if the conducting layers are so thick or the mismatch of the main dielectric so great that the effective skin depth[6] of the coaxial stack is small compared to its outer radius a. Then the mean radius of each layer may be taken equal to a and the problem is equivalent to calculating the surface impedance of a stack of parallel plane layers. For a plane stack several effective skin depths thick an expression for the surface impedance has already been given by Clogston.

[5] For the case of plane layers, compare *op. cit.*, Proc. I.R.E., eqs. (15) and (48); also *Bell Sys. Tech. Jour.* (III-7) and (III-10).
[6] *Ibid.*, Proc. I.R.E., eq. (80); also, *Bell Sys. Tech. Jour.* (III-12).

The numerical values used in the calculations were those given in Appendix I, except for the insignificant difference that we took $W = 2.00 \times 10^{-4}$ inches instead of the experimental value 1.97×10^{-4} inches, and $t = 1.060 \times 10^{-3}$ inches instead of 1.063×10^{-3} inches. The relative dielectric constant ϵ_{1r} of the main dielectric (assumed lossless) was varied from 1.0 to 4.6, the Clogston[7] value being

$$\epsilon_{1r} = (1 + W/t)\epsilon_r' = 2.85, \tag{15}$$

where ϵ_r' is the real part of the relative dielectric constant of the insulation. The resonance frequency f was calculated as a function of ϵ_{1r} from

$$f = 31.36/\sqrt{\epsilon_{1r}} \text{ mc per sec.} \tag{16}$$

For the Clogston value, $\epsilon_{1r} = 2.85$, the surface impedance of the laminated conductor was calculated by essentially the exact method described above, the hundred steps being performed on the Bell Laboratories Model VI Relay Computer. The effective skin depth for $\epsilon_{1r} = 2.85$ was 19.2 mils, according to the exact calculation. For other values of ϵ_{1r}, the parallel-plane approximation was used, since the effective skin depth for the nearest off-match value ($\epsilon_{1r} = 3.0$) was found to be only 6.8 mils. The use of the parallel-plane approximation for off-match calculations leads to slightly low values of surface resistance, but the error is not likely to be more than about 3 per cent in the worst case ($\epsilon_{1r} = 3.0$), and decreases rapidly as ϵ_{1r} departs further from the Clogston value.

The surface resistance of the outer conductor is easily obtained from (3) for the resonance frequency corresponding to any ϵ_{1r}, and then the total attenuation constant in nepers per meter at this frequency is given by (1). For plotting, the attenuation constant has been expressed in decibels per 94.1 inches, and the final results are shown in Fig. 1.

Acknowledgment

Messrs. C. E. Scheideler and W. F. Wolfertz constructed the hand-made laminated conductor and Mr. Wolfertz made most of the measurements. Many members of the Chemical Department of these Laboratories contributed to the solution of the difficult fabrication problems involved in laminating the conductor and producing a suitable variable dielectric.

[7] *Ibid.*, Proc. I.R.E., eq. (90); also, *Bell Sys. Tech. Jour.* (III-52). The surface impedance looking *into* the stack would be the negative of Clogston's R.

Single-Crystal Germanium*

G. K. TEAL†, M. SPARKS†, AND E. BUEHLER†

Summary—Significant advances have been made in the development of new types of transistors, photocells, and rectifiers and in the improvement of the reproducibility and reliability of the point-contact transistor. A key factor in this development has been the use of single-crystal germanium having a high degree of lattice perfection and compositional control.

Of particular interest to the device-development engineer is the fact that the rectifying barriers between the *p*-type and *n*-type sections behave in a manner predictable from the measured properties of each section. The exceptionally long lifetime of injected carriers observed in the material and the high degree of control over its chemical composition make it ideally suitable for the production of *p-n* structures. The ranges of properties of germanium single crystals which are now realizable are given, as well as their present degree of control.

SINCE THE ANNOUNCEMENT of the discovery of the transistor in 1948,[1] its development has produced marked advances. Working models of a new type, the *n-p-n* transistor, have been made and found to have exceptional properties, particularly with respect to high gain, low noise figure, and spectacularly low power operation.[2,3] Greatly improved reliability and reproducibility have been attained for the original point-contact structure.[4] These improvements result from a number of factors but primarily from (1) a better understanding of the phenomenon of transistor action and of the semiconducting properties of germanium, (2) better processes for the fabrication of the units, and, a key item, (3) better germanium crystals. This paper will discuss the single-crystal germanium now being employed for transistors and for the parallel development of a variety of rectifier and photocell devices.[5]

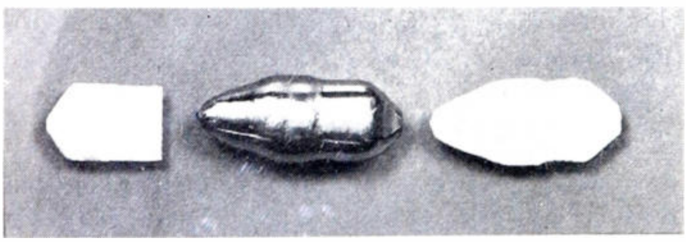


Fig. 1—Polycrystalline germanium.

A typical ingot of the polycrystalline germanium previously used for diodes and transistors is shown at the left in Fig. 1. The grain boundaries produce uncontrolled electrical variations that disturb the purposeful flow of current through the material. They, and other imperfections, such as vacant lattice sites and mechanical strains within the crystal, limit the properties of the ultimate devices. The single-crystal germanium discussed here has a high degree of lattice perfection, which is favorable to the unimpeded flow of carriers (holes and electrons) in the solid in much the same manner that a high vacuum facilitates and simplifies the flow of electrons in an electronic tube. In addition to improvement of lattice perfection in the crystals, *p*-type and *n*-type conductivity regions have been produced within a single crystal. This is represented schematically by Fig. 2 in which *p*-type and *n*-type regions are separated by a *p-n* barrier.

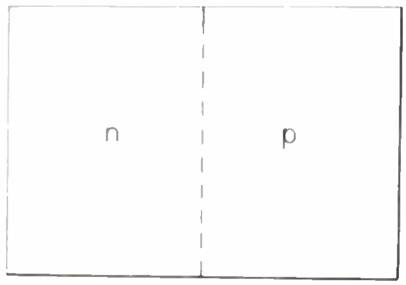


Fig. 2—*N*-type and *p* type regions in single-crystal germanium.

In the crystal-growing process the germanium is melted in a crucible by suitable heaters and a small germanium single-crystal seed is dipped into the melt. The heat input and losses are then adjusted to facilitate withdrawal of the desired size of crystal at the desired rate of growth. Germanium single crystals of a variety of shapes and sizes have been grown. The diameters have been varied from about 0.025 inch to about 1.5 inches. The maximum diameter is limited primarily by the diameter of the crucible and the size of the charge. The maximum length is similarly an arbitrary function of apparatus design and is about 8 inches in the work reported here. The center of Fig. 1 is a photograph of a typical product. It is $2\frac{1}{4}$ inches long and is roughly square around the growing direction. The right of Fig. 1 shows a photograph of a longitudinal cross section of such a crystal, the surface having been treated in the same way as the polycrystalline sample to bring out grain boundaries. In this case, however, the absence of grain boundaries is clearly evident.

When germanium solidifies, it is well known[6] that segregation of the impurities occurs so that most impurities concentrate in the liquid phase. Thus the growing crystal solidifies from a melt of continuously increasing impurity concentration, and shows a corresponding continuous decrease in resistivity. In Fig. 3 resistivity along the growth direction is shown for one type of ger-

* Decimal classification: R282.12. Original manuscript received by the Institute, February 28, 1952.
† Bell Telephone Laboratories, Inc., Murray Hill, N. J.
[1] J. Bardeen and W. H. Brattain, "Physical Principles Involved in Transistor Action," *Phys. Rev.* vol. 75, pp. 1208–1225; 1949.
[2] W. Shockley, M. Sparks, and G. K. Teal, "*p-n* Junction Transistors," *Phys. Rev.* vol. 83, pp. 151–162; 1951.
[3] R. L. Wallace and W. J. Pietenpol, "Some Circuit Properties and Applications of *n-p-n* Transistors," *Bell Sys. Tech. Jour.* Vol. 30, pp. 530–563, July, 1951; and PROC. I.R.E. Vol. 39, pp. 753–767; July, 1951. This article deals with the circuit properties and applications of *n-p-n* transistors.
[4] J. A. Morton, "Present Status of Transistor Development," *Bell Sys. Tech. Jour.*, vol. 31, pp. 411–442; May, 1952.
[5] W. J. Pietenpol, "*p-n* Junction Rectifier and Photocell," *Phys. Rev.*, vol. 82, p 120; 1951.
[6] G. L. Pearson, J. D. Struthers, and H. C. Theuerer, *Phys. Rev.*, vol. 77, p. 809 1950.

manium single crystal. The resistivity initially is 18-20 ohm-cm and decreases smoothly along the crystal to about 1 ohm-cm near the other end of the crystal. A similar plot for another type of single crystal is shown in Fig. 4. In this case the resistivity has been controlled

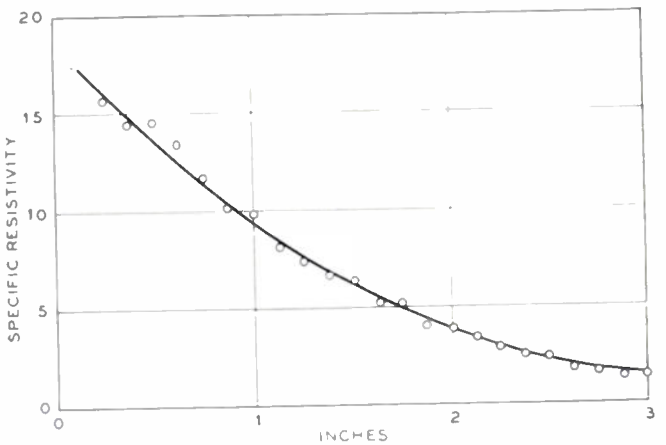


Fig. 3—Single-crystal germanium with resistivity changing in a given direction.

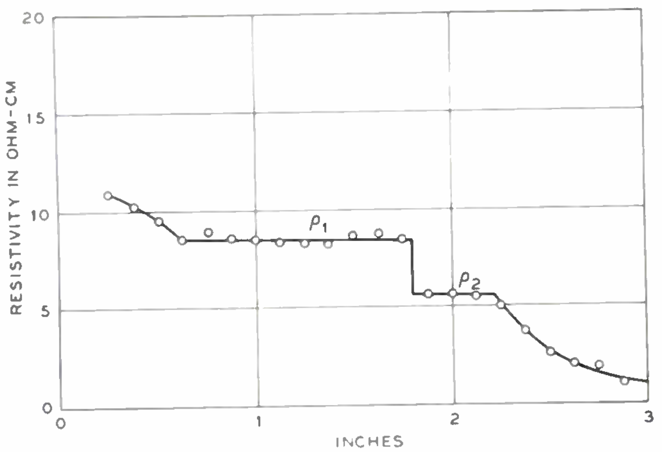


Fig. 4—Single-crystal germanium with adjacent regions of constant resistivity.

so that two adjacent constant resistivity regions are produced. For this example $\rho_1 = 8$ ohm-cm and $\rho_2 = 6$ ohm-cm. Other types of crystals have been grown in which constant resistivity regions have been produced at other levels. The order of constancy that has been obtained in slices of germanium cut perpendicular to the crystal length is indicated in Fig. 5. Results on four

CRYSTAL NO.	TYPE	ρ OF SLICE	RANGE OF ρ OVER CENTRAL 80% OF SLICE
1	n	12.8 10.3 3.56 1.15	±4% (OVER 1CM²) ±5% ±<1% (OVER 1CM²) ±<3% " "
2	n	19.4 13.0	±1% ±5%
3	p	37 30	±5% ±3%
4	p	6.8 4.2	±5% ±5%

Fig. 5—Constancy of resistivity in slices of single-crystal germanium.

typical crystals, two *n*-type and two *p*-type, are given. The smallest variation of specific resistivity is seen to be less than ±1 per cent and the largest is ±5 per cent. This represents a very great improvement in control of this important property.

Resistivity along the direction of growth for a crystal containing both *p*-type and *n*-type sections is shown in Fig. 6. Measurements were made with a two-point probe which was moved along a small, sandblasted germanium bar through which a constant current was passed. This type of measurement gives accurate results in a region of single-conductivity type, and is indicative of the barrier width at a junction. When the point probes are on opposite sides of the *p-n* junction, a part of the voltage drop between them occurs at the rectifying barrier, and thus gives an excessive value when interpreted in terms of resistivity. The probes used in the measurements shown here were 0.010 inch apart. The resistivities on the two sides of the junction may be independently controlled, and they usually lie in the approximate range 20 ohm-cm to 0.001 ohm-cm. The transition from *n*-type to *p*-type may be abrupt or gradual, as shown in the two curves of Fig. 6. A measure of the thickness of the transi-

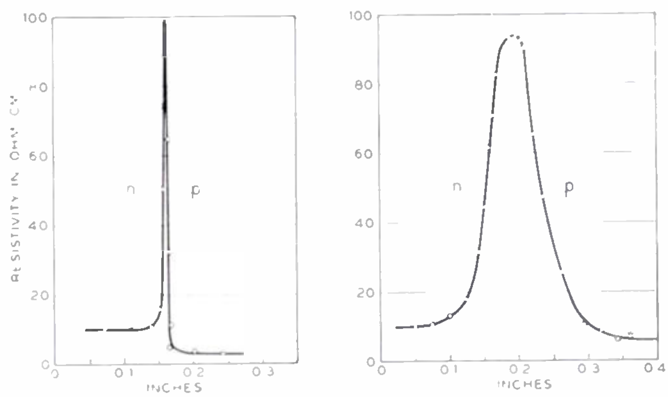


Fig. 6—Resistivity versus distance in *np* structures in single-crystal germanium (double-point measurements).

tion region may be obtained from the capacitance of the *p-n* barrier as a function of applied voltage, or by observing the voltage at which the Zener effect begins.[7] The Zener voltage is that necessary to produce the critical potential gradient in the *p-n* barrier just sufficient to extract electrons from the germanium-germanium bonds, and thus is the maximum voltage which the rectifying junction can support in the blocking direction without a large increase in the current. Zener voltages in germanium have been observed from 8 volts to 2,500 volts.

An important feature of the single-crystal germanium is the long lifetime of injected minority carriers.[8] This quantity is the average time which an injected carrier exists in the crystal before it disappears by combining with a majority carrier. The method of measurement of lifetime employed is one developed by Morton and Haynes, and is one often used in these Laboratories

[7] K. B. McAfee, E. J. Ryder, W. Shockley, and M. Sparks, "Observations of Zener Current in Germanium *p-n* Junctions," *Phys. Rev.*, vol. 83, p. 650; 1951.

[8] The minority carrier in *n*-type material is the hole and in *p*-type material is the electron.

since it involves relatively little cutting of the crystal.[9] It consists of scanning a small area of the surface of a massive piece of a single crystal with a narrow band of chopped light and of measuring the alternating current produced at a metal point at a fixed place in the scanned area of the surface. The point is held at a fixed potential in the blocking direction, and functions like a transistor collector point. The structure is in essence a phototransistor, and the variation of signal with distance between the incident light and the collector point may be interpreted in terms of lifetime. The rate of recombination of holes and electrons (related inversely to lifetime) is greater at the surface than in the interior of the crystal. A special treatment is used to reduce the surface recombination of the sample for the lifetime measurement.[10] The lifetime measurement is repeated at intervals along the crystal axis.

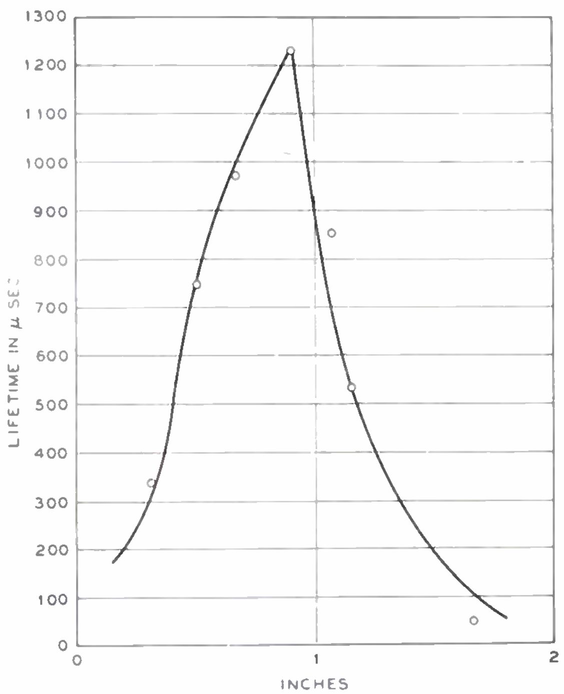


Fig. 7—Carrier lifetime in n-type germanium single crystal.

Two typical curves showing the variation of lifetime along a single crystal are given in Figs. 7 and 8. Fig. 7 shows the results for an n-type crystal in which the hole lifetime as a function of crystal length passes through a maximum of about 1,200 μsec with values of about 100 to 200 μsec near the two crystal ends. These values are in marked contrast to the usually low values of about 5 μsec or less observed in polycrystalline germanium. Fig. 8 shows the lifetime values for a crystal in which there is a transition from n-type to p-type conductivity. It will be noted that this transition does not significantly alter the trend of life-time in the crystal. This is quite often true. In some cases, however, there is an abrupt discontinuity of lifetime at the p-n barrier. Factors which influence recombination rates are not well understood at present.

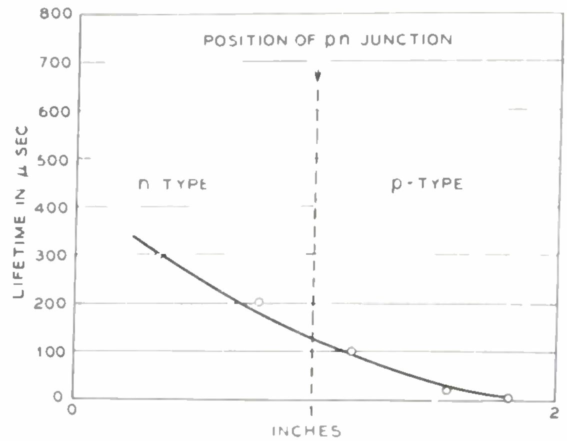


Fig. 8—Carrier life in germanium single crystal of mixed conductivity types.

The lattice perfection of the single crystals, the resultant long lifetime of carriers in them, and their chemical homogeneity cause them to be very well suited as the medium in which to construct complex electrical structures involving several p-n barriers in sequence. About two years ago, Shockley published the theory of a transistor made from a single piece of germanium in which an n-type conductivity layer lies enclosed between two p-type regions.[11,12] One of the objects of the single-crystal work has been to develop specific methods of modifying the chemical composition which would be generally applicable to the preparation of a variety of complicated structures.

Both p-n-p and n-p-n structures have been made, and recently the operating properties of an n-p-n transistor were compared with theory.[2,3] A single crystal of germanium in which a thin p-type layer is interposed between two n-type regions is represented schematically in Fig. 9. In preparing such structures one has the problem of controlling the resistivities of the n-type and p-type regions, and also that of controlling the width of the p-type layer. Structures of the n-p-n type with a variety of resistivity combinations have been made. The combination shown in Fig. 9 is typical. The width of the

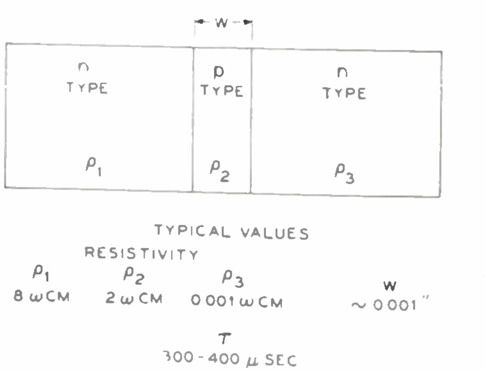


Fig. 9—Single-crystal germanium with p region interposed between two n regions.

[9] This method is to be described in detail in articles to be published soon by L. B. Valdes in PROC. I.R.E. and *Bell Sys. Tech. Jour.*

[10] J. R. Haynes and W. Shockley, "The Mobility and Life of Injected Holes and Electrons in Germanium," *Phys. Rev.*, vol. 81, pp. 835–843; 1951.

[11] W. Shockley, *Bell Sys. Tech. Jour.*, vol. 28, pp. 435–489; 1949.

[12] W. Shockley, "Electrons and Holes in Semiconductors," D. Van Nostrand Co., Inc., New York, N. Y. 1950.

p-type layer may be made either greater or less than the indicated value of about 0.001 inch, to suit the particular application. The hole lifetime in the n-type region on the left side may be 300 to 400 μsec. Lifetime in the other sections, in which direct measurements are more difficult, are probably somewhat less.

Fig. 10 illustrates the potential gradient across an n-p-n structure due to passage of current through a

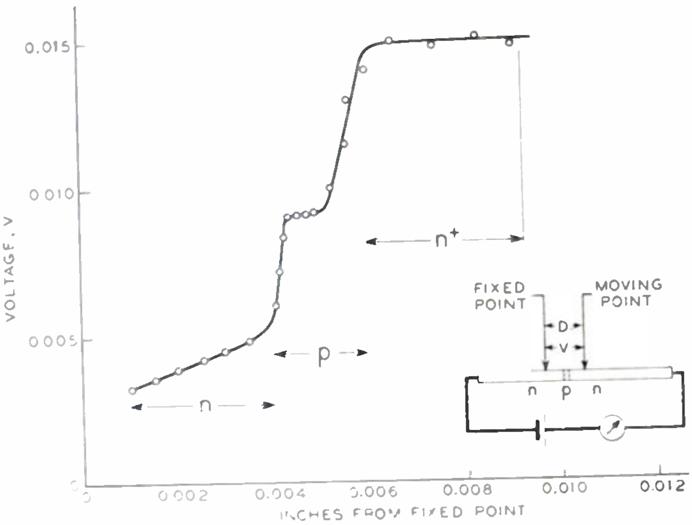


Fig. 10—Potential gradient across npn structure in single-crystal germanium.

sandblasted unit by application of a low voltage. The potential along the structure is plotted as a function of distance. From such a curve the width of the p-type layer and the resistivities of the n-type regions may be determined. In the construction of such p-n structures it is important that the junctions be approximately planar and parallel.

In conclusion, the improved materials and techniques have been one of the important factors in the development of a series of new transistor, photocell, and rectifier devices and greatly improved old ones. Of particular interest to the design engineer is the fact that the rectifying barriers between the p-type and n-type sections behave in a manner predictable from the measured properties of each section. Furthermore, the availability of uniform and reproducible single-crystal germanium has permitted the recognition of correlations between the physical properties of the semiconductor and the electrical properties of point-contact devices.

Acknowledgment

The authors are glad to acknowledge their indebtedness to many of their associates. Lasting contributions were made by J. B. Little, who collaborated in the initial single-crystal experiments. W. L. Bond gave much help in experimental arrangements. J. A. Burton and J. D. Struthers gave valuable aid with radioactive tracer techniques. R. M. Mikulyak did much of the crystal growing. J. R. Haynes, R. J. Kircher, F. S. Goucher, W. R. Sittner, G. L. Pearson, and Miss A. D. Mills contributed to the measurements.

We are particularly indebted to J. A. Morton and W. Shockley.

Requisite Color Bandwidth for Simultaneous Color-Television Systems*

KNOX McILWAIN†, FELLOW, IRE

Summary—It has been known for many years that the visual acuity of the human eye for color differences is less than that for changes in brightness. It has been shown that this fact can be used to reduce the bandwidth required for simultaneous color television systems. The experiments reported here relate to psychophysical measurements made by both skilled and lay observers to determine just how far this reduction can be carried without objectionable deterioration of the reproduction. It is shown that under the conditions tested, approximately 1 mc is sufficient for most color reproduction, provided 4 mc are available for brightness detail.

Introduction

FOR ALMOST A CENTURY observations have been made that the visual acuity of the human eye for color detail is markedly less than that for brightness detail. Color printers have used this knowledge in reducing the cost and complexity of printed reproductions; seamstresses have long known that attempts to match cloth against single threads were doomed to failure; and generations of children have found that color fill-ins required much less skill than pictures drawn in their entirety.

The recognition that this incapacity of the eye could be useful in reducing the bandwidth of a color-television system is due first to Goldsmith[1] and Bedford.[2] Goldsmith patented a television system having low resolution in the blue primary, while Bedford conceived the "mixed-highs" methods of transmission in which the color information is restricted to 2 mc (or less) for color and all of the fine brightness detail is carried in a single channel. This last is quite similar to the technique of color printing which employs three colors plus black to achieve its effects.

* Decimal classification: R583×535.6. Original manuscript received by the Institute, November 15, 1951; revised manuscript received April 1, 1952.

† Hazeltine Electronics Corp., 58-25 Little Neck Pkwy., Little Neck, L. I., N. Y.

[1] A. N. Goldsmith, U. S. Patent 2,335,180; issued November 23, 1943.

[2] A. V. Bedford, "Mixed highs in color television," Proc. I.R.E., vol. 38, p. 1003; September, 1950.

Just how much the bandwidth could be restricted without serious deterioration of the picture was not, however, realized until the report by Loughren and Hirsch,[3] who showed that restricting the bandwidth for color to one-tenth mc did not completely destroy the value of most color pictures. If it is granted that 4 mc are desirable for brightness detail (present U. S. standard), then reducing the color bandwidth to 2 mc saves one-third of the spectrum required (2 mc for each color plus two for mixed highs, a total of eight as against twelve if each color uses four). Similarly, restricting the color to one-tenth mc means that color pictures could be sent in 4.2 mc rather than in the 12 previously deemed necessary. As Fink[4] has pointed out, scarcity of spectrum space makes it imperative that "a television system should never be called upon to reproduce and image that is 'more than pleasing'," and certainly it is wasteful to transmit information which the eye cannot see. It was obviously desired in setting up standards by the National Television System Committee, that this point be resolved. Panel 11 of this committee requested the Hazeltine Corporation to conduct a psychophysical test on a group of mixed observers to determine the permissible contraction of the color bandwidth.

Conditions of Test

The demonstration utilized a flying-spot scanner activated by Kodachrome slides. The pick-up from the scanner was split by dichroic mirrors and fed through appropriate filters to three photocells, one for each color. The output of these photocells was fed to a display unit comprising red, green, and blue kinescope images superimposed, in register, by dichroic mirrors.

Two displays were available:

(1) A parallel dichroic mirror display giving a picture $9 \times 6\frac{3}{4}$ inches. With this display, participants could view the pictures from any distance greater than four times picture height.

(2) A crossed dichroic mirror display giving a picture $7\frac{1}{2} \times 5\frac{5}{8}$ inches. This display is normally used as a monitor, but was used during the tests by some of the observers. With this monitoring display the picture could be viewed from any distance greater than three times picture height.

In general, the observers tended to get as close to the screen as possible. Since the groups were kept small—six being the greatest number tested at one time—the measurements taken represent observations from about three to five times picture height.

The test color pictures could be varied all the way from the picture with 4 mc used for each color to a condition where all information from the three pickups was

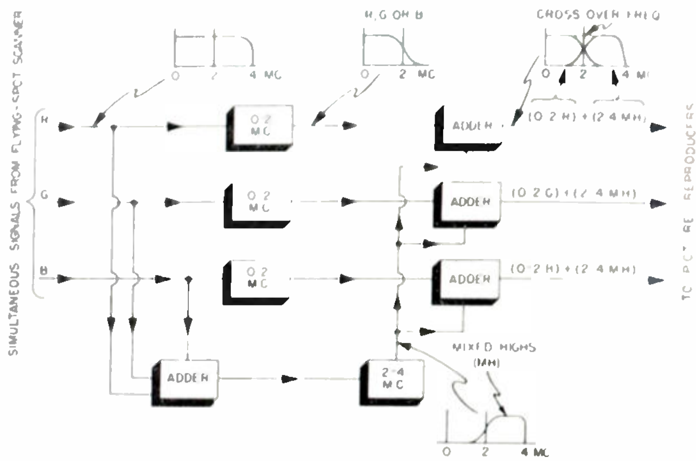


Fig. 1—Experimental set-up for transmission of mixed-highs signal.

mixed together and fed to all three tubes to give a monochrome picture. As shown in Fig. 1, filters introduced into the system controlled the frequency below

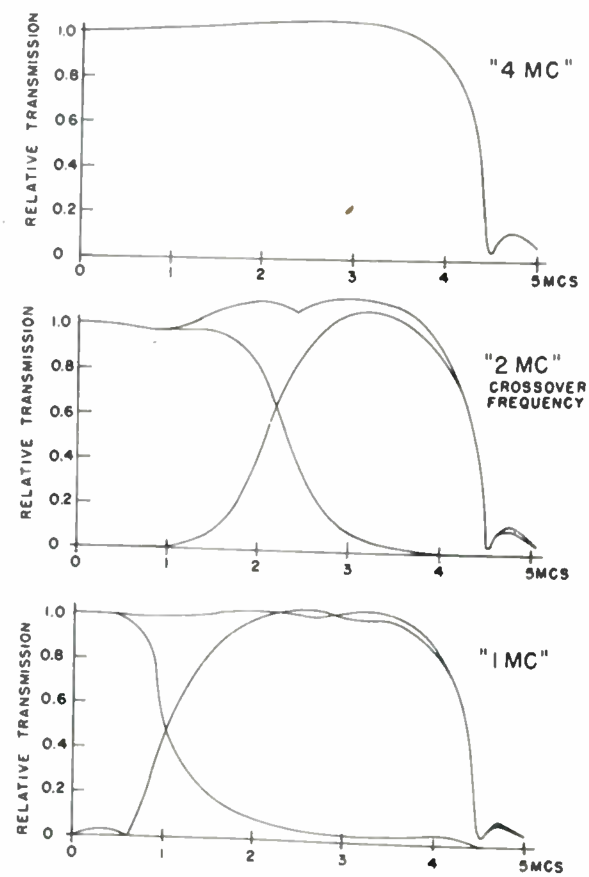


Fig. 2—Filter characteristics. Color cutoff as modified by the bandwidth limiting filter.

which the information was fed to the tubes as separate color information and above which the information from all three tubes was mixed together and supplied

[3] A. V. Loughren and C. F. Hirsch, "Comparative analysis of color-television systems," *Electronics*, pp. 92–96; February, 1951 and cover for December, 1950.

[4] D. G. Fink, "Alternative approaches to color television," Proc. I.R.E., vol. 39, p. 1124; October, 1951. "The word 'pleasing' is here used in its psychometric sense of 'giving pleasure in general'; it does not refer to the emotional reaction to particular program material. 'Convincing' and 'having the appearance of reality' are implied. 'Realistic' is *not* meant, since this means a characterization of things 'as they really are'."

as a brightness signal to all three tubes. The 6 db-down frequency of a pair of these filters is called the "cross-over frequency." As shown in Figs. 2 and 3, the crossover

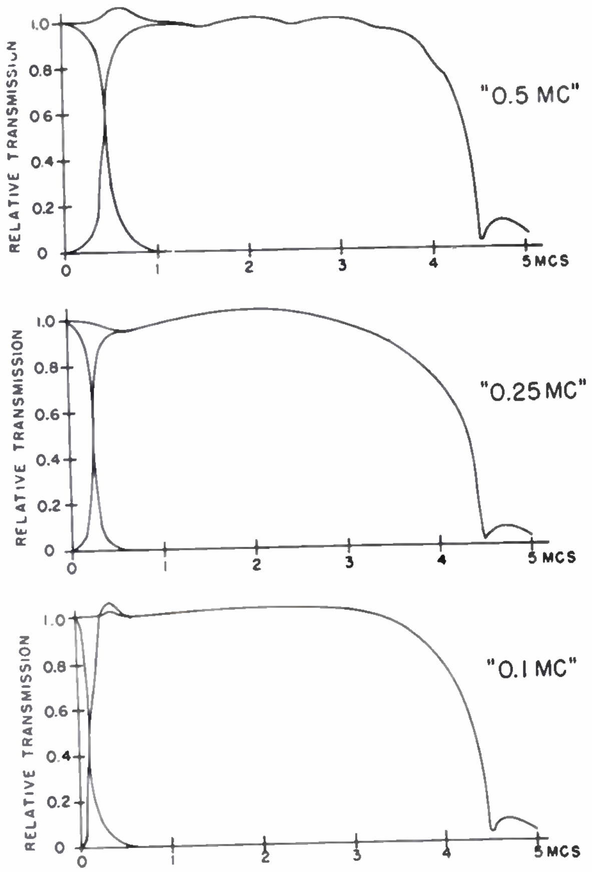


Fig. 3—Filter characteristics. Color cutoff as modified by the bandwidth limiting filter.

frequencies available were nominally 0.1, 0.25, 0.5, or 1.0 mc, and also 2 and 4 mc.

Highlight brightnesses of about 25- to 30-foot lamberts were available. About 7 or 8 seconds were required to switch from one picture to the next. Each picture was allowed on the screen for at least 15- to 20-seconds viewing time. The order of showing was as follows: A full 12-mc picture using 4 mc each for red, green, and blue was shown as a standard. Next would be shown a picture having some smaller amount (narrower frequency band) of color and a full 4-mc brightness signal. Then a return was made to the standard and a further return to the test picture. Thus, each picture was shown twice per setting, and the audience was free to, and on occasion did, request a review of certain pictures.

The test pictures were presented in a completely random fashion, unknown to any of the participants, and selected by drawing numbers from a hat, to give as nearly as possible a random distribution and one differing for each slide. Furthermore, the orders of showing differed for the several groups tested.

Participation

Three groups of individuals were tested: a group of completely uninstructed observers, six men, six women; the same group after they had received about 15 minutes of instruction in the type of detail degradation to watch for; and eleven members of Panel 11 of the National Television System Committee, many of whom were skilled color observers. In order to get all observers close to the display, each group was split into two subgroups. Instructions were given to each viewer as follows:

"During this test you will be shown several slides in which the amount of color information will be varied, while the amount of brightness information will remain constant.

"Please indicate on the accompanying sheet those settings in which the change is noticeable but satisfactory; settings in which the change is noticeable with marginal picture (still plausible); and settings in which you would judge the picture to be unsatisfactory—worse than black and white.

"Kindly note that, to avoid influencing the viewers, the individual settings may include repetitions of a given setting and will, in general, not be given in any regular order. Accordingly, you should try not to be influenced by the mere order of the sequence, but only by your free judgement, uninfluenced by other observers or any extraneous factors."

Results of Tests

Figs. 4 to 8 record the most pertinent results of the tests. Fig. 4 records the number of completely uninstructed laymen who thought that each particular slide

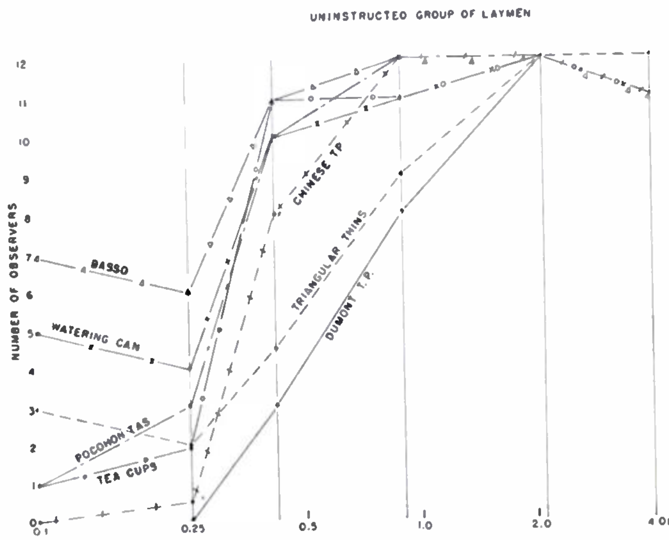


Fig. 4—Satisfactory for individual slides (laymen). Frequency at 6 db point of filters.

was satisfactory for various crossover frequencies. Fig. 5 gives the same data for Panel 11 observers. These slides are given to indicate the spread of the data for a particular group and to show that the sophisticated group is little more consistent than complete novices. In particular, note that on both of these figures points are plotted representing one-half of an observer. These

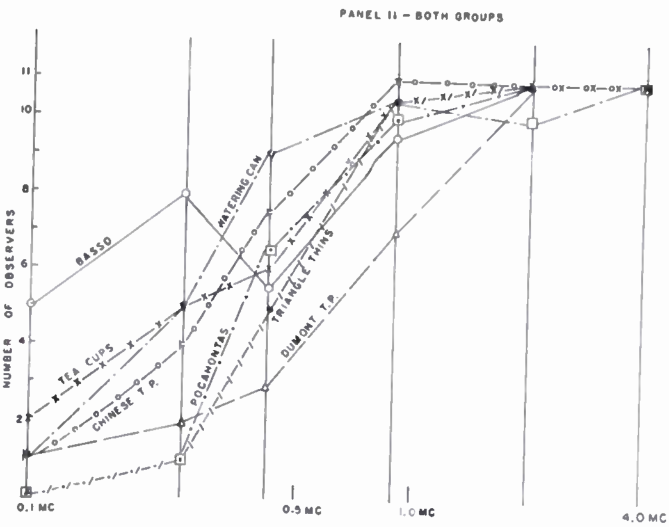


Fig. 5—Satisfactory for individual slides. Frequency at 6-db point of filters (Panel 11 experts.).

resulted from two showings of a particular slide at the same crossover frequency. The results of the two tests differed and the plotted point is the average.

Fig. 6 gives averages for the viewers of Panel 11 for all patterns to show the ranges in which pictures were judged unacceptable, satisfactory, and completely acceptable. These curves are cumulative in nature and should be read "as least as good as." The corresponding data were quite similar for the lay groups, the differences being slightly less regular curves and slightly more tolerance for color degradation.

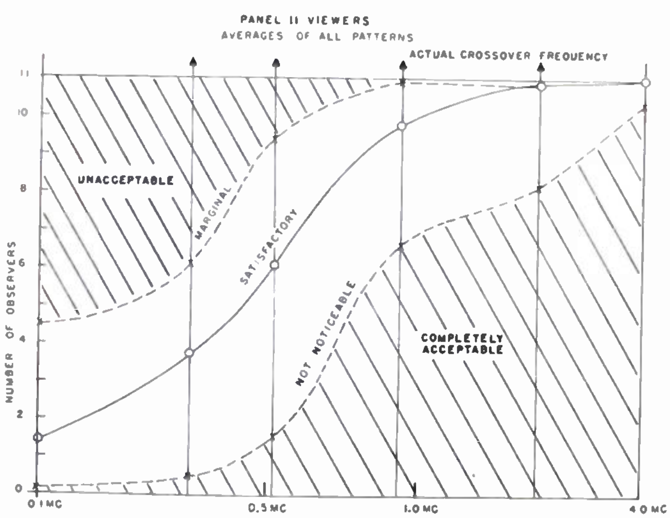


Fig. 6—Required color bandwidth.

Fig. 7 gives a comparison of the "satisfactory" curves for the various groups. The differences are not marked. Because of the spread of the data indicated on Figs. 4 and 5 and the small size of the samples, they may not even be significant. Fig. 8 gives the same curves for men and women. Again the small difference renders its significance doubtful.

It will be noted that some pictures are submarginal to a fair percentage of the observers with anything less than 1 mc of color. It will also be noted that, with 1 mc of color, no pictures are submarginal, 90 per cent of all combinations of viewers and scenes are satisfactory, and that about 70 per cent of the viewers notice no degradation at all. It will be further noted that by the time 2 mc of color are included, all pictures are satisfactory to all viewers and only a small percentage of

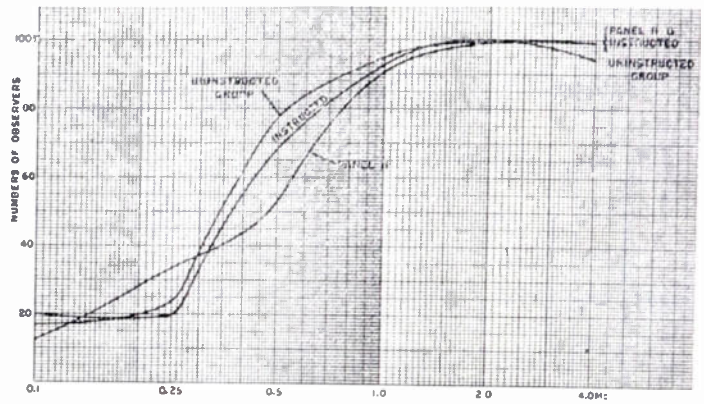


Fig. 7—Comparison of "satisfactory" curves for various skills.

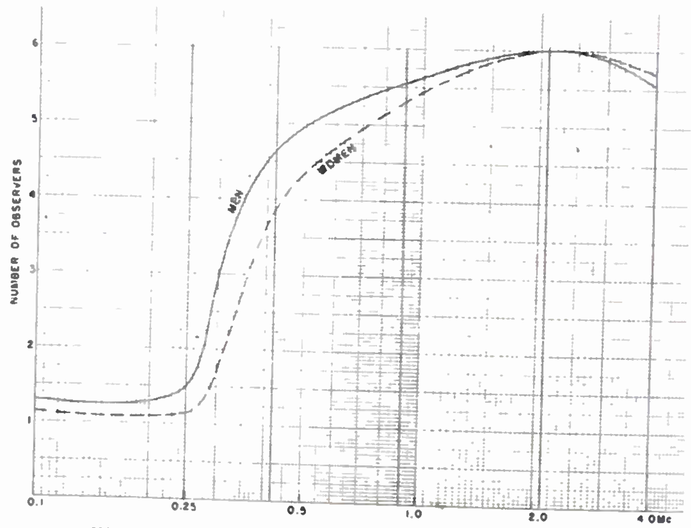


Fig. 8—Comparison of "satisfactory" curves. Men versus women (uninstructed groups).

viewers can notice any degradation in any scenes including test patterns. The teaching from this seems to be that at least 1 mc of color is a minimum requirement and that any additional color which can be obtained at not too great a cost should be included. Above 1 mc the principle of diminishing return has set in, but there is still some gain apparently all the way up to 4 mc.

This conclusion was used by the National Television System Committee, and is effectively incorporated in its proposed standards.

Acknowledgment

My thanks are due R. P. Burr and C. E. Page for assistance in performing the experiments and computations.

Multiple-Channel Telephony on VHF Radio Links*

B. R. TUPPER†, SENIOR MEMBER, IRE AND P. B. PATTON‡, ASSOCIATE, IRE

Summary—**This paper describes a single-sideband multiplexing system which has brought multiple-channel radio telephony into a new economic range for toll-circuit use. Problems of interchannel modulation and noise have been solved with simple, economical carrier and radio equipment. While this paper is concerned primarily with the multiplexing method, sufficient specifications on the radio transmitter and receiver are included to permit comprehension of the system as an integrated whole.**

Introduction

Not many years ago the expression "radio telephony" conveyed the thought of a single talking circuit on a radio channel. Early multiplexing systems, equipment of which was intricate, specialized, and costly, were limited to the few applications which would have been unusually difficult to achieve by wire-line methods.[1,2] With the exploitation of higher frequencies in the vhf and uhf regions and corresponding antenna gains, the cost of radio equipment has been greatly reduced. Present-day carrier radio is founded on the same basic principles as the early systems, but over-all equipment costs have been brought down so drastically that multiple-channel radio links compete with wire lines under practically any conditions of terrain.

Single-Sideband Suppressed-Carrier Multiplex

Multiple-channel telephony on wire lines was long since put on a single-sideband basis to save precious spectrum space and to reduce the power-handling requirements of equipment intervening the two terminals.[3] These two advantages of single-sideband operation are of as great if not greater advantage today than when first employed. The necessity for making most effective use of each radio-channel assignment increases daily, even in the uhf and shf regions. Minimum modulation levels for each multiplexed telephone channel are demanded more and more as the number of communication circuits on the radio channel is increased.

The single-sideband method is, in effect, a stacking process by which each added 3- or 3.5-kc band of frequencies required for good speech reproduction is lifted to a higher allocation. Still the bandwidth required for each speech channel in its multiplex allocation is never more than in the original voice location.

The multiplexing equipment described here achieves accepted international standards in spectrum economy through the use of a band translation method. This method permits the most modern core materials for filter coils to be employed in their most effective frequency range. By previous techniques, 40 or 50 kc was about the upper limit for which single-sideband image suppression and receiving channel discrimination could be made adequate with lc filters. In fact if single direct translations are used, that upper limit has still not been extended much by the use of powdered metal inductor cores. Systems involving more than 6 or 8 bands stacked above one another previously used crystal filters too expensive to permit the economies required for short-haul links.

Double Translation

The translation method employed in this equipment requires suppression for the voice-frequency image only at the relatively low frequency of 8–12 kc; a second translation stage puts each channel in a final transmitting position. The filters involved in selecting the wanted from the unwanted frequencies in that final positioning operation need only attain a high rejection ratio for frequencies 16 kc removed from the wanted band. Fig. 1 reveals the principle.

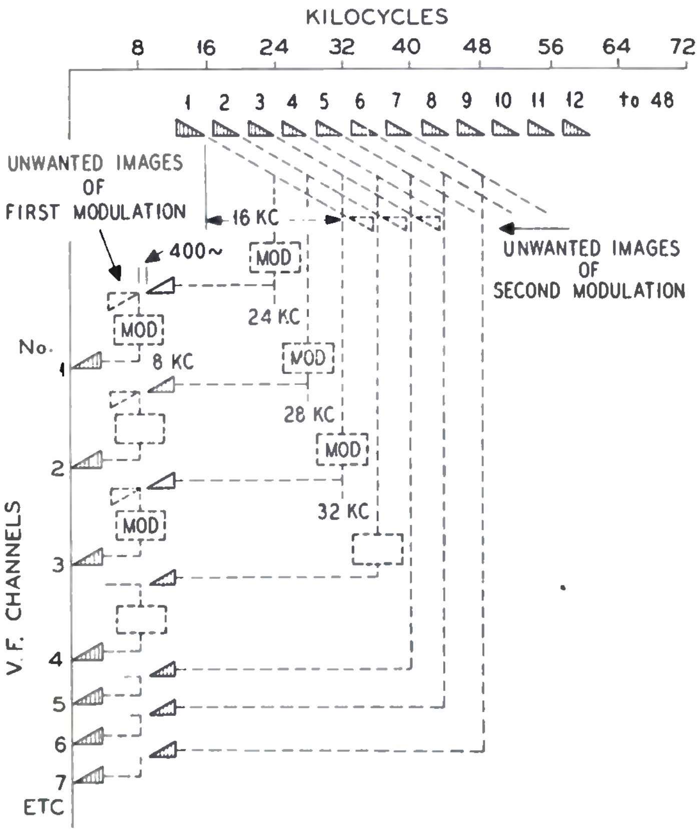


Fig. 1—Single-sideband translation method and channel allocation chart.

* Decimal classification: R460. Original manuscript received by the Institute, October 15, 1951; revised manuscript received, April 14, 1952. This paper contains the essential substance of two presentations made before the Vancouver Section of the Institute of Radio Engineers on March 19, 1951.

† Northwest Telephone Co., 1955 Wylie St., Vancouver, B. C., Canada.

‡ Lenkurt Electric Co., Inc., 1105 County Rd., San Carlos, Calif.

[1] N. F. Schlaack and A. C. Dickieson, "Cape Charles-Norfolk ultra-short-wave multiplex system," Proc. I.R.E., vol. 33, pp. 78–83; February, 1945.

[2] W. S. Marks, O. D. Perkins, and W. R. Clark, "Radio-relay communication system in the United States Army," Proc. I.R.E., vol. 33, pp. 502–522; August, 1945.

[3] E. H. Colpitts and O. B. Blackwell, "Carrier current telephony and telegraphy," *Trans. AIEE*, vol. 40, pp. 205–300; 1921.

The method of translating the channels allocated to the range from 12 to 204 kc is identical for all 48 channels. The band of frequencies, those between about 200 and 3,500 cycles, representing a typical channel, is used to modulate an 8-kc carrier.

The modulator, a doubly balanced device, is so arranged that neither the modulating signals nor the carrier appear in its output, only the two sidebands. The lower sideband appears in the range from 4,500 to 7,800 cycles, and the upper, in the range from 8,200 to 11,500 cycles. The upper sideband group is separated from the lower group by a band-pass filter whose characteristics are shown in Fig. 2. The output of this filter is then

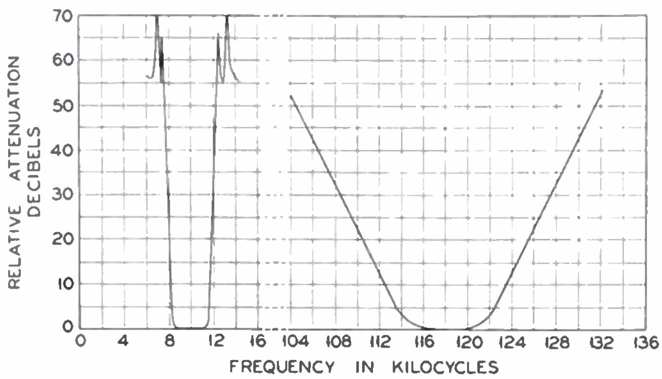


Fig. 2—Discrimination of intermediate-frequency and high-frequency filters.

fed to the final modulator, a signal and carrier-suppressing device almost exactly like the first. The carrier feed to this modulator for the channel, designated here as No. 1, is 24 kc. The lower sideband group of this modulation is chosen for transmission. The problem of separating the wanted group from the unwanted image in this stage of modulation is only one of discriminating adequately against frequencies at least 16.4 kc removed; this is readily done with an lc filter of conventional design. The characteristics of a typical high-frequency filter are also shown in Fig. 2. Succeeding channels up the scale are translated by carriers each 4 kc higher than the last, and the unwanted image is rejected in each case by a filter of similar characteristics, but having a pass band 4 kc higher than the last. The outputs of the so-called "line-frequency" filters are all brought together at a common transmitting bus which feeds the radio transmitter.

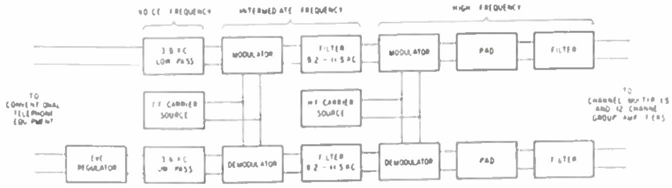


Fig. 3—Block diagram of typical channel.

The block diagram of Fig. 3 indicates the essential elements of a complete sending and receiving unit for a typical channel of this equipment.

In the outgoing path, the speech currents encounter first a 3.6-kc low-pass filter which insures that no unessential speech frequencies get into the intelligence side of the modulator. The modulator element itself is a bridge arrangement of germanium diode rectifier units not unlike those employed as mixers in RF practice. Differentially wound transformers at either end of the bridge insure that the carrier frequency does not appear at a significant level in either the input or output circuits. In this modulator configuration, the input signal frequencies are also cancelled and do not appear in the output circuit.

Following the first modulator are the intermediate frequency filter, the line-frequency modulator, and attenuator and the line-frequency filter. On the incoming side, the identical circuit elements are encountered in the reverse order with only a simple voice-frequency amplifier added. This rebuilds the power level to that value determined by established conventions as adequate to feed the subscriber's set via the telephone switchboard. The carrier supplies, intermediate and final, are derived from simple though stable oscillators, the latter being crystal-controlled for frequencies above 50 kc. The necessity for close relative-frequency accuracy between sending carrier supply and receiving carrier supply in single-sideband operation is well recognized.

The simplicity of the whole channelizing assembly is shown in Fig. 4. The vf amplifier in the receiving leg

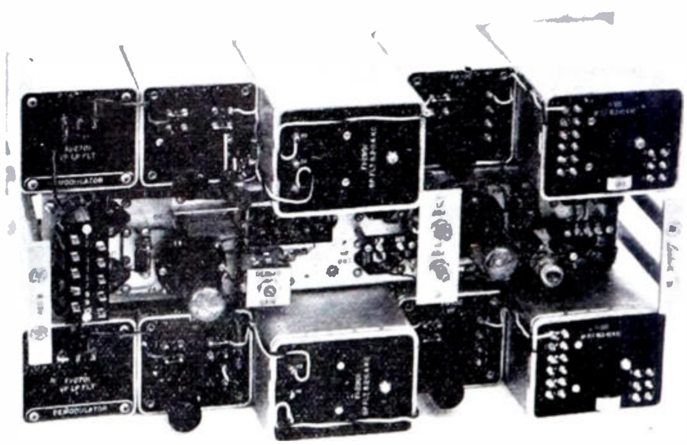


Fig. 4—Typical channel apparatus assembly.

has only one vacuum tube and the two oscillators have one each, the only electronic equipment involved in a complete sending and receiving channel. The crystal-diode elements have been shown to render such long trouble-free life in this service that they have come to be classed with the "inert" equipment components. Customarily the talking channels and associated signaling channels are assembled in groups of twelve in an arrangement such as shown by Fig. 5. The unitized design and the self-sufficiency of the units permit the installation initially of any desired number. Additions required to satisfy further load demands may be made as occasion arises.

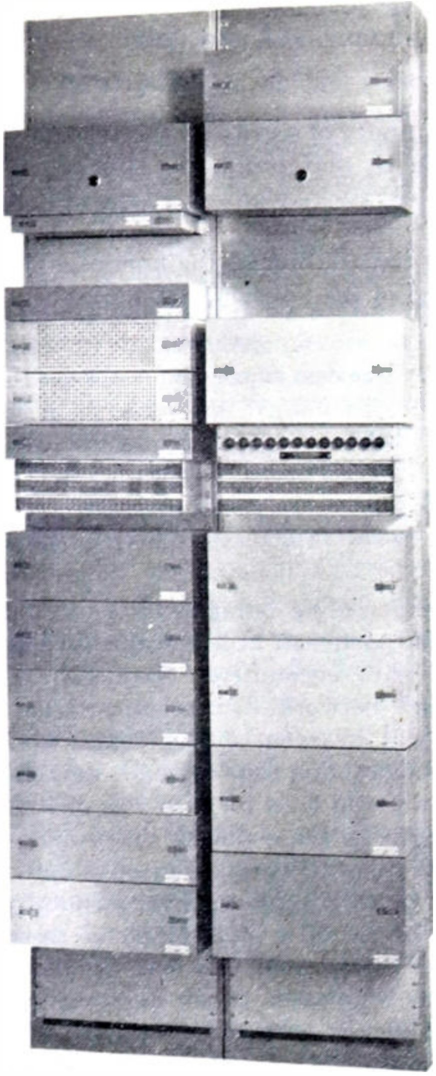


Fig. 5—Twelve-channel group assembly.

Signaling

Telephone multiplexing equipment for radio is incomplete without a system for signaling which is wholly compatible with similar facilities employed in other parts of the telephone plant. For this reason, the terminals here described were so designed that the exact nature of signaling provisions might be chosen to fit the existing plant to which the derived channels are added. A signaling method adaptable to present-day automatic and semi-automatic telephone switchboard requirements is the narrow-band vf carrier channel method. Under this system, one derived speech band from each two groups of twelve is set aside to provide signaling circuits. Another method of signaling, "speech plus signaling," is now gaining favor. This utilizes a keyed tone (for example, 3,400 cps) in the high end of the speech band. Filters are provided so that speech occupies the 200–3,100-cps portion of the band, reserving the higher frequencies in each channel for signaling.

Radio Equipment

Operating within the 240–250-mc range, the radio equipment used for multiple-channel telephone service is in extensive use by the Northwest Telephone Company of Vancouver, B. C., and gives highly satisfactory results. Power available to the transmitting antenna feed line is 30 watts, frequency modulation being employed with a final RF deviation of 100 kc. Successive doubler stages finally drive a type 4X-150A tube, which also acts as a doubler.

The receiver has a sensitivity of 5 μv for 20-db quieting with a noise figure of 9 db. It employs a two-stage, grounded-grid, tuned RF amplifier. The intermediate frequency is derived by mixing the RF with the output of multipliers driven from a 40-mc crystal. A five-stage

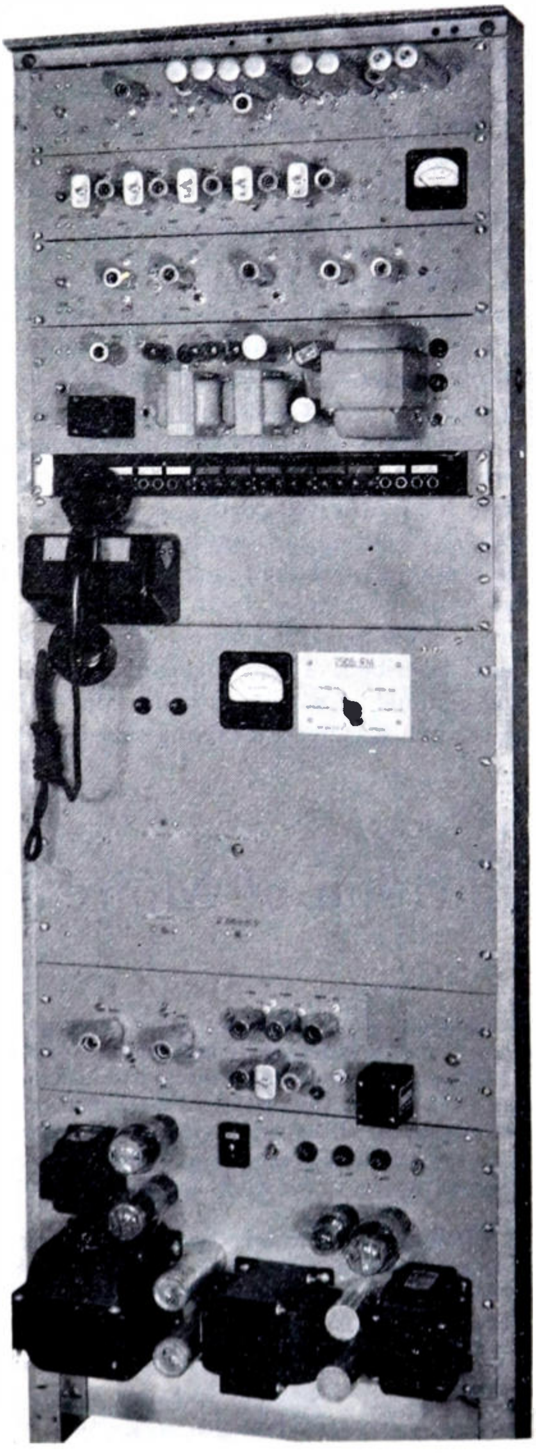


Fig. 6—250-mc FM radio terminal.

IF amplifier at 10.7 mc is followed by a two-stage limiter and a discriminator. The latter feeds an audio amplifier which delivers an aggregate signal of +10 dbm to the multiplexing equipment. The squelch circuit of the receiver is equipped with a relay which, upon failure of a satisfactory incoming signal, transmits an alarm to the connected telephone equipment and at the same time prevents noise from reaching the subscribers.

Compactness and simplicity of the radio equipment are indicated in Fig. 6. Having a combined power consumption of only 375 watts, the transmitter and receiver occupy only 4 feet of vertical space on a 19-inch rack.

The antennas used for both transmitting and receiving were designed for pole mounting, and are approximately 6 feet wide and 12 feet high. Both are multiple arrays of twelve half-wave elements horizontally polarized, backed by reflectors. The gain of each is 14 db over a half-wave dipole.

Performance

While a considerable number of radio-telephone installations of this type have been made recently, there are as yet few of these on which carefully kept transmission measurements can be presented. Over a link between Vancouver and Nanaimo, a distance of 40 miles, ten trunk telephone channels have now been operated for more than a year. For this system, precise operational data have been recorded.

The standard acceptance figures for toll circuits are that the total disturbance measured on a speech-weighted basis must be at least 60 db below average voice level. This performance is obtained with the equipment described. These measurements are made when one channel at a time is made idle while other channels are in use as normal talking circuits. The levels must be properly adjusted to obtain the best compromise between noise and cross talk.

Continuity

The continuity of service on such systems is probably comparable to that ordinarily obtained on an equal mileage of open-wire plant over favorable wire-line terrain.

The lost-time record for the first year of operation is summarized in the following tabulation:

TABLE I

PER CENT OF TIME OUT-OF-SERVICE

Multiplex equipment	Radio equipment	Radio tubes	Propagation	Supply	Total
0.01	0.26	0.11	0.014	0.01	0.404

The fact that fading outages are almost insignificant points to the particular utility of the vhf frequencies for a type of path encountered frequently in the large "Northwest" network. On this particular link with barely optical clearance at the center horizon and a mountain obstruction near one end, the received signal fell below a useful level for intervals of a few minutes' duration only 11 times during the year. Four similar paths in the same network, operated at 250 mc with the equipment described, have shown no propagation outages for a period of a year. Many of these paths are largely over water, as is this one. On the other hand, when efforts were made to use microwaves (6,600 mc) for 30- and 40-mile over-water hops, the continuity was wholly unsatisfactory for toll trunk telephone service. In these trials, paths were arranged to afford more than first Fresnel-zone clearance, and diversity reception was employed.

Elimination of Moiré Effects in Tri-Color Kinescopes*

E. G. RAMBERG†

Summary—**Moiré effects which may arise in aperture-mask tri-color kinescopes are spurious intensity variations in the picture in the nature of beat patterns between the scanning lines and the aperture array in the shadow mask. The visibility of these effects depends on the relative magnitudes of the scanning-line width, mask-aperture size, aperture spacing, and line separation, on the orientation of the scanning pattern relative to the mask, and, finally, on the picture content.**

For the narrow aperture spacing normally employed (e.g., 215,000 apertures in a rectangular picture area of 104 square inches or 195,000 apertures in the somewhat smaller area defined by the framing mask) and for the preferred orientation of the scanning pattern, however, the moiré effects are negligible. They may become noticeable, in the form of dot or bar patterns, if the aperture spacing is increased or the orientation of the scanning pattern relative to the mask is changed.

The variation in the line transmission of the mask indicates directly the degree to which the mask may distort transmitted intensity values. It increases with a reduction in the ratios of line width and aperture diameter to aperture spacing and with a departure from the preferred orientation of the scanning pattern relative to the mask. Again, for the preferred orientation and within the range of spot sizes required for optimum resolution, the variation in line transmission is negligible—1 per cent or less. Furthermore, since the increase in the variation in transmission with departure from the preferred orientation is quadratic, the picture quality is insensitive to small deviations from the optimum orientation.

* Decimal classification: R583.6×535.6. Original manuscript received by the Institute, October 28, 1951; revised manuscript received April 9, 1952.

† Radio Corporation of America, RCA Laboratories Division, David Sarnoff Research Center, Princeton, N. J.

The Nature of the Moiré Introduced by the Mask

The introduction of the mask, with its regular hexagonal array of apertures, into the aperture mask tri color kinescope[1] causes intensity variations in the image which are not present in a kinescope for monochrome reproduction. As long as these variations are limited to areas of the order of a picture element—as they would be, for example, if the mask were simply bombarded by a uniform spray of electrons—they do not affect the picture; at the normal viewing distance the eye recognizes only the average brightness of a picture element, and this, by assumption, is unaffected (in relative measure) by the mask.

It will be shown that this condition applies also for actual television pictures which are formed by a set of equally spaced scanning lines, even though for a uniform picture signal the intensity distribution on a kinescope screen, without mask, is not uniform. The brightness fluctuation, which is repeated identically for every line spacing, is greatest if the line width is small compared to the line spacing. Since, however, the latter lies close to the limit of visual resolution, this brightness fluctuation is scarcely perceived under normal circumstances.

The introduction of the mask, with its rows of apertures with a periodicity which differs, in general, from that of the scanning lines, will give rise to brightness fluctuations on the screen with a period which may be much greater than the spacing of either the mask-aperture lines or the scanning lines. It will be assumed, to begin with, that scanning lines and mask-aperture lines are aligned as shown in Fig. 1. Minimizing, in this manner, the vertical spacing of the mask-aperture lines serves to minimize the effect of the mask on vertical resolution and the prominence of spurious intensity fluctuations. The transmission of electrons through a row of apertures will be a maximum when a scanning line is centered on it; it will be least when its center line falls midway between two scanning lines. Thus, if, for example, the spacing of the rows of apertures, $a/2$, is but slightly less than the line separation h, the brightness of the field for uniform signal will be a maximum for the first condition and a minimum for the second, which will be reached $n-\frac{1}{2}$ scanning lines down the picture:

$$n\left(h-\frac{a}{2}\right)=\frac{h}{2}; \qquad n=\frac{1}{2-\dfrac{a}{h}}. \tag{1}$$

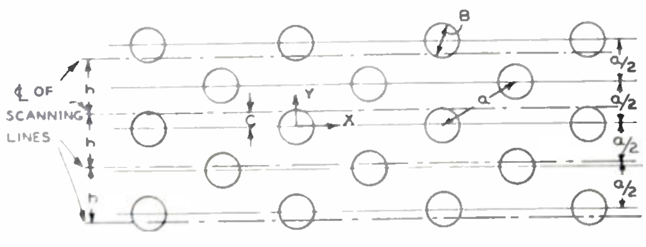


Fig. 1—Preferred orientation of scanning lines relative to mask.

Hence there may be gross intensity fluctuations, or moiré, in the vertical direction, with a period of $2n-1$ scanning lines.

It is seen from (1) that the peridicity of the moiré is determined exclusively by the ratio of the dot separation a on the mask to the separation h of the scanning lines. The relative amplitude of the fluctuation, on the other hand, is determined only by the ratios of the line width d_s (for a given form of intensity distribution in the line) and the aperture diameter B to the line separation h.

If the scanning pattern is rotated relative to the mask by an angle θ, both the periodicity and direction of the moiré changes. Essentially, the intensity distribution in the field is given by the superposition of sinusoidal intensity variations of constant amplitude and differing frequency and direction. The amplitude is determined, as before, by the ratios of the line width and aperture diameter to the line separation. If the period of one of the sinusoidal components is much larger than that of any other, a line moiré, consisting of a sequence of broad bars, is obtained for uniform signal. If there are two components of nearly equal period, a dot moiré (with periods greater than the line separation) will be present.

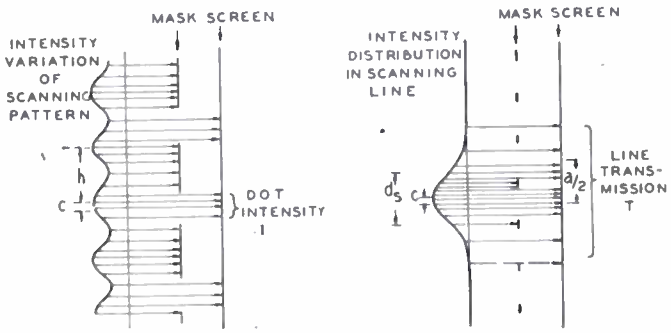


Fig. 2—Schematic representation of dot intensity and line transmission.

The intensity distribution on the screen obtained for uniform signal does not, however, tell the full story of the effect of the mask on the picture. The picture signal contains, in effect, information regarding the variation in brightness of the sequence of equally spaced lines forming the scanning pattern. Hence, a proper measure of the error introduced by the mask is the variation in the transmission of the mask for these scanning lines, depending on the relative position of the scanning line and the aperture array in the mask. Fig. 2 compares

[1] H. B. Law, "A three-gun shadow-mask color kinescope," Proc. I.R.E., vol. 39, pp. 1186–1194; October, 1951.

schematically the intensity of a dot in a uniform-signal field and the line transmission. The dot intensity is the transmission of a single mask aperture for the whole scanning pattern, whereas the line transmission is the transmission of the entire aperture mask for a single scanning line. The fluctuation in the dot intensity will be larger than that of the transmission as long as the spacing of aperture lines is less than the spacing of scanning lines. This condition is always satisfied in practice for the optimum orientation of the scanning pattern relative to the mask (Fig. 1).

This smaller fluctuation of the transmission is to be expected since here the electron current is, in effect, integrated over a coarser unit (the scanning line rather than the aperture line). In return for the greater coarseness of the information supplied, the transmission yields information not only regarding fluctuations in a uniform-signal field, but also for a field in which the line brightness may vary in some prescribed fashion (e.g., a succession of dark and light horizontal bars). If the scanning pattern is disoriented with respect to the mask, the transmission fluctuation rapidly increases, reaching a maximum for a displacement of 30° relative to the optimum position; here the fluctuation in transmission greatly exceeds that in (dot) intensity for uniform signal since the spacing of aperture rows (for, e.g., $a=0.023$ or 0.030 inch) is materially greater than the scanning line spacing h (placed at 0.018 inch).

Basic Assumptions

The following calculations are made for a tube with a 9×12-inch scanning raster of 500 lines[2] and a line separation of 0.018 inch. A sequence of values, from 0.009 to 0.018 inch and from 0.0075 to 0.013 inch, are assumed for the line with d_s and the aperture diameter B. The separation of nearest neighbors in the aperture array, a, is made equal, in general, to 0.023 and 0.030 inch, although most of the results can readily be extended to any other values. The following exponential distribution is assumed for the scanning line:

$$f(y) = \frac{1}{b\sqrt{\pi}} e^{-y^2/b^2} = \frac{1.668}{d_s\sqrt{\pi}} e^{-(1.668y/d_s)y^2}. \qquad (2)$$

Here the "line width" d_s is defined as the transverse distance between two points at which the brightness of the line has dropped to $\frac{1}{2}$ of its maximum value. A "flat field" would correspond approximately to $d_s/h=1$. The results reported here do not, of course, depend on the absolute dimensions of mask and scanning pattern, but only on the ratios d_s/h, B/h (or B/a), and a/h.

Intensity Distribution for Uniform Signal

The intensity transmitted by any one mask aperture may be written as a cosine series in terms of c/h, c being the distance between the center of the aperture and the center of the nearest scanning line.

$$I\left(\frac{c}{h}\right) = \frac{\pi B^2}{2\sqrt{3}\,a^2}\left(1 + \sum_{n=1}^{\infty} k_n \cos\frac{2\pi nc}{h}\right). \qquad (3)$$

Here, for convenience of comparison, the external coefficient has been set equal to the average transmission of the mask; in other words, the intensity falling on the total mask area per aperture has been set equal to unity. The values of k_n depend on the ratios of the line width and aperture diameter to the line separation, d_s/h and B/h. Tabulated below, they were calculated by the formulas derived in the Appendix. It is seen that even the second Fourier coefficient is quite insignificant, being 0.02 in the least favorable case considered. (The last column represents the Fourier coefficients of the intensity distribution without mask.) Thus the intensity variation, as function of c/h, is practically sinusoidal and the relative amplitude of its fluctuations is given directly by the values of k_1.

TABLE I

Values of k_1

B in B/h		0.013 0.7222	0.0115 0.6389	0.009 0.5	0.0075 0.4167	0 0
d_s in	d/h					
0.018	1	0.0277	0.0331	0.0416	0.0461	0.0576
0.0127	0.707	0.1631	0.1950	0.2450	0.2717	0.3394
0.009	0.5	0.3962	0.4732	0.5948	0.6596	0.8240

TABLE II

Values of k_2

B in B/h		0.013 0.7222	0.0115 0.6389	0.009 0.5	0.0075 0.4167	0 0
d_s in						
0.018	1	-1.5×10^{-7}	-5×10^{-8}	2.5×10^{-7}	4.9×10^{-7}	1.38×10^{-6}
0.0127	0.707	-1.76×10^{-4}	-5.9×10^{-5}	3.02×10^{-4}	5.9×10^{-4}	1.66×10^{-3}
0.009	0.5	-0.00612	-0.00204	0.01045	0.02050	0.0576

Whereas the amplitude of the fluctuations is given by k_1, their periodicity in a vertical direction, for the alignment between aperture rows and scanning lines shown in Fig. 1, is given by (1). In particular for the values $a=0.030$ in ($a/h=1.6667$) and $a=0.023$ in ($a/h=1.2778$) the period of the fluctuation becomes 5 and 1.78 scanning lines. Since, in the second case, the period is of the same order as the dot separations, the effect would be quite unnoticeable here. With the larger spacing employed in some earlier experimental tubes, a modulation of the vertical intensity should be readily visible even at viewing distance large enough so that the line structure and the dot patterns are no longer resolved. It may be noted that, for the narrower spacing, a fluctuation with a somewhat larger period (1.99 h) is observed in a direction forming an angle of about 26° with the vertical (see the following).

[2] Or, for example, a 8.45×11.27-inch raster of 470 lines. This refers to the raster on the mask. The picture area is some 5 per cent larger.

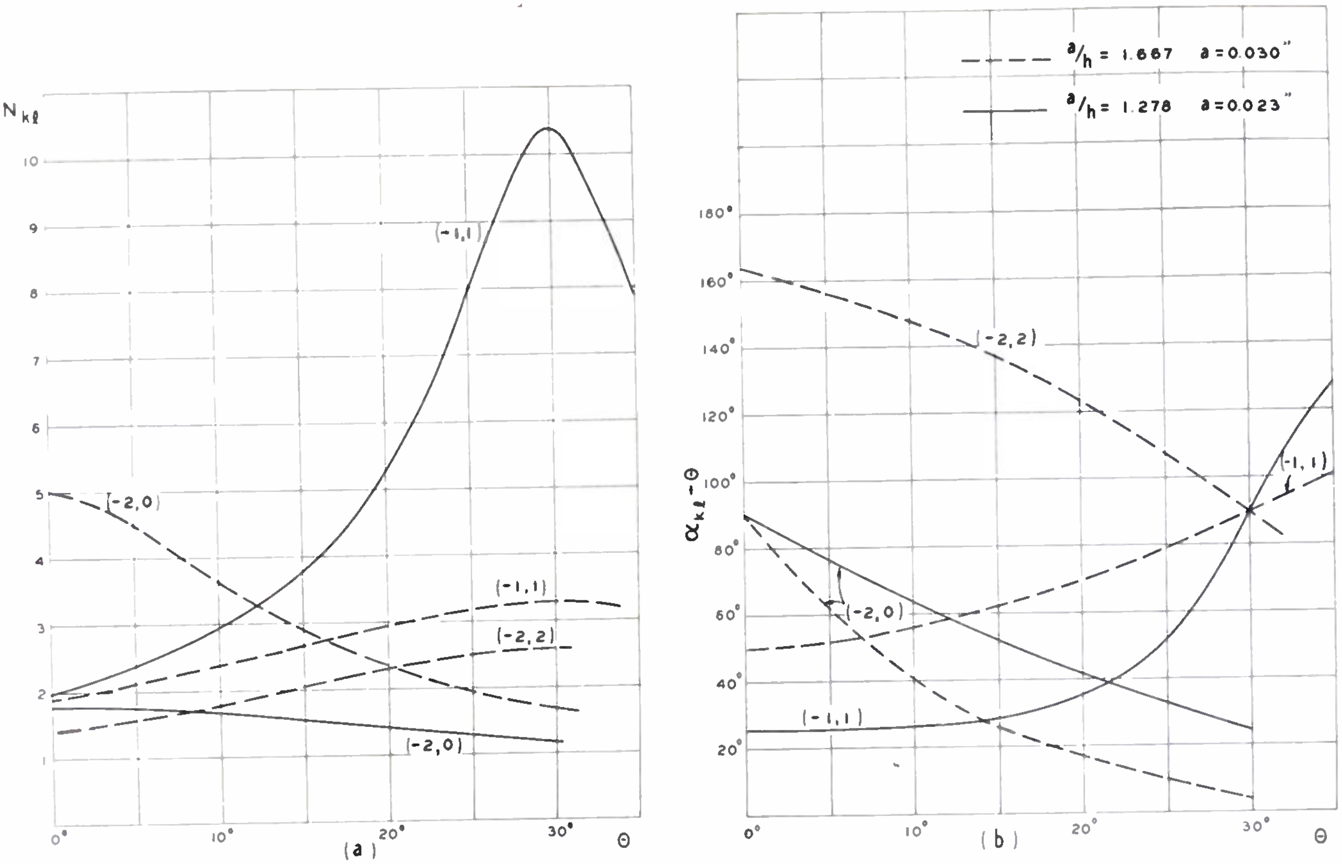


Fig. 3—Variation (a) of moiré periods N_{kl} measured in multiples of the scanning-line separation and (b) of directions of corresponding intensity variations relative to the horizontal scanning lines as function of angle θ between actual and preferred orientation of scanning lines relative to mask.

The moiré for uniform signal when an angle θ is formed between the scanning lines and the x-axis (Fig. 1), which is imagined as fixed on the mask, may be represented by the following expression, derived in the Appendix:

$$J(x, y; \theta) = \frac{B^2}{2\sqrt{3}\,a^2}\Bigg\{1 + \sum_{k,l=-\infty,\,k+l=2n,\,(k,l)\neq(0,0)}^{\infty} \cdot k_1 \cos \frac{2\pi}{N_{kl}h}(x \cos \alpha_{kl} + y \sin \alpha_{kl})\Bigg\}. \quad (4)$$

This represents the full solution for the intensity pattern obtained with uniform picture signal. The fluctuation components are all of the same amplitude k_1, but vary in the ratio N_{kl} of their period to the scanning-line separation as well as in the direction of intensity variation α_{kl}. Referred to the horizontal scanning lines, this direction is $\alpha_{kl}-\theta$.

Only the terms of lowest order in h and k are significant since for higher order, N_{kl} rapidly becomes small. Fig. 3 shows the variation of N_{kl} and $\alpha_{kl}-\theta$ as a function of θ for the components which yield, for small values of θ, the largest values of N_{kl}. Here, again, it is assumed that $a=0.030$ $(a/h=1.667)$ and 0.023 $(a/h=1.2778)$ inches.

Transmission of Mask for a Scanning Line

The calculation of the transmission of the mask for a scanning line is indicated in detail in the Appendix. It is, of course, independent of h and, instead, a function of the two ratios B/d_s and B/a, namely of the ratios of the aperture diameter to the line width and aperture separation. An amplitude factor k_T, corresponding to k_1 in Table I, may be introduced to describe the variation in transmission as the scanning line is displaced relative to the center line of a row of apertures by a distance varying from 0 to $a/4$. This factor is given, for comparison with Table I, in Table III. It is also plotted, more extensively, in Fig. 7. As already mentioned, these fluctuation amplitudes are materially smaller than values of k_1 listed in Table I. Identical val-

TABLE III

Values of k_T

a in	B in / d_s in	0.013	0.0115	0.009	0.0075	0
0.030	0.018	0.004	0.005	0.008	0.007	0.012
0.030	0.01273	0.050	0.066	0.095	0.108	0.155
0.030	0.009	0.179	0.242	0.344	0.408	0.551
0.023	0.018					0.0003
0.023	0.01273		0.005	0.011	0.014	0.026
0.023	0.009		0.042	0.100	0,127	0.228

ues of k_T and k_1 are reached for $a = 0.036$ $(a/2 = h)$, for which moiré effects are also most prominent, the period of the vertical intensity variation approaching infinity.

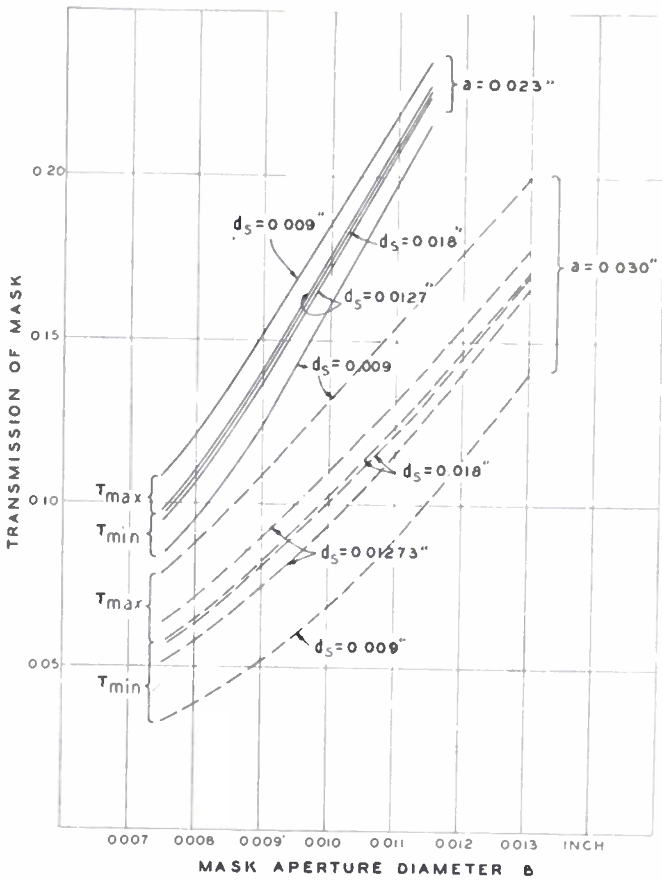


Fig. 4—Maximum and minimum mask transmissions as function of the mask-aperture diameter.

A plot of the actual transmissions as function of the aperture diameter B is given in Fig. 4.

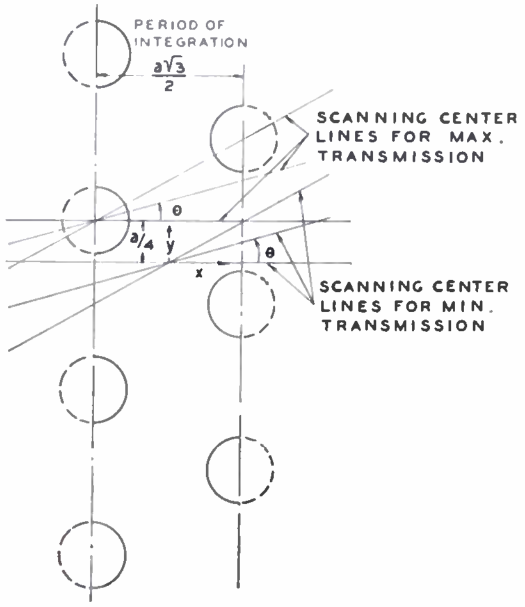


Fig. 5—Determination of mask transmission when scanning line forms angle θ with preferred direction of scanning.

The amplitude k_T increases rapidly with angular displacement of the scanning pattern relative to its optimum orientation (Fig. 1). This amplitude must now be defined as the fluctuation of transmission with scanning-line displacement for some prescribed "period" of the mask in the x-direction (Fig. 5). For the results shown in Fig. 6, the period was taken to be $a\sqrt{3}/2$, which is the minimum period which may be employed. It is readily seen that for both $\theta = 0$ and $\theta = 30°$ the maximum value of the transmission is obtained for rays passing through the center of an aperture within the period, whereas the minimum is obtained for rays passing through a point midway between both horizontal and vertical aperture rows. The same should apply, to a close approximation, for intermediate values of angle; k_T, plotted in Fig. 6, is then given by

$$k_T = \frac{T_{max} - T_{min}}{T_{max} + T_{min}}. \tag{5}$$

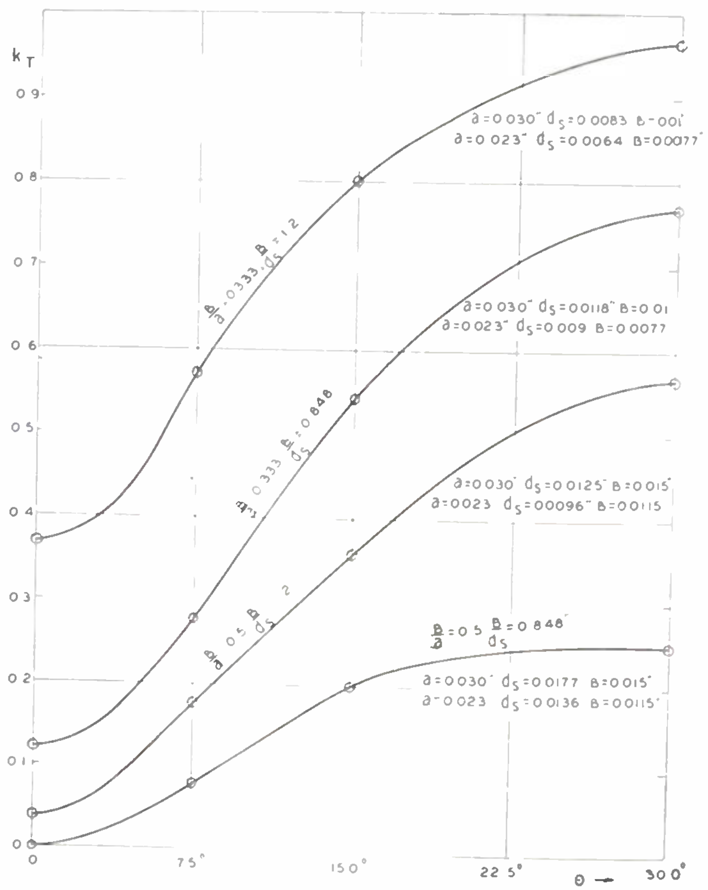


Fig. 6—Fluctuation in transmission as function of angle θ between actual and preferred scanning direction.

The choice of the period of integration—in effect, the distance in a horizontal direction over which intensities are summed—will affect the values of k_T only for angles between 0 and 30°, not at the end points; the greater the period of integration, the less the fluctuation amplitude at intermediate points. The fluctuation amplitudes of the transmission for the end points, i.e., the most and the least favorable orientations of the scanning pattern ($\theta=0°$ and $\theta=30°$), are plotted as a function of the number of apertures in the picture area in Fig. 7.

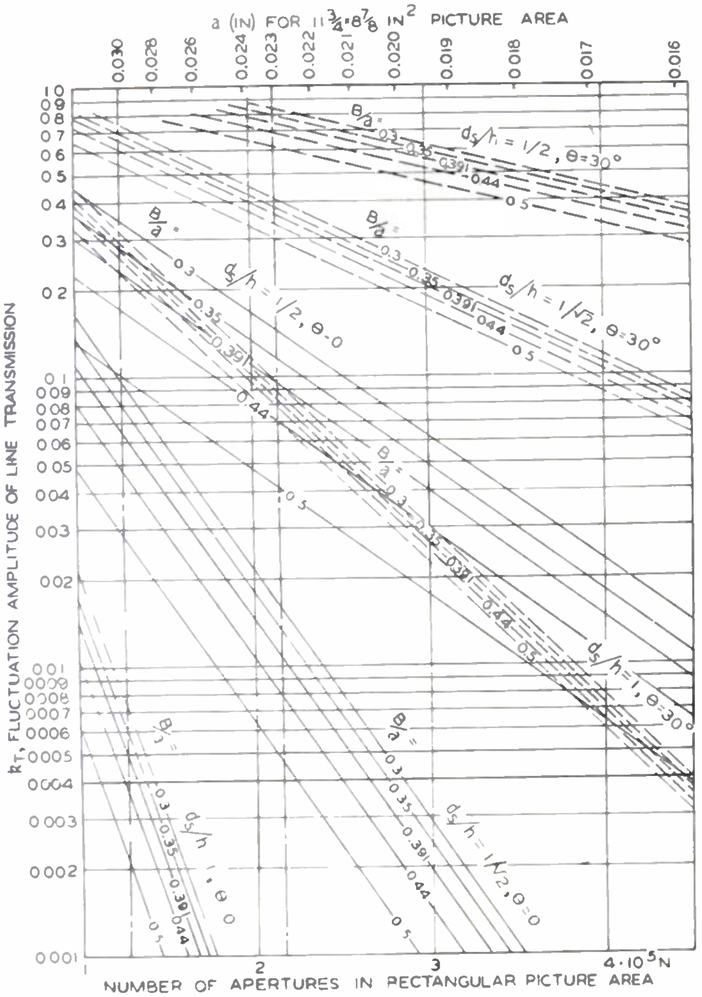


Fig. 7—Fluctuation amplitude of line transmission as function of number of apertures in mask.

Discussion of Results

The results for the moiré which is present with uniform signal are given by the data in Table I and Fig. 3. Table I shows that the amplitude of the moiré increases very rapidly as the scanning line is narrowed and more slowly as the mask apertures are narrowed. For a scanning line width of 0.0127 inch and an aperture diameter of 0.009 inch, for example, the modulation factor of the intensity is 0.25. With the larger aperture spacing (0.030 inch) this modulation has a period of 5 scanning-line separations for the preferred orientation of the scanning pattern relative to the mask (Fig. 1). The intensity variation is in the form of horizontal bars. In more detail, the intensity distribution given by the three terms with the largest periods in the range $\theta=0°$ to $\theta=30°$ may be represented by an intensity contour pattern, such as that shown in Fig. 8. Not much weight can be attached to the finer structure of the pattern since the higher-frequency components of the intensity distribution have been omitted. The predominant importance of the term $N_{-2,0}$, with a vertical periodicity of 5 scanning-line separations is, however, clearly evident.

As the scanning pattern is disoriented with respect to the mask, the system of horizontal bars also changes its orientation and contracts, eventually merging into a dot moiré with a periodicity of two to three line separations.

For the smaller aperture spacing (0.023 inch) the behavior is quite different. For the preferred orientation of the scanning pattern with respect to the mask, an unobjectionable dot moiré close to the limit of resolution (i.e., with periodicities of 2.0 and 1.8 scanning-line separations) exists. With disorientation this gives way to a bar mosaic which reaches a maximum period of over 10 scanning-line separations when the disorientation attains 30°. Whereas, for uniform signal, the moiré pattern becomes less objectionable with disorientation for the tube with wide dot spacing, it is scarcely evident in the tube with narrow dot spacing unless there is a considerable departure from the preferred orientation.

The fluctuation in the transmission of the mask for a scanning line with vertical displacement of the line, as given by Table III and Figs. 4, 6, and 7, may serve as an index for the degree to which the mask permits the reproduction of fine (vertical) picture detail. In effect,

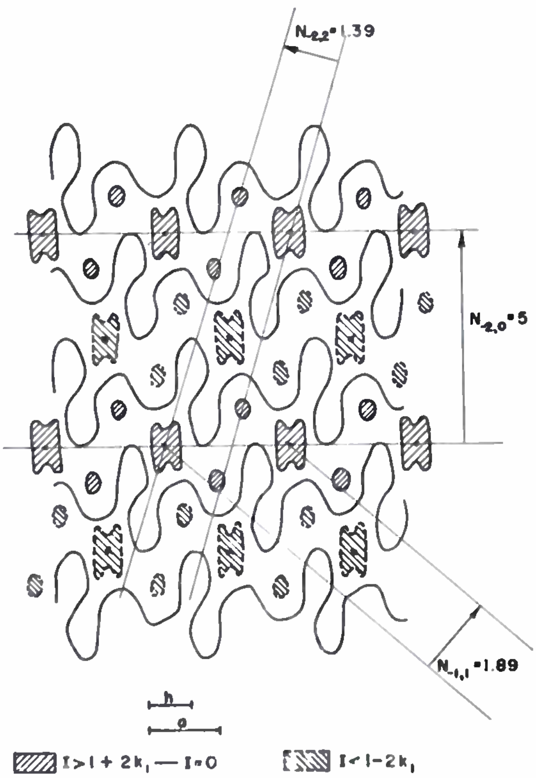


Fig. 8—Approximate intensity contours for uniform signal, computed for $a/h=1.667$, $\theta=0$ from the Fourier terms $(-2, 0)$, $(-1, 1)$, and $(-2, 2)$.

it establishes a lower limit for the contrast between the picture signals for two scanning lines which will assure qualitatively correct reproduction. Thus if the ratio of the two picture signals (or, more precisely, the resulting beam currents) exceeds $(1+k_T)/(1-k_T)$, the larger signal will always give rise to the brighter scanning line, the smaller signal to the less bright scanning line. If the signal contrast is less than this, it will happen, for certain positions of the scanning lines, that the mask effects an inversion of the order of light and dark. Whereas the amplitude factor k_T specifies the maximum amplitude in the intensity distortion which may be caused by the mask, its extent in any specific case must be determined individually, taking due account of the regular displacement of successive scanning lines with respect to the mask-aperture array.

The fluctuation in transmission rapidly increases as the scanning pattern is rotated with respect to the preferred position (Fig. 6). This corresponds simply to the resulting increase in the ratio of the spacing of successive aperture rows to the scanning-line width; the spacing of the aperture rows increases by a factor of $\sqrt{3}$ as the scanning pattern is rotated through 30°. Similarly for equal line width and total mask transmission, the fluctuation decreases rapidly as the aperture spacing is reduced. This is evident from Fig. 7, where the fluctuation amplitude is plotted as function of the number N of apertures in the image area (or the aperture spacing a) with the mask transmission (determined by B/a) and the relative line width d_s/h as parameters. From the point of view of obtaining the maximum sharpness in the image, it is pointless to lower d_s/h beyond $1/\sqrt{2}=0.71$, so that only the two lower sets of full and broken lines in Fig. 7 and the two lowest curves in Fig. 6 need be considered. It is seen that for 215,000 apertures[3] corresponding to an aperture spacing of 0.023 inch, and an aperture diameter of 0.009 in ($B/a=0.391$), the fluctuation amplitude for the preferred direction of scanning ($\theta=0$) is just about 1 per cent for $d_s/h=1/\sqrt{2}$, less for greater line widths. It is hence quite negligible. On the other hand, for the least favorable orientation of the scanning pattern ($\theta=30°$) it is as much as 35 per cent for $d_s/h=1/\sqrt{2}$. However, since the initial variation of the fluctuation amplitude with angle is quadratic, as shown by the curves in Fig. 6, a disorientation by at least 2° should be permissible before the fluctuation amplitude becomes even as much as 2 per cent.

Mathematical Appendix

I. Calculation of the Dot Intensity

If $I_0(c/h)$ is the intensity distribution of the scanning pattern for unit current falling or unit area,

$$I\left(\frac{c}{h}\right)=\frac{4}{a^2\sqrt{3}}\int_{-B/2}^{B/2}\sqrt{\left(\frac{B}{2}\right)^2-y^2}\, I_0\left(\frac{y-c}{h}\right)dy, \quad (6)$$

with

$$I_0\left(\frac{y}{h}\right)=\frac{h}{b\sqrt{\pi}}\sum_{m=-\infty}^{m=\infty}e^{-(y+mh)^2/b^2}$$

$$=1+\sum_{n=1}^{\infty}a_{0n}\cos\frac{2\pi ny}{h}$$

$$=1+2\sum_{n=1}^{\infty}e^{-\pi^2n^2b^2/h^2}\cos{2\pi ny/h}. \quad (7)$$

Substitution of (9) in (8) leads to

$$I\left(\frac{c}{h}\right)=\frac{\pi B^2}{2a^2\sqrt{3}}\left(1+\frac{4}{\pi}\sum_{n=1}^{\infty}e^{-\pi^2n^2b^2/h}\int_{-1}^{1}\sqrt{1-s^2}\cos\frac{\pi nBs}{h}ds\cos\frac{2\pi mc}{h}\right)$$

$$=\frac{\pi B^2}{2a^2\sqrt{3}}\left(1+\sum_{n=1}^{\infty}k_n\cos\frac{2\pi nc}{h}\right). \quad (8)$$

The Fourier coefficient in this equation becomes

$$k_n=2e^{-\pi^2n^2b^2/h^2}\sum_{m=0}^{\infty}\frac{(-1)^m}{(m+1)(m!)^2}\left(\frac{\pi Bn}{2h}\right)^{2m}. \quad (9)$$

II. Moiré for Uniform Signal and Angle θ between Scanning Lines and x-Axis

If the x-axis (Fig. 1) is imagined as fixed on the mask, the moiré is determined as follows: The intensity function I (3) for the individual dot, with c replaced by $y\cos\theta-x\cos\theta$, is multiplied by a Fourier expansion which represents the array of dot centers.

$$J(x,y;\theta)$$

$$=\frac{B^2}{2\sqrt{3}a^2}\left\{1+\sum_{n=1}^{\infty}k_n\cos\frac{2\pi n}{h}(y\cos\theta-x\sin\theta)\right\}$$

$$\cdot\left\{1+2\sum_{k=1}^{\infty}\cos\frac{4\pi ky}{a}+2\sum_{l=1}^{\infty}\cos\frac{4\pi lx}{a\sqrt{3}}\right.$$

$$\left.+2\sum_{k,l=1}^{\infty}(1+(-1)^{k+l})\cos\frac{2\pi ky}{a}\cos\frac{2\pi lx}{a\sqrt{3}}\right\}. \quad (10)$$

This formula assumes, in effect, that the intensity transmitted by each aperture is concentrated at the center of the corresponding dot. Since the dot size lies beyond the range of visual resolution, this leads to no erroneous conclusion regarding any visible, macroscopic, intensity fluctuations. If the dot structure itself is omitted and only k_1 is retained from (3) in view of the smallness of k_2 and higher terms, (10) assumes the form of (4) of the text, with the following values for the Fourier component parameters:

$$N_{kl}=\left\{\left(\frac{kh}{a}+\cos\theta\right)^2+\left(\frac{lh}{a\sqrt{3}}-\sin\theta\right)^2\right\}^{-1/2}$$

$$\tan\alpha_{kl}=\left(\frac{kh}{a}+\cos\theta\right)\Big/\left(\frac{lh}{a\sqrt{3}}-\sin\theta\right). \quad (11)$$

III. Calculation of the Transmission

If the scanning pattern is in the preferred orientation (Fig. 1) relative to the mask, the transmission of the mask for the scanning line is given by

[3] Since the framing mask cuts off approximately 10 per cent of the picture area (mainly at the corners), the corresponding number of apertures contributing to the framed picture is 195,000.

$$T\left(\frac{c}{a}\right) = \frac{2}{ab\sqrt{3\pi}} \sum_{n=-\infty}^{\infty} \int_{na-\beta/2}^{na+\beta/2} \sqrt{\left(\frac{B}{2}\right)^2 - \left(y - \frac{na}{2}\right)^2}\, e^{-(y-c)^2/b^2}.$$

$$= \frac{2}{ab\sqrt{3\pi}} \sum_{n=-\infty}^{\infty} \int_{-B/2}^{B/2} \sqrt{\left(\frac{B}{2}\right)^2 - y^2}\, e^{-(y-c+na/2)^2/b^2}. \tag{12}$$

A comparison of (12) with (8) shows that the transmission T is obtained from the dot intensity I if, in the latter, h is replaced by $a/2$. It follows, as already mentioned, that line transmission and dot intensity become identical for $a=2h$.

For purposes of integration it is convenient to introduce the two variables

$$\beta = 2c/a \quad \text{and} \quad \alpha = [B/(2b)]^2 = (0.834B/d_s)^2.$$

Then (12) becomes

$$T\left(\frac{\beta}{2}\right) = \frac{B}{a}\sqrt{\frac{\alpha}{3\pi}} \int_{-1}^{1} \sqrt{1-s^2} \sum_{n=-\infty}^{\infty} e^{-\alpha(s-\beta+na/B)^2} ds. \tag{13}$$

If the scanning line is inclined by an angle θ to the preferred direction, the transmission, calculated for the period shown in Fig. 5, becomes

$$T\left(\frac{B}{2}\right) = \frac{B}{a}\sqrt{\frac{\alpha}{3\pi}} \cos\theta \sum_{n=-\infty}^{\infty} \left\{ \int_{2na/B-1}^{2na/B+1} ds \int_{-\sqrt{3}a/B}^{-\sqrt{3}a/B+\sqrt{1-s^2}} d\xi + \int_{2na/B-1}^{2na/B+1} ds \int_{\sqrt{3}a/B-\sqrt{1-s^2}}^{\sqrt{3}a/B} d\xi \right\}$$

$$\cdot e^{-\alpha[(s-\beta)\cos\theta - \xi\sin\theta]^2} \tag{14}$$

with

$$s = \frac{2y}{B}, \qquad \xi = \frac{2x}{B}.$$

This integral can be brought into a form convenient for quadrature by introducing polar co-ordinates with origin at the centers of the several apertures over which the integration is carried out. The result is

$$T\left(\frac{\beta}{2}\right) = \frac{B}{2a}\sqrt{\frac{\alpha}{3\pi}} \cos\theta \left\{ \int_0^{\pi/2} \frac{d\phi}{\sin^2\phi} [F_1(\phi) + F_1(-\phi) + F_2(\phi) + F_2(-\phi)] + \int_{\pi/2-\theta}^{\pi/2} \frac{d\phi}{\sin^2\phi} [F_1(\phi) - F_1(-\phi) - F_2(\phi) + F_2(-\phi)] \right\} \tag{15}$$

with

$$F_j(\phi) = \sum_{n=-\infty}^{\infty} \left\{ \frac{1}{\alpha} \left[e^{-\alpha q_{nj}^2} - e^{-\alpha(q_{nj}+\sin\phi)^2} \right] + \sqrt{\frac{\pi}{\alpha}}\, q_{nj} [P(\sqrt{\alpha}\, q_{nj}) - P(\sqrt{\alpha}\,(q_{nj} + \sin\phi))) \right.$$

$$q_{n1} = \left[\beta - \frac{2a}{B}\left(n + \frac{1}{4}\right)\right] \cos\theta - \frac{\sqrt{3}}{2}\frac{a}{B}\sin\theta,$$

$$q_{n2} = \left[\beta - \frac{2a}{B}\left(n + \frac{3}{4}\right)\right] \cos\theta + \frac{\sqrt{3}}{2}\frac{a}{B}\sin\theta,$$

$$P(y) = \frac{2}{\sqrt{\pi}} \int_0^y e^{-x^2} dx.$$

Correction

H. H. Davids, of the General Electric Company, Electronics Park, Syracuse, N. Y., has brought the following errors to the attention of the editors.

On page 413 of the article entitled, "Radio Progress during 1951," which appeared in the April 1952 issue, there is a reference (389), T. S. Eader, "Equipment performance specs," *FM-TV*, vol. 11, pp. 22–24, 38–39; September, 1951. The author of this article should be H. H. Davids.

Similarly, in the section, "Abstracts and References," the author of the item (745), "A New Method of Predicting the Adjacent Channel Performance of Mobile Radio Equipment by Graphical Analysis," is also H. H. Davids, not T. S. Eader.

Standards on Gas-Filled Radiation Counter Tubes: Definitions of Terms, 1952*

COMMITTEE PERSONNEL

Standards Committee, 1951–1952

A. G. JENSEN, *Chairman*

M. W. BALDWIN, JR., *Vice Chairman* L. G. CUMMING, *Vice Chairman* ERNST WEBER, *Vice Chairman*

J. Avins
R. R. Batcher
H. G. Booker
J. G. Brainerd
M. R. Briggs
F. T. Budelman
P. S. Carter
A. G. Clavier
J. L. Dalke
A. G. Fox
F. J. Gaffney
W. D. Goodale, Jr.
R. A. Hackbusch
E. A. Laport
R. A. Miller
C. H. Page
W. M. Pease
W. J. Poch
A. F. Pomeroy
N. Rochester
H. E. Roys
A. L. Samuel
P. C. Sandretto
R. F. Shea
R. E. Shelby
R. A. Sykes
W. G. Tuller
J. P. Veatch
L. E. Whittemore
R. J. Wise

Definitions Co-ordinating Subcommittee

M. W. BALDWIN, JR., *Chairman*

P. S. CARTER
P. J. HERBST
E. A. LAPORT
ERNST WEBER—*Measurements Co-ordinator*

Committee on Electron Devices, 1947–1952

A. L. SAMUEL, *Chairman*

G. D. O'NEILL, *Vice Chairman* L. S. NERGAARD, *Past Chairman* R. S. BURNAP, *Past Chairman*

E. M. Boone,
J. W. Clark
K. C. DeWalt
W. J. Dodds
W. G. Dow
G. A. Espersen
C. E. Fay
W. R. Ferris
M. S. Glass
A. M. Glover
J. E. Gorham
J. W. Greer
L. B. Headrick
L. A. Hendricks
T. J. Henry
E. C. Homer
S. B. Ingram
R. B. Janes
J. W. Kearney
S. J. Koch
H. L. Krauss
D. E. Marshall
L. Malter
R. L. McCreary
J. A. Morton
I. E. Mouromtseff
O. W. Pike
P. A. Redhead
H. J. Reich
H. J. Reilly
A. C. Rockwood
R. M. Ryder
W. G. Shepherd
R. W. Slinkman
J. R. Steen
H. L. Thorson
C. M. Wheeler

Subcommittee on Gas Tubes

HARRY L. THORSON, *Chairman* D. E. MARSHALL, *Past Chairman*

J. H. Burnett
W. A. Depp
C. Greef
V. L. Holdaway
J. P. Hagen
E. J. Handley
R. A. Herring
J. D. Le Van
D. S. Peck
G. H. Rockwood
H. H. Wittenberg
Wm. Schmitt
W. W. Watrous
W. Widmairer

Subcommittee on Radiation Counter Tubes

D. E. MARSHALL, *Chairman*

N. G. Anton
H. Bryant
R. L. Butenhoff
D. L. Collins
H. J. DiGiovanni
T. Fahrner
T. Fields
A. M. Glover
L. T. Jansen
S. A. Korff
J. B. H. Kuper
S. H. Liebson
G. L. Locher
W. H. Metten
E. Osterland
W. E. Posey
D. Rigney
J. A. Schoke
E. H. Wakefield
L. C. White

Avalanche. A cascade multiplication of ions.

Background Counts. Counts caused by radiation coming from sources other than that measured.

Count (in a Radiation Counter). A single response of the counting system.

Note—See also **Tube Count.**

Counting-Rate Versus Voltage Characteristic. Counting rate as a function of applied voltage for a given constant average intensity of radiation.

Dead Time. The time from the start of a counted pulse until an observable succeeding pulse can occur.

Note—See also **Recovery Time.**

Efficiency (of a Radiation Counter Tube). The probability that a tube count will take place with a specified particle or quantum incident in a specified manner.

Externally Quenched Counter Tube. A radiation counter tube that requires the use of an external quenching circuit to inhibit reignition.

* Reprints of this Standard, 52 IRE 7.S3, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.

Gas Amplification. The ratio of the charge collected to the charge liberated by the initial ionizing event.

Note—See also Standards on Gas-Filled Radiation Counter Tubes: Methods of Testing, section 12.

Gas-Filled Radiation Counter Tube. A gas tube, in a radiation counter, used for the detection of radiation by means of gas ionization.

Geiger-Mueller Counter Tube. A radiation counter tube designed to operate in the Geiger-Mueller region.

Geiger-Mueller Region (of a Radiation Counter Tube). The range of applied voltage in which the charge collected per isolated count is independent of the charge liberated by the initial ionizing event.

Geiger-Mueller Threshold. The lowest applied voltage at which the charge collected per isolated tube count is substantially independent of the nature of the initial ionizing event.

Hysteresis (of a Radiation Counter Tube). The temporary change in the counting rate versus voltage characteristic caused by previous operation.

Initial Ionizing Event. An ionizing event that initiates a tube count.

Ionizing Event. Any interaction by which one or more ions are produced.

Multiple Tube Counts (in Radiation Counter Tubes). Spurious counts induced by previous tube counts.

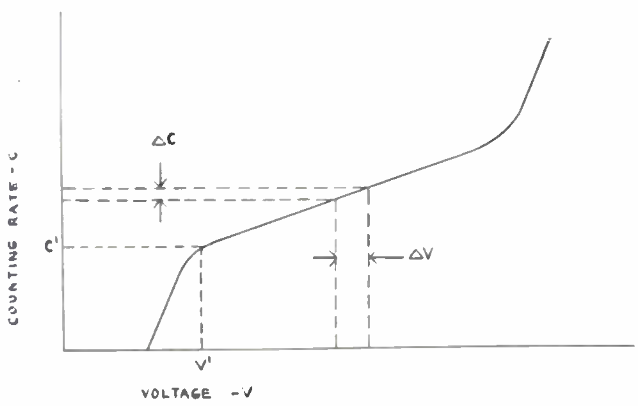


Fig. 1—Counting rate-voltage characteristic, in which

$$\text{Relative plateau slope} = 100\,\frac{\Delta C/C}{\Delta V}$$

$$\text{Normalized plateau slope} = \frac{\Delta C/\Delta V}{C'/V'} = \frac{\Delta C/C'}{\Delta V/V'}$$

Normalized Plateau Slope. The slope of the substantially straight portion of the counting rate versus voltage characteristic divided by the quotient of the counting rate by the voltage at the Geiger-Mueller threshold.

Note—See Fig. 1.

Overvoltage. The amount by which the applied voltage exceeds the Geiger-Mueller threshold.

Plateau. The portion of the counting rate versus voltage characteristic in which the counting rate is substantially independent of the applied voltage.

Plateau Length. The range of applied voltage over which the plateau of a radiation counter tube extends.

Predissociation. A process by which a molecule that has absorbed energy dissociates before it has had an opportunity to lose energy by radiation.

Proportional Counter Tube. A radiation counter tube designed to operate in the proportional region.

Proportional Region. The range of applied voltage in which the gas amplification is greater than unity and is independent of the charge liberated by the initial ionizing event.

Note—The proportional region depends on the type and energy of the radiation.

Quenching (in a Gas-Filled Radiation Counter Tube). The process of terminating a discharge in a radiation counter tube by inhibiting reignition.

Radiation. In nuclear work, the term is extended beyond its usual meaning to include moving nuclear particles, charged or uncharged, and electrons moving with sufficient speed to enter into nuclear processes.

Radiation Counter. An instrument used for detecting or measuring radiation by counting action.

Recovery Time (of a Radiation Counter). The minimum time from the start of a counted pulse to the instant a succeeding pulse can attain a specific percentage of the maximum value of the counted pulse.

Region of Limited Proportionality. The range of applied voltage below the Geiger-Mueller threshold, in which the gas amplification depends upon the charge liberated by the initial ionizing event.

Reignition (of a Radiation Counter Tube). A process by which multiple counts are generated within a counter tube by atoms or molecules excited or ionized in the discharge accompanying a tube count.

Relative Plateau Slope. The average percentage change in the counting rate near the mid-point of the plateau per increment of applied voltage.

Note—Relative plateau slope is usually expressed as the percentage change in counting rate per 100-volt change in applied voltage. See Fig. 1.

Resolving Time (of a Radiation Counter). The time from the start of a counted pulse to the instant a succeeding pulse can assume the minimum strength to be detected by the counting circuit. (This quantity pertains to the combination of tube and recording circuit.)

Self-Quenched Counter Tube. A radiation counter tube in which reignition of the discharge is inhibited by internal processes.

Sensitive Volume (of a Radiation Counter Tube). That portion of the tube responding to specific radiation.

Spurious Tube Counts (in Radiation Counter Tubes). Counts in radiation counter tubes other than background counts and those caused by the source measured.

Note—Spurious counts are caused by failure of the quenching process, electrical leakage, and the like. Spurious counts may seriously affect measurement of background counts.

Tube Count. A terminated discharge produced by an ionizing event in a radiation counter tube.

Standards on Gas-Filled Radiation Counter Tubes: Methods of Testing, 1952*

1. GENERAL

The Committees responsible for the following standards are shown under the title "Standards on Gas-Filled Radiation Counter Tubes: Definitions of Terms, 1952," on page 924 of this issue.—*The Editor.*

1.1 Environment

In setting up test procedures for radiation counter tubes, it is necessary to consider not only the test circuits but also the physical arrangements of test, such as nature of source, distance between source and counter, scattering by objects near the source or counter, shielding of the counter against unwanted radiations, and the background due to cosmic rays and general contamination. The relative importance of these various factors will depend on the type of counter under test.

In selecting a room for counter-testing purposes, it is advisable to pick one as far as possible from strong gamma-ray sources. It is particularly difficult to work in a room adjacent to one in which experiments are being performed involving the moving of a gamma-ray source from one location to another. A corner room is desirable from this standpoint.

Attention should be paid to the fact that the intensity of cosmic radiation depends strongly on the altitude and on the amount of material above the counters. Also, in testing counter tubes that are physically long there may be a noticeable difference in the background counting rate, depending on whether they are tested in a horizontal or vertical position. When measuring background counting rates, it is preferable that the tube be in a lead shield to reduce the effects of local contamination.

1.2 Radiation Sources[1]

A source is standardized with respect to the total number of disintegrations per unit time, and it is necessary to compute the fraction of the emitted radiation intercepted by the sensitive portion of the counter tube. In the case of the alpha and beta sources, the material and thickness of the source support and/or matrix and the distance from the counter are important in this calculation since particles may reach the counter after having been scattered through a large angle by the support material. Also, the thickness of the source itself must be considered because of self-absorption and the attendant straggling.[2,3]

These factors are not important in relative measurements if the distance from source to counter and disposition of surrounding material remain unchanged. However, self-absorption may make it impossible to get a good plateau on an alpha proportional counter.

Where gamma-ray sources can be used for calibration purposes, their use will be convenient because sufficient shielding of the counter to minimize interference from softer background radiation is possible. However, counters intended for beta or neutron detection should be tested on appropriate sources.

1.2.1 Gamma-Ray Sources

For gamma rays it is customary to use a source consisting of a radium salt of sufficient age (greater than one month) to insure that equilibrium has been attained with the decay products responsible for the greater part of the gamma radiation. Because of the long life of radium, a source of this type will remain constant in time, but for some purposes the complexity of the gamma-ray spectrum may be objectionable. In this case, it will probably be necessary to use an artificially radioactive isotope. The problem then is to find an isotope emitting the desired radiation and possessing

* Reprints of this Standard, 52 IRE 7.S2, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.75 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.

[1] It is suggested that those who propose to handle radiation sources familiarize themselves with the possible hazards involved. See, for example, the National Bureau of Standards Handbook 42, "Safe Handling of Radiation Isotopes."

[2] A. H. Jaffee, T. P. Kohman, and J. A. Crawford, "A Manual on the Measurement of Radioactivity," MDDC-388, U.S.A.E.C., 36 pp.

[3] D. W. Wilson, A. O. C. Nier, S. P. Reimann, "Preparation and Measurement of Isotopic Tracers," J. W. Edwards, Ann Arbor, Mich. 108 pp.; 1946.

a sufficiently long life to avoid the nuisance of daily correction for decay of the source.

One such isotope, cobalt 60, has a half life of five years and decays with a successive emission of a soft beta ray followed by two gamma rays of 1.1 and 1.3 mev, respectively. This substance is frequently used for counter standardization purposes. The National Bureau of Standards furnishes weak radium sources in the form of glass ampules containing a standard quantity of radium in solution, and furnishes cobalt 60 standards as well.

Discussion of the units customarily used in measurements of radioactivity is contained in the paper "Radioactive Units and Standards," by Evans.[4] This paper also carries an extensive bibliography.

In testing counters with gamma rays, it is important to use sufficient distance between the source and counter so that all parts of the volume of the counter receive approximately equal amounts of radiation.

1.2.2 Alpha- and Beta-Ray Sources

The Bureau of Standards supplies sources containing a standard amount of lead 210+ bismuth 210 (radium $D+E$) which can be used conveniently for testing the ordinary beta-ray counters, but counters with specially thin windows for detecting soft radiation may require the use of carbon 14 or sulphur 35. For a long-lived alpha particle source, either plutonium or uranium is satisfactory, but the security restrictions on these substances may make them impractical for general use. In this case, polonium 210 (radium F) may be satisfactory in spite of its rather short half life (138 days).

1.2.3 Statistics of Counting

Since any radioactive decay is a random process, the accuracy of determining the counting rate depends on the total number of counts recorded, and it is necessary to state the total number of counts or the statistical accuracy in all cases. Assuming a Poisson distribution (which is ordinarily safe in work involving radioactivity since the fraction of atoms disintegrating and detected per unit time is usually a small fraction of the whole), the fractional standard deviation of the result is the reciprocal of the square root of the number of counts. Where the number of observations is sufficient so that the distribution around the mean shows a normal Gaussian curve, the probable error, which in this case is 0.67 times the standard deviation, may be used. For a derivation of the Poisson distribution and a discussion of statistical errors and the smoothing effect of scalers, see the articles on probability theory as applied to particle detection by Rainwater and Wu in the October, 1947 and January, 1948 issues of *Nucleonics*.[5]

[4] R. D. Evans, "Radioactive units and standards," *Nucleonics*, vol. 1, no. 2, p. 32; October, 1947.

[5] L. J. Rainwater and C. S. Wu, "Applications of probability theory to nucleonic particle detection," pt. I, *Nucleonics*, vol. I, no. 2, pp. 60–69; October, 1947; pt. II, *Nucleonics*, vol. II, no. 1, pp. 42–49; January, 1948.

1.3 Counter-Tube Output Circuits

The circuits of Fig. 1 indicate the recommended methods of coupling the counter tube to subsequent circuits.

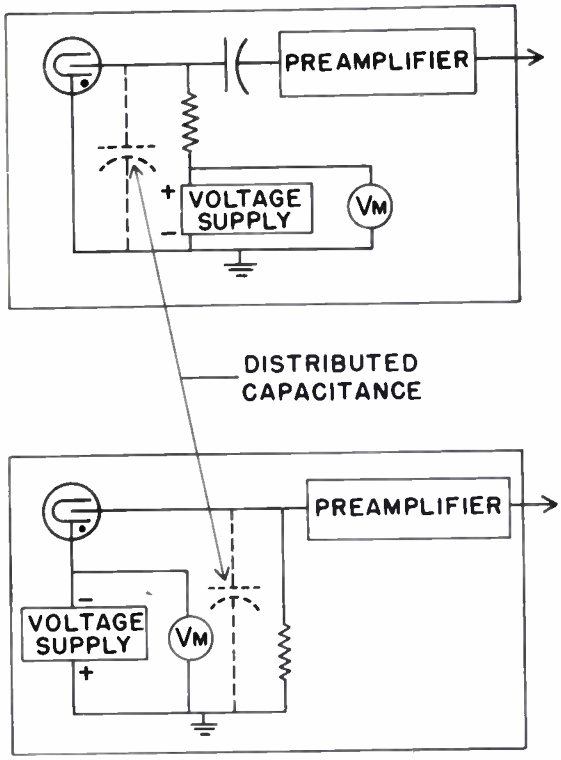


Fig. 1—Counter-tube output circuits.

2. COUNTING-RATE MEASUREMENTS

In counting-rate measurements, the parameters of the measuring circuits, the resolving time of the counting equipment, and the setting of the pulse-height discriminator are most important in determining statistical accuracy, and must be stated if the data are to be significant.

The counter tube may be connected as shown in Fig. 2. The pulses received from the counter tube are usually amplified, scaled down, and registered. One of two methods may be used to determine the counting rate: (1) The number of counts obtained within a predetermined time interval is recorded. (2) The time is observed for the accumulation of a predetermined number of counts.

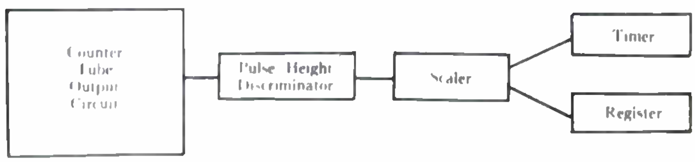


Fig. 2—Circuit for counting rate measurements.

2.1 Background Counting Rate

The counting rate of the radiation counter tube at a specific operating voltage is measured with no source within effective range. Because the background can be decreased by the use of sufficient shielding to minimize the softer components of local radiation and because the cosmic-ray component of the background varies with altitude and shielding above the counter and counter orientation, it is necessary to specify the test conditions when citing the background count.

2.2 Counting Rate Versus Voltage

The counting rate as a function of voltage may be measured by a point-by-point method using the circuit of Fig. 2. The radiation through the counter tube must be held constant throughout the test.

2.3 Automatic Curve-Tracer Method

Alternatively, this characteristic may be recorded automatically by the circuit of Fig. 3. In this circuit the shaper produces pulses of desired charge content which are integrated by the counting-rate meter. The current output actuates a continuous recording device whose chart displacement is directly proportional to voltage increments.

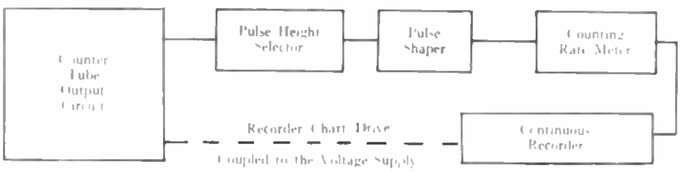


Fig. 3—Circuit for automatically recording counting rate versus voltage.

2.4 Operating Plateau Length

In practice, the operating plateau length is the voltage range (Fig. 4) over which the counting rate remains within stated limits, and is measured on the curve obtained from measurement 2.2 or 2.3.

Note—Ordinarily, a sufficient indication of satisfactory quality of a counter tube will be given by the fact

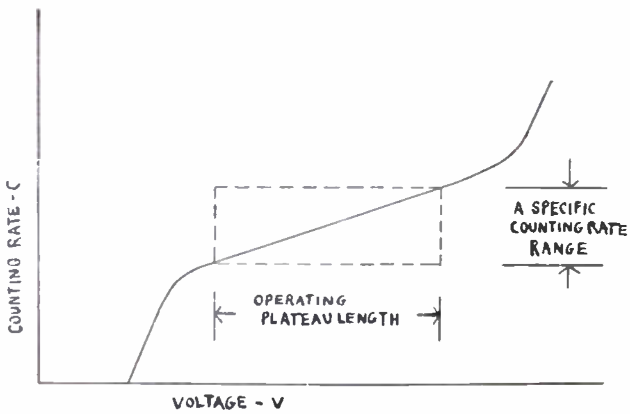


Fig. 4—Operating plateau length.

that the counting rate remains within specified limits over a specified range of operating voltage under constant radiation flux.

2.5 Plateau Slope

The relative plateau slope and the normalized plateau slope can be calculated from the counting rate versus voltage data obtained from test 2.2 or 2.3.

3. MEASUREMENT OF MULTIPLE COUNTS

Multiple counts are assumed to be present when the proportion of short time intervals between counts is greater than that to be expected from a Poisson or Bernoulli distribution. See Section 1.2.3. The number of short time intervals between counts may be measured in several ways. An example is illustrated in Fig. 5.

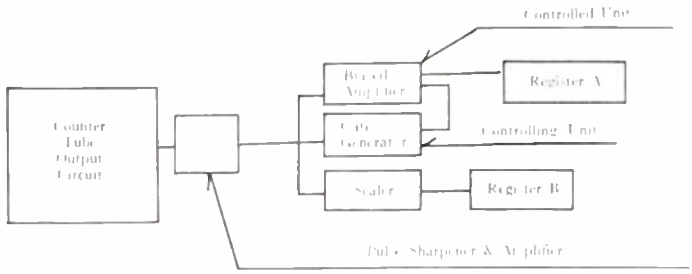


Fig. 5—Circuit for the measurement of multiple counts.

The output of the counter is connected to a pulse sharpening stage and amplified. The amplified voltage pulses are fed to a scaler and two units, one of which controls the other. The controlling unit consists of a gate generator which produces a rectangular dc voltage for a given time after the pulse is received. This dc voltage is applied, as a positive bias, to the controlled unit, which is a biased amplifier stage, or coincidence circuit. The output of the controlled amplifier is fed to a register. When a count occurs, the gate circuit is triggered, permitting operation of the biased amplifier for a selected time interval after the pulse occurs. Any pulses that occur in this interval are recorded on register *A*. The controlled amplifier then remains insensitive until the next count occurs. A comparison is made of the counting rates of registers *A* and *B*. The test voltage and gate width should be recorded.

3.1 Average Charge Per Isolated Tube Count

A meter is connected in series with the counter-tube voltage supply in the circuit of Fig. 6. A suitable circuit

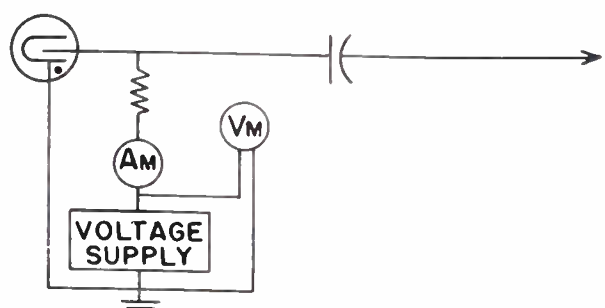


Fig. 6—Circuit for determining average charge per isolated tube count.

is used to record the counts. With the tube counting at the desired voltage and at a rate low with respect to the reciprocal of the counter-tube dead time, the current versus counting rate is determined and the charge per count calculated.

3.2 Charge Per Isolated Tube Count

Using the circuit of Fig. 7 with a calibrated amplifier nd cathode-ray oscillograph, and with the time contant RC large compared with the pulse duration, and mall compared with the pulse interval, the charge per ount can be computed from the peak voltage across he capacitor as measured on the oscillograph.

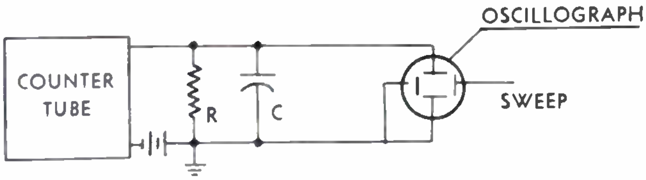


Fig. 7—Circuit for determining charge per isolated tube count.

4. DETERMINATION OF THE GEIGER-MUELLER THRESHOLD

4.1 Oscillograph Method

The Geiger-Mueller threshold may be measured using he apparatus of test 3.2. The operating voltage is varied intil the minimum voltage at which all pulses appear o be of the same charge is found.

4.2 Graphical Method

Alternatively, the threshold may be determined from in average charge per isolated tube-count measurement by plotting charge per count as a function of voltage on a linear scale, starting with a voltage high on the plateau and reducing the voltage until the average charge per tube count is no longer linear with respect to the voltage. The operating voltage at this point is the Geiger-Mueller threshold.

In either method, it is advantageous to use a source of heterogeneous radiation.

5. DETERMINATION OF SENSITIVE VOLUME

The sensitive volume may be explored by scanning the counter with a collimated beam of radiation of specified energy. The data may be presented as curves of counting range versus position of the beam. The sensitive volume will vary with the type of radiation used in the measurement and the applied voltage, both of which should be stated as conditions of the test.

6. TEMPERATURE DEPENDENCE

Since the threshold voltage, plateau length, plateau slope, and other characteristics may all vary with temperature, it is desirable to measure the temperature dependence of the counter tube by taking counting rate versus voltage data at various temperatures.

7. LIFE TEST

7.1 Normal Life Test

The counter tube is operated at the desired operating voltage with a source of radiation in such proximity to the tube under test as to cause it to count at the desired rate.

7.2 Recommended Life-Test End Point

When the counting rate over the specified operating voltage range under the standard radiation flux no longer falls within the desired operating range, the tube shall be considered to have reached its life end point. (See test and note for plateau length.)

The life of the tube is commonly expressed in terms of number of counts recorded.

7.3 Accelerated Life Test

The tube is operated in the corona discharge region with enough series resistance to limit the tube current to a value equal to that which would be obtained at a given counting rate at the desired operating voltage. The counting-rate voltage characteristic is measured at intervals during the test. This test should not be used unless correlation data with normal life test are available. No satisfactory method of accelerating the life test of proportional counters has been proposed.

8. PULSE MEASUREMENTS

The counter tube is placed in a circuit similar to that of Fig. 8. The output current of the tube is amplified by the pulse amplifier. The sweep of the oscillograph is triggered at the beginning of the count and the delay

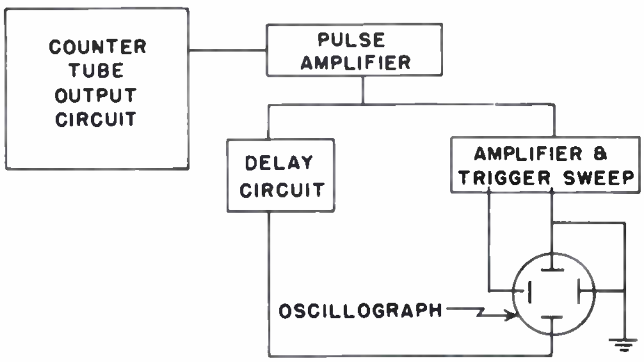


Fig. 8—Circuit for pulse measurements.

circuit insures that the pulse appears on the oscillograph screen in its entirety. The horizontal sweep of the oscillograph is calibrated with an audio oscillator. From the pattern on the oscillograph screen, the rise time, the recovery time, the dead time, and the resolving time may be determined. A typical pattern is shown in Fig. 9. The pulse shown by the heavy line results from the superposition of a great many pulses. The pulses (a), (b), and (c) shown by lighter lines result from photons or ionizing particles which arrive at the counter before the counter and the oscillograph sweep circuit have recovered completely from a previous count.

8.1 Rise Time

The rise time is defined as the time interval between instants at which the instantaneous amplitude first reaches specified lower and upper limits, namely, 10 per cent and 90 per cent of the peak-pulse amplitude unless otherwise specified. In counter-tube work, the specified upper and lower limits are as shown in Fig. 9.

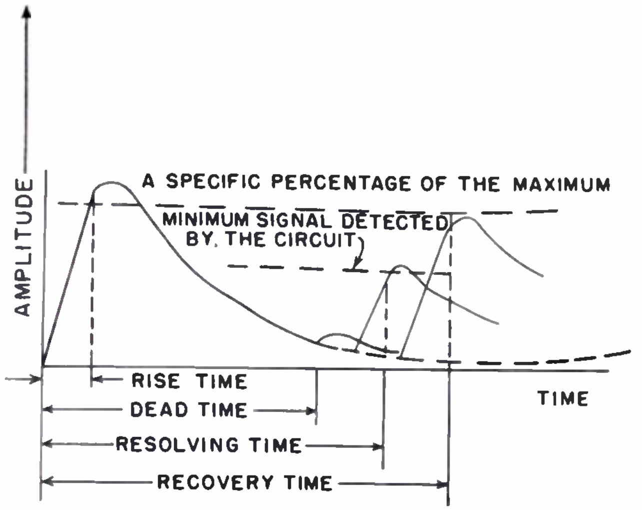


Fig. 9—Graphical representation of the terms defined in sections 8.1–8.4 inclusive.

8.2 Recovery Time

The recovery time is defined as the minimum time from the start of a counted pulse to the instant a succeeding pulse can attain a specific percentage of the maximum value of the counted pulse (Fig. 9).

Note 1—The present definition and measurement of recovery time differ from those proposed previously in that the results are independent of the gain of the amplifier used in the measurement.[6]

Note 2—It is recommended that "specific percentage" be $100\,(1-1/\epsilon)$ times full maximum value.

8.3 Dead Time

The dead time is defined as the time from the start of a counted pulse until an observable succeeding pulse can occur. (This quantity pertains to the combination of tube and recording circuit.) (Fig. 9.)

8.4 Resolving Time

The resolving time is defined as the time from the start of a counted pulse to the instant a succeeding pulse can assume the minimum strength to be detected by the counting circuit. (This quantity pertains to the combination of tube and recording circuit.) (Fig. 9.)

Resolving time may also be determined by measuring the counting rate, using two comparable separate sources separately and together. The details of the method are described in "A Precision Method of Measuring Geiger Counter Resolving Time," by Beers.[7]

9. HYSTERESIS

Hysteresis of a counter tube is indicated by a change in counting rate versus voltage characteristic when traversed in opposite directions, or under certain circumstances by a change in background counting rate as a result of a given history.

10. OVER-ALL EFFICIENCY

The efficiency of a counter tube depends not only on its construction and the operating voltage but also on the type and energy of the radiation being measured, the particular area of the tube on which radiation is incident, and the angle of incidence of the radiation. Hence, the test data should include electrical conditions of test, the type and energy of the radiation, the particular area of the tube on which radiation is incident, and the angle of incidence.

To measure the over-all efficiency of a radiation counter tube, place it in the desired position with respect to a standardized source of the particular radiation to be detected. The efficiency is, for the time of observation, the observed number of counts, minus the number of background counts, divided by the number of particles or quanta incident on the particular tube area.

11. PHOTOELECTRIC EFFECT

Photoelectric effect in counter tubes is indicated by a change in background counting rate on exposure of the tube to light of the desired incident intensity and spectral range.

12. GAS AMPLIFICATION OF A PROPORTIONAL COUNTER TUBE

With a radiation source present and starting at a low voltage on the counter tube, so that the gas amplification is equal to unity, measure the current with a shunted electrometer or other suitable means (Fig. 10). Raise the voltage to a desired value in the operating region and again measure the current. The gas amplification is the ratio of these currents.

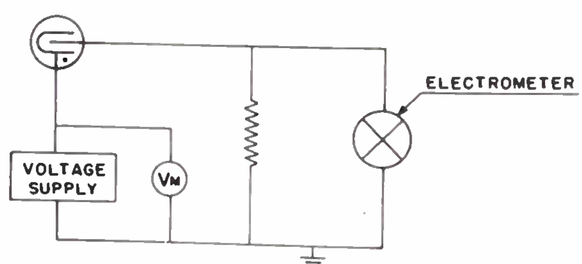


Fig. 10—Circuit for measuring gas amplification.

13. ELECTRICAL LEAKAGE

Geiger-Mueller tubes may be tested for electrical leakage by measuring current below threshold voltage with no radiation source present. Other types may need to be tested before being filled with gas or vapor.

14. INDUCED RADIOACTIVITY

Counter tubes may become radioactive as a result of neutron capture. This effect will be observed as an increase in background counting rate after exposure to a source of neutrons.

[6] H. G. Stever, *Phys. Rev.*, vol. 61, p. 40; 1940.
[7] Y. Beers, "A precision method of measuring Geiger counter resolving time," *Rev. Sci. Instr.*, vol. 13, p. 72; 1942.

Miniature Rectifier Computing and Controlling Circuits*

AN WANG†, SENIOR MEMBER, IRE

Summary—General types of rectifier computing circuits are described and analyzed. Stress is placed upon the realization of the highest speed of operation. Selenium-rectifier elements are used throughout, although the same analysis applies to all other types of rectifiers. A speed of 100 kc obtainable from selenium rectifiers. Megacycle operation is possible. Special techniques of circuit construction are developed which result in a simple, compact, and economical system, applicable not only to large-scale digital computing machines but also to complicated control systems.

INTRODUCTION

THE USE of rectifiers as elements in gating and switching circuits is well known. It is based on the fact that the ratio of the backward and forward impedances is very high.

Germanium crystal rectifiers have this property: Their usefulness in gating circuits has already been established.[1,2] Selenium rectifiers, which can be obtained in units that are smaller and cheaper than germanium rectifiers, should not be overlooked. Units consisting of many individual selenium rectifiers can be made very small and compact with the techniques described below. Although the shunt capacitance of selenium rectifiers determines an upper limit of the frequency at which they can be operated, reliable operation at 100 kc has been obtained by careful circuit design in accordance with the formulas developed below.

The analysis which follows was undertaken in connection with selenium-rectifier applications.[3] However, the analysis is general and can be applied to other types of rectifiers.

BASIC RECTIFIER CIRCUITS

Fig. 1 shows the two basic circuits. X and Y will take high or low voltages of E_0 or E_1 to represent 1 or 0, respectively; f_1 will be high if either X or Y is high and f_2 will be high only when both X and Y are high. Symbolically, the rectifier circuits can be written as follows:

$$f_1(X, Y) = 1 - X'Y', \qquad f_2(X, Y) = XY$$
$$= B_2(X, Y) \qquad\qquad = F_2(X, Y), \quad (1)$$

where

$$X' = 1 - X, \; Y' = 1 - Y, \text{ and so on.}$$

Assume here that the forward resistance of the rectifier is much smaller than the load resistance and the back resistance of the rectifier is much larger than the load resistance.

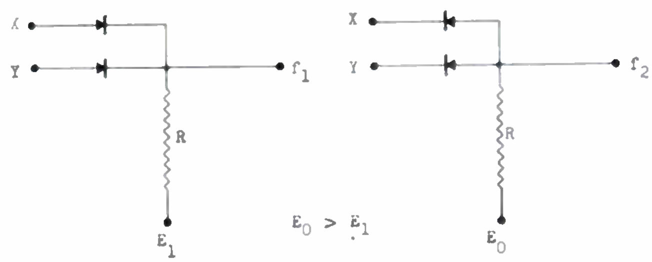


Fig. 1—Basic rectifier computing circuits.

It can be shown[4] that any function of n variables can be represented by either of the two following expressions:

$$f(X_1, X_2, \cdots, X_n) = B_m(F_n^i, F_n^j, \cdots, F_n^q) \qquad (2)$$

$$f(X_1, X_2, \cdots, X_n) = F_{\nu-m}(B_n^i, B_n^j, \cdots, B_n^r), \qquad (3)$$

where F_n^i, F_n^j are terms of the canonical form for $f(X_1, X_2, \cdots, X_n)$ and B_n^i, B_n^j, . . . are terms of the canonical form for $f'(X_1, X_2, \cdots, X_n)$.

The subscript n is the number of input variables. The superscripts $i, j, \ldots$ are the decimal equivalent of the binary number formed when the primed and the unprimed variables are replaced by 0 and 1, respectively. For example,

$$X_1', X_2, X_3, X_4' = F_4^6(X_1, X_2, X_3, X_4).$$

Here 0110 is a decimal equivalent of 6 in binary form. The corresponding basic circuits can be drawn as shown in Fig. 2.

Fig. 2(a) shows a circuit which represents

$$f_1(X_1, X_2, X_3, X_4) = F_4(B_4^0, B_4^3, B_4^5, B_4^6).$$

Fig. 2(b) shows a circuit which represents

$$f_2(X_1, X_2, X_3, X_4) = B_4(F_4^5, F_4^9, F_4^{12}, F_4^{15}).$$

With either of these two basic rectifier circuits any function can be realized.

In the circuit shown in Fig. 2(a) the current delivered to an input line is much larger when the corresponding input voltage is high than when it is low. In the circuit shown in Fig. 2(b) the reverse is true. The circuit of Fig.

* Decimal classification: R366.35. Original manuscript received by the Institute, June 20, 1951; revised manuscript received April 8, 1952.

† Wang Laboratories, 296 Columbus Ave., Boston, Mass.

[1] C. F. West and J. E. DeTurk, "A digital computer for scientific applications," PROC. I.R.E., vol. 36, pp. 1452–1460, December, 1948.

[2] T. C. Chen, "Diode coincidence and mixing circuits in digital computers," PROC. I.R.E., vol. 38, pp. 511–514, May, 1950.

[3] "Investigation for Design of Digital Calculating Machinery," Progress Report No. 10, Computation Laboratory, Harvard University, Cambridge, Mass., chaps. 2 and 3; May–August, 1950.

[4] "Investigation for Design of Digital Calculating Machinery," Progress Report No. 5, Computation Laboratory, Harvard University, Cambridge, Mass., chap. 2; May–August, 1949.

2(a) is therefore much better suited for use with a cathode-follower driver than the circuit of Fig. 2(b), since the output impedance of a cathode follower is lower when it is conducting than when it is cut off. For this reason, the remainder of the analysis will be concerned

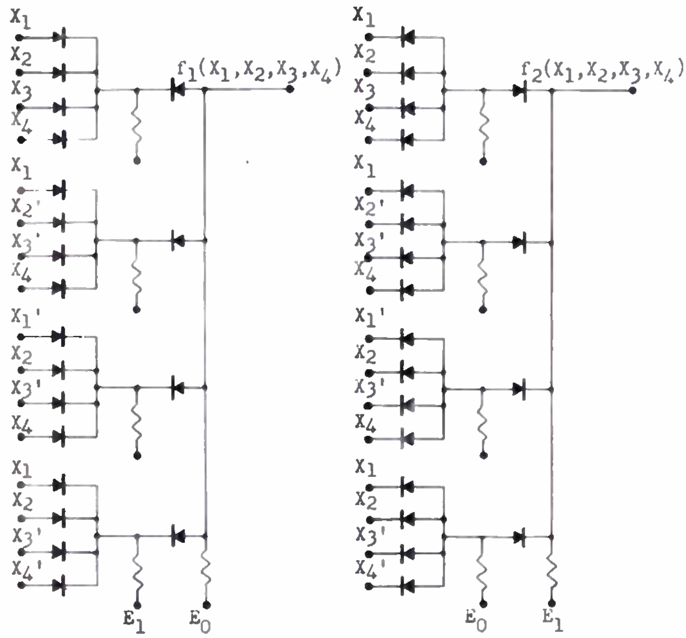


Fig. 2 (a) and (b)—Basic rectifier computing combinations.

with the circuit of Fig. 2(a) only. The analysis for the circuit of Fig. 2(b) would be similar, the essential differences being reversals in the polarities of voltages.

Analysis of the DC Characteristic of the Circuit

Consider the circuit shown in Fig. 3. The input voltages are either E_0 or E_1, and $E_0 > E_1$. Let R_f be the forward resistance of the rectifier and R_b its back resistance.

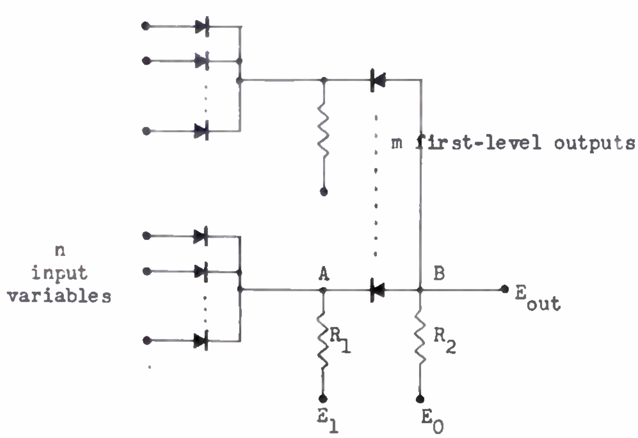


Fig. 3—General rectifier circuit.

In order for the output E_{out} to be high, all the n first-level outputs must be high. To make a first-level output high, any one of its inputs must be high. More high inputs will make the first-level output higher, and therefore need not be considered. The equivalent circuit is as shown in Fig. 4.

Assuming

$$R_f \ll R_1, R_2$$

$$R_b \gg R_1, R_2,$$

$$E_{out} \doteq E_0 - (E_0 - E_1)\frac{R_f}{R_f + \dfrac{R_1R_b}{R_b + (n-1)R_1}}. \quad (4)$$

In order for the output E_{out} to be low, any one of the n first-level outputs must be low. E_{out} is made lower as

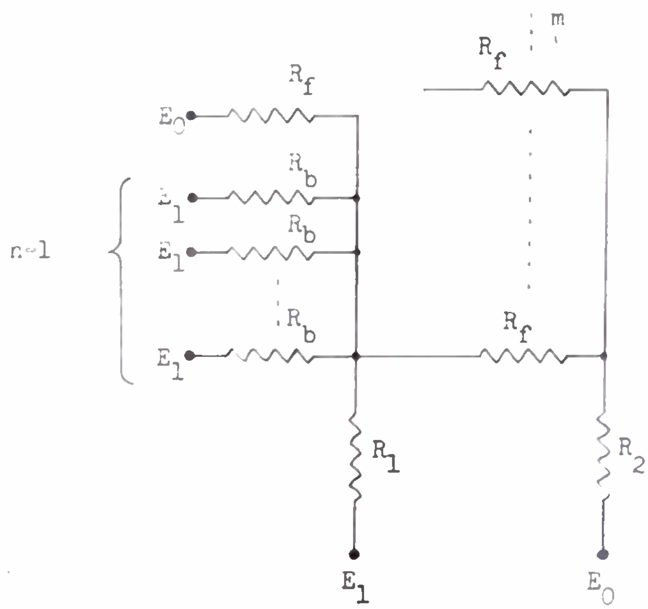


Fig. 4—Equivalent rectifier circuit corresponding to high output.

additional first-level outputs are made low. Therefore, the analysis need only cover the case when one of the m first-level outputs is low and all the others are at their highest possible voltage. This occurs when the corresponding first-level inputs are all high. Under these conditions the equivalent circuit is as shown in Fig. 5.

$$E_{out} \doteq E_1 + (E_0 - E_1)\frac{R_f + R_1}{R_f + R_1 + \dfrac{R_2R_b}{R_b + R_2(m-1)}}. \quad (5)$$

The difference of these two outputs is the minimum difference between high and low outputs.

$$E_{out} - E_{out}' = (E_0 - E_1)\left\{1 - \frac{R_f}{R_f + \dfrac{R_1R_b}{R_b + (n-1)R_1}} - \frac{R_f + R_1}{R_f + R_1 + \dfrac{R_2R_b}{R_b + R_2(m-1)}}\right\}. \quad (6)$$

The factor in the parentheses of the right side of (6) determines the attenuation of the signal through the rectifier network. It is desirable that this attenuation be kept small. This suggests the following relation:

$$\frac{R_b}{m-1} \approx R_2 > R_1 > R_f.$$

Since the ratio of R_b/R_f is limited by the rectifier characteristics, the attenuation is usually rather high es-

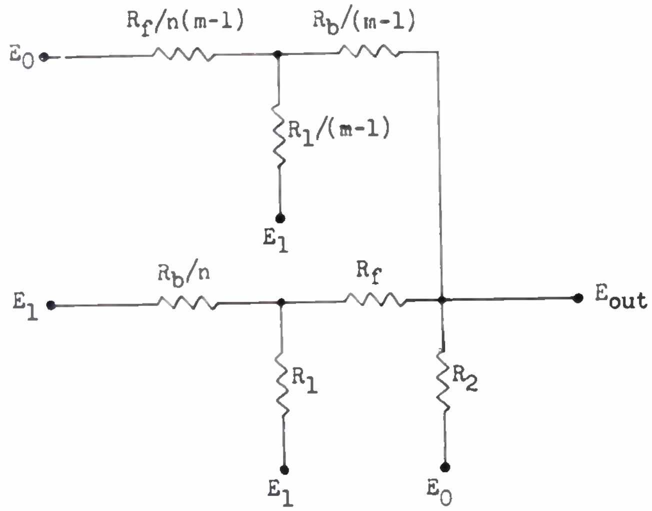


Fig. 5—Equivalent rectifier circuit corresponding to low output.

pecially when m is large. In practical cases m may often be very much larger than n. This attenuation can be largely eliminated with the improved circuits described later.

Transient Analysis

Because of the shunt capacitance associated with selenium rectifiers, circuits using these rectifiers must be carefully designed if the maximum possible operating frequencies are to be obtained. The remainder of the analysis will deal with the considerations which are relevant to the problem of high-speed operation.

In analyzing the transient response, the circuit of Fig. 6 will be found convenient.

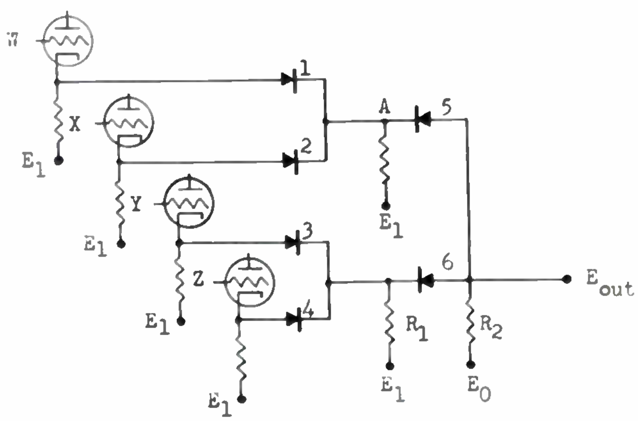


Fig. 6—A simple rectifier circuit.

Assume a capacitance C to be associated with each rectifier. The capacitance can be omitted when the rectifier is conducting in the forward direction since the small forward resistance of the rectifier essentially short circuits the capacitance.

First consider the case when the output is low. This is true when either or both of the first-level outputs are low. Let W, X be high and Y, Z be low. The output is low at a potential near E_1. The potential of A is nearly E_0. The potential of B is nearly E_1. The capacitance of rectifier 5 is charged. If the input Z is now raised high to a potential E_0, the input Z charges the capacity of rectifier 3 and causes point B to reach a potential E_0. Since the driving impedance of Z is usually very small, the potential at point B rises fast as shown in Fig. 7(a). As soon as the potential of point B rises, the output potential will rise accordingly. But the previous charge remaining in the capacity of rectifier 5 prevents the output voltage from following the potential B. The output will rise from E_1 to E_0 with a time constant equal to R_2C.

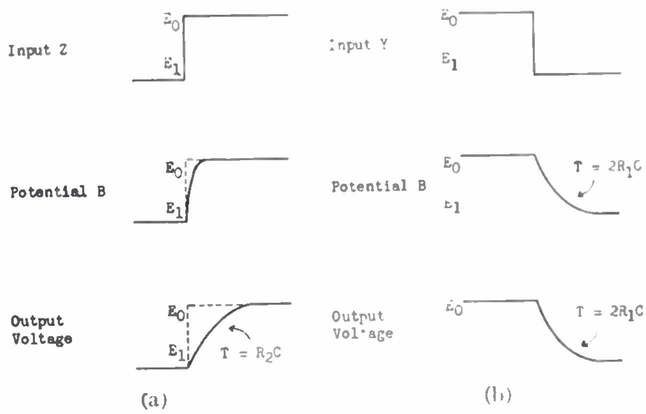


Fig. 7—Transient response.

Consider the case in which the output is normally high, such as when W, Y are high and X, Z are low. Capacities of rectifiers 2 and 4 are charged. Now let Y change from high to low. The potential of point B will decrease from E_0 to nearly E_1, only to be retarded by the discharge of the capacity of rectifier 4 and the charge of the capacity of rectifier 5. Assuming R_2 large compared to R_1 and the forward resistance of the rectifier negligible, the time constant equals $2R_1C$, as shown in Fig. 7(b).

If the analysis is extended to a circuit with n input variables and m canonical forms, the rise time constant will be $(m-1)R_2C$ and the decay time constant will be $(m+n-2)R_1C$.

Since the values of n and m increase in more complicated circuits, the time constants increase and the output response is slowed down. The attenuation is also increased.

Cascading of Rectifiers

One effective way of reducing the effects of the large capacitance which result when many rectifiers are connected in parallel is to connect additional rectifiers in cascade to form pyramids. Fig. 8 shows a portion of such a circuit in which n and m are equal to 8.

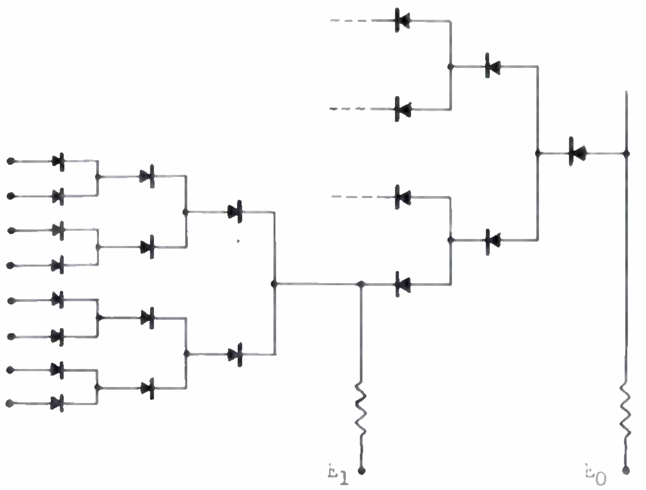


Fig. 8—Cascade connections of rectifiers.

It can be shown that the rise time constant will be $m'R_2C$ and the decay time constant will be $n'R_1C$. Table I shows the value of m', n' as a function of m and n.

TABLE I

		Cascaded rectifiers		Rectifiers connected in parallel	
n	m	m'	n'	$m'=m-1$	$n'=m+n-2$
2	2	1	2	1	2
	4	1.667	2.667	3	4
	8	2.238	3.238	7	8
4	4	1.667	3.334	3	6
	8	2.238	3.905	7	10
	16	2.771	4.438	15	18
8	8	2.238	4.476	7	14
	16	2.771	5.009	15	22
	32	3.287	5.525	31	38
	64	3.795	6.033	63	70

The table shows that for large values of m and n the speed is increased considerably by the use of extra rectifiers. It should be noted that there will be more series drop in the forward direction of the rectifiers. But the reduction of the shunt-leakage current through the backward direction of the rectifiers more than compensates for this drop. As a result, the de-attenuation of the rectifier network will be smaller.

Improved Circuits with Large Bias Voltage

In the circuit of Fig. 6, assume that A is high. Then the output, E_{out}, will follow the potential of B, but will lag it because of the need of charging the capacitance of rectifier 5. This is shown in Fig. 7(a). If the voltage available for charging the capacitance of rectifier 5 were increased, then the lag between the potential at E_{out} and the potential at B should be decreased. Therefore, if E_3 in Fig. 9 is made larger than E_0 in Fig. 6, the output voltage will rise faster. Fig. 10 shows this output voltage. It follows the line OC instead of OL because of the higher voltage E_3 applied. When the output voltage reaches E_0, it nearly equals the potential of B, and rectifier 6 conducts. The output voltage is clamped to this level, and changes abruptly to follow path CD. A slight

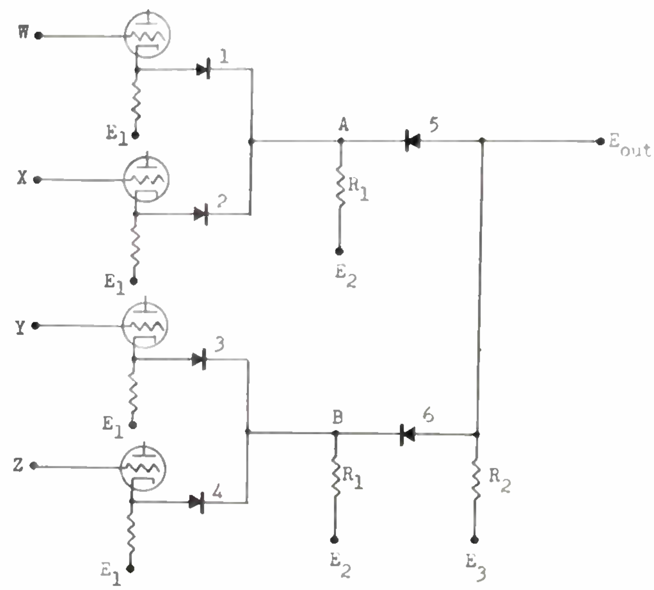


Fig. 9—A simple rectifier circuit with high bias voltage.

increase in the supply voltage E_3 over E_0 will considerably speed up the rise time.

Similarly, if the voltage E_2 is made somewhat lower than the voltage E_1, the decay of the output will be faster.

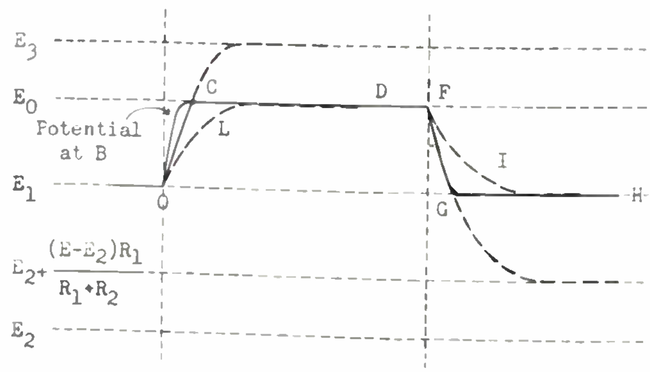


Fig. 10—Transient response with high bias voltage.

From the curve in Fig. 10, one can calculate the time lag approximately. Assuming operation in the linear charging portion of the curve, the rise time lag, T_r is given by

$$T_r = m' \frac{E_0 - E_1}{E_3 - E_1} R_2C. \qquad (8)$$

The decay time lag T_d is given by

$$T_d = n' \frac{E_0 - E_1}{(E_0 - E_2)\frac{R_2}{R_1} - (E_3 - E_0)} \cdot R_2C, \tag{9}$$

where m' and n' are the factors for complex networks given in Table I.

In order to lower the decay time lag when the voltage E_3 is increased, the following relation must hold:

$$E_2 < E_1 - (E_3 - E_1)\frac{R_1}{R_2}$$

or

$$\frac{E_1 - E_2}{R_1} > \frac{E_3 - E_1}{R_2}. \tag{10}$$

The value of the right-hand side can be increased to reduce the rise time lag. But the left-hand side has to be increased to reduce the decay time. This is limited by the forward current rating of the rectifier. It is possible to adjust the values of E_3, E_2, R_1, and R_2 so that the rise and decay lag will be nearly equal. This is a very desirable condition.

Let the maximum forward current of the rectifier be I.

$$I = \frac{E_0 - E_2}{R_1} > \frac{E_1 - E_2}{R_1} > \frac{E_3 - E_1}{R_2}$$

$$T_r = m' \frac{E_0 - E_1}{E_3 - E_1} R_2C$$

$$> \frac{m'(E_0 - E_1)C}{I}. \tag{11}$$

It can be seen from (11) that the minimum rise time lag depends on the complexity of the network m', the output-voltage difference $E_0 - E_1$, and the capacity of the rectifier per unit maximum forward current. This coefficient of capacity per unit forward current is the figure of merit of a rectifier for high-frequency operation. For a value of $m' = 3$, $E_0 - E_1 = 10$ volts, $C = 70 \times 10^{-12}$, and $I = 5 \times 10^{-4}$ amperes. The rise time T_r will be a minimum of four microseconds.

Another good point of this new circuit of Fig. 9 is that the attenuation of the signal through the network is very small. The output voltage will differ from the input at most by the voltage drop across one of the rectifiers. This voltage drop is usually kept small.

It is interesting to note that there is a maximum limit of speed attainable for fixed values of m', $(E_0 - E_1)$ and C/I. It is independent of the voltage E_3, E_2, R_1, and R_2. It is therefore sufficient to choose a value E_3 which will give just enough voltage difference to make the rising portion of the output voltage nearly linear. The values R_1, R_2, and E_2 should be chosen to give the decay time lag nearly the same value.

Practical Rectifier Elements

The rectifier computing circuits described may require a large number of rectifier units connected in complex, series-parallel combinations. The usefulness of the circuits depends to a large extent on the speed at which they can be operated.

The number of rectifiers which can be connected in a circuit of this type is limited by the ratio of the back-to-forward resistance of the individual rectifiers. The speed at which the circuits can be operated depends on the ratio of the forward-current carrying capacity of the rectifiers to their electrostatic capacitance, measured in the back direction. As pointed out previously, the larger this ratio the higher the speeds attainable. Experience up to this time indicates that this ratio is larger in rectifiers of small size. Small rectifiers can dissipate heat more effectively, and hence carry relatively larger current. Small rectifiers have the further advantage that the absolute magnitudes of the currents required are smaller, and therefore less power is needed from the driving source.

The above considerations, and commercial availability, have led to a choice of selenium-rectifier units having an effective area of 0.0005 square inch. With a bias voltage of zero, the electrostatic capacitance is about 70 micro-microfarads. In addition, a large number of small rectifiers may be placed in a small space.

Fig. 11 is a typical current-voltage characteristic.

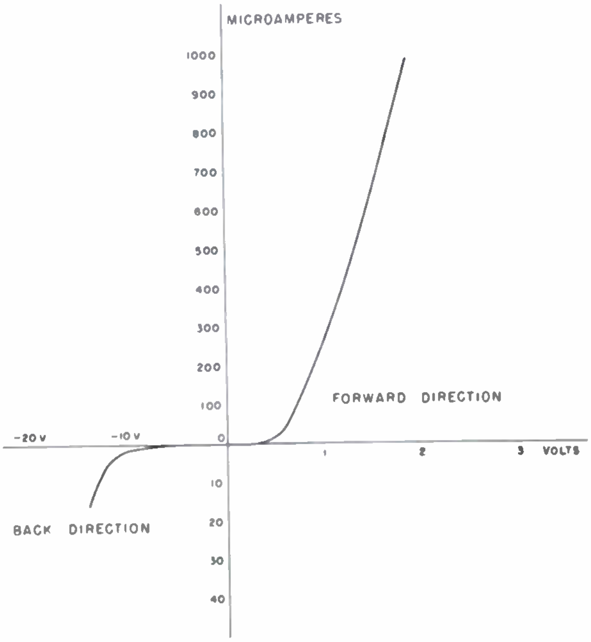


Fig. 11—Selenium rectifier characteristic.

Mounting Technique of Rectifier Elements —Printed Circuits

Fig. 12 shows the method of mounting selenium disc rectifiers. A hole slightly larger than the diameter of the disc is drilled through the bakelite plate. A brass button less than $\frac{1}{8}$ inch long is forced into one end of the hole. The rectifier disc is placed in the hole on top of the button. A metal washer is then placed in the hole. The brass button shown at the top of Fig. 12 is then forced into the hole. The spring provides enough pressure that the contacts between the rectifier and the brass buttons are very good. Conducting paint is used to connect the rectifiers and other components on the bakelite plate.

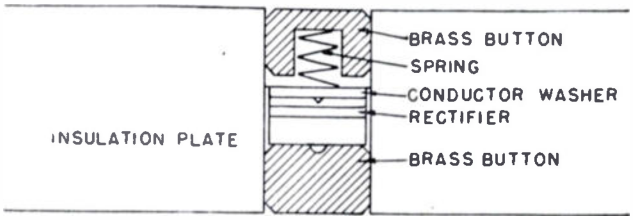


Fig. 12

Fig. 13 shows a complete decimal accumulator with carry. It uses 5, 4, 2, 1 coded decimal system. It has nine input variables corresponding to the two coded decimal numbers to be accumulated and a carry from the previous accumulation. There are five output functions, the low-order sum and a carry. The whole circuit uses around 500 small selenium cells. Although a considerable number of rectifiers are to be driven by the input signal, only a single miniature twin triode, acting as cathode followers, is needed for every input variable.

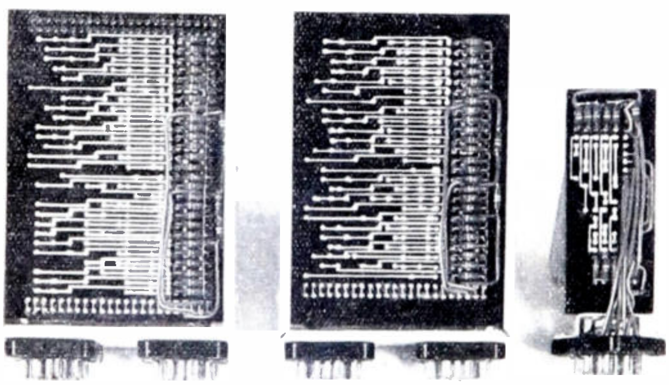


Fig. 13

The input signals vary from 0 to -15 volts with a period of 20 microseconds. One output function is shown in Fig. 14. The output is clamped between 0 and -7 volts. Since the output voltage lags the input voltage by only a few microseconds, the selenium rectifier circuit described can be expected to give satisfactory operation at frequencies up to 100 kc.

Discussion and Conclusion

As derived above in (11), the speed of operation of the computing circuit is inversely proportional to three factors, namely, m', $E_0 - E_1$, and C/I. The value of m' depends on the complexity of the circuit. For reasonable complexities as encountered in actual practice, this value is nearly a constant. The value of C/I is a figure of merit of the rectifiers being used. For point-contact rectifiers, this value is very small. In the case of germanium rectifiers, the frequency of operation may easily reach 10 mc. For selenium and copper oxide rectifiers, the value of C/I is rather large. But improvements to reduce this value could be made. The value of $E_0 - E_1$ represents the signal level. This level can be reduced at the extra cost of amplification. With standard selenium rectifiers at a relative large-signal level, an operating frequency of 100 kc has been reached. It is therefore highly possible that even megacycle operations can be achieved with slight improvement in selenium rectifiers and some reduction in signal level.

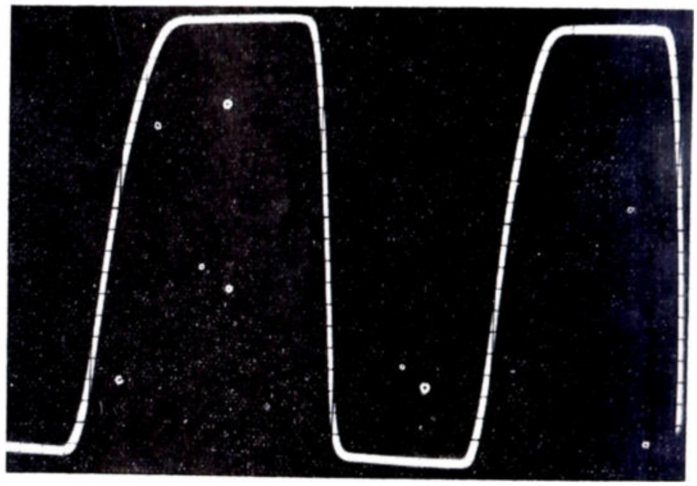


Fig. 14—Output wave form of rectifier computing circuit.

The effects of temperature, aging, creeping, and reforming of rectifier characteristics have not been exhaustively tested. So far, no serious trouble of this nature has been reported.

The complete rectifier circuit is simple and compact. It is relatively cheap to assemble both in limited specialized applications and in mass production. It will save thousands of vacuum tubes in a large-scale digital computing machine. It might be a great help in the development of low-cost computing machines. Its use may also simplify considerably complicated control systems.

Acknowledgment

The author wishes to express his indebtedness to Professor Howard H. Aiken for his suggestion that selenium rectifiers be used in computing circuits, and for his encouragement in the development. Credit is also due Mr. Robert E. Wilkins in regard to the design of the mechanical construction of the rectifier assemblies.

The research reported in this paper is supported by Contract W 19-122-AC-24 between the U. S. Air Force and Harvard University.

Cathode-Ray Picture Tube with Low-Focusing Voltage*

C. S. SZEGHO†, FELLOW, IRE

Summary—**Wide-angle picture tubes, high anode voltages (reduced magnification and beam divergence), and the need for material conservation prompt replacement of magnetic with electrostatic focusing. There are considerable advantages in a unipotential lens focusing system designed to obtain optimum focus with the focusing electrode operating at 0 to 5 per cent of the anode voltage instead of at the customary 20 per cent. The focusing electrode may then be connected to the cathode or to a potentiometer across the low voltage dc supply of the television set. A successful tube design is discussed and operating characteristics given. The voltage breakdown problem, which is stringent owing to the high-voltage gradients at prevalent high anode voltages, is considered.**

Introduction

In early picture tubes the superior imaging qualities of the magnetic lens had to be enlisted to keep the spot size small because, with the customary deflection angle not exceeding 50 degrees, the lens magnification, defined as the ratio of the lens-screen to cathode-lens distance, was large. In the interest of space conservation together with the increase in screen size of picture tubes, the deflection angle was increased to 70 degrees and more. This had the beneficial effect of providing smaller magnification or a smaller spot at the center of the fluorescent screen, but the deflection defocusing at the edges of the raster became more pronounced. As these larger tubes required higher anode voltages, resulting in a smaller beam divergence angle and improved edge focus, the stage was set to introduce electrostatic focusing, a step also prompted by the need to save materials which go into the manufacture of magnetic focusing devices.

Electrostatically Focused Picture Tubes

The first electrostatically focused picture tubes of 70-degree deflection angle employed a 3-electrode lens, with the outer electrodes maintained at the anode potential and the middle electrode, the focusing electrode proper, at $22\frac{1}{2}$ per cent of the anode potential.[1] Such lenses, called "unipotential lenses," were known at least as early as 1932[2] and have since been thoroughly investigated.[3] In cathode-ray tube applications, the electrodes of the unipotential lens are usually formed as short cylinders. The refractive power of such a lens, for a given geometrical configuration of the electrodes, depends on the focusing ratio $(V_f - V_a)/V_a$, where V_f is the potential of the middle electrode and V_a the common potential of the outer electrodes. With $V_f = 0$, the lens becomes independent of the anode voltage; the focusing ratio remains -1 for all anode voltages. If the focusing voltage is different from 0, it must be varied in proportion to changes in the anode voltage to maintain a given focusing ratio. Consequently, if the focusing voltage for a given construction is near 0, only slight variation is required to compensate for anode-voltage changes.

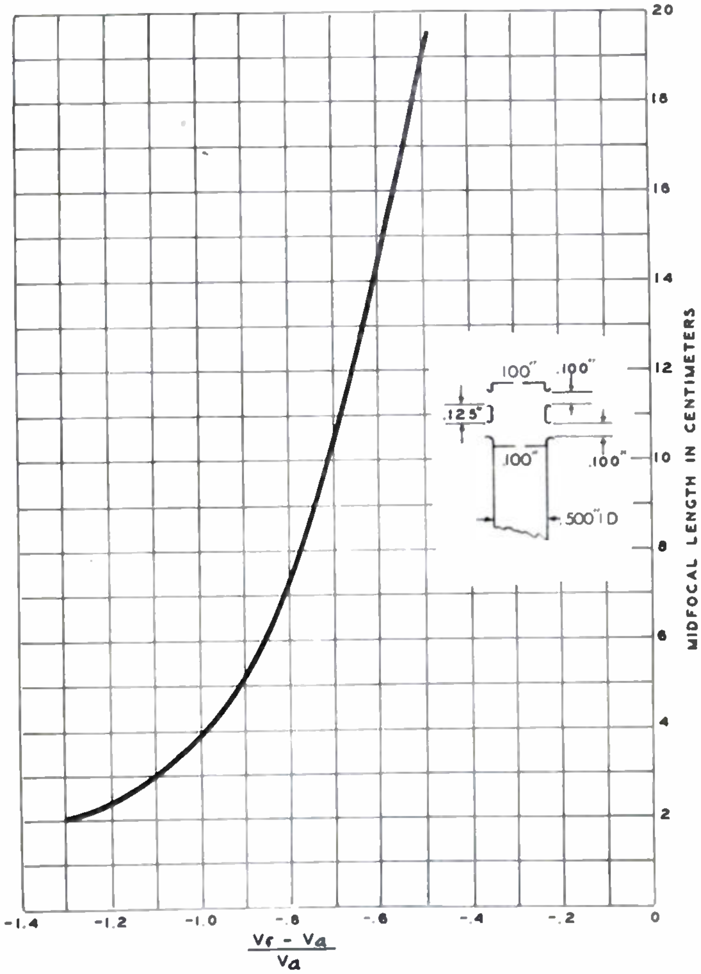


Fig. 1—The midfocal length of the unipotential lens versus the focusing ratio.

Fig. 1 is a plot of the midfocal length defined as the distance between the focal point and the center of the focusing electrode, as a function of the focusing ratio of a unipotential lens consisting of 3 cylindrical electrodes of $\frac{1}{2}$-inch diameter. The midfocal lengths were measured in a tube in which a parallel electron beam passing

* Decimal classification: R583.6. Original manuscript received by the Institute, July 13, 1951; revised manuscript received March 31, 1952.

† The Rauland Corp., 4245 N. Knox Ave., Chicago 41, Ill.

[1] L. E. Swedlund and R. Saunders, Jr., "Material-saving picture tube," *Electronics*, vol. 24, pp. 118–120; April, 1951.

[2] E. Brüche and O. Scherzer, "Geometrische Elektronenoptik," Julius Springer, Berlin, p. 90; 1934.

[3] A good bibliography up to 1948 can be found in P. Grivet, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. II, p. 98; 1950.

through paraxial holes is brought to focus on a sliding fluorescent screen. The parallel beam was obtained by placing the object (first beam crossover) at the focal distance from an auxiliary unipotential lens (condenser lens). The slope of the curve decreases as the focusing ratio approaches −1, from which it is apparent that this lens with the focusing electrode near cathode potential is still less sensitive to changes in anode voltage.

If the focusing electrode is maintained at exactly cathode potential, there is no need for a separate focusing voltage supply, and the focusing electrode may be connected to the cathode inside the tube.[4] It is true, however, that to achieve this objective the electrode system must be built to rather close tolerances. Also, to permit the same gun to be used in different size tubes, it is desirable to allow for slight variation of the focusing electrode voltage to accommodate varying focal distances. Consequently, the decision was made to bring out a lead for the focusing electrode and to design the gun so that the tube focuses at about 100 volts with an anode voltage of 12 kv. Then a potential divider across the low-voltage dc supply of a television set can serve as a means for adjusting the focus if small deviations in the electrode assembly or anode-voltage variations make this necessary. Since no operating voltage between the dc supply voltage and the final anode voltage is required, such a system may be termed a "univoltage" lens. This results in a number of other advantages besides economy. For example, high-voltage insulation is not required in the base and leakage currents and currents having their origin in field emission, treated later, have no effect on the lens performance because the low-voltage power supply has abundant current supplying capacity. If the focusing electrode is internally connected to the cathode, the tube becomes interchangeable with a magnetically focused tube.

Gun Design

A new electrostatically focused picture tube which fulfills the above objectives has been developed. The electron gun is illustrated in Fig. 2 and comprises a tilted offset indicator ion trap[5] and an axial univoltage lens. The required focusing voltage increases as the length of the middle electrode increases and it decreases with an increase in the diameter of the opening of grid 4 or in the separation between grid 4 and grids 3 and 5. Certain design limitations are imposed by practical considerations. If the diameter of grid 4 is made too small, excessive spherical aberration is encountered; the spherical aberration constant for unipotential lenses can be 3 to 4 times greater than that of a corresponding magnetic lens.[6] If the separation between grids 3 and 4 is made too great, the asymmetrical electrostatic field established between the focusing electrode lead wire and grid 3 and those attributable to the supporting pillars may become objectionable.

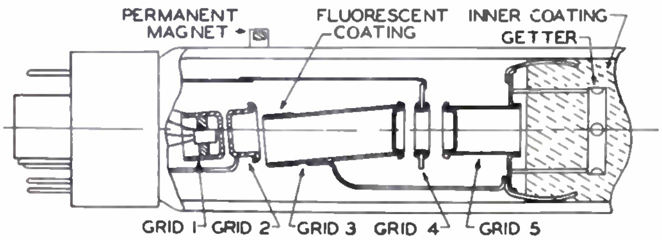


Fig. 2—The electron gun of the new low-voltage focus picture tube

For the same aperture ratios, the spherical aberration constant is proportional to the focal length of the lens. From this point of view, therefore, it is of advantage to shorten the cathode-to-lens distance and so increase the lens power. By placing the electrostatic focusing lens nearer the cathode, the beam diameter in the deflection field is reduced; this same measure is not feasible with a magnetic lens because of interference of the magnetic focusing field with the proper functioning of the ion trap. Because of the much smaller beam diameter in the deflection field, the new tube exhibits a uniformly focused raster from edge to edge. The line structure is made less apparent by the somewhat increased spot diameter, which is an advantage in the case of large tubes. This expedient cannot be carried to extremes, however, for two reasons: (1) The magnification of the lens increases and (2) a certain minimum beam travel is required for proper ion trapping. Large transverse electrostatic and compensating magnetic fields, together with a substantial gun tilt, are needed if the length of grid 3 is short. The offset ion-trap gun[5] of the new univoltage tube allows effective ion trapping without the just-mentioned drawback, because offsetting of the structure consisting of the cathode and the first and second grids with respect to grid 3, which establishes the transverse electrostatic field, keeps the beam close to the axis. The offset and gun tilt are so chosen that the beam, which is bent back by the magnetic field of the beam bender toward the axis, passes centrally through the unipotential lens, a condition which is necessary to avoid excessive coma and astigmatism. With the new gun, extremely accurate ion-trap alignment may be obtained by merely adjusting the position of the ion-trap magnet until a coating of fluorescent material on the outside of the third grid reaches a minimum glow.[5]

Voltage Breakdown

In unipotential lenses, the voltage gradient between the middle and outer electrodes assumes high values because the separation is in the order of only a few millimeters. In the univoltage lens, this gradient is about 25 per cent higher. Moreover, the present trend in larger tubes is toward using voltages in the vicinity of 15–20 kv. The question of voltage breakdown now merits more attention than it has heretofore received.

[4] H. Iams, "A fixed-focus electron gun for C. R. Tubes," Proc. I.R.E., vol. 27, pp. 103–105; February, 1949.
[5] C. S. Szegho and T. S. Noskowicz, "Improved electron-gun ion traps," *Tele-Tech*, vol. 10, pp. 45–47; June, 1951.
[6] G. Liebmann, "Measured properties of strong 'unipotential' electron lenses," *Proc. Phys. Soc.* (London), vol. 62, pp. 213–228, April, 1949.

Two types of breakdown may be observed: The vacuum space between the electrodes may break down or arcs may develop over the insulator supporting the electrodes. The field strength across the vacuum gap between the electrodes, even if they were plane, would be 10^5 volts per cm with a separation of 2 mm and an anode voltage of 20 kv. The wall thickness of the customary $\frac{1}{2}$-inch, diameter cylinders of which the electrodes are fabricated is around 0.010 inch; with this edge radius of 0.005 inch, the gradient is increased by a factor of 5.[7] Besides this "macroscopic" field, one must allow for the "microscopic" field caused by minute points or other irregularities on the edge; the so-called "roughness factor," which is between 10 and 100, must also be considered.[8] Near the Pyrex insulating supports, the field strength is increased by a factor corresponding to the dielectric constant, which is approximately 5. All told, assuming a roughness factor of 20, maximum field strength figures in the neighborhood of 50 million volts per cm may be encountered. It is not uncommon to measure field-emission currents in the neighborhood of 100 microamperes. From the Fowler and Nordheim field-emission formula[9] and the field strength values, one can calculate that the emitting areas must be in the order of 10^{-10} cm^2.

To minimize field emission, the univoltage lens of the new tube is equipped with corona rings which are preferably electropolished. The microscopic irregularities on the electrodes may be burned out during manufacture by the application of high over-voltages leading to arcs. It is possible that oxide layers on the electrodes operated at negative potential cause local lowering of the work function;[10] this may be overcome by depositing a pure metal of high work function on these parts.[11] Breakdown between the lead wire for the middle electrode and the conductive wall coating is eliminated by removing the coating from this vicinity altogether. This measure minimizes the danger of minute graphite particles shaking off and thereby causing arcs.

The ion-trap indicator of the new tube is also a sensitive detector of field emission; if it does not glow when the beam is cut off, there is no detrimental field emission present.

Performance

Operating characteristics are shown in Figs. 3 and 4. The diameter of the beam as it emerges from the deflection field, at the so-called reference line, is plotted as a function of beam current, curve "c" of Fig. 3, while curves "a" and "b", representing similar characteristics for magnetically focused and electrostatic tubes with high-focusing voltage, are shown for comparison purposes. On Fig. 4 one family of curves shows the diameter of the spot on the fluorescent screen as a function of anode voltage for various beam currents, for a tube of

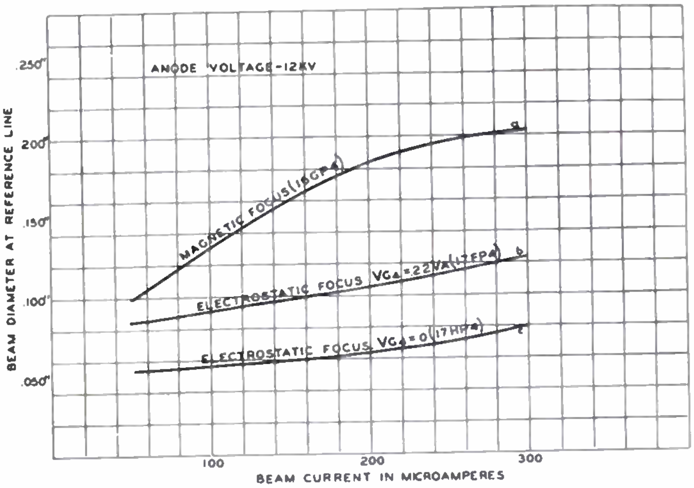


Fig. 3—The beam diameter at the reference line of magnetically and electrostatically focused tubes at various beam currents, with the beam focused at the fluorescent screen.

low-focusing voltage, the 17HP4. In contrast with the magnetically focused tubes, the beam diameter in the deflection field and the beam diameter at the fluorescent screen are approximately equal. This means that in the case of the new tube, the electron beam is nearly parallel. From the remaining curves on Fig. 4, it is apparent that the focusing voltage shifts to lower values with in-

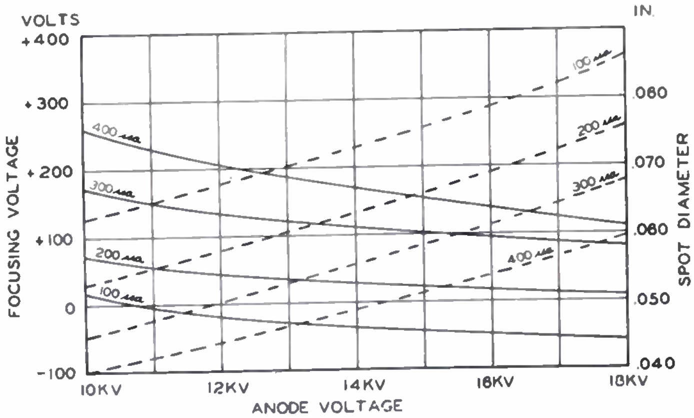


Fig. 4—Focusing voltage (broken lines) and spot size on the fluorescent screen (solid lines) as a function of the anode voltage at various beam currents for a low-voltage focus cathode-ray tube, the 17HP4.

creasing beam current. There is a marked improvement in spot size with increasing anode voltage. The spot diameters were determined with a microscope, a method which gives approximately twice the values measured by other methods like the shrinking raster method, for example. Thus it can be seen that the new tube is capable of resolution in excess of 450 lines.

Acknowledgment

Thanks are due to J. W. Stecker for making the measurements.

[7] A. Schwaiger, "Theory of Dielectrics," John Wiley and Sons, Inc., New York, N. Y.; 1932.
[8] P. H. Gleichauf, "Electrical breakdown over insulators in high vacuum," *Jour. Appl. Phys.*, vol. 22, pp. 535–551; May, 1951.
[9] R. H. Fowler and L. Nordheim, "Electron emission in intense electric fields," *Proc. Roy. Soc. A*, vol. 119, p. 173; 1923.
[10] F. Llewellyn Jones and C. G. Morgan, "Failure of Paschen's law and spark mechanism at high pressure," *Phys. Rev.*, vol. 82, pp. 970–971; June 15, 1951.
[11] U. S. Patent No. 2,408,894; October 8, 1946.

A Frequency Stabilization System for Microwave Gas Dielectric Measurements*

WILLIAM F. GABRIEL†, ASSOCIATE, IRE

Summary—A system is described for measuring the dielectric constant of a gas at uhf by observing the shift in resonant frequency of a cavity when filled with the gas to be measured. To do this, a klystron oscillator is stabilized with great precision on the cavity resonant frequency by means of a double-loop servo system. One loop is of the Pound type, which is effective against almost all frequency fluctuations. The other is a motor-type control which further reduces any small, steady, frequency error left by the electronic Pound loop. The two combine to give control that is better than one part in 10^8, and this is sufficient to obtain values of (n_0-1), where n_0 is the index of refraction, accurate to three places.

Introduction

THE STUDY of dielectric properties of gases at microwave frequencies has been intensified in recent years, and has resulted in the development of several different microwave measuring systems,[1-4] including the one to be presented here. The present system, which utilizes a resonant cavity in a high-gain, double-loop, frequency servo, was developed for the express purpose of obtaining greater precision in measurement of the index of refraction of low-loss gases.

A cavity may be used to measure the index of refraction of gases if the resonant frequency of that cavity can be accurately evaluated when it is evacuated and when it is filled with a gas. For a nonabsorbing gaseous medium, the relationship involved may be written as

$$n = \frac{1}{1 - \left(\frac{\Delta f}{f_0}\right)}, \tag{1}$$

where n is the index of refraction of the enclosed gas, f_0 is the resonant frequency of the cavity when it is completely evacuated, and Δf is the change in the cavity resonant frequency caused by introducing the gas into the cavity.

The value of f_0 can be readily determined to an accuracy of four or five places, but the value of Δf, which is smaller than f_0 by a factor of about 10^3 or 10^4, has heretofore been quite difficult to determine to an accuracy of more than two places. By the method presented in this paper, wherein the frequency of a klystron oscillator is automatically maintained at the resonant frequency of the cavity, the accuracy in the determination of Δf is limited by the tracking error of the control system, that is, by the amount that the oscillator frequency fails to follow a given change in cavity resonance frequency. A tracking error of less than one part in 10^8 has been achieved, thereby making it possible to obtain values of $(n-1)$ which are accurate to three or four places.

The Double-Loop Frequency Servo

The double-loop system to be described is not the first of its kind. To the best of the author's knowledge, a double-loop frequency stabilization system was first set up by Rideout at the Bell Telephone Laboratories in Holmdel, N. J., in 1945. In 1947, Huboi, at the University of Wisconsin, successfully engineered a low-gain system.[5]

The present double-loop system is shown in schematic form in Fig. 1. It is the result of combining two different

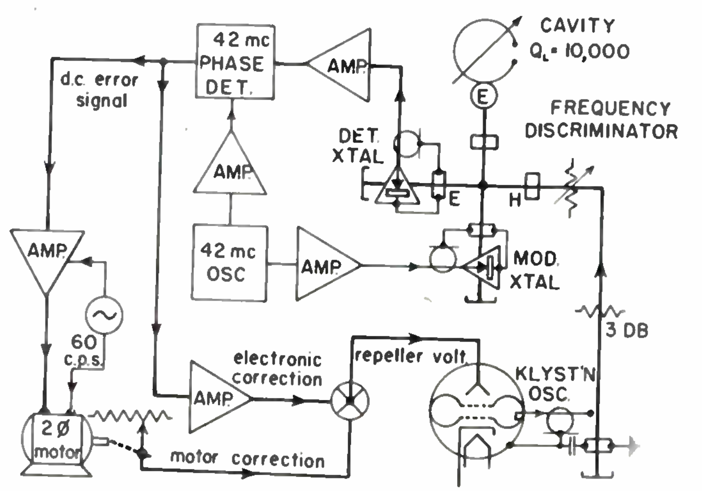


Fig. 1—Double-loop frequency stabilization system.

stabilization systems. One of them is a fast-acting, all-electronic system described by Pound,[6] and the other is a slow-acting, electromechanical servo motor system described by Rideout.[7] The microwave frequency discrim-

* Decimal classification: R216.3. Original manuscript received by the Institute, April 27, 1951; revised manuscript received February 29, 1952. This paper was presented at the IRE National Convention, New York, N. Y. on March 19, 1951. It is a slight modification of the thesis, "A Method for Accurately Measuring the Index of Refraction of Gases at Microwave Frequencies," submitted by the author in partial fulfillment of the requirements for the Ph.D. degree at the University of Wisconsin.

† Antenna Research Branch, Radio Division I, Naval Research Laboratory, Washington 25, D. C.

[1] G. Birnbaum, S. J. Kryder, and H. Lyons, "Microwave measurements of the dielectric properties of gases," *Jour. Appl. Phys.*, vol. 22, pp. 95–102; January, 1951.

[2] C. K. Jen, "A method for measuring the complex dielectric constant of gases at microwave frequencies," *Jour. Appl. Phys.*, vol. 19, pp. 649–653; July, 1948.

[3] C. M. Crain, "The dielectric constant of several gases at a wavelength of 3.2 cm.," *Phys. Rev.*, vol. 74, pp. 691–693; September, 1948.

[4] W. D. Hershberger, "The absorption of microwaves by gases," *Jour. Appl. Phys.*, vol. 17, pp. 495–500; June, 1946.

[5] R. W. Huboi, "Automatic Frequency Control of a Microwave Oscillator Using Combined Electrical and Electromechanical Servo Loops," M.S. Thesis, University of Wisconsin; 1947.

[6] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 35, pp. 1405–1415; December, 1947.

[7] V. C. Rideout, "Automatic frequency control of microwave oscillators," Proc. I.R.E., vol. 35, pp. 767–770; August, 1947.

nator employed is patterned after the type described by Tuller, Galloway, and Zaffarano.[8]

The important features, for our purposes, of each of these two systems may be described as follows: The Pound system can correct oscillator frequency deviations from cavity resonance occurring at the rate of from zero to a few thousand cycles per second; but, since it is an error-dependent servo, it cannot completely eliminate a given error and is susceptible to cumulative slow-drift error. The Rideout system, on the other hand, can integrate out a steady-state or slow-drift error to within the limits of its static friction starting sensitivity; but, since the servo motor possesses considerable inertia, it cannot correct frequency deviations more rapid than one or two cycles per second. The advantages to be gained in combining the two systems are not entirely obvious and can best be shown by considering the loop gain-phase relationships.

Let us represent the double-loop stabilization system shown in Fig. 1 by the simple servomechanism schematic diagram shown in Fig. 2. In this diagram, each

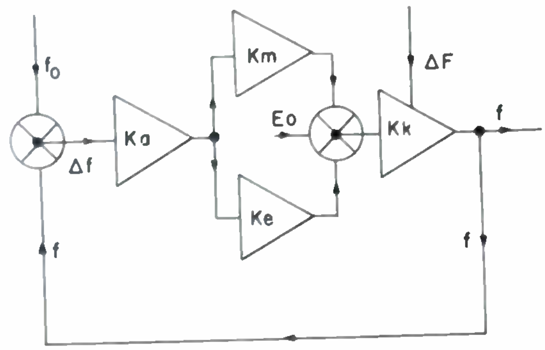


Fig. 2—Double-loop servomechanism schematic diagram.

physical component of the actual system is represented by its complex number transfer function. The symbols are defined as follows: f is the frequency of the klystron oscillator energy; f_0 is the true resonant frequency of the cavity; Δf is the frequency error of the servo ($\Delta f = f_0 - f$); K_a is the product of the transfer functions of the microwave-frequency discriminator, the 42-mc amplifier, and the 42-mc phase detector; K_m is the product of the transfer functions of the motor, the motor amplifier, a gear reduction unit, and potentiometers which control the klystron repeller voltage; K_e is the transfer function of the klystron-control amplifier; K_k is the transfer function of the klystron with respect to repeller voltage control; E_0 is a constant repeller voltage which is adjusted so that $E_0 K_k = f_0$; and ΔF represents the frequency deviations from f_0 which would occur in the klystron frequency if the loop were opened at the input to the microwave discriminator.

From the diagram, one can solve for the relationship between the closed loop error, Δf, and the open loop error, ΔF, obtaining,

$$\Delta f = \frac{\Delta F}{1 + A} \tag{2}$$

where

$$A = K_k K_a (K_e + K_m).$$

Equation (2) is the standard error relationship for a frequency control servo, and it shows that system performance is determined by the function A, which represents the total loop gain of the servo. But A is simply the complex sum of the electronic and motor loop gains taken separately; that is,

$$A = A_e + A_m \tag{3}$$

where

$$A_e = K_a K_k K_e = \text{electronic loop gain}$$
$$A_m = K_a K_k K_m = \text{motor loop gain.}$$

Therefore, if the complex transfer function terms can be evaluated, then the performance of the two servo loops, either separately or together, can be determined.

A convenient way of representing the physical components of a servo loop is to use their inverse time constant in complex terms of the type $(w_c + jw)$, where w_c is the value of the particular inverse time constant involved and w is the angular frequency rate $(2\pi f)$ at which the disturbance signal is assumed to vary. Using terms of this type and assuming the various components to be linear in operation, the transfer function terms for this system become:

$$K_a = \left(\frac{de}{df}\right)\frac{G_a}{(w_2 + jw)(w_a + jw)^6} \qquad w_0 = 6,666 \text{ sec}^{-1}$$
$$w_1 = 10 \times 10^6 \text{ sec}^{-1}$$
$$K_e = \frac{G_e(w_1 + jw)}{(w_0 + jw)(w_3 + jw)} \qquad w_2 = 25 \times 10^6 \text{ sec}^{-1} \tag{4}$$
$$w_3 = 35 \times 10^6 \text{ sec}^{-1}$$
$$K_m \cong \frac{G_m}{jw(w_m + jw)} \qquad w_a = 34.6 \times 10^6 \text{ sec}^{-1}$$
$$w_m = 20 \text{ sec}^{-1}$$
$$K_k = \left(\frac{df}{dE}\right)$$

G_a, G_e, and G_m are gain constants; w_0 represents the electronic loop cut-off network which is in the plate circuit of the klystron-control amplifier; w_1 and w_3 represent a lead network in that same amplifier; w_2 represents the 42-mc phase detector circuit; w_a represents any one of six similar circuits in the 42-mc amplifier; and w_m represents the armature of the servo motor.

Using the complex quantities given above, the expression for the electronic loop gain alone will be found to be

$$A_e = \frac{S_e(w_1 + jw)}{(w_0 + jw)(w_2 + jw)(w_3 + jw)(w_a + jw)^6}, \tag{5}$$

where S_e is a gain constant. Equation (5) is shown plotted in Fig. 3 in terms of the amplitude and phase versus frequency, with the amplitude scale normalized

[8] Tuller, Galloway, and Zaffarano, "Recent developments in frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 36, pp. 794–800; June, 1948.

to the value which A_e would have if w is set equal to zero. The effects of the various terms are easily recognized on a gain-phase diagram such as this and, also, one can quickly ascertain the maximum stable loop gain and the useful corrective frequency range of the servo. In Fig. 3, the former is about 63 db (1,400) and the latter is from zero cps to about 1,000 cps.

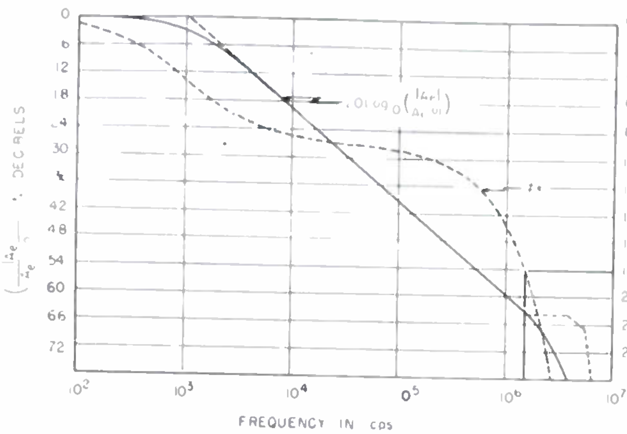


Fig. 3—Gain-phase diagram of electronic loop.

The motor loop-gain expression may be approximated for our purposes by a gain constant, S_m, and the complex terms contributed only by the motor itself.

$$A_m \cong \frac{S_m}{jw(w_m + jw)}. \tag{6}$$

The important term here is the first imaginary quantity, jw, in the denominator. This term is peculiar to all servo motors and indicates an integration operation, that is, continuous correction as long as an error signal exists. Equation (6) is shown plotted in Fig. 4. The gain scale

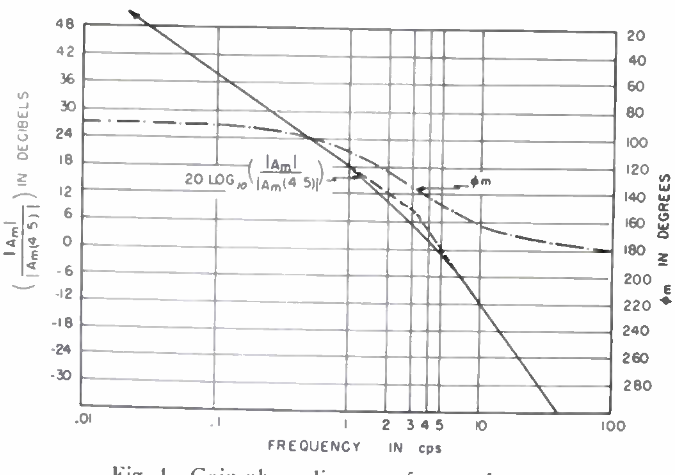


Fig. 4—Gain-phase diagram of motor loop.

is normalized to the 4.5-cps value because this is approximately the point at which the motor loop would break into unstable oscillations if the loop gain reached unity. Phase shift in the 60-cps motor amplifier is responsible for the loop phase shift reaching 180° at this point. The associated terms are not included in (6). Also, the effects of mechanical static friction are neglected in the above expression.

Turning now to combined loop operation, represented by (3), a very interesting aspect may be pointed out. Let us assume that the electronic loop and the motor loop are operating at their respective maximum loop gains. This would result in a double-loop gain-phase diagram depicted by the curves labelled $|A_1|$ and ϕ_1 in Fig. 5. From these curves, it will be noted that the

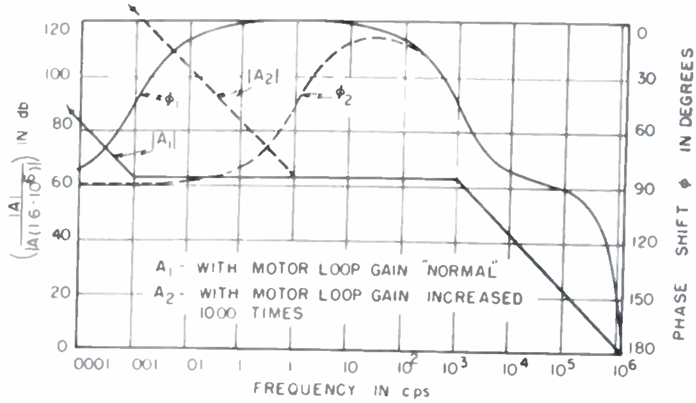


Fig. 5—Gain-phase diagram of double-loop system.

phase of the double-loop system at any particular frequency is dominated by the phase of the loop which has the highest gain at that frequency. Since this is true, then it should be possible to increase the gain of the motor loop so that it exceeds its normal maximum value of unity at 4.5 cps because the gain (and, therefore, the phase) of the electronic loop predominates at 4.5 cps. In this particular instance, the exact relationship for total loop gain would be

$$A(4.5) = |A_e| - |A_m| \tag{7}$$

where

$$|A_e| \leqq 1{,}400.$$

Equation (7) shows that, when operated in conjunction with the electronic loop, the motor loop gain at its normal point of instability can be increased to any value less than the electronic loop gain at that point, and stable double-loop operation will still be maintained. This is very important in obtaining added frequency stability, because it allows the gain of the motor amplifier to be greatly increased, thereby increasing the starting sensitivity of the motor loop and causing a reduction in the steady-state frequency error allowed by the system. The curves labeled $|A_2|$ and ϕ_2 in Fig. 5 illustrate the gain-phase diagram for double-loop operation after the motor loop gain has been increased by a factor of 1,000. It will be noted that the stabilizing effect of the electronic loop on the motor loop is similar, in certain respects, to the modification of a servo loop by the use of rate-circuits or tachometer feedback.

Double-Loop Performance

The system performed very well when it was properly adjusted. The electronic loop efficiently removed almost all of the audio-frequency modulation in the klystron signal and would follow a given change in cavity-resnance frequency to the extent of its loop-gain factor; hat is,

$$(\text{electronic loop } \Delta F) = (\text{cavity } \Delta F) - \frac{(\text{cavity } \Delta F)}{(\text{loop gain})}. \quad (8)$$

The motor loop would then slowly reduce the residual requency error until its input voltage dropped below he value necessary to overcome mechanical static friction. It would respond to a cavity-resonant frequency change of less than 75 cps, thus giving a nominal frequency control of better than one part in 10^8. Several actors relating to this control figure must be very carefully considered, however, lest too much be expected from the system: First, it is relative to the cavity-resonant frequency; that is, it would never be absolute unless the cavity-resonant frequency remained absolutely constant (no temperature drifts or other changes). Second, the frequency about which this control occurs is not necessarily the true resonant frequency of the cavity but is, instead, the "balance" frequency of the microwave discriminator, which point is dependent, to some extent, upon the effective reflection coefficient of the modulator crystal as well as the cavity itself. And third, this excellent control was disturbed in a random manner by small, spurious frequency jumps (usually several hundred cycles per second but occasionally a few thousand cycles per second) which apparently were caused by erratic electrical behavior of the 1N23B modulator crystal. However, after one of these spurious jumps, the frequency always returned to its original average value, so that their long-time effect did not appear to be particularly serious.

It is possible that the above-mentioned frequency jumps could be either eliminated or reduced to less than 100 cps by investigating the phenomenon thoroughly. If they can be eliminated, then the fact that this system removes the drift errors which are present in the all-electronic, Pound frequency-control system would indicate that it might provide a substitute for crystal-controlled oscillator and frequency-multiplier combinations, because the controlled frequency is dependent only upon the cavity itself.

Gas-Measuring System

The complete gas-measuring system is shown in Fig. 6. It consists of three separate parts:

(1) The double-loop, frequency stabilized, klystron oscillator system;

(2) The system for accurately measuring the frequency of the klystron-oscillator energy;

(3) The system for accurately measuring either the pressure of a gas in the cavity or the vacuum pressure in the cavity.

The first part has already been discussed and should be easily recognized in Fig. 6.

The second part functions in the following manner: A 40-mc harmonic generator, which is stabilized by a 5-mc quartz crystal oscillator, supplies its sixth harmonic, 240 mc, to a two-stage lighthouse tube amplifier. The output of that amplifier is fed into a 1N23B waveguide mounted crystal, and the thirty-ninth harmonic derived therefrom, 9,360 mc, is used as a reference microwave frequency. The cavity-resonant frequency is adjusted to about 9,420 mc so that a difference beat frequency of about 60 mc is obtained at the mixer crystal. After amplification, this 60-mc signal is then mixed with the signal from a calibrated signal generator whose fre-

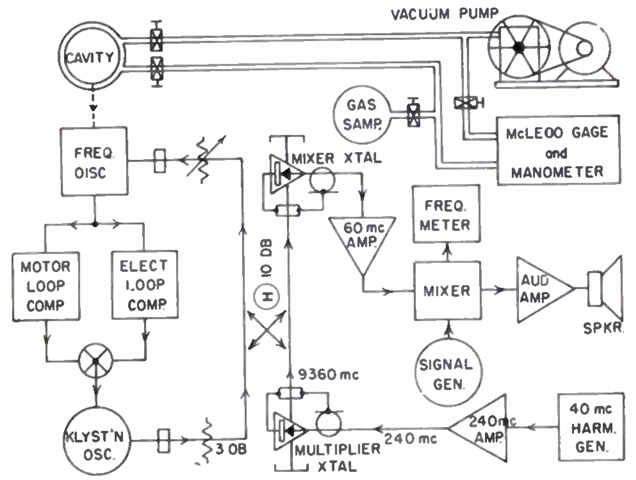


Fig. 6—Complete gas-measurement system.

quency is set so as to obtain either an audio beat note, which is amplified and sent to a loudspeaker, or a 1- to 10-mc beat frequency which can be accurately measured by a precision-frequency meter. The loudspeaker is used for listening to the klystron signal while making fine system adjustments, and the precision frequency meter is used for making gas measurements.

The third part of the gas-measuring system consists of a vacuum pump for evacuating the cavity, a McLeod gage for measuring the vacuum pressure, and a mercury manometer with a parallax-corrected, vernier slide for measuring the absolute pressure of gases introduced into the cavity.

Gas Measurements

Measurements on a gas are made as follows:

(a) The cavity is evacuated and its resonant frequency, f_0, plotted versus time.

(b) At a definite time, the gas is slowly admitted into the cavity.

(c) The new resonant frequency is plotted versus time, and several determinations of gas temperature and pressure are made.

(d) The gas is evacuated from the cavity and f_0 again plotted.

For the purpose of illustration, one of the actual data runs on helium is shown in Fig. 7. This particular run was chosen because it illustrates both the best and the worst type of operation that was experienced with the present measuring system. It will be noted that nearly perfect operation prevailed between 4:00 and 4:07 and again between 4:10 and 4:15, but at 4:08 a spurious frequency jump of 4,500 cps was detected. Minor frequency variations occurred between 4:15 and 4:25.

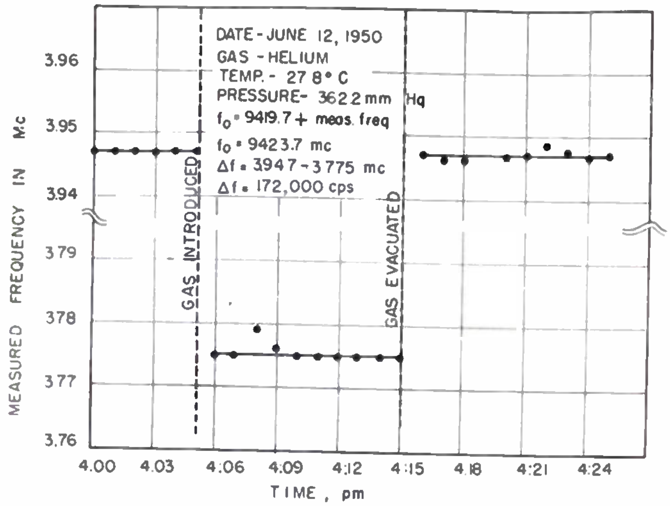


Fig. 7—Reproduction of actual data run.

Ambient cavity temperature did not vary during this data run since temperature drifts cause a corresponding slope in frequency plots.

Data from the measurements made in the above manner are inserted into the following relationships to get the value of n for the gas at standard-temperature-pressure, n_0:

$$n = \frac{1}{1 - \dfrac{\Delta f}{f_0}} \tag{9}$$

$$(n_0^2 - 1) = (n^2 - 1) \times \frac{760}{273.16} \times \frac{T}{P} \tag{10}$$

where T = temperature in degrees K and
P = pressure in mm Hg

If Δf is due solely to the index of refraction of the gas inside the cavity, then (9) will result in an absolute value of n, the accuracy of which is mostly dependent upon the inherent errors of the measuring system. In the present system, however, the changes in cavity-resonant frequency were due not only to the index of refraction of the enclosed gas but, also to atmospheric pressure forces which acted on the cavity tuning mechanism, so that absolute measurements could not be made with it. The effect could have been eliminated, of course, by constructing a new cavity which would be completely protected against atmospheric pressure stresses, but, unfortunately, a limited time schedule prevented that step. Therefore, in order to evaluate the system from the standpoint of gas measurements, it was necessary to run a pressure-frequency calibration on the defective cavity, and this was done by making careful measurements with helium, whose index of refraction is already accurately known from the extrapolated optical data of C. and M. Cuthbertson (see reference 5 in the data table). The pressure-frequency error was found to be linear and amounted to −84 cps per mm of mercury pressure of the gas inside the cavity. Using this Δf correction, the cavity was then used to measure the index of refraction of neon, argon, and dry air. The results obtained are shown in the data table, which also lists the most reliable comparison data available at the present time.

DATA TABLE

VALUES OF INDEX OF REFRACTION AND DIELECTRIC CONSTANT AT S. T. P.[9]
(FREQUENCY 9,423 MC)

Gas	No. of determ.	Index of refraction[12] n_0	Dielectric constant $K_e = \dfrac{n_0^2}{K_m}$	Comparison values of K_e
Neon[10]	6	1.00006666	1.0001333	1.00013345813
Argon[10]	2	1.00027739	1.0005548	1.00055464713
Dry air[11]	3	1.00028784	1.0005754[17]	1.0005754[14,18]
				1.000572[15]
				1.0005749[16,18]

[9] Standard temperature pressure (0° C. and 760 mm of Hg).
[10] Gases obtained from Linde Air Products (1 liter flasks).
[11] Air dried by passing it through five feet of tubing containing anhydrous P_2O.
[12] Helium was used as a reference gas in calibrating the cavity pressure error. Therefore, values given for neon, argon, and dry air are based upon the optical value of n for helium, 1.000034613, which was used in the calibration.
[13] Extrapolated from optical results of C. and M. Cuthbertson, *Proc. Roy. Soc. A*, vol. 135A, p. 40; 1932.
[14] G. Birnbaum, G. J. Kryder, and H. Lyons, of the National Bureau of Standards, at 9,280 mc, *Jour. Appl. Phys.*, vol. 22, p. 95; January, 1951.
[15] C. M. Crain, *Phys. Rev.*, vol. 74, p. 691; September, 1948.
[16] Value extrapolated from optical results of Barrel and Sears, *Roy. Soc. London Philos. Trans.*, vol. 238A, p. 1; 1939–40.
[17] Corrected for magnetic permeability of oxygen content, $(K_m - 1) \times 10^6 = 0.378$.
[18] Corrected for magnetic permeability of oxygen content and also for deviation from the perfect gas law.

On the basis of the results obtained, it seems reasonable to assume that, with a properly designed cavity and close temperature regulation, a careful operator should be able to achieve four-place accuracy in absolute measurements on nonabsorbing gases. The measurement of absorbing gases would require some additional equipment,[19] and would involve a decrease in accuracy roughly proportional to the decrease in the loaded Q of the cavity.

COMPONENT DESIGN

In a paper of this nature, space considerations severely limit the description of individual system components. For that reason, only the principal circuitry of the electronic control loop is included herein. Fig. 8 shows the 42-mc error signal amplifier, the balanced phase detector, and the klystron-control amplifier. The connector labeled "input" receives the 42-mc error signal output from the microwave-frequency discriminator. After amplification in the five-stage amplifier, this 42-mc error signal is converted to a dc error signal in the balanced phase detector and then fed to the motor amplifier and the klystron-control amplifier. The latter,

[19] R. V. Pound, "Electronic frequency stabilization of microwave oscillators," *Rev. Sci. Instr.*, vol. 17, pp. 503–505; November, 1946.

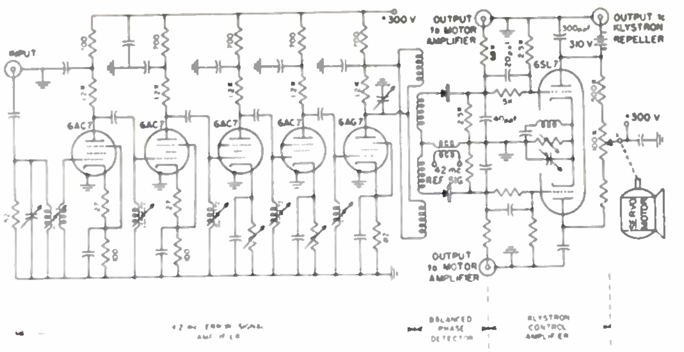


Fig. 8—Circuit diagram of 42-mc amplifier, phase detector, and klystron-control amplifier.

which is probably the most interesting circuit in the entire system, has the following features:

1. It contains the electronic loop "cut-off" circuit (300 $\mu\mu$f in parallel with 500,000 ohms).
2. It has a tuned cathode circuit which, in conjunction with the "cut-off" circuit, causes the 6SL7 stage to become a detector for 42-mc pulses on its grid, thus increasing the phase-detector efficiency for the electronic loop.
3. It has a lead network in its grid circuit which allows an increase in useful loop gain.
4. It combines the electronic and motor loop correction voltages which go to the klystron repeller electrode.

ACKNOWLEDGMENTS

The author is indebted to Associate Professor Vincent C. Rideout, who initiated and supervised this project at the University of Wisconsin, Madison, Wis.

The Response of RLC Resonant Circuits to EMF of Sawtooth Varying Frequency*

JEAN MARIQUE†, MEMBER, IRE

Summary—**By the use of Laplace transforms, the author derives the expression of the current flowing in an RLC series resonant circuit under the excitation of an emf, the frequency of which follows a sawtooth law. He shows that the response is composed of two terms, one arising during each sweep, the other resulting from the preceding sweeps. The bandwidth at −3 db of the circuit is introduced as a general parameter, and some curves obtained by graphical integration are given and compared with curves computed by other authors. The influence of the limits of exploration is discussed.**

I. INTRODUCTION

THE theoretical study of the response of a linear resonant system to an emf of constant amplitude, the frequency of which varies linearly with the time, has been made by several authors.

Some of them, as van der Pol[1] and Clavier,[2] have aimed to find the mathematical condition for which the dynamic response is nearly the same as the static response, which condition involves the use of very low exploration speeds.

Other authors, on the contrary, have tried to give the mathematical answer to the general case. Among the latter, the first one is probably Salinger,[3] who, by means of Fourier transforms, obtained the response of an ideal low-pass filter. Lewis[4] investigated the case of a mechanical shaft of linearly varying speed of rotation, but he limited himself to the determination of the over-all envelope curve of the response.

Hok[5] and Barber and Ursell[6] almost simultaneously gave, by different methods of calculation, the full solution of the problem in the case of a single exploration. The author has called attention to the importance of Hok's paper in a short note.[7] Barber has made an extension of the practical use of his results in a further paper.[8]

The purpose of the present paper is (1) to give the mathematical solution of the case encountered in the so-called spectrum analyzers or in panoramic receivers in which the exploration is continually repeated, and (2) to discuss the influence of the limits of the exploration.

As a matter of fact, in the present paper, we shall limit ourselves to the study of the response curve, as other aspects of the problem have been largely dealt with by Hok and Barber and Ursell. As Hok's paper seems to give an excellent approach to the problem, we shall make several references to it and use, as far as possible, the same notations in order to facilitate the comparison.

II. MATHEMATICAL SOLUTION

We consider a succession of intervals of time, each of duration h (Fig. 1(a)); during each of these intervals, the angular frequency of the emf varies linearly with

* Decimal classification: R140. Original manuscript received by the Institute, September 1, 1951; revised manuscript received April 21, 1952.

† Monitoring Centre for Marine and Aviation, Uccle, Bruxelles, Belgium.

[1] Van der Pol, "Fundamental principles of frequency modulation," *Jour. IEE* (London), pt. III, pp. 153–158; May, 1946.

[2] A. G. Clavier, "Application of Fourier transforms to variable-frequency circuit analysis," PROC. I.R.E., pp. 1287–1290; November, 1949.

[3] H. Salinger, "Zur Theorie der Frequenzanalyse mittels Suchtons," *Elek. Nach. Tech.*, pp. 293–302; August, 1929.

[4] F. M. Lewis, "Vibration during acceleration through a critical speed," *Trans. Amer. Soc. Mech. Eng.*, APM 54–24, p. 253; 1932.

[5] G. Hok, "Response of linear resonant systems to excitation of a frequency varying linearly with time," *Jour. Appl. Phys.*, pp. 240–250; March, 1948.

[6] N. F. Barber and F. Ursell, "The response of a resonant system to a gliding tone," *Phil. Mag.*; May, 1948.

[7] J. Marique, "Note sur la réponse d'un circuit soumis à un balayage linéaire de fréquence," *L'onde Elect.*, pp. 313–315; July, 1951.

[8] N. F. Barber, "The optimum performance of a wave analyzer," *Elec. Eng.*, pp. 175–179; May, 1949.

the time between two limits ω_0 and $(\omega_0+2\epsilon h)$, where ϵ is half the speed of the angular frequency variation in radian/sec² (or π times the speed of the frequency variation in cycles/sec²).

Then the instantaneous angular frequency is $\Omega=(\omega_0+2\epsilon\theta)$, where $0<\theta\leqq h$ and the emf itself is given by the well-known expression

$$e(t) = \exp\left[j(\omega_0\theta + \epsilon\theta^2)\right]. \tag{1}$$

This emf is repeatedly applied in series with a resonant RLC circuit of angular frequency ω_1, and we assume that

$$\omega_0 \leqq \omega_1 \leqq (\omega_0 + 2\epsilon h).$$

We will find the expression of the current $i(t)$ flowing through the circuit at the time t during the m^{th} interval, assuming that the circuit is at rest for $t=0$. We have

$$L\frac{di}{dt} + Ri + \int_0^t \frac{idt}{C} = e(t). \tag{2}$$

Using the symbol $\mathcal{L}[e(t)]$ for the Laplace transform of $e(t)$, and $I(p)$ for the transform of $i(t)$, we can write

$$I(p) = \frac{1}{L}\cdot\frac{p\mathcal{L}[e(t)]}{(p-p_1)(p-p_2)}, \tag{3}$$

where

$$\left.\begin{matrix}p_1\\p_2\end{matrix}\right\} = -\frac{R}{2L} \pm j\sqrt{\frac{1}{LC}-\frac{R^2}{4L^2}} = -\delta \pm j\omega_1.$$

The complete development of (3) is given in the appendix where it is shown that, with simplifications legitimate in the case of a high-frequency circuit of a relatively high Q, the final expression of the current in function of the time is

$$i(t) = \frac{e^{j\omega_1\theta}}{2L}\left\{e^{-\delta\theta}\int_0^\theta e^{\delta T + j(\omega_0-\omega_1)T + j\epsilon T^2}\cdot dT + e^{-\delta\theta}\cdot i(h)\right\}. \tag{4}$$

θ is the time in any interval $(0, h)$ $(h, 2h)$ $\cdots$ counted from the beginning of the interval, and $i(h)$ the value of the current at the end of the preceding interval.

The response thus comprises two terms, one arising during the m^{th} interval, the other due to the preceding sweeps. We shall now consider these terms separately.

III. Term Arising During the m^{th} Sweep

This term is the most significant one and deserves special consideration. In the following, we shall make several changes of variable, and their relations are tabulated in the following table. The reader is also referred to Fig. 1(c).

In order to obtain an expression which can be more easily computed, we first change the origin of the time, taking as new origin the moment when the exploration frequency coincides with the static resonance frequency of the circuit. This gives

$$\omega_0 + 2\epsilon t_0 = \omega_1.$$

As no confusion is possible with the previous mathematics, we call the new variable t with the new origin (see Fig. 1(c)); t is now >0 after the coincidence and <0 before.

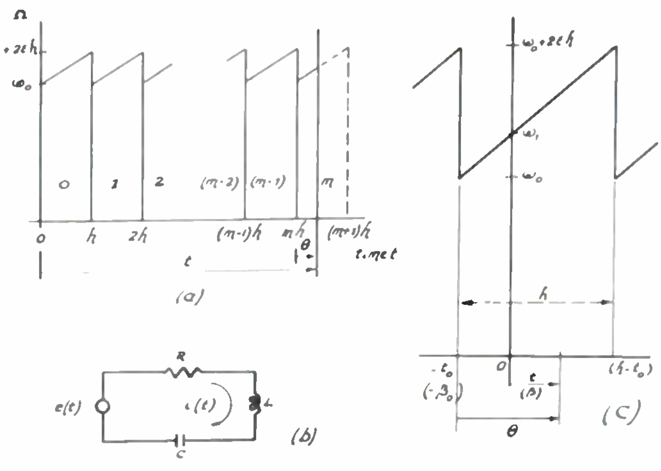


Fig. 1 (a) Schematic representation of the sawtooth-varying exploration frequency
(b) Diagram of the circuit
(c) Change of origin for the m^{th} sweep (§§3 and 4).

The term arising during the m^{th} sweep is thus

$$i(t) = e^{j(\omega_1 t+\Phi_0)}\left[\frac{e^{-\delta t}}{2L}\int_{-t_0}^{t} e^{\delta t + j\epsilon t^2}\cdot dt\right]; \tag{5}$$

$\exp\left[j(\omega_1 t+\Phi_0)\right]$ represents an oscillation having the resonance frequency of the circuit, with a constant phase shift. The argument of the term between brackets gives a supplementary phase, variable with the time, and the modulus gives the amplitude of the current.

Table of Symbols for §§ 3 and 4

Time

1. Variable outside the integration sign: $t = \theta - t_0$ $(0<\theta\leqq h)$; $\beta = \sqrt{\epsilon}\cdot t$
2. Variable under the integration sign: $t = T - t_0$; $\beta = \sqrt{\epsilon}\cdot t$
3. Origin for the interval under consideration: $-t_0 = -(\omega_1-\omega_0)/2\epsilon$; $-\beta_0 = -\sqrt{n}\,\pi B t_0$
4. Limits of the sweep with the new origin: lower: $t=-t_0$; upper: $t=(h-t_0)$

Other symbols

Exploration speed in cycles/sec²: $2\epsilon/2\pi = n\pi B^2$; $\epsilon = n\pi^2B^2$; $\delta = R/(2L) = \pi B$

Damping: $\delta/\sqrt{\epsilon} = 1/\sqrt{n}$

Phase: $\Phi_0 = (\omega_1^2-\omega_0^2)/4\epsilon$

Frequency deviation from the static resonance: $\Delta F = \sqrt{n}\,\beta B$

IV. Introduction of the Bandwidth B at −3 db as General Parameter

Hok had the idea of obtaining "universal curves," permitting a study of the response of any circuit, and introduced new variables. With the same idea in mind, but going a little farther in this direction, we introduce as general parameter the bandwidth B of the static selectivity curve of the circuit measured 3 db below the resonance. This element is generally known, and is given by $B=\omega_1/(2\pi Q)$. Introducing the new variable $\beta=\sqrt{\epsilon}\cdot t$ as Hok has done and taking into account the symbols listed in the table, we obtain as final expression of the current arising during the m^{th} sweep

$$i(\beta) = e^{j(\omega_1 t+\phi_0)}\left[\frac{e^{-\beta/\sqrt{n}}}{2\pi L B\sqrt{n}}\int_{-\beta_0}^{\beta} e^{\beta/\sqrt{n}+j\beta^2}\cdot d\beta\right]. \quad (6)$$

The amplitude $|i(\beta)|$ of the current is the modulus of the term between brackets.

It is worthwhile to note that the integral can be converted into a complementary error function between complex limits by putting, for example,

$$ju = \sqrt{j}\cdot\beta + 1/(2\sqrt{jn}),$$

which gives

$$\int_{-\beta_0}^{\beta} e^{\beta/\sqrt{n}+j\beta^2} d\beta = \sqrt{j}\, e^{j/4n}\int_{u_0}^{u} e^{-u^2}\cdot du$$

$$= \sqrt{j}\cdot\frac{\sqrt{\pi}}{2}\, e^{j/4n}[\mathrm{erfc}(u_0) - \mathrm{erfc}(u)].$$

Convergent and asymptotic series of such a function are known,[9] but as no tables were available to us, we have made graphical integrations of (6), the form of which is very convenient for that kind of computation.

We calculated by this method the response curves for $n=1$, $n=4$, and $n=9$, assuming that $|\beta_0|$ is large. This condition allows for a direct comparison with the results of Hok and Barber and Ursell as these authors took $-\infty$ as the lower limit of integration. Fig. 2 gives the three curves we have computed, with some published by Hok and Barber and Ursell, and transformed them to conform to our own notations. In Fig. 2, we have taken as ordinates the values of $|i(\beta)|2\pi LB$ and as abscissas $\Delta F/B$, where ΔF is the frequency deviation with reference to the static resonance frequency of the circuit.

Another graphical representation of the response curve which is more in conformity with what is normally seen on an oscilloscope is given for two values of n on Fig. 3, for which the time is taken as abscissas. These curves can be direcetly compared with the experimental curves of Fig. 4(b) and (c).

V. Conclusions Concerning the Term Arising During the m^{th} Sweep

Bearing in mind that the curves were computed for $|\beta_0|$ very large, the examination of Fig. 2 shows that when a dynamic exploration of the type considered here is used the response of the circuit differs from the static response by the following phenomena:

1. The apparition, after the maximum amplitude of the current, of a series of fluctuations (the so-called "ringing") which are the result of the beating between oscillation of the circuit on its own frequency and the forced oscillation due to the exploring emf.

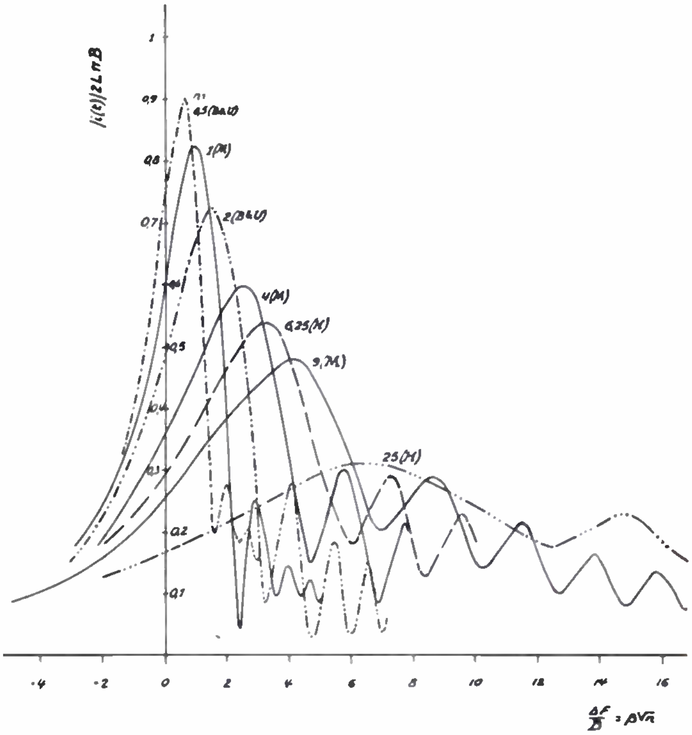


Fig. 2—Response curves in function of the frequency deviation in terms of the bandwidth B, for various values of the exploration speed parameter n. The curves in interrupted line are transposed from Hok's paper or from Barber and Ursell's. Those in full line were calculated by graphical integration by the author.

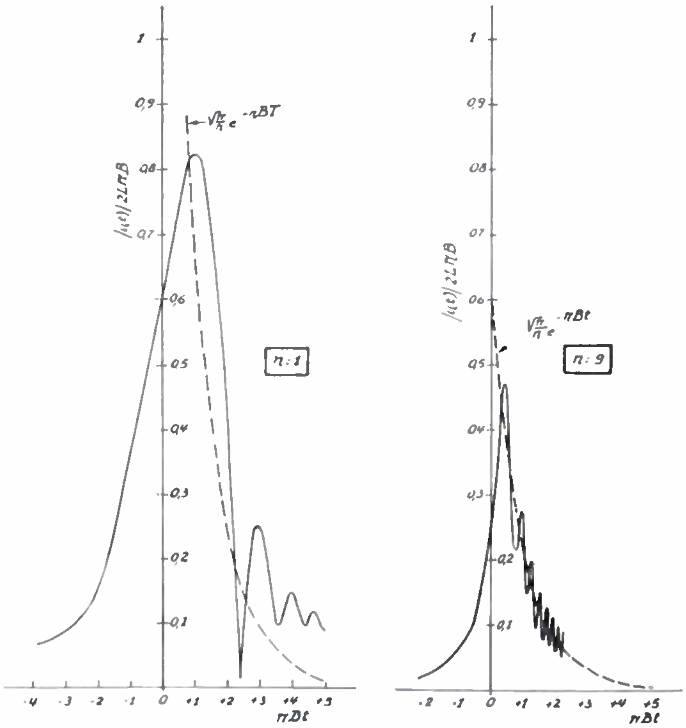
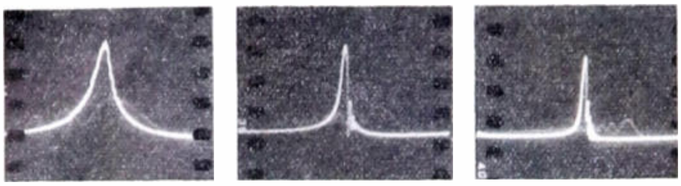


Fig. 3—Response curves for $n=1$ and $n=9$ in function of the time showing the aspect of the response curve generally observed in practice (compare with Fig. 4).

2. A reduction of the maximum amplitude with reference to the static response.

[9] For example, Jahnke-Emde, "Tables of higher functions," p. 24; 1948.

3. A shift of the position of the maximum toward the high frequencies when the exploring emf is of increasing frequency. The reverse would, of course, happen if the angular exploring frequency were $\Omega=(\omega_0-2\epsilon\theta)$.
4. An apparent reduction of the selectivity of the circuit which results in an increase of the apparent bandwidth.



(a) (b) (c)

Fig. 4—Three experimental records of a 225 -kc/s quartz-crystal response (a) below the critical speed ($n \ll 1$), (b) for the critical speed ($n=1$), (c) above the critical speed ($n \approx 4$). The flyback responses are visible and correspond to much larger values of n. The small ripples in the curves are due to the ac power supply.

These conclusions are well-known, but we would call the reader's attention to the following interesting facts:

As the curve flattens when the exploration speed is increased, at least when $n>1$, the frequency corresponding to the maximum is not so well defined, and this is an example of the indetermination principle, or of the "reciprocal spreading" of time and frequency. This could be overlooked in practice when using the usual method of observation with an oscilloscope when the abscissas are proportional to the time; but, in Fig. 2, the abscissas being proportional to the frequency deviations, the fact is clearly apparent (as in the curve for $n=25$).

One can distinguish three different kinds of response curves according to the value of the exploration speed:

1. At very low speeds, there is no ringing.
2. Between $n=0$, 5 and $n=3$ or about, very severe and characteristic distortion is present, the largest occurring in the neighborhood of $n=1$ which we consider as a *critical speed*. The highest maximum is followed by a deep minimum, but the amplitude of the oscillations is rapidly damped.
3. When the exploration speed is increased to $n=4$ and above, the shape of the response curve becomes different: There is no more deep minimum, but a succession of spaced maxima and minima. These maxima and minima are situated respectively above and below an exponentially decreasing curve. Hok had pointed out the fact. This can be verified on Fig. 3(b) for $n=9$, and on the original curves published by Hok. In the co-ordinates used in Fig. 3(b), the exponential curve corresponds to [mean decreasing value of $|i(t)2L\pi B|$] $=\sqrt{\pi/n}\cdot\exp[-\pi Bt]$.

On the Fig. 3(a), which corresponds to the critical speed $n=1$, it is seen that the response curve does not follow that law.

The experimental curves of Fig. 4(a), (b), and (c) correspond respectively to the three cases. The flyback traces which are visible give a good idea of what happens at very high speeds (compare with the curve $n=25$ on Fig. 2).

VI. Influence of the Limits of Exploration

As we have pointed out hereabove, the curves of Fig. 2 correspond to very large values of $(-t_0)$, and so do not take into account the finite lower limit of integration in (5) or (6).

Moreover, they do not take into account the second term of (4), which is due to the existence of an upper limit in the exploration. We shall now consider the influence of these limits.

A. Influence of the Lower Limit

When $(\omega_1-\omega_0)$ is small (Fig. 1(c)), t_0 and β_0 are small and the response curve is distorted. For example, on Fig. 5 two curves for $n=1$ are drawn. The one in full line

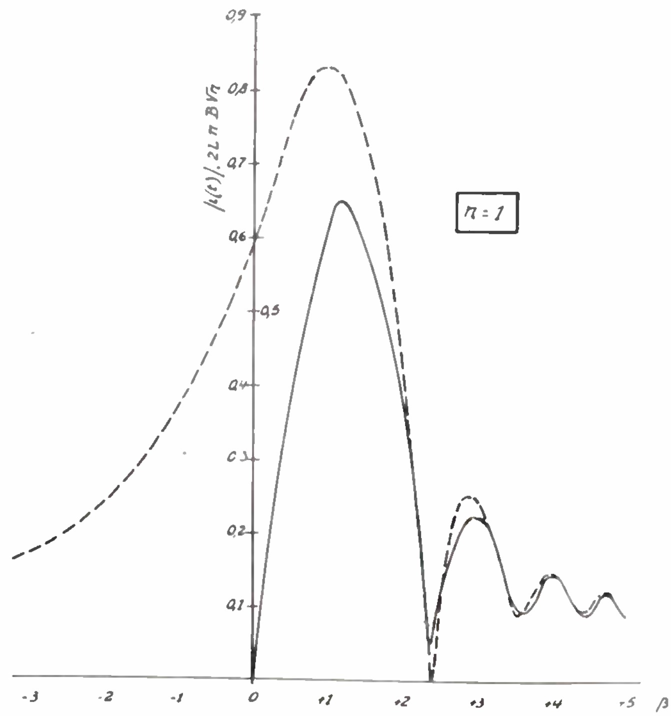


Fig. 5—Influence of the lower limit of exploration. Curve in full line: $\beta_0=0$ (the static resonance frequency of the circuit is equal to the initial frequency of the exploration $\omega_1=\omega_0$). Curve in interrupted line: $-\beta_0=-\infty$ (the resonance frequency of the circuit is much higher than the initial exploring frequency).

corresponds to $\beta_0=0$, the one in interrupted line corresponds to $-\beta_0=-\infty$. It is immediately seen that the position and the value of the maximum are not at all the same for both curves. This is very important in the case of a spectrum analyzer; one can then draw the conclusion that if a component of a spectrum lies too near the beginning of the exploration neither its amplitude nor its position on the frequency scale are correct with respect to other components more remote from the initial frequency.

B. Influence of the Upper Limit

It is obvious that when $\omega_1 \approx (\omega_0 + 2\epsilon h)$ the response curve may be suddenly cut down just when the exploration ceases. The response curve is not complete; but if the maximum occurs before the end of the sweep, neither its value nor its position are affected. However, since the exploration has not been extended far enough, the value of $i(h)$ in (4) is not zero, and may, in practice, have a large value.

During the following sweep, the current decreases exponentially with the time according to the law $\exp[-\delta\theta]\cdot i(h)$, but is superimposed on the term arising during the new sweep, at least up to the time when $\exp[-\delta\theta]$ becomes negligible (Fig. 6).

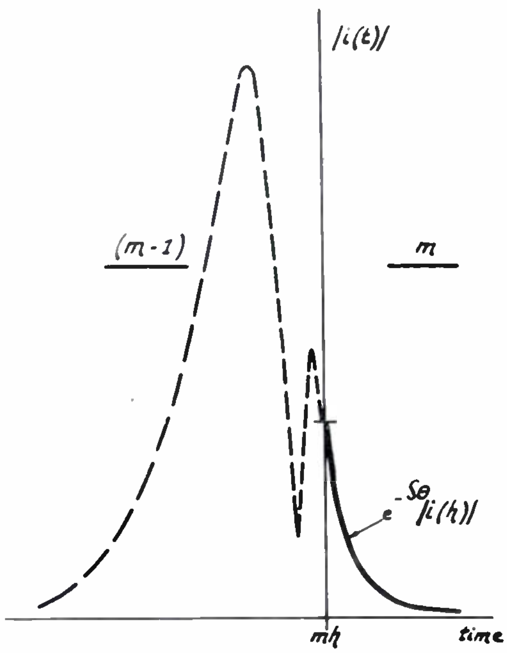


Fig. 6—Influence of the upper limit of exploration. The current existing at the end of one sweep decreases exponentially during the following one.

As shown in (4), $i(h)$ comprises terms resulting from all the preceding sweeps, but in practice only the last one needs to be considered.

This case is of importance when analyzing a spectrum with several components. Some of these are in the immediate neighborhood of both limits of the exploration, the response curves of those occurring near the lower limit ω_0 being affected by those occurring near the upper limit $(\omega_0 + 2\epsilon h)$.

C. Influence of the Flyback

During the sudden return of the sweeping angular frequency from $(\omega_0 + 2\epsilon h)$ to the initial value ω_0, a spurious exploration takes place. Theoretically, the exploration speed is infinite, giving rise to no response at all; but actually this speed is finite although very large.

The response curve corresponding to the flyback interval takes a shape like the one shown in Fig. 2 for $n = 25$, but in the reverse direction (Fig. 4).

At the end of the flyback, the current has a finite value, and exactly as the term due to the preceding sweep, its amplitude decreases exponentially with the time during the m^{th} sweep. This current is also superimposed on the current arising during the m^{th} sweep.

VII. Conclusions on the Influence of the Limits

The use of a limited but repeated sweep involves distortions in the response curve, which ***all occur at the beginning of each sweep.***

These distortions are due to a too small difference between ω_0 and ω_1, to the residual current at the end of each sweep, and to the residual current at the end of the flyback. It must be understood that this is of special importance when analyzing a RF spectrum composed of several components scattered along the frequency range of the exploration.

VIII. General Conclusions

Van der Pol[1] after investigating the circumstances in which the dynamic response curve is approximately the same as the static resonance curve, gave the following condition

$$\epsilon \cdot |Y''(j\Omega)| \ll |Y(j\Omega)|,$$

where $Y(j\Omega)$ is the admittance of the circuit and $Y''(j\Omega)$ its second derivative. For the RLC series circuit considered in this paper, Van der Pol's condition becomes

$$\epsilon \ll 0.5\pi^2 B^2$$

or

$$n \ll 0.5.$$

This is in full agreement with the above results since Fig. 2 shows that for $n = 0.5$ there is a slight "ringing."

But the results exposed in this paper show that higher exploration speeds are quite usable provided that new phenomena due to the dynamic exploration be taken into account. Nevertheless, it must be pointed out that, in order to take full advantage of the selectivity of the circuit, the higher the selectivity, the lower the exploration speed must be.

Moreover, in extending the preceding results to other cases, one must keep in mind that the conclusions are valid only when the impedance of the selective system can be entirely assimilated to that of a RLC series circuit. When the selective system is of more complicated texture, some conclusions may become questionable.

Furthermore, so far as the effect of a repeated sweeping is concerned, we have shown that important differences arise which all introduce distortions at the beginning of the exploration, and the practical conclusion for this is that one must always disregard the very first part of the sweeping.

Appendix

A. Mathematical Development of (3)

In order to simplify the notations, we put

$$\left.\begin{matrix} a \\ b \end{matrix}\right\} = 1 \pm j\delta/\omega_1.$$

Then, Heaviside's expansion gives for (3)

$$I(p) = \frac{1}{2L}\left[a/(p-p_1) + b/(p-p_2)\right]\mathcal{L}[e(t)].$$

The emf $e(t)$ being successively repeated in the intervals $(0, h)$ $(h, 2h)\cdots(mh, (m+1)h)$, a well-known rule of the operational calculus gives, for the Laplace transform of $e(t)$,

$$\mathcal{L}[e(t)] = [1 + \exp(-hp) + \cdots + \exp(-mhp)]$$
$$\cdot\int_0^h \exp(-pT)\exp[j(\omega_0 T + \epsilon T^2)]dt, \qquad (8)$$

where T is the variable of integration.

The inverse transform of $I(p)$ is given by the Bromwich integral

$$i(t) = \frac{a}{2L}\cdot\frac{1}{2\pi j}\int_{Br}\frac{\exp[pt]\cdot\mathcal{L}[e(t)]dp}{p-p_1}$$
$$+ \frac{b}{2L}\cdot\frac{1}{2\pi j}\int_{Br}\frac{\exp[pt]\mathcal{L}[e(t)]dp}{p-p_2}. \qquad (9)$$

In order to take immediate advantage of an important simplification which would otherwise appear only at the end of this discussion, we point out at this stage that the term containing p_2 can eventually be disregarded (see note below). It is thus not necessary to continue considering it, and we shall only work on the term in p_1.

Putting $\sum_{n=0}^{m}\exp[-nhp]$ for the summation in (8) and changing the order of integration, (9) gives

$$i(t) = \frac{a}{2L}\sum_{n=0}^{m}\int_0^h \exp[j(\omega_0 T + \epsilon T^2)]$$
$$\cdot\frac{1}{2\pi j}\int_{Br}\frac{\exp[p(t-nh-T)]dp}{p-p_1}dt. \qquad (10)$$

In this expression, the Bromwich integral has the value $\exp[p_1(t-nh-T)]$ when $(t-nh-T)>0$, and is zero otherwise. We thus have

$$i(t) = \frac{a}{2L}\sum_{n=0}^{m}\int_0^h e^{p_1(t-nh-T)+j(\omega_0 T+\epsilon T^2)}\cdot dT, \qquad (11)$$

provided, when using this expression, we take into account the above remark concerning the value of the Bromwich integral.

We now consider what happens during the m^{th} sweep for which we have

$$mh < t \leqq (m+1)h,$$

and we put for this interval

$$t = mh + \theta,$$

where

$$0 < \theta \leqq h.$$

θ is thus the time counted from the beginning of the m^{th} interval (Fig. 1(a)).

B. Contribution of the m^{th} Exploration Itself

In order to obtain this contribution, we consider separately the terms for $n=m$ in (11); the condition $(t-mh-T)>0$ shows that in this case the upper limit of integration h can be changed into θ for this interval without changing the value of the integral. Equation (11) gives for this term

$$\frac{a}{2L}e^{p_1\theta}\int_0^\theta e^{(j\omega_0-p_1)T+j\epsilon T^2}dT, \qquad (12)$$

which is the current arising during the m^{th} sweep.

C. Contribution of the Preceding Sweeps During the m^{th} Sweep

In (11), this contribution corresponds to $n=0, 1, 2\cdots(m-1)$. Obviously, for these sweeps the upper limit of integration is always h, so that (11) gives

$$\frac{a}{2L}e^{p_1\theta}\left[e^{(m-1)p_1h} + \cdots + 1\right]e^{p_1h}\int_0^h e^{(j\omega_0-p_1)T+j\epsilon T^2}\cdot dT,$$

which can be written in the form

$$\frac{a}{2L}e^{p_1\theta}\; i(h), \qquad (13)$$

where $i(h)$ is constant during the m^{th} sweep, as the only factor variable with the time is $\exp[p_1\theta]$.

D. Total Response During the m^{th} Sweep

The total response of the circuit during the m^{th} sweep is the sum of (12) and (13).

If we consider only a highly selective circuit, $\delta \ll \omega_1$, the imaginary term $j\delta/\omega_1$ in a, which produces only a small change of phase, can be neglected before unity.

Replacing p_1 by its value in the sum of (12) and (13), we obtain the following expression for the response:

$$i(t) = \frac{e^{j\omega_1\theta}}{2L}\left[e^{-\delta\theta}\int_0^\theta e^{\delta T + j(\omega_0-\omega_1)T + j\epsilon T^2}dT + e^{-\delta\theta}\cdot i(h)\right]. \quad (14)$$

E. Note

We have neglected the second right-hand term of (10) which has the same form as the first one but contains p_2 instead of p_1. The exponent of the corresponding integrand is

$$(j\omega_0 - p_2)T + j\epsilon T^2 = \delta T + j(\omega_0 + \omega_1)T + j\epsilon T^2.$$

In the case of a hf system, the exploration range $2\epsilon h$ being generally $\ll(\omega_0+\omega_1)$, an approximate value of the integral can be easily calculated assuming that ϵT^2 is negligible in front of $(\omega_0+\omega_1)T$. Such a calculation shows that the terms in p_2 which would appear in (11), (12), and (13) are negligible in front of those containing p_1, thus justifying the simplification made.

A Broad-Band Interdigital Circuit for Use in Traveling-Wave-Type Amplifiers*

R. C. FLETCHER†

Summary—**Because of its high-power handling capacity, the interdigital circuit has been considered for use in traveling-wave-type amplifiers. An analysis is presented here which indicates that this type of circuit can be arranged to give constant phase velocity over a wide bandwidth (30 per cent), as required to give constant gain. The analysis is qualitatively checked experimentally. The impedance parameter (proportional to the cube of the gain in db) is approximately the same as the flattened helix (such as has been used for the magnetron amplifier) and about one-third that of the conventional circular helix.**

Introduction

As has been shown,[1] in order to get the most gain from a traveling-wave-type amplifier, it is desirable to have a circuit with a high impedance parameter $(E^2/2\beta^2P)$, and in order to make it with a broad bandwidth, it is desirable that its phase velocity remain as constant as possible with changing frequency. The helix-type circuit is admirably suited in both of these respects, and has thus found wide application as a circuit in traveling-wave tubes. The helix has the disadvantage, however, of being unable to dissipate much power. This becomes a limiting factor in pushing traveling-wave amplifiers to higher power, particularly for magnetron-type traveling-wave amplifiers where the electrons are required to fall on the circuit for efficient operation. It would seem desirable to find a circuit which could dissipate more power than the helix without a significant sacrifice in impedance parameter or broadbandedness. A number of proposals have been made to this end,[2] but these have generally involved resonant elements which have either made the circuit rather narrow banded and somewhat sensitive to small circuit imperfections, or have had low impedance parameters. It is the purpose of this note to show that an interdigital structure with its inherent capability for dissipating more power can be arranged to be relatively broadbanded without too great a sacrifice in impedance parameter.

General Analysis

Let us consider first a general type of interdigital circuit, consisting of two sets of interlaced fingers of arbitrary cross section, each set attached to an opposite side of a base plate (Fig. 1). For a system such as this (in which (a) is uniform in the y direction, (b) has two or more unconnected regions of zero field (conductors), and (c) has boundary conditions in the $\pm y$ direction that can be confined to a small distance compared to the length of the circuit in that direction) a TEM mode, with no field components in the y direction, is sufficient to satisfy the boundary condition. This mode always

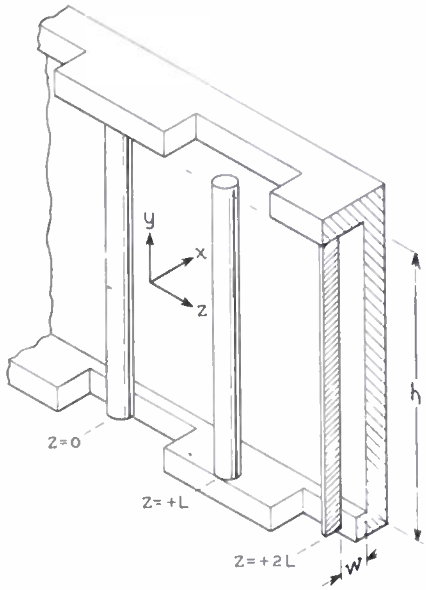


Fig. 1—Schematic drawing of an interdigital circuit with fingers of arbitrary cross section.

travels with the velocity of light so that the wavelength in the $\pm y$ direction is equal to the free-space wavelength λ. In a plane transverse to the y direction for such a mode, curl E vanishes so that a voltage can be defined which satisfies the two-dimensional Laplace equation. For a sinusoidal variation of fields with time, the voltage can be written

$$V(x, y, z, t) = F(x, z)(Ae^{-j(2\pi y/\lambda)} + Be^{+j(2\pi y/\lambda)})e^{j\omega t}, \quad (1)$$

where

$$\frac{\partial^2 F}{\partial x^2} + \frac{\partial^2 F}{\partial z^2} = 0. \quad (2)$$

Let us denote the voltage of the finger located at $(x=0, z=mL)$ by $V_m(y)$. A current $I_m(y)$ running along the mth finger in the y direction can be defined as the line integral of H around the periphery of the finger. These will be required to satisfy the interdigital boundary conditions in the y direction

$$V_{2n}\left(\frac{h}{2}\right) = V_{2n+1}\left(-\frac{h}{2}\right) = 0, \quad (3)$$

* Decimal classification: R339.2. Original manuscript received by the Institute, October 9, 1951; revised manuscript received, February 7, 1952.

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

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

[2] L. M. Field, "Some slow-wave structures for traveling-wave tubes," Proc. I.R.E., vol. 37; pp. 34–40; January, 1949.

$$I_{2n}\left(-\frac{h}{2}\right) = I_{2n+1}\left(\frac{h}{2}\right) = 0, \tag{4}$$

(n any integer).

We will look for a wave propagating in the z direction as $e^{-jm\theta}$. Because the boundary conditions have a period $2L$, we will have to add to this a wave propagating as $e^{-jm(\theta+\pi)}$. Note that since m is an integer, the addition of $2n\pi$ to θ does not change the exponential, and hence does not describe a different mode. Thus, these two waves are sufficient to satisfy any boundary conditions lumped at $[y=\pm(h/2),\ z=mL]$ with a z-direction periodicity of $2L$. The general form for V_m is therefore

$$V_m(\Phi) = (A_1 \cos\Phi + A_2 \sin\Phi)e^{-jm(\theta+\pi)} + (A_3 \sin\Phi + A_4 \cos\Phi)e^{-jm\theta}, \tag{5}$$

where Φ has been introduced for $2\pi y/\lambda$, and the time dependence has been omitted.

We will define the characteristic impedance, $K_c(\theta)$, in the usual fashion as the ratio of the voltage to the current of a particular finger for a single wave propagating in the $+y$ direction when the variation in the z direction consists of one mode varying as $e^{-jm\theta}$. Thus the current corresponding to (5) is

$$I_m = \frac{j}{K_c(\theta+\pi)}(-A_1\sin\Phi + A_2\cos\Phi)e^{-jm(\theta+\pi)} + \frac{j}{K_c(\theta)}(A_3\cos\Phi - A_4\sin\Phi)e^{-jm\theta}. \tag{6}$$

Conditions (3) on the voltage when placed in (5) give the relations

$$A_1 = -A_3\tan(\Phi_0/2) \tag{7}$$

$$A_2 = -A_4\cot(\Phi_0/2), \tag{8}$$

where Φ_0 equals $2\pi h/\lambda$. Conditions (4) on the current and the relations (7) and (8), when placed in (6), give two equations for A_1 and A_2, the simultaneous solution of which shows two sets of waves

$$1.\ A_2 = 0, \qquad \tan^2(\Phi_0/2) = \frac{K_c(\theta+\pi)}{K_c(\theta)} \tag{9}$$

$$2.\ A_1 = 0, \qquad \cot^2(\Phi_0/2) = \frac{K_c(\theta+\pi)}{K_c(\theta)}. \tag{10}$$

Actually, these waves are identical. For since the characteristic impedance is the same for a wave going in the negative y direction as in the positive (and thus the same for $+\theta$ as for $-\theta$), and since, as previously noted, adding $2n\pi$ to θ does not change the description of the mode, the following relations for K_c should be valid:

$$K_c(\theta) = K_c(-\theta) = K_c(\theta+2n\pi),\ (n \text{ any integer}) \tag{11}$$

$$K_c(\theta+\pi) = K_c(\theta-\pi) = K_c(\pi-\theta). \tag{12}$$

Thus, if we consider (9) as an equation determining $\theta_1(\Phi_0)$ versus frequency (subscript on θ refers to wave 1), and (10) as determining $\theta_2(\Phi_0)$ versus frequency, it is evident from (9), (10), and (11) that

$$\theta_1(\Phi_0) = \theta_2(\Phi_0) + \pi. \tag{13}$$

Examination of (5) and (6) shows that wave 2 has the same distribution of voltage and current for $\theta+\pi$ as wave 1 has for θ. The two waves are simply two different ways of describing the same wave. We will henceforth consider only wave 1, i.e., $A_2=0$.

Equations (5), (6), (7), and (9) thus represent a formal solution of the problem. To investigate further the phase velocity as a function of frequency from (9), we need to consider the dependence of $K_c(\theta)$ on θ for a particular shaped finger.

One general statement can be made which is independent of the finger shape. From (12) it can be seen that $\theta=\pi/2$, $K_c(\theta+\pi)=K_c(\theta)$, and thus (9) reveals that $\Phi_0=\pi/2$, independent of what K_c is; that is, when the length of the fingers is just one quarter wavelength, the phase constant, θ, equals $\pi/2$. This phase constant corresponds to a wave traveling with the velocity of light in the spaces between the fingers, up one space, across the open end of a finger, and down the next space, and so on.

Phase Velocity of Rectangular Fingers

The field equations have been solved for rectangular fingers (see Appendix) subject to the assumption that the field is uniform in the region between the wires. This assumption is probably a good one as long as the depth

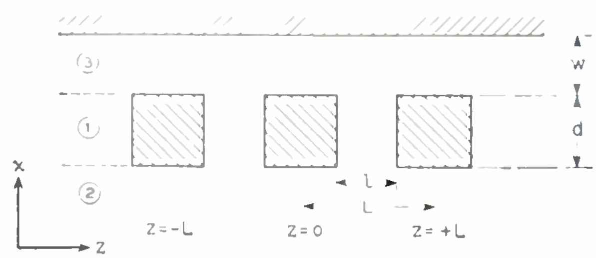


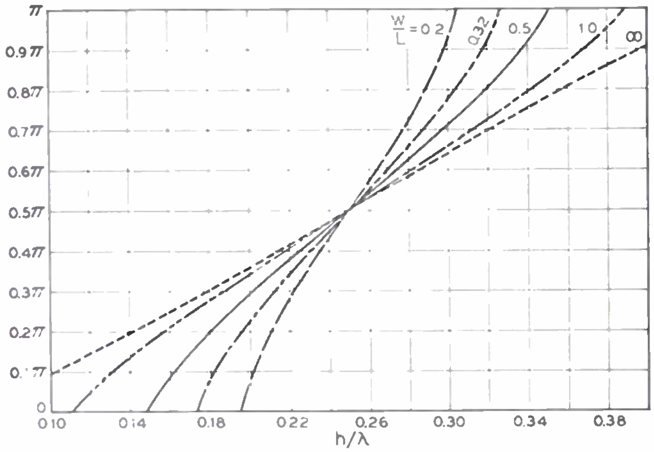
Fig. 2—Cross section of schematic interdigital circuit with rectangular fingers.

of the fingers, d, (Fig. 2) is large enough compared to the space between fingers, l. This analysis gives

$$\frac{1}{K_c(\theta)} = 2\sqrt{\frac{\epsilon}{\mu}}\sin\frac{\theta}{2}\left\{2\frac{d}{l}\sin\frac{\theta}{2} + \frac{L-l}{L}\sum_{n=-\infty}^{\infty}\left[1+\coth(\theta+2\pi n)\frac{w}{L}\right]\left[\frac{\sin(\theta+2\pi n)\frac{l}{2L}}{(\theta+2\pi n)\frac{l}{2L}}\right]\left[\frac{\sin(\theta+2\pi n)\frac{L-l}{2L}}{(\theta+2\pi n)\frac{L-l}{2L}}\right]\right\}. \tag{14}$$

quation (14) can be summed and substituted in (9)) obtain θ, the phase constant, as a function of h/λ, a uantity proportional to frequency. This has been done or typical values of d/l, l/L, and w/L, and the results re plotted in Fig. 3.

We may now ask how the phase velocity of the com-onent of the wave which is to interact with the elec-rons is related to θ. In order to decide which component) use, let us first write down an expression for the



ig. 3—The calculated phase constant as a function frequency for rectangular fingers ($d/l = 0.25$, $l/L = 0.4$).

oltage difference between adjacent fingers, which is roportional to the longitudinal field acting on the lectrons. This is obtained from (5), using (7) and (9)

$$E_z \sim V_{m+1} - V_m = -2A_1 e^{j(\theta/2)} \left[\cos\frac{\theta}{2} \cos\Phi e^{-jm(\theta+\pi)} - j\sin\frac{\theta}{2} \sin e^{-jm\theta} \right]. \quad (15)$$

We observe that E_z is composed of two sets of waves vith different z (i.e., m) variations, one with a phase onstant of θ and the other $\theta+\pi$. The θ component has a null at the center of the fingers ($\Phi = 0$), and thus varies a considerable amount with y, while the $\theta+\pi$ remains more nearly constant, particularly for small Φ_0. It would seem desirable therefore to use the $\theta+\pi$ component and operate at frequencies corresponding to small θ and hence small Φ_0.

The phase velocity for this component is

$$v_{\theta+\pi} = \frac{\omega L}{\theta + \pi} = c \frac{L}{h} \frac{\Phi_0}{\theta + \pi}. \quad (16)$$

In Fig. 4 this velocity divided by the velocity at $\theta = \pi/2$, $v_0 = c(L/3h)$ is plotted versus h/λ for rectangular fingers of different base plate distances, w. It can be observed that the velocity tends to pivot around $h = \lambda/4$, ($\theta = \pi/2$), so that for a particular value of w/L the velocity remains practically constant with changing frequency (within ± 1 per cent from $h/\lambda = 0.18$ to 0.25 or ± 16 per cent). The value of w/L, which makes the velocity independent of frequency, is dependent on the finger separation and thickness. This value has been

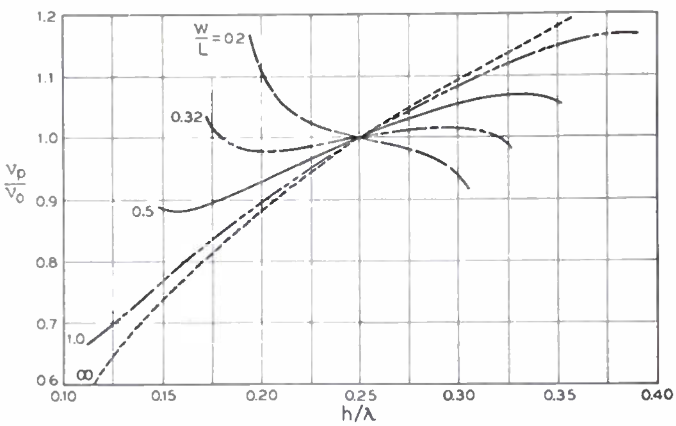


Fig. 4—The calculated phase velocity (divided by the velocity at $\theta = \pi/2$, $v_0 = cL/3h$) as a function of frequency for rectangular fingers ($d/l = 0.25$, $l/L = 0.4$).

calculated and plotted in Fig. 5 as a function of l/L for different values of d/L. (More exactly, this plot was calculated for $dv_{\theta+\pi}/d\Phi_0 = 0$ at $\theta = \pi/4$.)

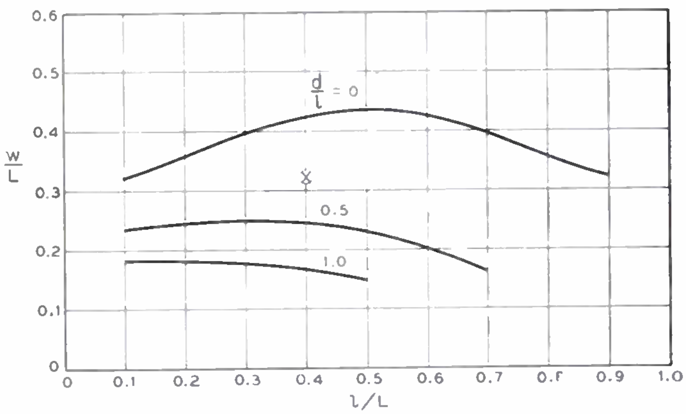


Fig. 5—The base-plate distance for broad-banded operation at $\theta = \pi/4$ as a function of finger width for rectangular fingers.

Experimentally Measured Phase Constants

In order to check the validity of the theoretical analysis presented above, a 300-mc model of an interdigital circuit was built with eight fingers and shorted at either end (z direction) so as to form a resonant cavity. This should have resonant frequencies corresponding to $\theta = n\pi/8$, $n = 1, 2, \cdots, 8$. The cavity was completely enclosed so that it had, effectively, two base plates. The fingers were made T-shaped in cross section for mechanical ruggedness. With two capacitive probes opposite the ends of the first and last fingers, the resonant frequencies of the cavity were measured for different spacings of one of the base plates. The field patterns were separately probed at each frequency to identify the mode. In Fig. 6 the phase velocity, normalized as in Fig. 4, is plotted against frequency. The similarity to the theoretical curve is apparent. Also, the resonant frequency for $\theta=\pi/2$ was such as to make the fingers very nearly a quarter wavelength.

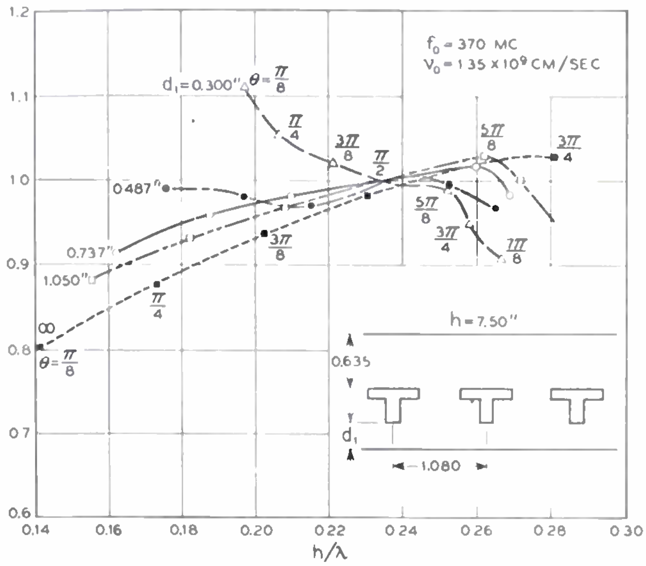


Fig. 6—Experimentally determined phase velocity as a function of frequency for fingers of T-shaped cross section.

Impedance Parameter

Since the power flowing down the circuit is equal to the product of the stored energy per unit length, W_s times the group velocity, v_g, the impedance parameter can be written as

$$K = \frac{E^2}{2\beta^2 v_g W_s}. \tag{17}$$

The group velocity can be obtained from the phase constant in the usual manner

$$v_g = \frac{d\omega}{d\beta} = \frac{L}{h}\frac{d\Phi_0}{d\theta}c; \tag{18}$$

v_g/v_p for rectangular fingers has been obtained from Fig. 3 and plotted in Fig. 7. At constant θ the group velocity decreases as w/L is decreased, as one would suppose from Fig. 4 where the slope of v_p versus h/λ decreases as w/L is decreased. The group velocity tends to zero at $\theta=0$, corresponding to a nonpropagating structure. At this point the fingers and the base plate act as a simple resonant circuit with the fingers mostly capacitive and the base plate inductive.

The stored energy per finger for a single TEM wave propagating in the y direction as $e^{-j2\pi y/\lambda}$ can be written as the product of the power flowing in the y direction times the length of the finger and divided by the velocity of light. Thus, for this type of wave, the stored energy per unit length is given by

$$W_s = \frac{1}{2}\frac{h}{cL}\frac{A_0^2}{K_c(\theta)}, \tag{19}$$

where A_0 is the voltage amplitude of the wave. For the wave consisting of a number of such TEM components in order to satisfy the boundary condition at $y=\pm h/2$, the stored energy is the sum of the energies in each of

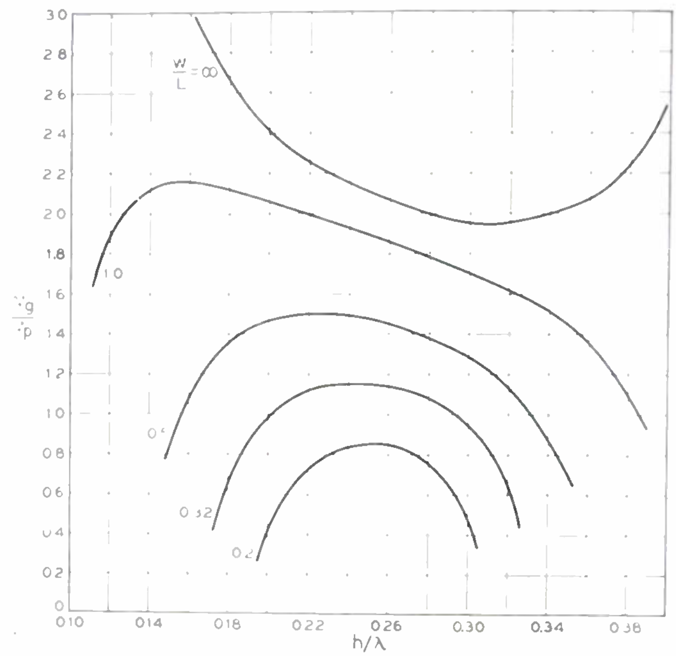


Fig. 7—The ratio of calculated group velocity to phase velocity as a function of frequency for rectangular fingers ($d/l=0.25$, $l/L=0.4$).

the TEM components. The stored energy per unit length of the interdigital circuit as represented by (5) (with $A_2=A_4=0$) is therefore

$$W_s = \frac{1}{4}\frac{h}{cL}\left[\frac{A_1^2}{K_c(\theta+\pi)} + \frac{A_3^2}{K_c(\theta)}\right]. \tag{20}$$

The first term inside the bracket of (20) is proportional to the energy of the $\theta+\pi$ component of the field, while the second is proportional to the θ component. Equations (7) and (9) indicate that these two are equal so that equal energies are stored in each component, independent of θ. This might at first seem contradictory to the conclusion drawn from (15), namely, that the $(\theta+\pi)$ component had a larger useful field than the θ component, the latter, in fact, vanishing at $\theta=0$. The answer is that around $\theta=0$, all the energy of the θ component is stored in the transverse field, and thus is not available for longitudinal interaction.

Since the amplitude of the voltage on the free end of a finger is $2A_3 \sin(\Phi_0/2)$, the stored energy at constant voltage is proportional to

$$\frac{\sqrt{\frac{\mu}{\epsilon}}}{K_c(\theta)\sin^2\frac{\Phi_0}{2}}.$$

his quantity is plotted in Fig. 8 for rectangular fingers. he stored energy is practically a constant independent frequency, but increases with decreasing w/L.

$$K = \frac{2}{\theta+\pi}\frac{v_p}{v_g}\cos^2\frac{\theta}{2}\tan^2\frac{\Phi_0}{2}(\Phi_0+\sin\Phi_0) \cdot \left[\frac{\sin(\theta+\pi)\frac{l}{2L}}{(\theta+\pi)\frac{l}{2L}}\right]^2 K_c(\theta). \qquad (22)$$

This is plotted in Fig. 9 as a function of h/L.

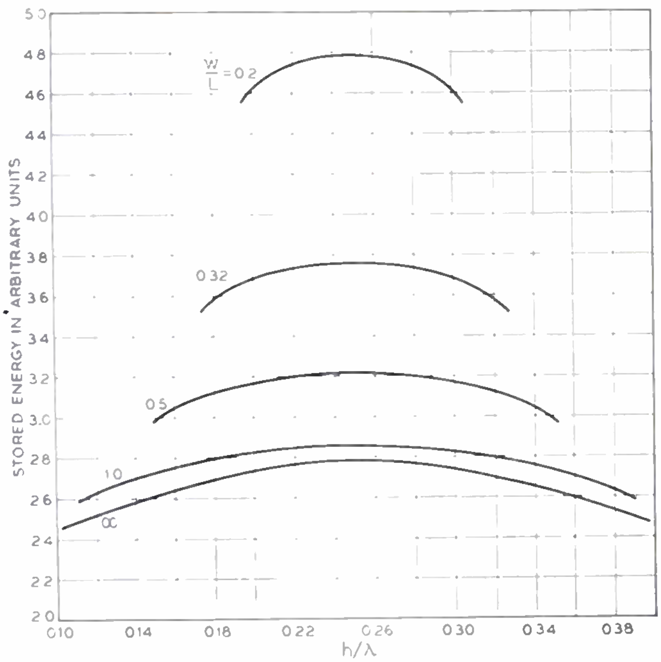


Fig. 8—The stored energy in arbitrary units as a function of frequency for rectangular fingers ($d/l=0.25$, $l/L=0.4$).

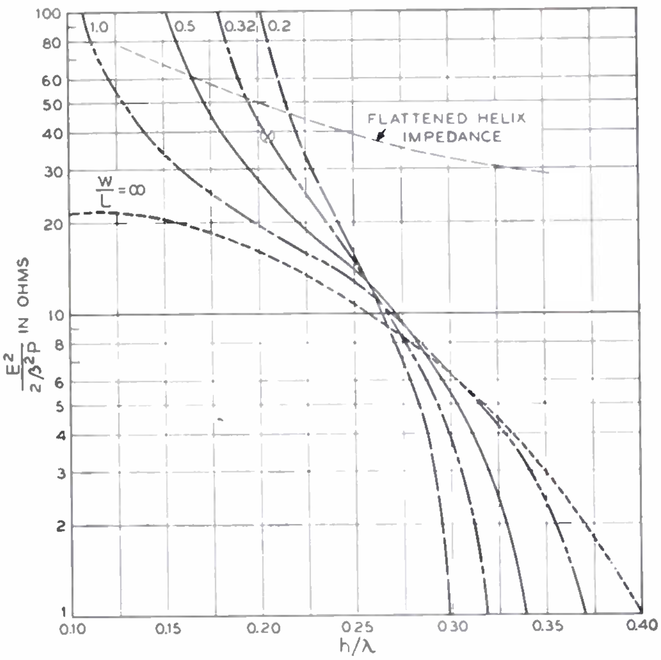


Fig. 9—The impedance parameter $E^2/2\beta^2P$ as a function of frequency for an interdigital circuit with rectangular fingers ($d/l=0.25$, $l/L=0.4$). The dashed curve is that for a flat helix which has the same interaction surface as the interdigital circuit.

The field E of (17) is the average field acting on the lectron, the averaging being done in a particular ashion such as described by Pierce[3] and Fletcher[4] for he helix. For a comparison of K we will compute the E appropriate to a thin beam just grazing rectangular ngers on one side. The appropriate field component is he $(\theta+\pi)$ harmonic of the field obtained from (15). This is easily found for rectangular fingers subject to he same assumptions made previously (by a similar rocess to that which led to (33) in the Appendix) to be

$$E = 2\frac{A_1}{L}\cos\frac{\theta}{2}\frac{\cos\Phi}{\cos\frac{\Phi_0}{2}}\frac{\sin(\theta+\pi)\frac{l}{2L}}{(\theta+\pi)\frac{l}{2L}}. \qquad (21)$$

The average of the square of this E over Φ must be substituted in (17) together with (20) to obtain K

[3] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., nc., New York, N. Y.; 1950.
[4] R. C. Fletcher, "Helix parameters used in traveling-wave tube heory," Proc. I.R.E., vol. 38, pp. 413–417; April, 1950.

The impedance parameter tends to ∞ at $\theta=0$ because of the vanishing group velocity. It decreases with increasing frequency apart from the effect of the group velocity because of the $\cos^2\theta/2$ term, coming originally from (21); that is, it decreases because the phase difference between adjacent fingers for the $\theta+\pi$ component approaches 2π as θ approaches π, and hence the voltage difference, and field, approach zero. At constant θ, the impedance parameter increases as w/L decreases even though the stored energy increases. This is brought about by the even more rapid decrease of the group velocity. We can thus see that if one wished to increase the gain the w/L could be decreased. But this results in a corresponding narrowing of the bandwidth, a quite general correlation, characteristic of all traveling-wave circuits whose group velocity can be varied.[3] This feature, which the developed helix does not have, may be useful if it is desired to exchange bandwidth for gain.

Let us now make a comparison of this impedance parameter with that of a helix. In order to compare them we will consider first a flattened helix, such as used by Warnecke and his co-workers for a magnetron amplifier,[5] and assume the same current density of electrons flowing past at the same distance. Since gain depends on the product of current and impedance parameter, we should compare the product of impedance parameter by the height, h, in the y direction. For a flattened helix, only about one-third of the circumference is available for electron interaction so that the effective height is $\frac{1}{3}\times 2\pi a$, where a is the effective radius. The impedance can be calculated, assuming a developed conducting sheet[3] to be

$$K_{\text{helix}} = \frac{\lambda}{8\pi^2 a}\sqrt{\frac{\mu}{\epsilon}}. \tag{23}$$

From this, the equivalent helix impedance is obtained by multiplying by

$$\frac{2}{3}\pi\frac{a}{h} = \frac{4\pi^2}{3\Phi_0}\frac{a}{\lambda};$$

thus

$$K_{\text{helix}} = \frac{1}{6\Phi_0}\sqrt{\frac{\mu}{\epsilon}}. \tag{24}$$

This impedance is plotted as a dashed curve in Fig. 9. Notice that at the point at which the phase velocity equals the group velocity for $\theta=\pi/4$ (indicated by a cross) the interdigital impedance is 0.78 times the equivalent helix impedance.

This comparison has favored the helix in that the energy stored in spatial harmonics of the helix has been neglected. For a round wire of optimum size, this additional stored energy knocks down the helix impedance parameter by a factor of 1.7 for four turns per wavelength, and by a factor of 3 for two turns per wavelength.[3] Thus, we see that the impedance parameter of the interdigital filter compares favorably with a flattened helix. A comparison with circular helices such as are found in ordinary traveling-wave tubes is not so favorable in this respect. The essential difference is that circular helices, interacting with circular cylindrical beams, may interact over the entire circumference. Thus circular helices have an effective impedance parameter three times that shown in Fig. 9. This would increase the gain in db by a factor of $(3)^{1/3}=1.45$ over the flattened helix for an ordinary traveling-wave tube.

Physical Interpretation of Impedance Parameter

At this point we might inquire as to the physical reasons for the impedance parameter of the interdigital circuit at the point where its phase velocity is constant, ($v_g=v_p$) being lower than that of the circular helix. This is simply a matter of stored energy. The highest possible impedance parameter for $v_g=v_p$ is obtained when all of the energy is stored where the electrons are located and in the single mode which is synchronous with the electrons. The circular helix is almost ideal in this respect (except for the small amount of energy stored in the TE mode and in higher spatial harmonics of the TM mode). The interdigital circuit, however, must store energy in at least one other mode (the θ component), and also must store additional energy in spatial harmonics of the same mode. This accounts for the factor of about 3 by which the circuit impedance is diminished over that of the circular helix.

One might suppose that, considering the additional energy stored, we should lose even more than the above-mentioned factor because of the close proximity of the base plate compared to the finger separation

$$\left(\frac{a}{L}\approx\frac{1}{3}\quad\text{for } v_g=v_p\right).$$

Actually, for the $\theta+\pi$ mode at small θ, the base plate is not very close compared to $\lambda_p/2\pi$. Most of the energy between the base plate and the fingers at small θ is stored in the θ mode, which energy has already been accounted for. Thus, although some additional energy is stored by having the base plate so close (as indicated in Fig. 8), it is not such a large factor as might at first be supposed.

About the same factor of 3 is lost by the flattened helix because of the energy stored in two-thirds of the circuit where there are no electrons. Thus for a magnetron amplifier where a flat circuit is generally required, the interdigital circuit should have almost identical impedance parameter, and hence identical gain as a flat helical circuit.

Acknowledgments

The writer is indebted to J. R. Pierce and L. R. Walker for several helpful discussions, to Mrs. M. E. Shannon for the numerical computations, and to W. W. Rigrod for criticisms of the manuscript.

Appendix

To find K_c for rectangular fingers we need consider only one of the TEM waves, which by itself need not satisfy any boundary condition in the $\pm y$ direction. Thus the voltage is given by

$$V_m = A_0 e^{-\phi_0}e^{-jm\phi}, \quad z = mL, \tag{25}$$

where the time variation is left off. K_c will be given by finding an expression for the current I_m and noting that

$$K_c = \frac{V_m}{I_m}. \tag{26}$$

To simplify the analysis, we will divide the region into three parts, as indicated in Fig. 2, and assume the electric field between the fingers, region 1, uniform.

[5] R. R. Warnecke, W. Kleen, A. Lerbs, O. Döhler, and H. Huber, "The magnetron-type traveling-wave tube," Proc. I.R.E., vol. 38, pp. 486–498; May, 1950.

$$E_{z1} = \frac{V_{m+1} - V_m}{l} = E_0 e^{-j\Phi} e^{-j(m+1/2)\theta},$$

$$\left(m + \frac{1}{2}\right)L - \frac{l}{2} < z < \left(m + \frac{1}{2}\right)L + \frac{l}{2}, \tag{27}$$

where

$$E_0 = -2j\frac{A_0}{2}\sin\frac{\theta}{2}. \tag{28}$$

The electric fields in regions 2 and 3 must be the superposition of waves having the correct periodicity in the z direction and satisfying the boundary conditions at $x = w$ and $x = -\infty$. (The origin of x is assumed to be one edge of the finger for region 2 and the other edge for region 3.)

$$E_{z2} = \sum_{n=-\infty}^{\infty} F_{n2} e^{+|\theta+2\pi n|x/L} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{29}$$

$$E_{z3} = \sum_{n=-\infty}^{\infty} F_{n3} \frac{\sinh(\theta+2\pi n)\left(\frac{W-x}{L}\right)}{\sinh(\theta+2\pi n)\frac{W}{L}} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L}, \tag{30}$$

where the subscripts indicate the respective regions. In order for the electric fields to be continuous across the boundaries of regions 1, 2, and 3,

$$\sum F_{n2} e^{-j(\theta+2\pi n)z/L} = \sum F_{n3} e^{-j(\theta+2\pi n)z/L}$$

$$= \begin{cases} E_0 e^{-j(m+1/2)\theta}, & \left(m+\frac{1}{2}\right)L - \frac{l}{2} < z < \left(m+\frac{1}{2}\right)L + \frac{l}{2} \\ 0, & mL - \frac{L-l}{2} < z < mL + \frac{L-l}{2}. \end{cases} \tag{31}$$

Multiplying both sides of (31) by $e^{+j(\theta+2\pi n)z/L}$ and integrating with respect to z from mL to $(m+1)L$, we find

$$F_n \equiv F_{n2} = F_{n3}$$

$$= \frac{1}{L} E_0 e^{-j(m+1/2)\theta} \int_{(m+1/2)L-l/2}^{(m+1/2)L+l/2} e^{+j(\theta+2\pi n)z/L} dz \tag{32}$$

$$= (-1)^n \frac{l}{L} E_0 \frac{\sin(\theta + 2\pi n)\frac{l}{2L}}{(\theta + 2\pi n)\frac{l}{2L}}. \tag{33}$$

The remaining field components are found with the aid of Maxwell's equations

$$H_{z1} = \sqrt{\frac{\epsilon}{\mu}} E_0 e^{-j\Phi} e^{-j(m+1/2)\theta},$$

$$\left(m+\frac{1}{2}\right)L - \frac{l}{2} < z < \left(m+\frac{1}{2}\right)L + \frac{l}{2} \tag{34}$$

$$E_{x2} = j\sum g_n F_n e^{|\theta+2\pi n|x/L} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{35}$$

$$H_{x2} = \sqrt{\frac{\epsilon}{\mu}} \sum F_n e^{|\theta+2\pi n|x/L} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{36}$$

$$H_{z2} = -j\sqrt{\frac{\epsilon}{\mu}} \sum g_n F_n e^{|\theta+2\pi n|x/L} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{37}$$

$$E_{x3} = -j\sum F_n \frac{\cosh(\theta+2\pi n)\frac{W-x}{L}}{\sinh(\theta+2\pi n)\frac{W}{L}} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{38}$$

$$H_{x3} = +\sqrt{\frac{\epsilon}{\mu}} \sum F_n \frac{\sinh(\theta+2\pi n)\frac{W-x}{L}}{\sinh(\theta+2\pi n)\frac{W}{L}} e^{-j\Phi} e^{-j(\theta+2\pi n)z/L} \tag{39}$$

$$H_{z3} = +j\sqrt{\frac{\epsilon}{\mu}} \sum F_n \frac{\cosh(\theta+2\pi n)\frac{W-x}{L}}{\sinh(\theta+2\pi n)\frac{W}{L}} e^{-j\Phi} e^{-j(\theta+2\pi n)}, \tag{40}$$

where $g_n = +1$ for $n \geqq 0$ and $g_n = -1$ for $n < 0$. Finally, the current flowing in the fingers can be obtained by integrating the tangential magnetic field around the periphery of a finger

$$I_m = d\left[H_{z1}\left(m + \frac{1}{2}\right) - H_{z1}\left(m - \frac{1}{2}\right)\right]$$

$$+ \int_{mL-(L-l)/2}^{mL+(L-l)/2} (H_{z2} - H_{z3})_{x=0} dz. \tag{41}$$

By using (34), (37), (40), and (28), integrating (41), and remembering the definition of K_c, (26), we obtain the equation for K_c in the text (14).

Glossary of Symbols

Rationalized mks units are used throughout; x, y, z, t, j, ω, λ, ϵ, μ, E_x, E_y, H_x, H_y have their usual meanings (c.f.,[3]).

β = propagation constant of wave which interacts with electrons $[e^{j(\omega t-\beta z)}]$
λ_p = wavelength of interacting circuit mode $= 2\pi/\beta$
$\Phi = 2\pi y/\lambda$
$\Phi_0 = 2\pi h/\lambda$
θ = phase difference between ends of fingers
A, A_0, A_1, A_2, A_3, A_4, B = arbitrary constants for voltage
c = velocity of light
d = thickness of finger (Fig. 2)
E = amplitude of longitudinal field at the position of the electrons
E_0 = amplitude of electric field between fingers (28)
F = transverse voltage parameter (1)
F_n, F_{n2}, F_{n3} = amplitude of spatial harmonic field components (33)
$g_n \equiv +1$ for $n \geqq 0$; $g_n \equiv -1$ for $n < 0$
h = length of fingers (Fig. 1)

I_m = current on mth finger
K = impedance parameter $= E^2/2\beta^2 P$
K_{helix} = impedance parameter for equivalent flattened helix
K_c = characteristic impedance of single wave of phase constant θ propagating in $+y$ direction
L = distance between centers of adjacent fingers (Figs. 1, 2)
l = spacing between adjacent edges of rectangular fingers (Fig. 2)
m = number of finger
n = arbitrary integer
P = average power flowing along filter
V = voltage at any point
V_m = voltage on mth finger
v_g = group velocity
v_p = phase velocity of wave which interacts with electrons
W = stored energy per unit length
w = distance separating rectangular fingers from base plate (Fig. 2).

The Frequency Spectrum of a Pulled Oscillator*

T. J. BUCHANAN†

Summary—**The spectrum of an oscillator is examined experimentally and theoretically under conditions where the oscillator frequency is disturbed by a stable signal injected into the oscillator circuit.**

Introduction

When a signal is injected into an oscillator circuit, the oscillator frequency is pulled toward the signal frequency. The pulling increases as the signal increases and as the signal frequency approaches the oscillator frequency, until the oscillator becomes locked to the signal. Adler has developed an equation giving the phase of the pulled oscillation as a function of time.[1] Using Adler's equation as a basis, the pulled oscillation is expressed as a Fourier series, each term of which corresponds to a frequency in the spectrum of the pulled oscillation.

The work was performed at a wavelength of 3.2 cm. The spectrum was displayed on the cathode-ray tube of a spectrometer. In Fig. 1 photographs are shown for various degrees of pulling. The incident signal (indicated by the arrow) was kept at constant power, and its frequency varied from a value greater than to a value smaller than the undisturbed oscillator frequency.

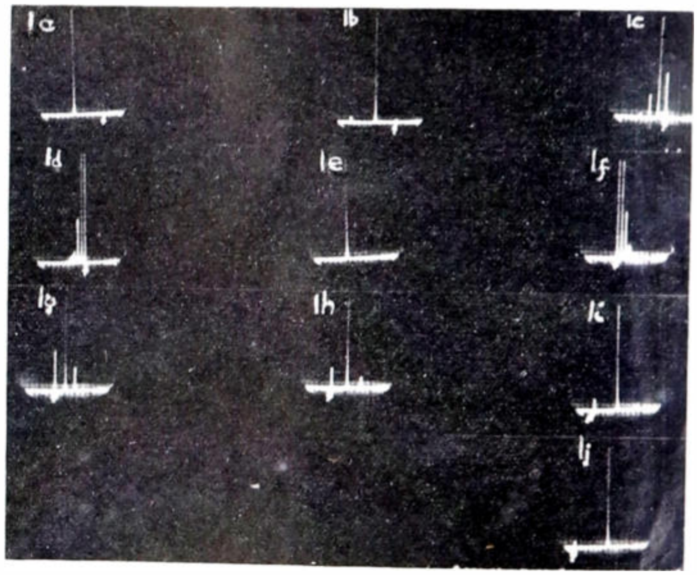


Fig. 1—Stages in the development of the spectrum of the pulled oscillator.

Apparatus

Power from a CV87 reflex klystron (Fig. 2) was fed into a waveguide. The tube under test, type 723A/B, was mounted at the other end. Two directional couplers monitored the power output of the tubes. A probe, loosely coupled to the waveguide, fed part of the signal and part of the output of the test tube to the spectrometer. To minimize mutual interaction the generator tube was chosen for its high power output and high-loaded Q -100 mw and 1,000 respectively. These quantities

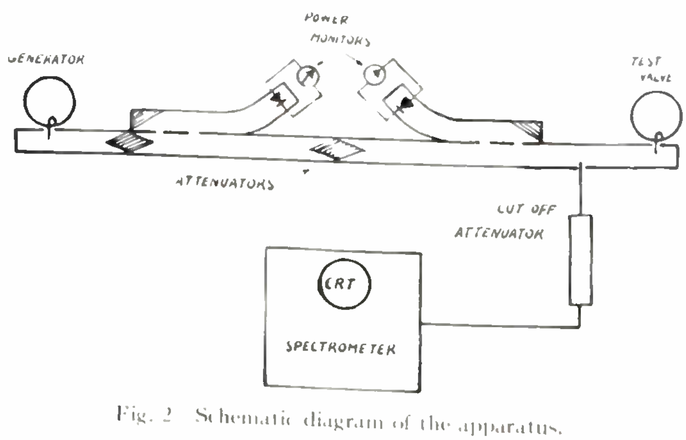


Fig. 2—Schematic diagram of the apparatus.

* Decimal classification: R355.912. Original manuscript received by the Institute, September 4, 1951; revised manuscript received February 21, 1952.
† Physicist, Middlesex Hospital Medical School, London, Eng.
[1] Robert Adler, "Locking Phenomena in Oscillators," Proc. IRE, vol. 34, pp. 351–357; June, 1946.

for the test tube were 15 mw and 100. This ensured that the pulling of the generator by the test tube was small compared with the pulling of the test tube by the generator.

Spectrum of the Pulled Oscillation

The spectrum is best described by reference to Fig. 1. In Fig. 1(a) the pip on the screen is due to the undisturbed oscillator. In Fig. 1(b) the arrow indicates the incident signal; the other small pip on the left of the trace is an additional frequency in the test oscillator spectrum arising from the pulling effect of the signal. In Fig. 1(c) the signal frequency has been increased and more components appear, equally spaced in frequency. The signal appears to have increased although the power monitor shows that the signal has remained constant. The increase in amplitude is due to a component of the pulled spectrum which has the same frequency as the signal. In Fig. 1(d) the spectrum develops rapidly until, Fig. 1(e), the oscillator becomes locked to the signal. There is always a component in the pulled spectrum of the same frequency as the signal. As the pulling increases, this component and the other additional components in the spectrum grow in amplitude. The primary component (Fig. 1(a)) diminishes as the pulling increases. Close to the locking point almost all the power in the spectrum is concentrated in the component at the signal frequency, and all the other components, having passed through a maximum, are becoming vanishingly small and finally disappear at the locking point.

Theory

Considering a conventional oscillator, Adler obtains, for conditions outside the locking range, and under certain conditions, the following equations:

$$\frac{d\alpha}{dt} = -B \sin \alpha + \Delta\omega_0, \tag{1}$$

$$\omega = -B \sin \alpha + \omega_0, \tag{2}$$

where

ω = the instantaneous angular frequency of the pulled oscillation,
ω_0 = the undisturbed angular frequency of the test oscillator,
ω_1 = the angular frequency of the incident signal,
$\Delta\omega_0 = \omega_0 - \omega_1$,
E_1 = voltage at the grid due to the incident signal,
E = voltage at the grid due to the pulled oscillation,
α = the phase difference, at the grid, between the pulled oscillation and the signal,
Q = the Q-factor of the oscillator circuit,
$B = E_1\omega_0/2EQ$.

Adler gives a solution of (1) in the form

$$\alpha = 2 \tan^{-1} \left\{ \frac{1}{K} + \frac{\sqrt{K^2 - 1}}{K} \tan \frac{Bt}{2} \sqrt{K^2 - 1} \right\}, \tag{3}$$

where

$$K = \frac{\omega_0 - \omega_1}{B}.$$

Adler's analysis is now applied to a reflex klystron oscillator. By using low signal powers and considering only those frequencies within the pass band of the resonator, all of Adler's conditions are satisfied.

The pulled oscillation can be written

$$y = \cos \int_0^t \omega dt = \cos \left(\omega_0 t - B \int_0^t \sin \alpha \cdot dt \right). \tag{4}$$

This is a general expression for phase- or frequency-modulated oscillations. For convenience, the amplitude has been taken to be unity. So long as $E_1 < E$, which is one of Adler's conditions, the assumption of a constant amplitude is justified. To develop (4) it is necessary to evaluate $\int_0^t \sin \omega \cdot dt$. Using (3), making the substitutions $\sin \beta = \sqrt{K^2 - 1}/K$ and $\phi = (Bt/2)\sqrt{K^2 - 1}$, and changing the variable from t to ϕ, one obtains

$$\int_0^t \sin \alpha \cdot dt$$

$$= \frac{2}{B \tan \beta} \int_0^\phi \frac{2 \cos \phi \cos (\phi - \beta)}{\cos^2 \phi + \cos^2 (\phi - \beta)} d\phi \tag{5}$$

$$= \frac{2\phi}{B \sin \beta} - \frac{2}{B} \int_0^\phi \frac{\sin \beta d\phi}{\cos^2 \phi + \cos^2 (\phi - \beta)} \tag{6}$$

$$= \frac{2\phi}{B \sin \beta} - \frac{2}{B} \int_0^\phi \frac{\tan \frac{\beta}{2} \cdot \sec^2 \left(\phi - \frac{\beta}{2}\right)}{1 + \tan^2 \frac{\beta}{2} \cdot \tan^2 \left(\phi - \frac{\beta}{2}\right)} d\phi \tag{7}$$

$$= \frac{2\phi}{B \sin \beta} - \frac{2}{B} \left[\tan^{-1} \tan^2 \frac{\beta}{2} + \tan^{-1} \left\{ \tan \frac{\beta}{2} \tan \left(\phi - \frac{\beta}{2}\right) \right\} \right]. \tag{8}$$

Simplifying (8) and substituting in (4),

$$y = \cos \left[\left(\omega_1 t + 2 \tan^{-1} \tan^2 \frac{\beta}{2} \right) + 2 \tan^{-1} \left\{ \tan \frac{\beta}{2} \tan \left(\phi - \frac{\beta}{2} \right) \right\} \right] \tag{9}$$

$$= \cos \xi \cos (2 \tan^{-1} \eta) - \sin \xi \cdot \sin (2 \tan^{-1} \eta), \tag{10}$$

where

$$\xi = \omega_1 t + 2 \tan^{-1} \tan^2 \frac{\beta}{2}$$

and

$$\eta = \tan \frac{\beta}{2} \tan \left(\phi - \frac{\beta}{2} \right).$$

Let

$$\cos(2\tan^{-1}\eta) = A_0 + \sum_{n=1}^{\infty}(A_n\cos n\psi + B_n\sin n\psi) \quad (11)$$

and

$$\sin(2\tan^{-1}\eta) = A_0' + \sum_{n=1}^{\infty}(A_n'\cos n\psi + B_n'\sin n\psi), \quad (12)$$

where

$$\psi = \phi - \frac{\beta}{2}.$$

The coefficients of these series are zero for all odd values of n. The only non zero coefficients are

$$A_0 = \frac{1 \pm \sin\beta}{\cos\beta} \quad (13)$$

$$A_{2m} = (-1)^{m-1}\cdot\frac{1}{2^{m-1}}\cdot\sin^2\beta\cos^{m-1}\beta F\left(\frac{m}{2}+1,\ \frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right), \quad (14)$$

and

$$B_{2m}' = \pm A_{2m}, \quad (15)$$

where

$$F\left(\frac{m}{2}+1,\ \frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right)$$

is a hypergeometric series. The positive sign in (13), and the negative sign in (15), is used when $\omega_0 > \omega_1$, and vice versa.

Equation (10) can now be written

$$y = \frac{1 - |\sin\beta|}{\cos\beta}\cos\xi + \sum_{m=1}^{\infty} A_{2m}\cos(\xi \pm 2mx), \quad (16)$$

the positive sign being taken when $\omega_0 > \omega_1$ and vice versa. Hence (16) represents a spectrum in which the frequencies, for $m > 1$, lie all above or all below the frequency of the signal.

The following deductions can be made from:

(a) The first term on the right of (16) excepted, all other terms vanish when $\sin\beta = 0$. The locking points are therefore given by the condition $\sin\beta = 0$, i.e. $K = \pm 1$. Since $K = (\omega_0 - \omega_1)/B$, the locking points are symmetrical with respect to ω_0 and the locking range is given by $2B$.

(b) The first term on the right of (16) can be written

$$\frac{1 - |\sin\beta|}{\cos\beta}\cos\xi = \frac{K - \sqrt{K^2 - 1}}{K}\cos\left(\omega_1 t + 2\tan^{-1}\tan^2\frac{\beta}{2}\right). \quad (17)$$

This component has the same frequency as the signal. The amplitude increases towards unity as K decreases towards unity; that is, as the locking point is approached, the energy in the spectrum tends to concentrate in this component. At the locking point all the other components vanish, leaving the energy concentrated in this single component. This is observed in the photographs.

(c) The frequency pulling of the primary component ($m = 1$) is readily shown to be given by

$$\frac{1}{2\pi}\left\{\omega_0 - (\omega_1 + B\sqrt{K^2 - 1})\right\}, \quad (18)$$

which agrees with the expression given by Adler.

(d) The amplitude A_{2m} of any component is determined by the value of $\cos\beta$. Since $\cos\beta = 1/K$, it follows that $\cos\beta$ increases from zero, in the undisturbed condition, to unity at the locking point. As $\cos\beta$ increases toward unity, the hypergeometric series increases monotonically, while the factor $\sin^2\beta\cos^{m-1}\beta$ increases, passes through a maximum, and then decreases to zero. When $\cos\beta$ tends to unity, the hypergeometric series becomes very slowly convergent and A_{2m} tends asymptotically to the expression

$$\frac{\sin\beta\cdot\cos^{m-1}\beta}{2^{m-1}}\cdot\frac{m!\sqrt{\pi}}{\left(\frac{m}{2}\right)!\left(\frac{m-1}{2}\right)!}, \quad (19)$$

which tends to zero as $\cos\beta$ tends toward unity. The value of A_{2m}, for $m > 1$, must then increase from zero, pass through a maximum, and then decrease to zero as the locking point is approached. It can be shown that A_2 ($m = 1$) decreases monotonically to zero as $\cos\beta$ increases from zero to unity.

(e) For $m \geqq 2$ the ratio of the amplitudes of successive components is approximately constant when $\cos\beta$ is small; but when $\cos\beta$ is large, the hypergeometric series increases rapidly and the amplitude ratio becomes less simple. When $\cos\beta$ is very close to unity, A_{2m} is given by (19). Examination of this expression shows that as $\cos\beta$ tends to unity each component tends to zero, and the ratio of successive components tends to $\cos\beta$, i.e., the components tend to become of equal amplitude while diminishing to zero.

In the immediate neighborhood of the locking point the pulling was so sensitive to small fluctuations in the signal that the spectrum was too unsteady to photograph. It was not possible therefore to show convincing agreement between theory and experiment in this region.

Acknowledgment

The work was performed while the author was a member of the Royal Naval Scientific Service. Acknowledgments are due to Mr. A. J. Sutton, who performed much of the experimental work; to Mr. O. L. Ratsey, who initiated the investigation; and to the Admiralty for permission to publish.

APPENDIX

The coefficients of the Fourier series are evaluated:

$$A_0 = \frac{1}{2\pi}\int_0^{2\pi} \cos\left\{2\tan^{-1}\left(\tan\frac{\beta}{2}\tan\psi\right)\right\} d\psi$$

$$= \frac{1}{2\pi}\int_0^{2\pi} \frac{\cos\beta + \cos 2\psi}{1+\cos\beta\cos 2\psi} d\psi$$

$$= \frac{1}{\cos\beta} - \frac{\sin^2\beta}{2\pi\cos\beta}\int_0^{2\pi}\frac{d\psi}{1+\cos\beta\cos 2\psi}$$

$$= \frac{1\pm\sin\beta}{\cos\beta}. \qquad (1a)$$

The negative sign is used when $\sin\beta$ is positive, and vice versa.

$$A_n = \frac{1}{\pi}\int_0^{2\pi}\frac{(\cos\beta+\cos 2h)\cos n\psi}{1+\cos\beta\cos 2\psi}$$

$$= -\frac{\sin^2\beta}{\pi\cos\beta}\int_0^{2\pi}\frac{\cos n\psi}{1+\cos\beta\cos 2\psi}d\psi$$

$$= \frac{-\sin^2\beta}{\pi\cos\beta}\int_0^{2\pi}\cos n\psi\{1-\cos\beta\cos 2\psi$$

$$+\cos^2\beta\cos^2 2\psi + \cdots$$

$$+(-1)^r\cos^r\beta\cos^r 2\psi + \cdots\}d\psi. \qquad (2a)$$

The series within the brackets are absolutely and uniformly convergent, when $\beta\neq 0$, and can be integrated term by term. The general term is

$$-\frac{\sin^2\beta}{\pi\cos\beta}(-1)^r\cos^r\beta\int_0^{2\pi}\cos n\psi\cdot\cos^r 2\psi\cdot d\psi.$$

Replacing $\cos^r 2\psi$ by its expansion in terms of cosines of multiples of 2ψ, the terms corresponding to odd values of n are found to vanish. If $n=2m$, one obtains finally

$$A_{2m} = (-1)^{m-1}\frac{\sin^2\beta\cos^{m-1}\beta}{2^{m-1}}F\left(\frac{m}{2}+1,\right.$$

$$\left.\frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right),$$

where

$$F\left(\frac{m}{2}+1,\ \frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right)$$

is a hypergeometric series which is absolutely and uniformly convergent for all values of m, if $B\neq 0$.

$$B_n = \frac{1}{\pi}\int_0^{2\pi}\frac{(\cos\beta+\cos 2\psi)\sin n\psi}{1+\cos\beta\cos 2\psi}d\psi$$

$$= 0, \text{ for all values of } n. \qquad (3a)$$

$$A_n' = \frac{1}{\pi}\int_0^{2\pi}\sin\left\{2\tan^{-1}\left(\tan\frac{\beta}{2}\tan\psi\right)\right\}\cos n\psi\cdot d\psi$$

$$= \frac{1}{\pi}\int_0^{2\pi}\frac{\sin\beta\sin 2\psi}{1+\cos\beta\cos 2\psi}\cos n\psi\cdot d\psi$$

$$= 0, \text{ for all values of } n. \qquad (4a)$$

$$B_n' = \frac{1}{\pi}\int_0^{2\pi}\frac{\sin\beta\sin 2\psi}{1+\cos\beta\cos 2\psi}\sin n\psi\cdot d\psi$$

$$= \frac{\sin\beta}{2\pi}\int_0^{2\pi}\frac{\cos(n-2)\psi-\cos(n+2)\psi}{1+\cos\beta\cos 2\psi}d\psi$$

$$= -\frac{\cos\beta}{2\sin\beta}(A_{n-2}-A_{n+2})$$

$$= 0, \text{ for odd values of } n,$$

$$= \frac{\cos\beta}{2\sin\beta}\{A_{2(m+1)}-A_{2(m-1)}\} \quad \text{when} \quad n=2m. \qquad (5a)$$

Now by a property of the hypergeometric function

$$F\left(\frac{m}{2}+1,\ \frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right)$$

$$= \frac{\pm 1}{\sin\beta}F\left(\frac{m}{2}+\frac{1}{2},\ \frac{m}{2},\ m+1,\ \cos^2\beta\right),$$

where the negative sign is used when $\sin\beta$ is negative, and vice versa. Now the following identity can be verified:

$$F\left(\frac{m}{2}+\frac{1}{2},\ \frac{m}{2},\ m,\ \cos^2\beta\right)$$

$$= F\left(\frac{m}{2}+\frac{1}{2},\ \frac{m}{2},\ m+1,\ \cos^2\beta\right)$$

$$+\frac{\cos^2\beta}{4}F\left(\frac{m}{2}+\frac{3}{2},\ \frac{m}{2}+1,\ m+2,\ \cos^2\beta\right),$$

that is,

$$A_{2(m-1)} = (-1)^{m-2}\frac{\sin^2\beta\cos^{m-2}\beta}{2^{m-2}}F\left(\frac{m}{2}+\frac{1}{2},\right.$$

$$\left.\frac{m}{2},\ m+1,\ \cos^2\beta\right)+A_{(2m+1)}$$

or

$$A_{2(m-1)} - A_{2(m+1)}$$

$$= (-1)^{m-2}\frac{\sin^2\beta\cos^{m-2}\beta}{2^{m-2}}F\left(\frac{m}{2}+\frac{1}{2},\right.$$

$$\left.\frac{m}{2},\ m+1,\ \cos^2\beta\right)$$

$$= \pm\frac{2\sin\beta}{\cos\beta}(-1)^{m-1}\frac{\sin^2\beta\cos^{m-1}\beta}{2^{m-1}}F\left(\frac{m}{2}+1,\right.$$

$$\left.\frac{m}{2}+\frac{1}{2},\ m+1,\ \cos^2\beta\right)$$

$$= \pm\frac{2\sin\beta}{\cos\beta}A_{2m}$$

$$\therefore\ B_{2m}' = A_{2m},$$

when $\sin\beta$ is positive and

$$B_{2m}' = -A_{2m},$$

when $\sin\beta$ is negative.

A Note on the Reproduction of Pulses*

DAVID K. CHENG†, SENIOR MEMBER, IRE

Summary—The present note aims to demonstrate the relative significance of rise time and percentage energy content in the reproduction of rectangular pulses. Formulas are derived and their dependence upon various factors discussed. It is also shown that the rise time for periodic rectangular pulses is in general not the same as that for a step signal or a single transient pulse; it depends upon the number of harmonic components, the pulse repetition period, and the pulse duration in a more complicated way.

I. Introduction

Theoretically, a transient pulse or a train of periodic pulses consists of an infinite number of frequency components. Consequently, an infinite bandwidth is required for perfect reproduction of either. In actual practice it is not possible to provide such a bandwidth, while problems often arise wherein a certain degree of faithfulness of reproduction is desired. It is therefore important to examine the various factors affecting the faithfulness of pulse reproduction.

Suppose a rectangular pulse is passed through a network which lets pass all frequencies lower than or equal to f_0, but blocks all those higher than f_0. Because of limited bandwidth, the pulse takes a certain length of time to build up its amplitude, i.e., it requires a certain "rise time." As one can easily realize, the rise time bears a close relation to the faithfulness of pulse reproduction. Since the resulting pulse from the network does not have strictly straight sides and it may overshoot at the top and undulate at the bottom, there is need to arbitrarily define the rise time in such a way that it is calculable. Two such ways of defining rise time have been used and are depicted in Figs. 1(a) and 1(b). In these

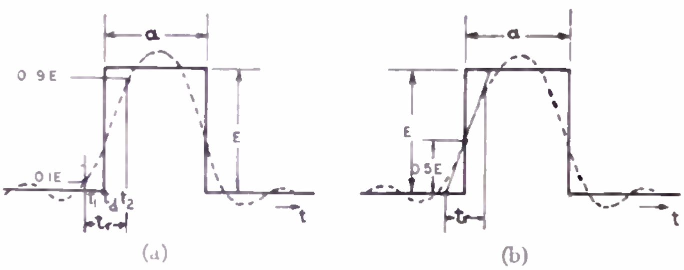


Fig. 1—Ways of defining rise time, t_r.

figures, solid lines form the original rectangular pulse and dashed lines represent the assumed response. In Fig. 1(a), the rise time t_r is defined as the time elapsed for the response to rise from $0.1E$ to $0.9E$, where E is the height of the original pulse. In Fig. 1(b), the rise time is considered as the time interval between the intercepting points on the original pulse of the tangent at $0.5E$. Both cases will be considered in the following.

Another way of judging the faithfulness of reproduction is to examine the energy content of the resulting pulse expressed in percentage of that of the original pulse. A relative standard for the rise time and percentage energy content may be obtained by correlation and comparison.

In this discussion attention is not directed to the question of best signal-to-noise ratio, or possible distortion due to nonideal response characteristics. Ideal amplitude and phase characteristics as shown in Fig. 2 will be assumed.

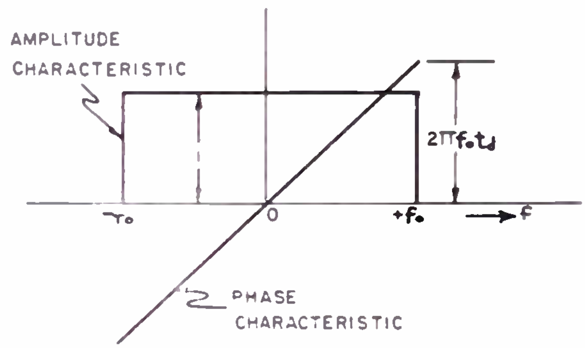


Fig. 2—Idealized amplitude and phase response characteristics of a selective network.

II. Rise Time of a Step Signal

It is well known[1] that the Fourier transform of a step signal of height E is

$$g(\omega) = \frac{E}{2\pi j\omega}, \tag{1}$$

and that the output after passing through a network of idealized response characteristics as shown in Fig. 2 can be represented by

$$e(t) = E\left\{0.5 + \frac{1}{\pi}\,\mathrm{Si}\,\left[2\pi f_0(t - t_d)\right]\right\}, \tag{2}$$

where t_d is the time delay.

If the rise time t_r is defined as the time required for the response to rise from $0.1E$ to $0.9E$ (see Fig. 1(a)), then $t_r = t_2 - t_1$ can readily be solved from (2) with the aid of a table for sine integrals such that

* Decimal classification: R143.5. Original manuscript received by the Institute, October 11, 1951; revised manuscript received, February 28, 1952.

† Electrical Engineering Department, Syracuse University, Syracuse, N. Y.

[1] E. A. Guillemin: "Communication Networks," John Wiley and Sons, New York, N. Y., vol. II, chapt. XI; 1935.

$$t_r = \frac{0.445}{f_0}. \quad (3)$$

If the rise time is defined according to Fig. 1(b), one as

$$t_r = \frac{E}{\left[\frac{de(t)}{dt}\right]_{t=t_d}}. \quad (4)$$

'sing (2), one obtains

$$t_r = \frac{0.5}{f_0}. \quad (5)$$

:omparing (3) and (5), it is seen that t_r defined accord-ıg to Fig.1(a) is less than that defined according to 'ig. 1(b); the latter case, therefore, represents a more onservative estimate. Both (3) and (5) show that the ise time t_r is inversely proportional to the width of the ransmission band f_0. Extending the bandwidth shortens he rise time. It is interesting to note that if one works ackwards on the assumption of $t_r=0.5/f_0$, it will be ound that this condition gives the time required for he pulse amplitude to rise from $0.064E$ to $0.936E$ in-tead of from $0.1E$ to $0.9E$.

Inasmuch as a single transient rectangular pulse can e considered as the superposition of a positive step ignal and a negative step signal of the same amplitude ut delayed by a time equal to the duration of the ulse, the rise time after passing through the idealized etwork may be calculated by the same formulas for a tep signal. This statement, however, must be qualified y the condition that the pulse is wide enough to allow he full amplitude to build up. Derivations of the effect f a selective network on the rise time for pulses of other hapes are difficult; improper integrals will be involved.

III. Relation between Energy Content and Bandwidth

If the available bandwidth is not wide enough to let hrough all the frequency components of a pulse, the esulting pulse would have less energy content than the riginal pulse. Specifying the percentage energy con-ent of the resulting pulse therefore puts a definite re-quirement on the width of transmission band and speci-ies a certain degree of faithfulness of reproduction.

Consider the case of a single transient rectangular ulse of height E and duration a. The Fourier trans-orm is then

$$g(\omega) = \frac{Ea}{2\pi}\left[\frac{\sin\frac{\omega a}{2}}{\frac{\omega a}{2}}\right]. \quad (6)$$

The positive half of $g(\omega)$ is plotted in solid lines in Fig. 3; the amplitude has been normalized to unity. In Fig. 3 is also plotted a curve in dashed lines showing the general shape of $g^2(\omega)$. Since energy content is proportional to the square of amplitude, use will be made of the energy spectrum. The bandwidth f_0 required to pass frequency components that represent p per cent of the total energy can be found by solving the following:

$$\int_0^{\omega_0}\left[\frac{\sin\frac{\omega a}{2}}{\frac{\omega a}{2}}\right]^2 d\omega = p\int_0^{\infty}\left[\frac{\sin\frac{\omega a}{2}}{\frac{\omega a}{2}}\right]^2 d\omega. \quad (7)$$

By carrying out the integrations and rearranging terms,

$$\mathrm{Si}\,(\omega_0 a) - \frac{1}{\omega_0 a}(1 - \cos\omega_0 a) = p\frac{\pi}{2}. \quad (8)$$

This equation is important. It shows the relation between percentage energy content, p, and angular bandwidth, ω_0.

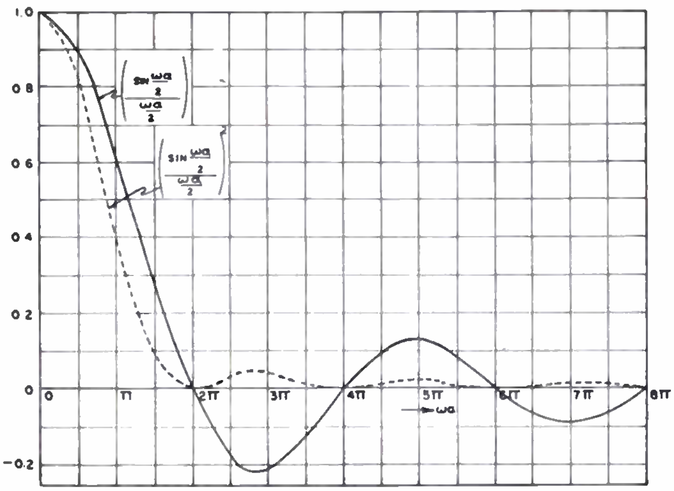


Fig. 3—Amplitude and energy spectra of a rectangular pulse of duration a.

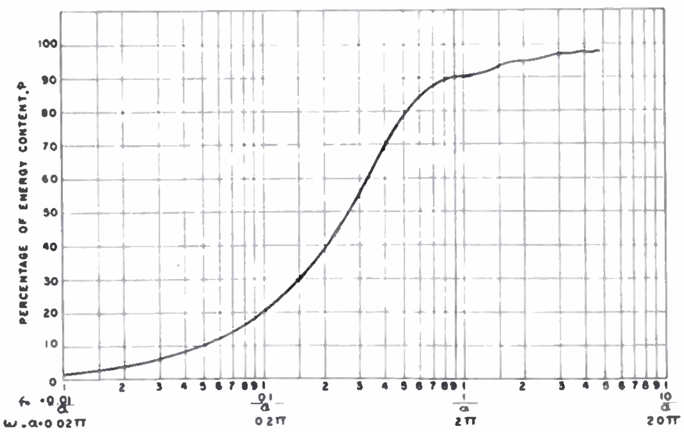


Fig. 4—Relations between energy content and bandwidth for a rectangular pulse of duration a.

If the bandwidth is made to be equal to the reciprocal of the pulse duration ($f_0=1/a$, or $\omega_0 a=2\pi$), (8) will show that the resulting pulse will contain 90.3 per cent of the original energy. From Fig. 3 it is also seen that with $\omega_0 a = 2\pi$, all the frequency components inside the main lobe of the function $g(\omega)$ up to the first zero are admitted. The rise time calculated by use of (5) is

$$t_r = \frac{0.5}{f_0} = 0.5a.$$

A bandwidth equal to the reciprocal of pulse duration is usually considered to be fairly satisfactory in reproducing rectangular pulses in practice. This condition represents an approximate rise time equal to one half the pulse duration and an energy content of 90.3 per cent. The curve plotted with percentage of energy content p versus bandwidth f_0 is shown as Fig. 4. The corresponding values of $\omega_0 a$ are also indicated on the abscissa. The oscillations of the curve after $f_0 = 1/a$ are due to the varying rate of energy increase with bandwidth as one goes up and down the humps of the dashed curve in Fig. 3.

IV. Rise Time of Periodic Rectangular Pulses

It is seen from (5) that the rise time for a step signal or a single transient rectangular pulse, the frequency spectrum of which is expressible in Fourier integral form, is equal to one half of the reciprocal of the bandwidth. For a train of period rectangular pulses, the frequency spectrum is a line spectrum instead of a continuous one. It can then be expressed in Fourier series form as a summation of discrete terms, and the relation between the rise time and the number of harmonic components can be established as follows:

Suppose the definition of rise time as depicted in Fig. 1(b) is used, which requires the differentiation of the original time function. The legitimacy of differentiating a Fourier series term by term demands that the original function $e(t)$ be continuous for all values of t, that $e'(t)$ has only a finite number of discontinuities in a period, and that both $e(t)$ and $e'(t)$ have limited total fluctuation throughout the period. If only a limited number of terms in the Fourier series is taken to represent a rectangular pulse train ($n = 1 \to N$, instead of $n = 1 \to \infty$), $e(t)$ will be a continuous function and the legitimacy for differentiating the series will not be violated. Thus,

$$e(t) = E\frac{a}{T}\left\{1 + 2\sum_{n=1}^{N}\frac{\sin\frac{n\pi a}{T}}{\frac{n\pi a}{T}}\cos n\omega t\right\} \tag{9}$$

$$\frac{de(t)}{dt} = -\frac{2E\omega}{\pi}\sum_{n=1}^{N}\sin\frac{n\pi a}{T}\sin n\omega t. \tag{10}$$

Equation (10) is equivalent to

$$\frac{de(t)}{dt} = -\frac{E\omega}{\pi}\left\{\frac{\cos\left[\frac{1}{2}(N+1)\left(\frac{\pi a}{T} - \omega t\right)\right]\left[\sin\left[\frac{1}{2}N\left(\frac{\pi a}{T} - \omega t\right)\right]\right]}{\sin\frac{1}{2}\left(\frac{\pi a}{T} - \omega t\right)} - \frac{\cos\left[\frac{1}{2}(N+1)\left(\frac{\pi a}{T} + \omega t\right)\right]\sin\left[\frac{1}{2}N\left(\frac{\pi a}{T} + \omega t\right)\right]}{\sin\frac{1}{2}\left(\frac{\pi a}{T} + \omega t\right)}\right\}. \tag{11}$$

It is easy to show that the second term in the bracket of (11) will have the value N as $t = t_a = -a/2$, or $[\pi a/T + \omega t]$ approaches zero. From (4), the rise time is

$$t_r = \frac{T}{2}\frac{1}{\left\{N - \frac{\cos\left[(N+1)\frac{\pi a}{T}\right]\sin\frac{N\pi a}{T}}{\sin\frac{\pi a}{T}}\right\}}. \tag{12}$$

Write

$$t_r = \frac{T}{2N}\frac{1}{\left[1 - \frac{1}{N}f(N)\right]} = \frac{T}{2N}F(N), \tag{13a}$$

where

$$f(N) = \frac{\cos\left[(N+1)\frac{\pi a}{T}\right]\sin\frac{N\pi a}{T}}{\sin\frac{\pi a}{T}}$$

$$= \frac{1}{2}\left\{\frac{\sin\left[(2N+1)\frac{\pi a}{T}\right]}{\sin\frac{\pi a}{T}} - 1\right\}. \tag{13b}$$

Note that the function $f(N)$ is not always equal to zero. Only when $f(N) = 0$, does one have the relation

$$t_r = \frac{T}{2N} = \frac{1}{2fN} = \frac{1}{2f_0},$$

which is equivalent to (5) for a step signal or a single transient rectangular pulse. $F(N)$ may be called the deviation factor. The condition for $f(N)=0$ is

$$\sin\left[(2N+1)\frac{\pi a}{T}\right] = \sin\frac{\pi a}{T}, \tag{14}$$

which may be split into two relations,

$$\text{(i)} \quad (2N+1)\frac{\pi a}{T} = \frac{\pi a}{T} + m2\pi$$

or

$$N = m\frac{T}{a}, \qquad m = 1, 2, 3, \cdots \tag{15a}$$

$$\text{(ii)} \quad (2N+1)\frac{\pi a}{T} = (2m-1)\pi - \frac{\pi a}{T}$$

or

$$N = (2m-1)\frac{T}{2a} - 1, \quad m = 1, 2, 3, \cdots \tag{15b}$$

Since $N/T = Nf = f_0 =$ bandwidth, (15a) and (15b) may also be written as

$$f_0 = \frac{m}{a} \tag{16a}$$

and

$$f_0 = \frac{(2m-1)}{2a} - \frac{1}{T}, \quad m = 1, 2, 3, \cdots. \tag{16b}$$

This means that the rise time of a train of periodic pulses is equal to one half of the reciprocal of the bandwidth only when (16a), which requires the bandwidth to be an integral multiple of the reciprocal of the pulse duration, or (16b), which depends upon both pulse duration and pulse repetition frequency, is satisfied. In general, $f(N) \neq 0$, and t_r may be greater or smaller than $T/2N$ or $1/2f_0$.

The fact that the rise time for a step signal or a single transient rectangular pulse after passing through an ideal low-pass filter of flat response over a limited bandwidth f_0 is different from that for a periodic rectangular pulse train is understandable, because a step signal or a transient pulse cannot be represented by a summation of discrete frequency components. Given a bandwidth f_0, it only means a certain finite number N of discrete harmonic components in the case of a periodic pulse train. The number of components N is evidently dependent on the pulse repetition period T; and the percentage of energy it represents, and therefore the faithfulness of reproduction, is dependent on the pulse duration a. The dependence of t_r upon T and a is small when the period T is long.

The maximum and minimum values of the deviation factor $F(N)$ can be determined from (13a) and (13b). The results are

$$\text{(a)} \quad \max F(N) = \frac{1}{1 - \dfrac{1}{\dfrac{T}{2a}(4s+1)-1}\left[\dfrac{1}{\sin\dfrac{\pi a}{T}} - 1\right]}, \tag{17a}$$

which occurs when

$$N = \frac{1}{2}\left[\frac{T}{2a}(4s+1) - 1\right], \quad s = 0, 1, 2, 3, \cdots. \tag{17b}$$

If $a \ll T$, (17a) can be simplified, and the biggest value $F(N)$ can have (for $S=0$) is approximately 2.75.

$$\text{(b)} \quad \text{Min}\, F(N) = \frac{1}{1 + \dfrac{1}{\dfrac{T}{2a}(4S+3)-1}\left[\dfrac{1}{\sin\dfrac{\pi a}{T}} + 1\right]}, \tag{18a}$$

which occurs when

$$N = \frac{1}{2}\left[\frac{T}{2a}(4s+3) - 1\right], \quad s = 0, 1, 2, 3, \cdots. \tag{18b}$$

If $a \ll T$, (18a) can be simplified, and the smallest value $F(N)$ can have (for $s=0$) is approximately 0.825.

It is then seen that the rise time of periodic rectangular pulses can vary from 0.825 to 2.75 times that calculated for a step signal or a single transient rectangular pulse depending on the particular value of N, or the bandwidth of the low-pass filter as compared with the pulse-repetition frequency. It can also be shown that the maximum and minimum values of the deviation factor $F(N)$ are nearly independent of the ratio a/T, provided the latter is small compared with 1.

V. Conclusion

The rise time and the percentage of energy content of the resulting pulse are two ways of describing the faithfulness of pulse reproduction. For a single transient rectangular pulse, the rise time is in the order of one half the reciprocal of the given bandwidth for idealized response. The duration of the pulse does not enter into calculation provided it is long enough to allow the full amplitude to build up. But, rise time alone does not carry too much significance unless the pulse duration is also specified. The relation between energy content and bandwidth furnishes more complete information.

The rise time of periodic rectangular pulses is shown to be dependent upon the number of harmonic components, the pulse-repetition period, and the pulse duration in a more complicated manner. It is not always equal to $T/2N$, but is nearly so when N is sufficiently large or when it assumes the value of (15a) or (15b).

Dominant Wave Transmission Characteristics of a Multimode Round Waveguide*

A. P. KING†, SENIOR MEMBER, IRE

Summary—**This paper presents some dominant wave transmission characteristics of multimode round waveguide lines in the 4-kmc range of frequencies. The use of such waveguide lines offers the advantages of lower transmission losses than obtainable with single-mode rectangular waveguide, and relative ease of making good joints.**

Possible mode conversion effects, including dominant mode elliptical polarization, have been examined and found to be innocuous. As a result, cross-polarized dominant waves can be used to provide two reasonably independent signalling channels at the same frequency in one pipe.

The experimental results obtained with a straight line 2.812-inch inside diameter and length of 150 feet are given.

Introduction

At microwave radio relay stations and in radar systems, it is current practice to employ rectangular waveguide transmission lines for the straight runs between the antenna and terminal equipment. As an alternate possibility, round waveguide lines may be substituted in these applications. By making the round waveguide slightly oversize, appreciably lower attenuations may be realized.

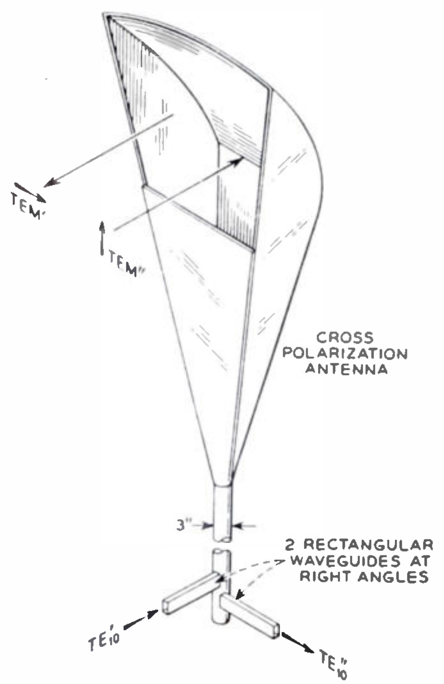


Fig. 1—Antenna and waveguide structure for transmitting two dominant mode signals whose electric fields are orthogonal in the round waveguide. The horizontally polarized signal mode is designated by single primes, the vertically polarized by double primes.

Round waveguide lines also possess the advantage that two cross-polarized dominant (TE_{11}) wave components may be transmitted to provide two moderately independent signalling channels. The combination of round line with an antenna,[1] which can transmit both horizontally and vertically polarized signals, suggests its possible use as a two-way repeater. Such a system is indicated in Fig. 1. The antenna is connected to the upper end of the line and the two mode selectors at the lower end provide terminal connections for two channels.

It is the purpose of this paper to discuss the transmission properties on a straight run of round waveguide. These measurements include attenuation, return loss, the effect of bowing the line, and various mode conversion effects. Measurements were conducted at the Holmdel Laboratory of the Bell Telephone Laboratories.

Description of Line

A round waveguide line comprising a straight horizontal run of 150 feet was set up for test purposes. It consisted of eight random lengths of phosphorous deoxidized copper pipe. The phosphorous content of the pipe was kept low in order to maintain a high dc conductivity, which was found to measure 0.95 that of the I.C.S. value. The dimensional tolerances of each piece of pipe were measured and it was found, in general, that the variation in the inside diameter was approximately one part per thousand. The pipe was not ordered with any special tolerances, but according to existing commercial waveguide practice. Soft-soldered sleeve joints were used.

Choice of Waveguide Size

Although the choice of diameter of a waveguide to be used in a particular frequency range is not critical, there are some broad limits in the selection of size. In commercially produced waveguide, the circularity and uniformity of dimensions of the pipe are limited by manufacturing tolerances; it is such irregularities which tend to set up spurious modes in a transmission line. Thus, for waveguide lines whose perfection is limited by such tolerances, the choice in size is a compromise between attenuation and mode conversion effects. A larger size waveguide will yield a lower attenuation, but it will also support a greater number of spurious modes.

For the 2.8125-inch ID chosen, the attenuation in the 3.7- to 4.2-kmc range is only half that of 2.290×1.145-inch ID(2.415- $\times1.260$-inch OD) rectangular copper

* Decimal classification: R118.2. Original manuscript received by the Institute, November 10, 1951; revised manuscript received, March 4, 1952.
† Bell Telephone Laboratories, Inc., Holmdel, N. J.

[1] H. T. Friis, "Microwave repeater research," *Bell Sys. Tech. Jour.*, vol. 27, pp. 202-204; April, 1948.

waveguide. This size of rectangular waveguide is in common use for this frequency range and is about as large as can be used if transmission is to be restricted to the dominant mode. This size of round waveguide supports only two spurious modes, the TM_{01} and the TE_{21}. Of these, the TE_{21} is limited to the higher end of the 3.7- to 4.2 kmc frequency band. A graph indicating the relation of guide wavelength as a function of frequency for the three modes is shown in Fig. 2 and the theoretical attenuation curves are plotted in Fig. 3.

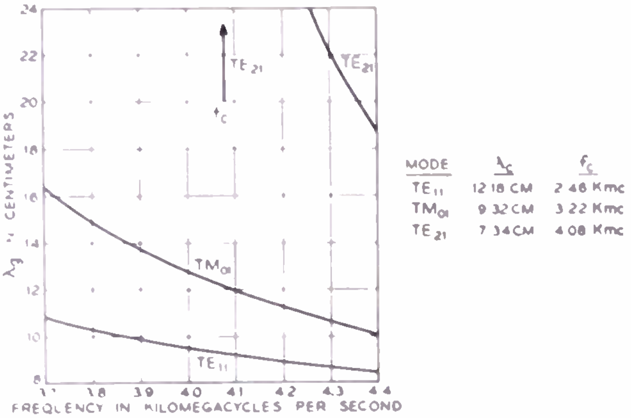


Fig. 2—Guide wavelength versus frequency of modes supported in 2.8125 inch inside diameter waveguide.

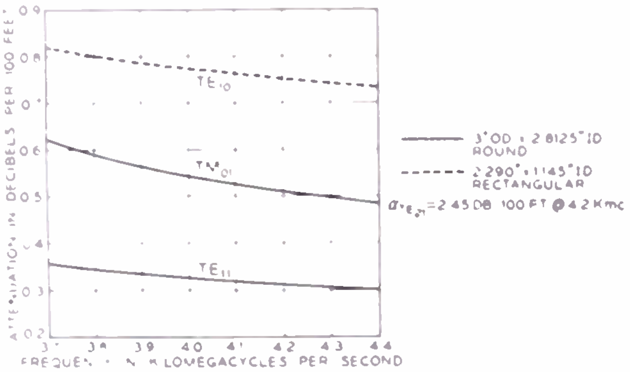


Fig. 3—Theoretical attenuation of modes in copper waveguide.

Measurements

Attenuation

The circuit employed for the measurement of attenuation is shown in Fig. 4. It consists of a hybrid junction in 1- × 2-inch rectangular waveguide, a screw-type tuning element which is indicated by *Y*, a $TE_{10} \rightleftharpoons TE_{11}$ wave transducer which comprises a tapered rectangular-to-circular transition of conventional form, and a transverse vane type shorting switch, *S*.

The function of the tuning element is to tune out the mismatch of the transducer and balance the bridge circuit. Under these conditions the transmitted component is balanced out so that only the signal which is reflected from the end of the line appears at the receiver. The switch provides a convenient means for shorting the line and reflecting the outgoing signal to the receiver to measure power level at the input of the line.

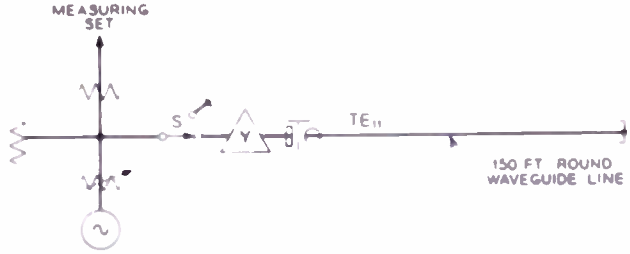


Fig. 4—Assemblage of waveguide components for the measurement of round-trip attenuation of 150 foot line.

With the bridge balanced and the far end of the line shorted, the attenuation was measured on a round trip basis. The level of transmitted power (switch-shorting line) relative to received power (switch open), has been plotted in db per 100 feet on a one way basis as shown in Fig. 5. Also plotted in this figure is the calculated

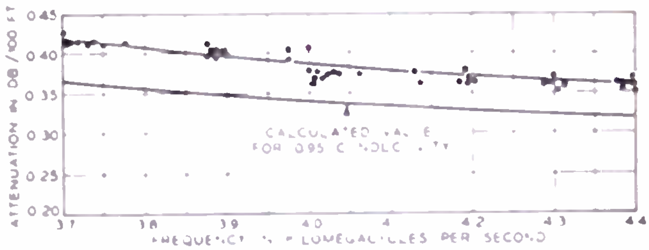


Fig. 5—Attenuation of line in db per 100 feet as a function of frequency. Upper curve shows the measured results and lower curve the calculated values.

value of attenuation. The ratio of measured to calculated value is 1.14 at 3.7 kmc and 1.13 at 4.4 kmc. The increase in transmission losses above the amount calculated is attributed, in substantial part, to surface roughness effects[2] and losses due to mode conversion.

Return Loss

The reflections which occur as a result of the irregularities present in the line were measured in terms of return loss by means of the arrangement shown in Fig. 6. The same bridge circuit arrangement was employed

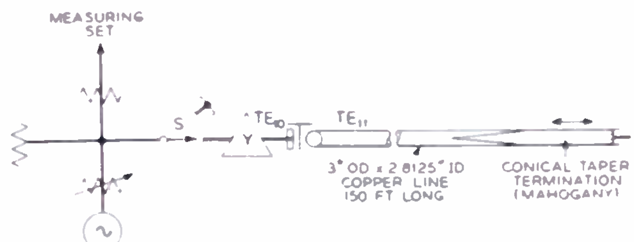


Fig. 6—Waveguide circuit employed for the measurement of the return loss of the 150-foot line.

[2] A. C. Beck and R. W. Dawson, "Conductivity measurements at microwave frequencies," Proc. I.R.E., vol. 38, pp. 1181–1190; October, 1950.

as that used for the attenuation measurements (Fig. 4). In this case, however, the far end of the line was terminated by means of a termination constructed of wood (mahogany) 29 inches in length and having a conical tapered section 21 inches long. Provision was made to permit movement of the termination along the line.

The return loss of the line was obtained in the conventional way using a movable termination. The measurement was made with the movable termination in the maximum and minimum return loss positions in order to isolate the return loss of the line from that of the termination. The results obtained in this way are plotted in Fig. 7 and indicate a range from -46- to -56-db return loss over the 3.7- to 4.2-kmc frequency band.

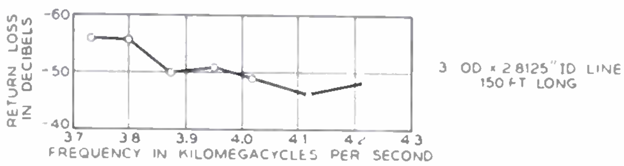


Fig. 7—Measured return loss of 150-foot line as a function of frequency.

Mode Conversion Effects

As has already been pointed out, spurious modes are possible in this oversize waveguide line. Some indication of the presence of the TM_{01} wave has been found, but its level is of the order of 50 db below that of the desired dominant wave. No reactive effects have been observed with the TM_{01} wave even when no mode suppressors were present in the line. No evidence has been found of the presence of the TE_{21} wave.

By far the predominant mode conversion effect is dominant mode elliptical polarization which is due to noncircular irregularities in cross section shape which occur as the result of the manufacturing process. The effect of these deformities is to produce dominant wave splitting, i.e., the generation of a second component which is orthogonal to the main component and travels along the waveguide at a slightly different velocity. The resultant of these two is an elliptically polarized wave.

The measurement of wave ellipticity was performed by means of the setup indicated in Fig. 8. In this arrangement, the transmitter was connected at one end of the line and the receiving and measuring equipment

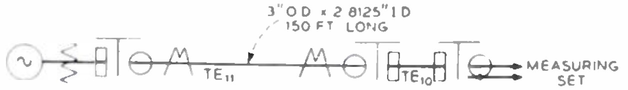


Fig. 8—Waveguide circuit employed in the measurement of dominant mode wave ellipticity.

at the other. The two transducers are the rectangular to circular transitions described in connection with Fig. 4. The undesired mode suppressors associated with the transducer at each end of the line consist of a flat resistive coated (200 ohms per square) card five inches long and mounted in the diametral plane of the round waveguide adjacent to each transducer. Each end of the card was V-notched to improve impedance matching. The function of these cards is to absorb the dominant wave component which is not transmitted by the associated transducer, thereby minimizing possible undesired mode resonance effects. In order to effect this result, the plane of the card is set perpendicular to the electric field. The transmission loss of each mode suppressor is about 40 db when the card is aligned with the electric field and less than 0.1 db when perpendicular to the electric field.

Dominant-mode wave ellipticity was measured for a given transmitted polarization by rotating the receiver (including transducer and mode suppressor) and noting the maximum received signal (main component). It was then rotated for the minimum received signal which corresponds to the level of the spurious cross component, and which occurs at an angle of 90° from the maximum signal position. The ratio of minimum to maximum signal is a measure of the elliptical polarization of the line. At the same frequency this measurement was repeated for a series of transmitted polarizations corresponding to 45° intervals which were obtained by rotating the transmitting transducer and mode suppressor together as a unit.

These data are shown in Fig. 9 where the ordinate values represent the energy level of the cross-component

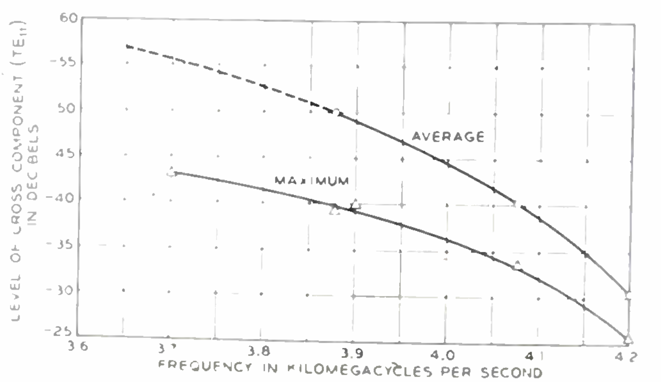


Fig. 9—Measured values of dominant wave mode ellipticity. The ordinate values indicate the wave ellipticity in terms of the cross component level relative to the main transmitted component as a function of frequency.

relative to that of the main component, and are plotted as a function of frequency. Of the two curves plotted, the upper curve represents the average cross-component measurement for the different transmitted polarizations and the lower curve shows the maximum value of cross component. The latter condition was found to occur in the vicinity of 45° relative to vertical polarization. The shape of both curves is dominated by the higher level of cross-component signal present at the 45° transmitted polarization; for the other polarizations, the levels show dissimilar variations with frequency. The difference between the two curves suggests that irregularities producing wave ellipticity are not entirely random.

The data of Fig. 9 applies to the transmission of a single dominant mode and shows the amount of cross-component wave power generated by this particular line as a function of frequency. Over most of the frequency band, the loss due to this mode conversion is less than 0.1 per cent of the transmitted power. While such a small loss is, in itself, quite negligible, this spurious mode can resonate if the ends of the line are not terminated. The function of the spurious mode suppressor is to absorb the cross component.

For the transmission of two cross-polarized dominant waves, the presence of the cross components generated by the line will introduce cross talk between the two signal channels. The effects of these cross components on system performance can, of course, be minimized by using a modulation system such as PCM, and this may be desirable when the two signals occupy identical frequency bands. As an alternative, the two signals may be separated in frequency. For this latter case, the degree of isolation provided by the cross-polarized waves reduces appreciably the frequency-filter requirements that would have to be imposed to use a single line and antenna for the same application.

Rotation of Polarization

The propagation of a dominant wave in round line which introduces an elliptically polarized wave also produces a rotation in polarization of the transmitted wave. Tests were performed to evaluate the angular shift in the 150-foot round waveguide line by measuring the angular difference between the transmitted wave polarization and that of the principal component arriving at the receiver. A small shift in the angle of polarization was noted; the magnitude of shift was, to the precision of measurement, in accordance with the resolution of the three vector fields, the transmitted wave being the resultant of the elliptically polarized wave components. On this basis, the angular change in polarization can be expressed in terms of the elliptical polarization measurement at the receiver by

$$\tan \theta = \frac{b}{a} \tag{1}$$

where θ is the angular shift in polarization, b the amplitude of the cross component, and a the amplitude of the principal component of the elliptically polarized wave.

Actual measurements showed that the angular shift in the transmission of dominant wave through a 150-foot line increased with frequency, the maximum shift being less than 1° for frequencies below 4.0 kmc, 2° to 3° for frequencies in the 4.2-kmc region. The rotation at other frequencies can be computed from the lower curve of Fig. 9 by means of (1).

No factor or mechanism which tends to shift the polarization of the wave, other than the one described above, has been observed.

Effect of Bending Waveguide Line

All measurements reported so far are for a straight run of waveguide. The question arises, quite naturally, as to the effect of an appreciable bend in the line. This effect has been examined by bowing the middle of the line for a range of vertical displacements S, as indicated in Fig. 10. In order to avoid any appreciable distortion of the cross section shape of the waveguide, the

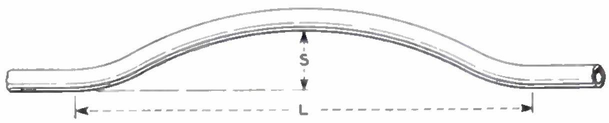


Fig. 10—Type of bend produced by vertical displacement of the line.

maximum vertical displacement was limited to 8 inches. The length L as a function of S in Fig. 10 was: $L=40$ feet at $S=2$ inches, $L=55$ feet at $S=4$ inches and $L=65$ feet at $S=8$ inches. The measurements were performed in the same manner as were the ellipticity tests of the line.

The results obtained in bowing the line are shown plotted in Fig. 11, where the ordinate indicates the level of the cross component in db below that of principal component and is plotted as a function of displacement

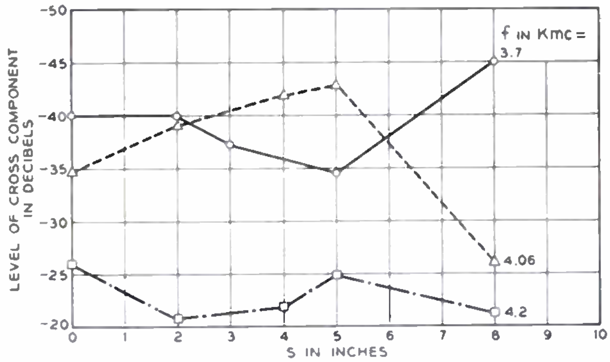


Fig. 11—Plot of measured dominant mode wave ellipticity generated by the line with a bend as a function of vertical displacement S.

S of the line. These are for the worst conditions, i.e., the maximum cross component appears when the transmitted wave polarization is in the vicinity of 45° relative to the plane of the bowed section.

For bends where the value of S is 4 inches or less, the maximum variation in the cross component is 5 db or less, but for values of S equal to 8 inches, variations of 10 db or more were observed. At some frequencies a given bend was found to reduce the value of cross component, and at other frequencies an increase was observed. This irregular behavior is believed to be due to the bend introducing differences in phasing between the elliptical wave components present in the line and those generated by the bend itself.

The results of the line bending tests indicate that the line need not be perfectly straight. In bends such that S deviates by as much as 4 inches, there is only a slight increase in mode conversion.

Network Synthesis by the Use of Potential Analogs*

R. E. SCOTT†, ASSOCIATE, IRE

Summary—**Two analog devices are described which are useful in solving the approximation problem of network synthesis. Both devices utilize a sheet of conducting Teledeltos paper to produce a two-dimensional potential field which simulates a required network function. The first device simulates the gain and phase of a network, and is primarily useful in designing networks for a prescribed steady-state response. The second device simulates the real part of the function, and is useful in designing networks for a prescribed transient response.**

I. Introduction

POTENTIAL ANALOGS have been discussed extensively in connection with the synthesis of linear, passive, lumped-parameter networks. Both the image-parameter method of Zobel[1,2] and the more general insertion-loss method of Darlington[3,4] have been expressed in terms of potential theory. Cauer[5] supplied a mathematical background for the methods in terms of Poisson's integrals, and later writers have extended his work.[6,7,8] The present paper describes two devices which are useful in obtaining approximate forms of required insertion-loss or gain functions. Standard methods exist[9,10,11] for developing these functions into physical networks. When the analog results are not accurate enough, they can be refined analytically with relative ease.[12] A few hours suffice to construct the analog devices and many days can be saved by their use in even a single problem.

* Decimal classification: R143. Original manuscript received by the Institute, May 15, 1951; revised manuscript received February 7, 1952. This work was supported, in part, by the Signal Corps, the Air Materiel Command, and the Office of Naval Research.

† Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge 39, Mass.

[1] J. F. Klinkhamer, "Empirical determination of wave-filter transfer functions with specified properties," Philips Research Reports, nos. 3 and 5; 1948.

[2] A. R. Boothroyd, "Design of electric wave filters with the aid of the electrolytic tank," *Proc. IEE* (London), vol. 98, pt. IV; 1951.

[3] W. W. Hansen and O. C. Lundstrom, "Experimental determination of impedance functions by the use of an electrolytic tank," Proc. I.R.E., vol. 33, pp. 528–533; August, 1945.

[4] W. H. Huggins, "A note on frequency transformations for use with the electrolytic tank," Proc. I.R.E., vol. 36, pp. 421–424; March, 1948.

[5] W. Cauer," Das Poissonsche Integral and seine Anwendungen auf die Theorie der linearen Wechselstromschaltungen," *Elektrische Nachrichten Technik 17*, pp. 17–30; 1940.

[6] R. E. Scott, "An analog device for solving the approximation problem of network synthesis," ScD. Thesis, M.I.T., Cambridge, Mass.; February, 1950.

[7] S. Darlington, "The potential analogue method of network synthesis," *Bell Sys. Tech. Jour.*, pp. 315–365; April, 1951.

[8] M. Cerrillo, "Generalized Relation between the Time and Frequency Domains," Research Laboratory of Electronics, M.I.T., Cambridge, Mass., to be published.

[9] O. Brune, "Synthesis of a finite two-terminal network whose driving-point impedance is a prescrib·d function of frequency," *Jour. Math. Phys.*, vol. 10, pp. 191–233; 1931.

[10] C. M. Gewertz, "Synthesis of finite 4 terminial network from its prescribed d·iving-point functions and transfer function," *Jour. Math. Phys.*, vol. 12, pp. 1–257; 1933.

[11] S. Darlington, "Synthesis of reactance four-poles," *Jour. Math. Phys.*, vol. 18, pp. 257–353; 1939.

II. The Gain-Phase Potential Analog

The gain function of a physically realizable network is expressible in the form[13]

$$H(s) = A\frac{(s - s_1)(s - s_3)\cdots}{(s - s_2)(s - s_4)\cdots}, \tag{1}$$

in which $s_1, s_2 \cdots$ are the zeros and $s_2, s_4 \cdots$ are the poles of the network. Alternatively,

$$\ln H(s) = G + j\phi = \ln A + \ln |s - s_1| + \cdots + j[\arg (s - s_1) + \cdots], \tag{2}$$

where G is the gain function and ϕ is the phase function of the network.

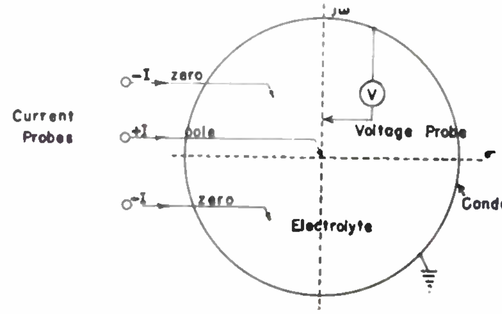


Fig. 1—The basic form of the potential analog.

The basic form of the potential analog is shown in Fig. 1. Positive and negative currents, introduced at the appropriate points, represent the poles and zeros, and the voltage measured with respect to a ground rim at infinity has the same form as the gain function of (2) (see Appendix 1). A practical machine for network synthesis based on this analog is shown in Fig. 2. To minimize errors caused by the finite size of the plane, a conformal mapping of the s plane on a logarithmic strip has been used.[4] For convenience, a mechanical commutator is provided, which scans a set of fixed probes along the center of the strip and displays the gain function versus logarithmic frequency on a cathode-ray tube. Poles and zeros are adjusted experimentally until the required form of the gain function is obtained.

The phase ϕ is obtained from the Cauchy-Riemann equation $\partial\phi/\partial\omega = \partial G/\partial\sigma$. The quantity $\partial G/\partial\sigma$ is meas-

[12] J. G. Linvill, "The selection of network functions to approximate prescribed frequency characteristics," Technical Report No. 145, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; March 14, 1950.

[13] H. W. Bode, "Network analysis and feedback amplifier design," C. Van Nostrand Co., Inc., New York, N. Y.; 1945.

ured by placing two rows of probes along the $j\omega$ axis. The voltage between two probes on opposite sides of the axis is proportional to $\partial G/\partial\sigma$ at that value of ω. An analog integration from zero to ω_1 yields phase at ω_1.[14]

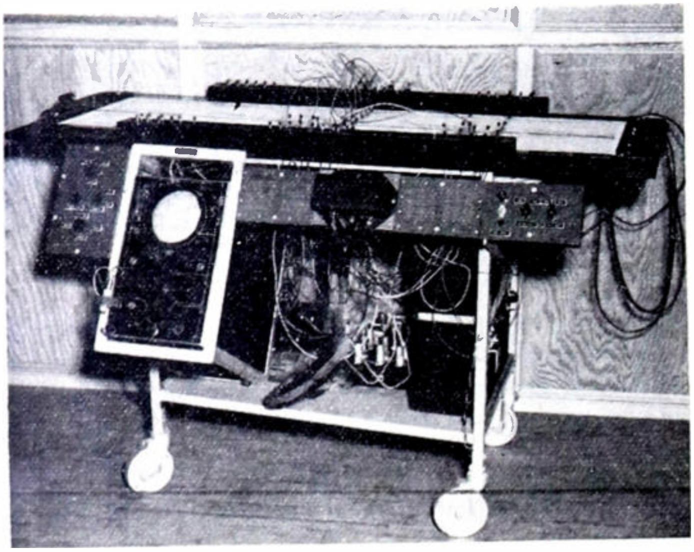


Fig. 2—A practical machine for network synthesis.

III. Experimental Results from the Gain-Phase Analog

As a consequence of the logarithmic transformation of the plane, the curves obtained from the machine are of the "ln-db" type.[15] The relation between the analog and the standard "ln-db" curves can be seen from Fig. 3. The current from a pole or zero flows toward the conductor at infinity and eventually achieves a uniform distribution. The region of uniform flow corresponds to a straight-line asymptote. In the neighborhood of the pole the exact behavior depends on the damping ratio (or the Q) of the pole.

The calibration of the machine is extremely simple.

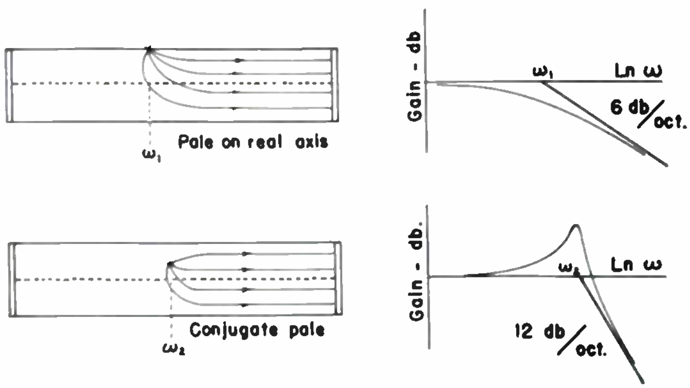


Fig. 3—The relation between the analog and the "ln-db" plots.

[14] For a driving-point impedance, the phase along the axis must remain between −90 and +90 degrees.

[15] G. S. Brown and D. Campbell, "Principles of Servomechanisms," John Wiley and Sons, New York, N. Y.; 1948.

The commutator scans one decade of the $j\omega$ axis, and thus fixes the frequency scale. The gain is calibrated in terms of the 6 db per octave slope produced by a pole at the origin. The phase is calibrated in terms of the 180-degree phase shift introduced by a complex pole.

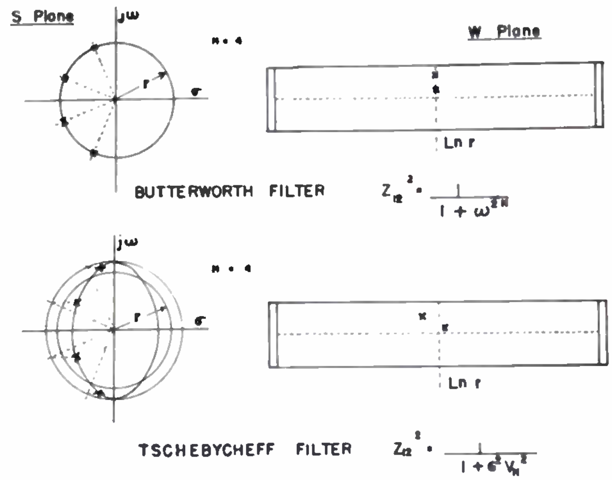


Fig. 4—Pole locations for Butterworth and Tschebyscheff filters.

The locations of the poles for Butterworth and Tschebyscheff filters are shown in Fig. 4 and experimental values of the gain functions are shown in Fig. 5 for poles on the circle and on three ellipses of increasing ellipticity. The measured accuracy of the machine for these filters is within 3 per cent for the gain and 5 per cent for the phase. The errors, of course, vary with the problem.

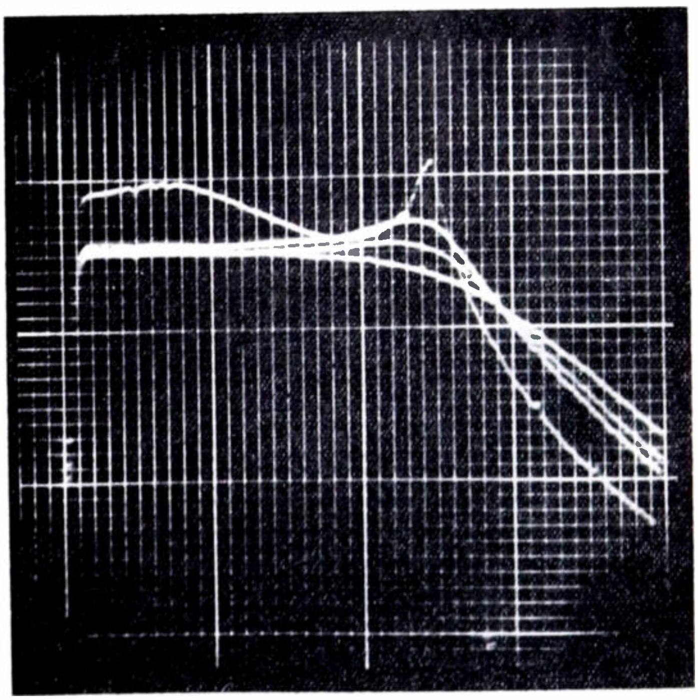


Fig. 5—Gain functions for Butterworth and Tschebyscheff filters as observed on the machine.

IV. The Real-Part Analog

A potential analog which displays the real part (or the imaginary part) of the network function $H(s)$ was first described by Guillemin.[16] A practical form is described here which is useful for network synthesis when the transient response is of primary interest.[17]

The impulse transient response $h(t)$ is the inverse Laplace transform of the frequency function $H(s)$. Providing there are no poles in the right half-plane or on the $j\omega$ axis, the real part of $H(s)$ along the $j\omega$ axis and $h(t)$ are cosine transforms.

$$h(t) = \frac{2}{\pi}\int_0^\infty H_r(\omega)\cos\omega t\, d\omega \tag{3}$$

$$H_r(\omega) = \int_0^\infty h(t)\cos\omega t\, dt. \tag{4}$$

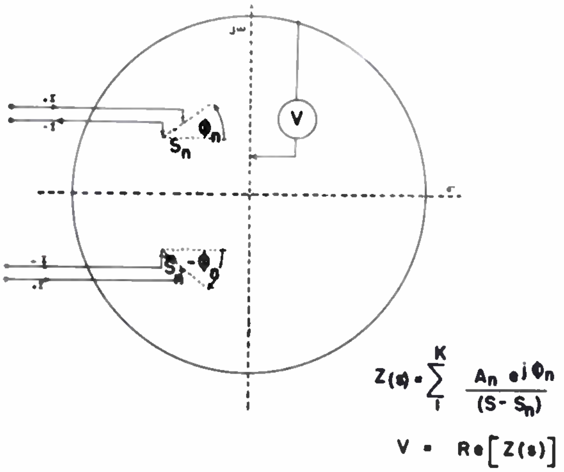


Fig. 6—The basic form of the dipole analog.

The basic form of the analog is shown in Fig. 6. It represents $H(s)$ in the partial fraction form,

$$H(s) = \sum_1^K \frac{A_n e^{j\phi_n}}{s - s_n}, \tag{5}$$

where s_n = poles

A_n = magnitude of residues

ϕ_n = angle of residues.

Dipole currents are introduced at the points s_n. The dipole moments are proportional to the magnitudes of the residues and the angles of orientation of the dipoles are equal to the angles of the residues. The resulting voltage along the $j\omega$ axis is proportional to the real part of the function $H(s)$ (see Appendix II). For synthesis the poles and residues are adjusted until the desired form of $H_r(\omega)$ is obtained. The practical device is shown in Fig. 7. The errors due to the finite size of the plane are re-

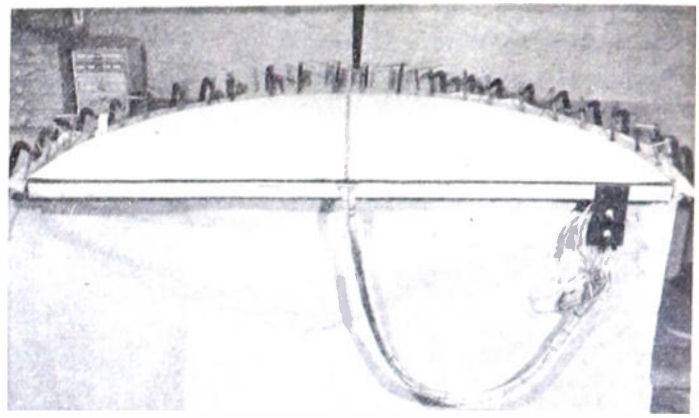


Fig. 7—A practical dipole analog.

moved by using a double-layer scheme.[18] Accuracies within approximately 5 per cent are attainable. The low voltage level associated with the dipole field is the limiting factor.

Appendix I

Derivation of the Formulas for the Gain-Phase Analog

A general form of the gain-phase formulas can be derived by considering a region A enclosed by a contour C, in an infinite sheet of conductivity ρ. Sources of current I are distributed through the region with a density m. Since the current originating in the region must equal the current crossing the boundary,

$$\int_C J\, n ds = \int_A m(I) da. \tag{6}$$

Hence by Gauss's theorem,

$$\text{div}\, J = m(I); \tag{7}$$

and hence

$$\text{div grad}\, \phi = -m\rho I. \tag{8}$$

This is Poisson's equation. A solution in polar co-ordinates is[19]

$$\phi(r, \theta) = \frac{-\rho I}{2\pi}\int_A m \ln r da - B, \tag{9}$$

where $\phi(r, \theta)$ is the potential at the point (r, θ) and B is a constant.

If the continuous distribution m is now made finite, the integral becomes a sum and the potential is

$$\phi(r, \theta) = \frac{-\rho I}{2\pi}\sum_m \ln |s - s_n| - B. \tag{10}$$

This equation has the same form as the gain function in (2).

[16] E. A. Guillemin, "The Mathematics of Circuit Analysis," John Wiley and Sons, New York, N. Y., pp. 351–360; 1949.

[17] D. D. Holmes, "The Dipole Potential Analog and Network Synthesis," Master's Thesis, Dept. of Electrical Engineering, M.I.T., Cambridge, Mass.; 1950.

[18] A. R. Boothroyd, E. C. Cherry, and R. Makar, "An electrolytic tank for the measurement of steady-state response, transient response, and allied properties of networks," *Proc. IEE* (London), vol. 96, 99, pp. 163–177; May, 1949.

[19] J. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y.; 1941.

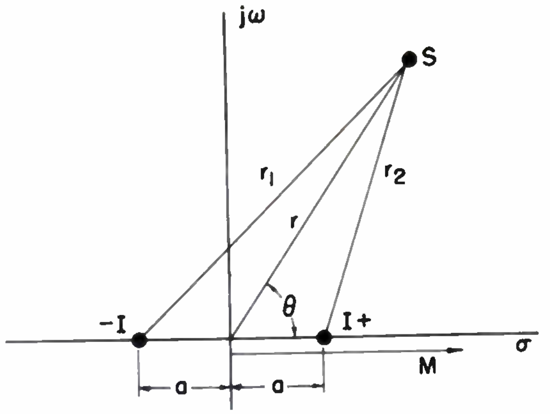


Fig. 8—The field from a dipole.

Appendix II

Derivation of the Formulas for the Real-Part Analog

The real-part-analog formulas are derived by considering the field of a dipole as shown in Fig. 8. Equation 10 gives the field at s as

$$\phi(s) = \frac{\rho I}{2\pi}\left[\ln r_1 - \ln r_2\right] = \frac{\rho I}{2\pi}\ln\left[1 + \frac{r_1 - r_2}{r_2}\right]. \qquad (13)$$

In the limit as a tends to zero $r_1 = r_2 = r$ and

$$\phi = \frac{\rho I}{2\pi}\,\frac{2a\cos\theta}{r} = \frac{\rho M}{2\pi}\,a\cos\theta, \qquad (14)$$

where M is the dipole moment.

If the dipole is moved to a point s_n and rotated through an angle ψ, the potential at any point s becomes

$$\phi = \frac{\rho M}{2\pi}\,\frac{\cos(\theta - \psi)}{|s - s_n|}. \qquad (15)$$

This equation has the same form as the real part of a term in the partial fraction expansion of (5). Higher-order poles result in higher-order multipoles.[16]

A Calculation Method for Sky-Wave Field Strength*

K. Rawer†

An analytic method for the calculation of sky-wave field strength had been introduced by the author as early as 1941. It distinguishes the different transmission modes like $1\times E$, $2\times E$, $1\times F$, $2\times F$, and so forth. The field strength corresponding to each mode depends on geometrical optics, absorption in lower layers, and on blanketing phenomena. Introducing the parameters of the different ionospheric regions, we can calculate these influences with sufficient accuracy. With predicted parameters the field strength can be calculated, and one immediately obtains the lowest usable frequency (LUF), if the properties of sender and receiver are known.[1] In Germany during the last war, radio propagation predictions were made in this way. Since that time the method has been improved in many respects and is now used by the French prediction service, SPIM.[2]

Geometrical optics of reflections between an ionospheric layer and the earth are essentially astigmatic. One has a caustic for rays leaving the earth in horizontal direction. Thus the field strength corresponding to flat rays is relatively high. This result, first obtained in the case of a thin layer,[1] is also true for a thick refracting layer.[3] Two other effects of caustic focussing, namely at the antipode point and at the skip-distance, should not be used for practical predictions and are eliminated.[4]

During daytime absorption occurs in the D- and E-layer. The absorption decrement decreases rapidly with frequency so that this variation is most effective for the LUF.[1] The geometrical factor for oblique incidence is obtained from a calculation supposing parabolic variation of electron density with height in the reflecting layers F and E, respectively.[4] The fundamental absorption value is taken from observations at vertical incidence.[5]

With the decrements of geometrical optics and of absorption, the field strength can be calculated for each mode. But a mode obtained by reflection on the upper layer will be "cut off," if the lower layer is sufficiently ionized, so that it reflects itself the rays in question. This blanketing effect just cuts away the modes, which should have a low attenuation for low frequencies. Thus it is extremely effective for the LUF and for the effective angle of incidence, too.[2]

Recently, refraction effects in both the E- and F-layers have been calculated in detail. Blanketing is now marked by a high divergency of the rays, while selective absorption in the E-layer turns out to be very effective. We thus arrive at a new method of calculation combining the three influences of geometrical optics (for two refracting regions now), of the nonselective absorption in the D- and the selective in the E-layer.[6]

For distances exceeding 3,000 kilometers the same principles are used. We have to sum up only the influences of the different contact points with the layers. On the other hand, combined modes of the form $mE+nF$ should also be considered here.[7]

During nighttime absorption is very low. Thus, at night the variation of external noise with frequency effectively determines the LUF.[8] But instead of using median results, it is preferable to introduce the same propagation influence for noise as is predicted for the actual MUF.[9]

In 1948 the CRPL replaced their former method for distances between 400 and 3,200 kilometers by a new one,[8] which follows ideas similar to our former one. Numerical values are somewhat lower because the CRPL supposes that the gain of geometrical optics is more than compensated for by losses at the earth. An experimental test on this point would be very interesting.

* Decimal classification: R112.6. Original manuscript received by the Institute, January 4, 1951, abstract received, March 31, 1952. The paper will be published in full in the *Wireless Eng.*

† Service de Prévision Ionosphérique Militaire (SPIM), France.

[1] K. Rawer, "Der Einfluss der Dampfung auf die Kurzwellenausbreitung," Deutsche Luftfahrtforschung FB No 1872, 1943

[2] K. Rawer, "Methode de Prévision du SPIM," rapport SPIM—R 7; 1948.

[3] K. Rawer, "Optique géometrique de l'ionosphère, *Revue Scientifique*, vol. 86, pp. 481–485, 585–600; 1948.

[4] K. Rawer, "L'influence de l'optique géometrique de l'ionosphère sur la propagation des ondes décamétriques," rapport SPIM—R 6; 1949.

[5] K. Rawer, "Les paramètres de l'absorption des ondes par la couche D et leur prévision," rapport SPIM—R 5; 1949.

[6] K. Bibl, K. Rawer and E. Theissen, "An improved method for the calculation of the fieldstrength of waves reflected by the ionosphere," *Nature*, vol. 169, pp. 147–148; 1952. See also report SPIM—R 11; 1951.

[7] E. Theissen, "Conditions de propagation des ondes courtes pour la transmission à grande distance par réflexions intermédiaires dans l'ionosphère," rapport SPIM—R 8; 1950.

[8] "Ionospheric Radio Propagation," Circular no. 462, Nat. Bur. Stand., Washington, D.C ; 1948.

[9] K. Rawer, "Ausbreitungsvorhersage fur Kurzwellen mit Hilfe von Ionospharenbeobachtungen," *Arch. Elekt. Übertragung*, vol. 5, pp. 154-167; 1951.

Interpretation of High-Frequency CW Field-Intensity Records with the Aid of Simultaneous Pulse Data*

R. SILBERSTEIN†, SENIOR MEMBER, IRE

Summary—Several causes of changes in field intensity of cw signals are outlined and their identification by the use of pulses discussed. Results of comparisons of cw and pulse records are given for a winter-type record and a summer-type record and characteristics due to several of the listed causes are identified.

Introduction

An ionospherically propagated skywave from a distant point usually consists of a group of several waves, each with its own changing phase and amplitude characteristics, coming down at different angles to the horizontal. These waves and their ground-reflected components combine at the antenna to give a resultant voltage which is amplified and detected.

In a record of the field intensity of a continuous-wave signal in which more than one mode is present at any one time, it is not easy to tell what combinations are present. Identification of arriving modes by determination of their angles of arrival is difficult when several modes are present simultaneously; the resultant arriving wave does not have a plane phase front, does not provide uniform illumination of the antenna aperture, and undergoes violent time fluctuations of all its characteristics.[1]

With pulsed signals the delay differences between the various observed modes can be measured and correspondence obtained between them and the delay differences calculated for combinations of possible modes. Some ambiguities can still occur, but may often be resolved if enough is known about ionospheric conditions over the path.

It is possible to enumerate several different causes of field-intensity changes appearing on radio-field-intensity records, among which are

(a) focusing effects as conditions approach or recede from skip,
(b) failure and recovery due to ionospheric irregularities,
(c) interference between modes,
(d) interference within a mode,
(e) changes in absorption in or shielding by a lower layer,[2]
(f) antenna-pattern effects.

A study of simultaneously observed pulses transmitted over the same path and on about the same frequency should make it possible to identify effects such as those of (a) and (b) with a fair degree of certainty. It should also be possible to obtain an idea of what modes of propagation are present and thus identify a given depth and rate of fading with the modes responsible for it (c). The effect of (d) will probably always be present because of ionospheric roughness, but will show as a fast fading rate usually superimposed upon slow-fading cycles due to other causes.

The effects of (e), not considered in this work, should be identifiable by a very slow change in the intensity of one mode of a group on A-scope pulse records.

One effect under (f) is that of changing polarization or angle of arrival resulting in a change of pickup as determined by receiving-antenna characteristics. Angle-of-departure changes also result in changes in received field intensity because of the directional patterns of the antennas. The effects of (f) are not always identifiable without auxiliary polarization- and angle-of-arrival measuring equipment, but should be easy to demonstrate in certain cases, such as near the limit of one-hop transmission, where angles of departure for the lowest order mode are very small. In the case to be considered the path was fairly short so that the lowest order mode was not severely attenuated.

Instrumentation

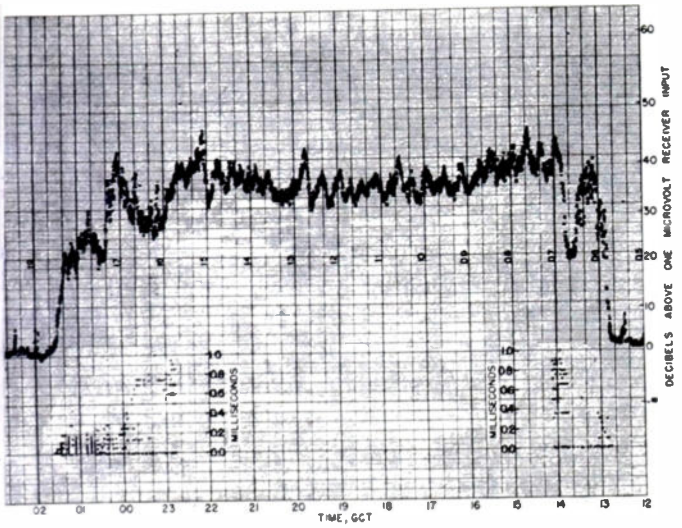
The path of which a study was made was that from Beltsville, Md. to the White Sands Proving Ground, N. M., a distance of about 2,700 km. Continuous recordings were made at White Sands of the field strength of WWV 15 mc, the over-all time constant of the receiving and recording system being 18 seconds. Occasionally, simultaneous high-power pulse transmissions were made on a frequency between 15.00 and 15.09 mc from Sterling, Va. and received at White Sands on a loran-type indicator, pulse groups being photographed every minute during the failure and recovery period. The cw power at WWV was 9 kw into a half-wave vertical antenna and the peak power at Sterling was 700 kw into a rhombic antenna, a 40-microsecond pulse being used.

Results

An examination was made of a number of field-strength records for the summer and late fall in conjunction with accompanying pulse records. With regard to the field-strength records, in general, there were large differences between individual days and between seasons. However, certain common characteristics were

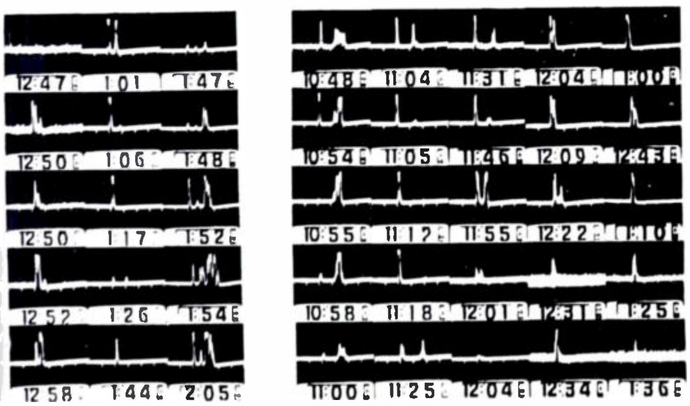
* Decimal classification: R271.4. Original manuscript received by the Institute, July 11, 1951; revised manuscript received, April 9, 1952. The material is adopted from the National Bureau of Standards Report No. 1085 of the same title, dated July 27, 1951.
† National Bureau of Standards, Washington, D. C.
[1] F. W. Schott, "On the response of a directive antenna to incoherent radiation," Proc. I.R.E., vol. 39, p. 677; June, 1951.
[2] E. V. Appleton, W. J. G. Beynon, and W. R. Piggott, "Anomalous effects in ionospheric absorption," *Nature*, vol. 161, p. 968; June 19, 1948.

ɔted and made more readily identifiable by the use of ılses. The simpler records were those for late fall beıuse of an absence of the F_1-layer and because the rate change of ionization at the beginning and end of skip as very fast.



ig. 1—Field Intensity of WWV, 15 mc, received at White Sands Proving Ground, N. M., with concurrent pulse delay times for an adjacent frequency arriving over almost the same path, December 1–2, 1950.

Fig. 1 shows a field-strength recording made on Deːember 1, 1950 which illustrates a typical late fall or winter record. The time is Greenwich Civil Time and



Path recovery 1247 to 1405 GCT, December 1

Path failure 2248 GCT, December 1 to 0136 GCT, December 2

Fig. 2—Typical pulse patterns received at White Sands, December 1–2, 1950.

reads from right to left. Below the periods of rise and fall are superposed scalings of relative delay times of the various modes as shown by the pulse positions on the loran indicator. Zero delay is taken as the time of arrival of the first pulse, presumably propagated by the one-hop F_2-mode; occasional solid lines indicate groups of many peaks, or strong peaks with indefinite separation. Fig. 2 is a group of A-scope photographs of the pulse groups, the time here being GCT on a 12-hour basis.

In Fig. 1 the field intensity of WWV is seen rising above the noise level at 1246. At 1247 a single dot on the record indicates a pulse seen in the noise. At 1250 two pulses are seen separated by about 0.1 millisecond. In the succeeding minutes there are usually two pulses separating in range as time increases. The first pulse is the one-hop f_2 low ray and the second pulse the one-hop f_2 high, or Pedersen, ray. The phenomenon of the low and high rays may be explained by the theory advanced by Appleton and Beynon[3] or by application of Smith's transmission curves to vertical-incidence ionospheric data.[4] A continuous photographic plot of echoes showing both low and high rays appears in a memorandum by Pierce.[5]

As the high ray increases in delay and dies out in intensity the cw signal increases in intensity, rising to about 20 db above the noise in the first 4 minutes and reaching a peak of about 35 db above the noise by 1310. This peak represents a focusing effect pointed out by Eckersley[6] and others.

After 1310 the field intensity drops about 20 db, with the high ray presumably delayed even more, and falls below the noise level. A minimum intensity of the cw field is reached at 1340 after a fall of about 15 db. Up to this time the cw signal is seen to fade at a rapid rate. It seems reasonable to say that these fades are partly caused by interference between waves propagated by different modes, such as the low and the high modes seen on the pulse record, (c), and partly because of the effect of interference within a single mode (d).[7]

At 1348 a pulse group is first seen at 0.65 millisecond spreading upward and downward as time progresses. This pulse group is most probably the start of two-hop propagation, the spread being due to the formation of low and high rays. Coincidentally, with its appearance the field-intensity curve rises to a maximum which it reaches within two or three minutes. The group of echoes starting and persisting at about 0.35 millisecond is possibly an M-type reflection.

The presence of waves coming via a two-hop mode along with those coming via one hop is identifiable directly on the field-intensity records by coarse fading cycles of duration of the order of 30 minutes.

These fading cycles are most probably caused by beats between the one- and two-hop modes due to

[3] E. V. Appleton and W. J. G. Beynon, "The application of ionospheric data to radio communication problems," *Proc. Phys. Soc.* (London), vol. 52, pt. I, p. 518; July, 1940; vol. 59, pt. II, p. 58; January, 1947.

[4] N. Smith, "The relation of radio sky-wave transmission to ionospheric measurements," PROC. I.R.E., vol. 27, pp. 332–347; May, 1939.

[5] J. A. Pierce, "The frequency dependence of ionospheric maximum speeds," Cruft Laboratory, Harvard University, Technical Memorandum No. 4, Contract N5ORI-76 Task Order No. 28, Office of Naval Research, May 5, 1949.

[6] T. L. Eckersley, "Studies in radio transmission," *Jour. IEE*, vol. 71, pp. 405–459; September, 1932.

[7] J. A. Ratcliffe, "Diffraction from the ionosphere and the fading of radio waves," *Nature*, vol. 162, p. 9; July, 1948.

changes in the relative path lengths for the two modes. The existence of swells in the F_2 region of the ionosphere of 5 to 30 minutes in length has been fairly well demonstrated by Ross[8] in lateral-deviation experiments with precision direction finders and by other experimenters in other ways.[9] Such swells, having greater effect in changing total path length for the relatively high-angle two-hop mode at its two reflecting regions than for the relatively low-angle one-hop mode at its one reflecting region, could well produce beats of the order of magnitude of 30 minutes. A characteristic beat-wave form has sharp spikes at the bottom and is rounded on top. The peaks appearing at the tops of the two-hop fading-pattern maxima in Fig. 1 are probably due to momentary maxima of the individual modes.

The middle of the day is characterized on the cw record by a slight drop of the average curve to a minimum at about noon over the path or around 1800 or 1900 GCT. No pulse patterns were available for the period.

At 2248 pulse delay times are recorded again with multiple 2-hop echo groups at delay times starting at 0.5 millisecond. By 2310 the last 2-hop pulse echo is seen at 0.75-millisecond delay. Sometimes a 2-hop focus effect is seen just before 2-hop failure, but does not appear on this record; ionospheric irregularities probably masked the effect. A build-up to one-hop focus now begins. A high ray is first seen at 2315 on the pulse plot. As the rays merge the focus field intensity rises, aided by the decrease in ionospheric absorption at the end of the day, the peak of focus being reached at about 0005. At 0031 the pulses observed were weakest, well merged, and had a scattered appearance. The field-intensity record at this point is at the end of a 15-db drop. A field-intensity rise beyond this point is followed by an apparent spreading of the pulse groups, indicating a partial recovery of propagation, such as would be caused by the appearance of another high-ionization area affording partial regular-layer propagation, perhaps because of motion of irregularities or because of pressure waves in the F_2-region. This field-intensity rise might also be caused by a drop of virtual height, resulting in a decrease of skip distance. Since the pulses appearing after this first failure were fairly clean but with an occasional appearance of a large number of modes, it is definite that the partial recovery is not a scatter mode. The peak of the partial recovery of WWV appears to be at 0045 although the signal was partially contaminated by the presence of fields from WWVH in Hawaii. Beginning at 0120 the signal makes a final rapid drop of 18 db to the noise level. The last pulse was seen at 0138 and WWV was last heard at 0144.

Fig. 3 is similar to Fig. 1 for a summer day, June 13–14, 1950. Pulse data were available for the failure period only.

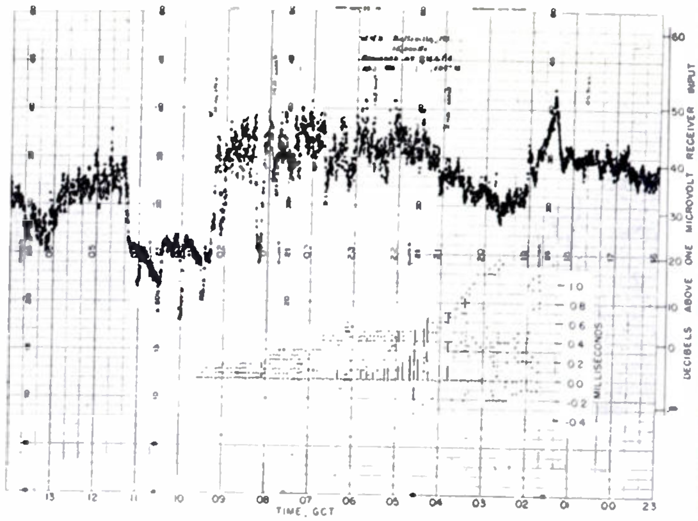


Fig. 3—Field intensity of WWV, 15 mc, received at White Sands Proving Ground, N.M., with concurrent pulse delay times for an adjacent frequency arriving over almost the same path, June 13–14, 1950.

The June 13–14 record is harder to interpret than that for December 1 because of the slower rates of change of ionization at path failure, general propagation characteristics approaching those for summertime ionospheric conditions. However, the pulse pattern shows the same general agreement with the cw pattern as was the case for the December record. The group of dots at a delay time of 0.9 to 1.1 milliseconds disappearing at 0200 appears to be the nose of the 3-hop curve. A sharp dip of 0812 agrees with a merging of the pulsed modes and probably represents a temporary path failure. An immediate recovery follows, corresponding to a spreading of the modes and a subsequent second failure corresponding to another decrease of delay time between pulse groups. The final disappearance of the pulses into the noise was at 0936, but the exact time of signal disappearance could not be determined because of contamination of the record by weak signals from WWVH in Hawaii, identifiable by interruptions on the hour and half hour.

Identification of the modes on the pulse patterns was accomplished in part by geometric calculations for the path length and reasonable ionospheric heights, and in part by considerations of the changes of the pulse patterns with time. In general, measured delay differences were smaller than expected, indicating lower effective ionospheric heights.

Acknowledgment

The author wishes to express his thanks to Mr. E. J. Wiewara whose careful operation of the recording gear and meticulous annotation of the log sheets were an invaluable aid to this study.

[8] W. Ross, "Lateral Deviation of Radio Waves Reflected at the Ionosphere," Department of Scientific and Industrial Research, Radio Research Special Report No. 19, London; 1949.

[9] "Winds and turbulence in the upper atmosphere," *Nature*, vol. 167, pp. 626–628; April 21, 1951.

An Optimization Theory for Time-Varying Linear Systems with Nonstationary Statistical Inputs*

RICHARD C. BOOTON, JR.†, ASSOCIATE, IRE

Summary—The mean-square optimization problem is stated for time-varying systems with nonstationary statistical input functions. Correlation functions are defined for nonstationary ensembles. The mean-square error is calculated in terms of these correlation functions. The integral equation defining the optimum system is determined by minimization of the mean-square error.

Introduction

As NETWORK THEORY developed, the desirability of extending the now well-known techniques of analyzing systems on the basis of sinusoidal and transient input functions became evident. In answer to this need, a statistical theory leading to the optimum design of constant-coefficient linear systems was created, primarily by Wiener.[1] During recent years, the increasing importance of time-varying systems accentuates the desirability of extending statistical analysis to include time-varying systems. To complete the generalization, the analysis should include nonstationary input functions as well.

Although most of the system-analysis techniques were developed in electrical theory, no mathematical reason exists to distinguish between linear systems with different physical natures. In this paper, a system is defined by a mathematical transformation between two functions. One of the functions is referred to as the "input function," and the result of the transformation is called the "response function," but the system need not exist physically in the form of a "box." The functions can be distances or angles, as in the case of aircraft-control problems, or mechanical (or hydraulic) forces, as in servo problems. In any case, a linear-system problem exists if one variable in a physical situation is derived from another variable by means of a linear transformation.

In Wiener's theory the assumption is made that the input to a system consists of a random signal plus a random noise, each of which is stationary in the sense that its statistical properties do not vary with time. Response of the actual system is compared with the result of a desired, time-invariant, linear operation upon the signal component of the input, and the difference is called the "error." The mean-square value of this error is used as the criterion for optimization. The fundamental mathematical considerations involved in the statistical approach are first, that the input is not one function but one of an ensemble of functions, and second, that the measure of error used to define the optimum system is not the error resulting from one input function but the mean-square error with respect to the ensemble of input functions. Even though the importance of the ensemble concept has been recognized,[2] Wiener's treatment is based upon averages, with respect to time, for a single input function. Unfortunately, the impression seems to have arisen that for stationary ensembles "time averages are equivalent to ensemble averages," meaning that a time average for any function of the ensemble is both independent of the particular function used and also equal to the corresponding ensemble average. This statement is not true, however, for the general stationary ensemble, but only for the special case known as an "ergodic ensemble," for which any function is statistically equivalent to any other function. As Wiener stated his results,[3] they apply only to the ergodic case. Nevertheless, for the general stationary input, the ensemble approach, as presented here, can be used to derive for the optimum system an equation essentially the same as Wiener's, except that all the expressions must be defined by ensemble averages.

This paper generalizes the optimization theory to apply to nonstationary input functions and time-varying desired operations. The input ensemble is described by a set of three correlation functions. The desired response is the result of a time-varying linear operation upon the signal component of the input. For each input function, the error is the difference between this desired response and the actual system response. The optimization criterion is taken as the mean-square value (ensemble average) of the error. Because the input to the system contains noise, a filtering problem exists, even though the desired operation is merely the reproduction of the signal component. If the result of the desired operation depends upon future values of the signal, then the system must be both a filter and a predictor. Other problems, such as the optimum evaluation of derivatives, are specified by suitable choices of the desired operation.

To avoid possible misunderstanding, the range of applicability of the results presented here should be clearly understood. Physical reasoning makes evident, and the equations developed later in this paper prove mathematically, that the characteristics of the optimum sys-

* Decimal classification: 510. Original manuscript received by the Institute, August 15, 1951; revised manuscript received, April 3, 1952.

† Dynamic Analysis and Control Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.

[1] N. Wiener, "Extrapolation, Interpolation and Smoothing of Stationary Time Series, with Engineering Applications," Technology Press, Cambridge, Mass.; 1949.

[2] H. W. Bode and C. E. Shannon, "A simplified derivation of linear least square smoothing and prediction theory," Proc. I.R.E., vol. 38, pp. 417–425; April, 1950.

[3] N. Wiener, *op. cit.*, p. 57.

tem depend upon the statistics of the input during the entire time of interest. This fact is of basic importance in prediction problems. If the future behavior of the statistics of the ensemble is not definitely known, then optimum prediction, in the sense of this paper, is impossible. Under these conditions, mathematical treatment of the prediction problem, if at all possible, is certainly of an order of difficulty greater than in the problem considered in this paper. One of the nonstationary problems to which the results of this paper have already[4] been applied relates to an aircraft-control system in which the ensemble statistics are completely known, although the individual functions are random. Here, there is no ambiguity concerning the determination of the optimum predictor. At the other extreme are problems such as the prediction of the value of a variable associated with the stock market, where separation of the individual variation from an ensemble variation is perhaps indeterminant (or meaningless). Between these extremes problems exist in which the statistical variation can be estimated and approximately optimum predictors determined.

Superposition Integrals

A general expression for the response of any linear system is needed for the construction of an optimization theory. One of the most useful of these expressions is the superposition integral,[5] which, for the constant-coefficient linear system, assumes the well-known form

$$x_R(t) = \int_0^\infty f(\tau)x_I(t-\tau)d\tau, \tag{1}$$

where x_R and x_I denote the response of the system and the input to the system, respectively. The function $f(\tau)$ can be interpreted as the response, after a time τ, to a unit-impulse input function.

Although the superposition integral for time-varying systems is not so well known as the corresponding expression for constant-coefficient systems, the fact that the system response x_R can be expressed in terms of the input x_I as

$$x_R(t) = \int_0^\infty h(\tau, t)x_I(t-\tau)d\tau \tag{2}$$

can be established in essentially the same manner as for the constant-coefficient equation. The principal difference is that the time-varying system is characterized by a function of two variables $h(\tau, t)$, known as the "impulse-response function," whereas for the constant-coefficient system, the impulse response is independent of t and is a function of the one variable τ. The function $h(\tau, t)$ can be shown to be the response, evaluated at a time t, to a unit-impulse input applied at a time $t-\tau$.

Only linear operations are considered in this theory, and thus the desired operation upon the signal component of the input can be expressed in the superposition-integral form

$$x_D(t) = \int_{-\infty}^{\infty} g(\tau, t)x_S(t-\tau)d\tau, \tag{3}$$

where x_D and x_S denote the desired result and the signal component, respectively. The function $g(\tau, t)$ can be interpreted as an impulse response, in a manner similar to the interpretation of $h(\tau, t)$, except that $g(\tau, t)$ is not necessarily the impulse response of a physical system. Because operations such as prediction, which are operations upon future values of x_S, should be included, the lower limit of the integral in (3) must be minus infinity. In contrast, the response of a physical system, as shown by (2), can depend only upon past values of the input x_I, and thus the lower limit of (2) is always zero.

Correlation Functions

For the purpose of mean-square calculations, the input ensemble is completely specified[6] by three correlation functions. These correlation functions are defined by ensemble averages of products of the input and the signal component of the input. To be precise, if x_I and x_S denote the input function and its signal component, respectively, then correlation functions are defined as

$$\left.\begin{aligned} \gamma_{II}(t_1, t_2) &= \overline{x_I(t_1)x_I(t_2)} \\ \gamma_{IS}(t_1, t_2) &= \overline{x_I(t_1)x_S(t_2)} \\ \gamma_{SS}(t_1, t_2) &= \overline{x_S(t_1)x_S(t_2)} \end{aligned}\right\}, \tag{4}$$

where the bar denotes an averaging operation with respect to the ensemble of input functions.

An alternate set of correlation functions is sometimes used to describe the input ensemble. This set of correlation functions is defined by ensemble averages of products of the signal and noise components of the input. With x_N denoting the noise component of the input, these correlation functions are given by

$$\left.\begin{aligned} \gamma_{SS}(t_1, t_2) &= \overline{x_S(t_1)x_S(t_2)} \\ \gamma_{SN}(t_1, t_2) &= \overline{x_S(t_1)x_N(t_2)} \\ \gamma_{NN}(t_1, t_2) &= \overline{x_N(t_1)x_N(t_2)} \end{aligned}\right\}. \tag{5}$$

The first set of correlation functions can be calculated easily from the second set. Because

$$x_I(t) = x_S(t) + x_N(t), \tag{6}$$

correlation function γ_{II} is obtained by first evaluating

$$\begin{aligned} x_I(t_1)x_I(t_2) &= [x_S(t_1) + x_N(t_1)][x_S(t_2) + x_N(t_2)] \quad (7) \\ &= x_S(t_1)x_S(t_2) + x_S(t_1)x_N(t_2) \\ &\quad + x_S(t_2)x_N(t_1) + x_N(t_1)x_N(t_2). \end{aligned}$$

[4] Unpublished work at the Dynamic Analysis and Control Laboratory, M.I.T., Cambridge, Mass.

[5] S. Goldman, "Transformation Calculus and Electrical Transients," Prentice-Hall, Inc., New York, N. Y., pp. 112–118; 1949.

[6] This statement should not be taken as an implication that the ensemble of input functions is completely described by the correlation functions defined here. These correlation functions suffice only for mean-square calculations.

An average (with respect to the ensemble) of each side of (7) yields

$$\gamma_{II}(t_1, t_2) = \gamma_{SS}(t_1, t_2) + \gamma_{SN}(t_1, t_2) + \gamma_{SN}(t_2, t_1) + \gamma_{NN}(t_1, t_2). \tag{8}$$

In a similar manner γ_{IS} is evaluated by first multiplying each side of (6) by x_S to yield

$$x_I(t_1)x_S(t_2) = [x_S(t_1) + x_N(t_1)]x_S(t_2) = x_S(t_1)x_S(t_2) + x_S(t_2)x_N(t_1). \tag{9}$$

An average of each side of (9) results in

$$\gamma_{IS}(t_1, t_2) = \gamma_{SS}(t_1, t_2) + \gamma_{SN}(t_2, t_1). \tag{10}$$

Either of the preceding sets of correlation functions, (4) or (5), can be used to describe the input ensemble.

The input ensemble has been described by the correlation functions γ_{II}, γ_{IS}, and γ_{SS}. The statement of the optimization problem is completed by the function $g(\tau, t)$ that defines the desired operation, as shown by (3). In the calculation of the mean-square error that is presented later in this paper, the desirability of defining still another set of correlation functions is indicated. These correlation functions are γ_{ID}, the cross-correlation of the input and the desired response, and γ_{DD}, the autocorrelation of the desired response. They are defined in terms of ensemble averages as

$$\gamma_{ID}(t_1, t_2) = \overline{x_I(t_1)x_D(t_2)} \tag{11}$$

and

$$\gamma_{DD}(t_1, t_2) = \overline{x_D(t_1)x_D(t_2)}. \tag{12}$$

Expressions for these two correlation functions in terms of the given correlation functions and the desired operation are easily derived. Substitution of t_2 for t in (3) and multiplication of each side of the resulting equation by $x_I(t_1)$ yield

$$x_I(t_1)x_D(t_2) = \int_{-\infty}^{\infty} g(\tau, t_2)x_I(t_1)x_S(t_2 - \tau)d\tau. \tag{13}$$

If each side of this equation is averaged,[7] then the correlation function γ_{ID} is given by

$$\gamma_{ID}(t_1, t_2) = \int_{-\infty}^{\infty} g(\tau, t_2)\gamma_{IS}(t_1, t_2 - \tau)d\tau. \tag{14}$$

Similarly, the correlation function γ_{DD} is obtained as

$$\gamma_{DD}(t_1, t_2) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} g(\tau_1, t_1)g(\tau_2, t_2)\gamma_{SS}(t_1 - \tau_1, t_2 - \tau_2)d\tau_1\, d\tau_2. \tag{15}$$

Minimization of Error

The error in system performance has been defined as the difference between the desired response and the actual system response, that is,

$$e(t) = x_D(t) - x_R(t). \tag{16}$$

Use of (2) gives the error as

$$e(t) = x_D(t) - \int_0^{\infty} h(\tau, t)x_I(t - \tau)d\tau. \tag{17}$$

The square of the error then is given by

$$[e(t)]^2 = x_D(t)x_D(t) - 2\int_0^{\infty} h(\tau, t)x_I(t - \tau)x_D(t)d\tau + \int_0^{\infty}\int_0^{\infty} h(\tau_1, t)h(\tau_2, t)x_I(t - \tau_1)x_I(t - \tau_2)d\tau_1 d\tau_2. \tag{18}$$

The ensemble-averaged mean-square error is obtained by averaging both sides of (18) with respect to the ensemble of input functions. In terms of the correlation functions previously defined, this mean-square error can be written as

$$M(t) = \gamma_{DD}(t, t) - 2\int_0^{\infty} h(\tau, t)\gamma_{ID}(t - \tau, t)d\tau + \int_0^{\infty}\int_0^{\infty} h(\tau_1, t)h(\tau_2, t)\gamma_{II}(t - \tau_1, t - \tau_2)d\tau_1 d\tau_2, \tag{19}$$

where the assumption is made that the averaging operation commutes with the integration. The mean-square error given by (19) is the error resulting from use of the system with an impulse response of $h(\tau, t)$. In order that this error be the minimum error (and thus $h(\tau, t)$ the impulse response of the optimum system), replacement of $h(\tau, t)$ by $h(\tau, t)+f(\tau, t)$, where $f(\tau, t)$ is any impulse response, must result in a larger value for the mean-square error. Replacement of $h(\tau, t)$ by $h(\tau, t)+f(\tau, t)$ in (19) results, after some simplification, in

$$\left.\begin{aligned} N(t) &= \gamma_{DD}(t, t) - 2\int_0^{\infty} h(\tau, t)\gamma_{ID}(t - \tau, t)d\tau \\ &+ \int_0^{\infty}\int_0^{\infty} h(\tau_1, t)h(\tau_2, t)\gamma_{II}(t - \tau_1, t - \tau_2)d\tau_1 d\tau_2 \\ &-2\int_0^{\infty} f(\tau, t)\gamma_{ID}(t - \tau, t)d\tau \\ &+2\int_0^{\infty}\int_0^{\infty} f(\tau_1, t)h(\tau_2, t)\gamma_{II}(t - \tau_1, t - \tau_2)d\tau_1 d\tau_2 \\ &+ \int_0^{\infty}\int_0^{\infty} f(\tau_1, t)f(\tau_2, t)\gamma_{II}(t - \tau_1, t - \tau_2)d\tau_1 d\tau_2 \end{aligned}\right\}, \tag{20}$$

where $N(t)$ denotes the mean-square error corresponding to impulse response $h(\tau, t)+f(\tau, t)$. Reference to (19) shows that the sum of the first three terms in (20) is the mean-square error $M(t)$ corresponding to the impulse response $h(\tau, t)$. Substitution of (19) into (20) and use of the fact that last term of (20) is nonnegative[8] yield

[7] The statistical problem is meaningful only if the averaging process commutes with all the operators considered.

[8] The last term of (20) is the ensemble average of $\left[\int_0^{\infty} f(\tau_1, t)x_I(t - \tau_1)d\tau_1\right]^2$, and is thus nonnegative.

$$N(t) \geqq M(t) - 2\int_0^\infty f(\tau, t)\gamma_{ID}(t-\tau, t)d\tau$$

$$+ 2\int_0^\infty \int_0^\infty f(\tau_1, t)h(\tau_2, t)\gamma_{II}(t-\tau_1, t-\tau_2)d\tau_1 d\tau_2. \quad (21)$$

Therefore, the minimizing condition $N(t) \geqq M(t)$ is satisfied if

$$0 = -2\int_0^\infty f(\tau, t)\gamma_{ID}(t-\tau, t)d\tau$$

$$+ 2\int_0^\infty \int_0^\infty f(\tau_1, t)h(\tau_2, t)\gamma_{II}(t-\tau_1, t-\tau_2)d\tau_1 d\tau_2. \quad (22)$$

Because (22) can be written as

$$0 = 2\int_0^\infty f(\tau_1, t)\left[-\gamma_{ID}(t-\tau_1, t) + \int_0^\infty h(\tau_2, t)\gamma_{II}(t-\tau_1, t-\tau_2)d\tau_2\right]d\tau_1, \quad (23)$$

the minimizing condition is satisfied for all impulse response functions $f(\tau, t)$ if

$$\gamma_{ID}(t-\tau_1, t) = \int_0^\infty h(\tau_2, t)\gamma_{II}(t-\tau_1, t-\tau_2)d\tau_2 \quad (24)$$

for $0<\tau_1<\infty$.

This equation defines the impulse response of the optimum system. The development here proves the sufficiency of (24) for a minimum. The usual calculus-of-variations methods easily establish its necessity. If (24) has more than one solution, several optimum systems exist, each yielding the same mean-square error. Analysis of this situation is relatively complicated, but the uniqueness of the optimum system can be shown to be equivalent to the completeness of the ensemble of input functions x_I. Fortunately, most physical problems yield unique solutions, and physical reasoning usually makes clear when the optimization procedure merely places a constraint on the system instead of uniquely determining it.

Determination of the Optimum System

At this point, a solution of (24) yielding an explicit expression for $h(\tau, t)$ in terms of γ_{ID} and γ_{II} would complete the determination of the optimum system. Because the general form of (24) is the general form of the integral equation of the first kind, no such explicit solution can be written. Except for the special cases that can be solved analytically, recourse must be made to numerical methods (including machine computation) if actual solutions are desired. Since discussion of these methods is outside the scope of this paper, (24) essentially completes the analysis of the optimization problem presented here.

One special case of the general problem should be mentioned. If the input is stationary, an explicit expression for $h(\tau, t)$ can be obtained in a form similar to the solution of the completely stationary case. The result is a generalization of the completely stationary problem since the desired operation can still be time-varying. If the input x_I is stationary, then a function ϕ exists such that

$$\gamma_{II}(t_1, t_2) = \phi(t_2 - t_1), \quad (25)$$

and (24) can be written as

$$\gamma_{ID}(t-\tau_1, t) = \int_0^\infty h(\tau_2, t)\phi(\tau_1 - \tau_2)d\tau_2. \quad (26)$$

Since (26) can be solved by a method that follows closely the technique used by Levinson[9] in discussing the completely stationary problem, the final solution will be presented without details.

The spectral-factorization theorem proved by Wiener[10] implies that ϕ can be factored in the form

$$\phi(t) = \int_{-\infty}^\infty \phi_+(t-\tau)\phi_-(\tau)d\tau, \quad (27)$$

where

$$\left.\begin{aligned}\phi_+(t) &= 0 \quad \text{if} \quad t<0\\ \phi_-(t) &= 0 \quad \text{if} \quad t>0\end{aligned}\right\}. \quad (28)$$

The correlation function γ_{ID} can be factored as

$$\gamma_{ID}(t_1, t_2) = \int_{-\infty}^\infty \psi(t_2 - t_1 - \tau, t_2)\phi_-(\tau)d\tau. \quad (29)$$

Then the impulse response of the optimum system, that is, the solution of (26), can be shown to be

$$h(\tau, t) = \frac{1}{2\pi i}\int_{-i\infty}^{i\infty} e^{s\tau}\frac{\Psi(s, t)}{\Phi_+(s)}ds, \quad (30)$$

where

$$\left.\begin{aligned}\Psi(s, t) &= \int_0^\infty e^{-s\tau}\psi(\tau, t)d\tau\\ \Phi_+(s) &= \int_0^\infty e^{-s\tau}\phi_+(\tau)d\tau\end{aligned}\right\}. \quad (31)$$

Conclusions

The statistical optimization theory developed by Wiener is extended to include the general case involving a linear time-varying system with a nonstationary input. Statistical characteristics of nonstationary ensembles are defined by ensemble-averaged correlation functions. The mean-square error in the performance of a time-varying linear operation upon the signal component of the input is calculated, and the integral equation defining the optimum system is derived. This integral equation is expressed in terms of functions of time, that is, the impulse response of the system and the cor-

[9] N. Levinson, "A heuristic exposition of Wiener's mathematical theory of prediction and filtering," *Jour. Math. Phys.*, vol. XXVI, pp. 110–119; July, 1947.
[10] N. Wiener, *op. cit.*, p. 53.

relation functions that define both the input functions and the desired operation.

Note should be taken of the fact that, although the optimum system is completely specified by the integral equation derived here, this equation cannot be solved directly in the general case. If actual results are desired, numerical or machine methods of computation must be employed. The development of approximation methods would increase the applicability of the optimization technique. Because little practical application of this optimization theory has been made, no intuitive understanding of the relative importance of the factors involved exists. In some cases a rough solution yields acceptable results, whereas in others, careful determination of the optimum system is essential. Some basis on which to estimate the accuracy required in a particular application would be useful. In work at the Dynamic Analysis and Control Laboratory, the nonstationary optimization theory has been applied to an aircraft-control-system design problem and to the analysis of suppressed-carrier modulation systems.

ACKNOWLEDGMENT

This research, which is a portion of a doctoral thesis,[11] was sponsored by the Bureau of Ordnance of the Navy Department, Bureau of Ordnance Contract NOrd 9661, and was conducted at the Dynamic Analysis and Control Laboratory of the Massachusetts Institute of Technology. The author gratefully acknowledges the support given by Dr. J. A. Hrones, Director of the D.A.C.L., and the benefit of discussions with Dr. W. W. Seifert and M. H. Goldstein, Jr., also of the D.A.C.L.

[11] R. C. Booton, Jr., "Nonstationary Theory Associated with Time-Varying Linear Systems," Thesis (Sc.D.) Dept. of Electrical Engineering, M.I.T., Cambridge, N. Y.; June, 1952.

Self- and Mutual Impedances of Parallel Identical Antennas*

RONOLD KING†, SENIOR MEMBER, IRE

Summary—**A method is described for obtaining approximate second-order self- and mutual impedances of parallel identical antennas. Extensive groups of curves of the newly determined impedances are provided. Since they lead to the second-order King-Middleton impedances for isolated antennas at infinite separation, and these are known to be in excellent agreement with experimental measurements, they are necessarily more accurate than the first-order or zeroth-order values thus far available, which do not lead to experimentally verifiable impedances at infinite separation except for extremely thin antennas.**

INTRODUCTION

ANALYSIS of two parallel identical antennas was carried out by King and Harrison[1] under the general conditions that the antennas may be center-driven by arbitrary voltages or one antenna driven and the other center-loaded with an arbitrary impedance. Essentially the same problem was solved later, but independently, by Bouwkamp.[2] The formulation of King and Harrison was re-examined by Tai,[3] who introduced an improved kernel and the expansion parameter of King and Middleton[4] into the series solution, which is based on the original work of Hallén[5] for the single isolated antenna. Tai evaluated the distributions of current and the self- and mutual impedances for antenna half-lengths corresponding to $\beta_0 h = 2\pi h/\lambda_0 = \pi/2$ and π for three ratios of half-length h to the radius a as given by $\Omega = 2 ln(2h/a) = 10$, 15, and 20. Tai's computations are based on a so-called first-order theory, since the leading or zeroth-order terms in the series and the first-order terms which have $(1/\Psi)$ as coefficient are retained while higher powers of $(1/\Psi)$ are neglected.

The analysis of Tai, just as the earlier work of King and Harrison, is based on the simple two-element form of the method of symmetrical components. The general case is constructed by superimposing the solutions of the symmetrical problem with equal currents in phase, and those of the antisymmetrical problem with equal currents in opposite phase. The symmetrical and anti-symmetrical impedances are given by

$$Z_0{}^s = \frac{-j\zeta_0\Psi_s}{2\pi}\left[\frac{\cos\beta_0 h + A_{1s}/\Psi_s + A_{2s}/\Psi_s{}^2 + \cdots}{\sin\beta_0 h + B_{1s}/\Psi_s + B_{2s}/\Psi_s{}^2 + \cdots}\right] \quad (1)$$

$$Z_0 = \frac{-j\zeta_0\Psi_a}{2\pi}\left[\frac{\cos\beta_0 h + A_{1a}/\Psi_a + A_{2a}/\Psi_a{}^2 + \cdots}{\sin\beta_0 h + B_{1a}/\Psi_a + B_{2a}/\Psi_a{}^2 + \cdots}\right], \quad (2)$$

where $\zeta_0 = 120\pi$ ohms, Ψ_s and Ψ_a are the expansion factors, and the A's and B's are complex coefficients. Formulas (1) and (2), as written, include second-order

* Decimal classification: R120×R221. Original manuscript received by the Institute, November 28, 1951; revised manuscript received February 4, 1952.

† Cruft Laboratory, Harvard University, Cambridge, Mass.

[1] R. King and C. W. Harrison, Jr., "Mutual and self-impedance for coupled antennas," *Jour. Appl. Phys.*, vol. 15, pp. 481–495; June, 1944.

[2] C. J. Bouwkamp, "On the theory of coupled antennae," *Philips Res. Rep.*, vol. 3, pp. 213–226; June, 1946.

[3] C. T. Tai, "Coupled antennas," Proc. I.R.E., vol. 36, pp. 487–500; April, 1948.

[4] R. King and D. Middleton, "The cylindrical antenna; current and impedance," *Quart. Appl. Math.*, vol. 3, pp. 302–325; January, 1946.

[5] E. Hallén, "Theoretical investigations into the transmitting and receiving qualities of antennae," *Nova Acta Uppsala.*, vol. 77, pp. 1–44; November, 1938.

terms. Whereas the first-order coefficients A_1 and B_1 can be expressed in terms of generalized sine and cosine integrals which have been tabulated,[6] the more intricate second-order coefficients can be evaluated only by numerical methods. Self- and mutual impedances for the coupled antennas are expressed in terms of the symmetrical and antisymmetrical impedances

$$Z_{s1} = Z_{s2} = (Z_0^s - Z_0^a)/2, \tag{3}$$

$$Z_{12} = Z_{21} = (Z_0^s - Z_0^a)/2. \tag{4}$$

If the zeroth- and first-order terms are retained in the numerators of (1) and (2) but only the zeroth-order term in the denominators, the solution reduces to the conventional one obtained by the so-called emf method.[7,8] This form will be designated the *zeroth-order solution* even though it includes more than the zeroth-order terms in the numerators of (1) and (2).

Since the zeroth-order formulas for Z_{s1} and Z_{12} are sufficiently accurate quantitatively to be of practical value *only for very thin antennas* ($\Omega > 15$) *of half-lengths near* $\lambda_0/4$, it is necessary to evaluate impedances of higher order for antennas of greater thickness ($\Omega \geqq 10$) and arbitrary length. For this purpose Tai's first-order computations have been extended from the two values $\beta_0 h = \pi/2$ and π to the range of $\beta_0 h$ from 1.0 to 4.0 in steps of 0.5 for $\Omega = 10$, 12.5, 15, and 20. However, while much more accurate than zeroth-order values, even first-order self- and mutual impedances cannot be in good agreement with experimental measurements. This is concluded from the fact that first-order impedances of *isolated antennas* are quantitatively unsatisfactory. Careful experimental studies have shown that the *second-order* King-Middleton impedances of isolated antennas are in excellent agreement with experimental results if account is taken of terminal-zone effects. But this is not true of first-order values. Clearly, therefore, since the impedance of an antenna oscillates about the value for isolation when a second antenna is moved nearer from an infinite distance, all first-order self- and mutual impedances lead to input impedances that oscillate about a first-order value that is a rather poor approximation, instead of a second-order value known to be a good approximation.

The analysis for two coupled antennas necessarily includes the isolated antenna as a special case for infinite separation; the same degree of approximation is required as for a single antenna. Unfortunately, the evaluation of second-order symmetrical and antisymmetrical impedances, and from these the calculation of second-order self- and mutual impedances, involves such an extensive series of numerical integrations that their determination is not contemplated. There is an alternative. A method has been devised for determining *approximate second-order* self- and mutual impedances from the computed *first-order* values of Z_{s1} and Z_{12} and the known *second-order* values of the self-impedance Z_0 of an isolated antenna. In effect, the first-order variation due to a coupled antenna is superimposed onto the second-order self-impedance of an isolated antenna. The justification for this procedure is that the interaction of current filaments within the small diameter $2a$ of an antenna requires second-order accuracy, whereas the interaction of currents separated the relatively large distance b between antennas, even when closely coupled ($b^2 \gg a^2$), is represented satisfactorily by a first-order approximation. Accomplishing the transfer of first-order coupling effects from a first- to second-order impedance is described next.

Approximate Second-Order Self- and Mutual Impedances

The first-order values of symmetrical and antisymmetrical impedances, and hence of the self- and mutual impedances determined by using (3) and (4), are not quantitatively accurate for a particular length of antenna, primarily because the first-order formula for the impedance Z_0 of an isolated antenna *is, in effect, distorted in its* $\beta_0 h$ *scale* as compared with the quite accurate second-order formula. This is shown in Fig. 1 where first-

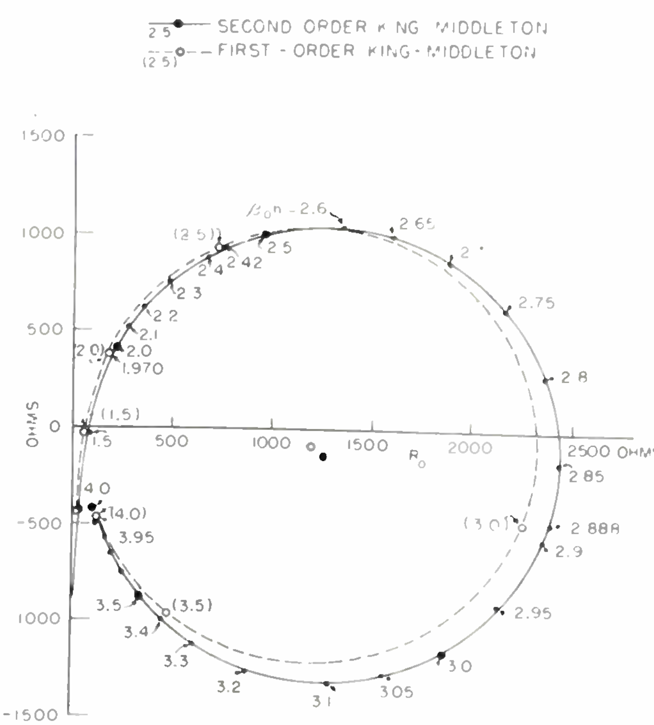


Fig. 1—Spiral plot of self-impedance Z_0 of isolated antenna $\Omega = 15$.

and second-order values of $Z_0 = R_0 + jX_0$ are plotted in the complex RX-plane. The parametric values of $\beta_0 h$ are indicated along each spiral by points and numbers; the first-order values are in parentheses. Whereas the two spirals agree well in magnitude, the $\beta_0 h$ scale for the

[6] Harvard University Computations Laboratory, "Tables of Generalized Sine and Cosine Integrals," vols. I and II, Harvard University Press; 1949.

[7] P. S. Carter, "Circuit relations in radiating systems and applications to antenna problems," Proc. I.R.E., vol. 20, pp. 1004–1042; June, 1932.

[8] G. H. Brown, "Directional antennas," Proc. I.R.E., vol. 25, pp. 78–145; January, 1937.

first-order curve is shifted greatly from that of the second-order curve, especially in the range near antiresonance. Note particularly the points marked 3.0 and (3.0) on the curves. The corresponding impedances are $(Z_0)_1 = 2{,}250 - j474$, $(Z_0)_2 = 1{,}870 - j1{,}159$. On the other hand, if the first-order impedance at $\beta_0 h = (3.0)$ is associated with the second-order impedance at an adjacent point on the second-order spiral (for example, at $\beta_0 h = 2.888$) the corresponding second-order impedance is $(Z_0)_2 = 2{,}390 - j474$. Thus it is possible to correct the first-order impedances merely by associating each numerical value of $(Z_0)_1$ with an appropriately modified electrical half-length $\beta_0 h$. In Fig. 1 the first-order value at $\beta_0 h = (3.0)$ is associated with the second-order value at $\beta_0 h = 2.888$; the first-order value at $\beta_0 h = (2.5)$ is associated with the second-order value at $\beta_0 h = 2.42$, and so on. The new $\beta_0 h$ is always the second-order value at an adjacent point on the second-order impedance spiral. Evidently any first-order value of Z_0 may be converted *exactly* into an adjacent second-order value by applying a scale factor near unity to $(R_0)_1$ and another scale factor to $(X_0)_1$. In order to maintain a scale factor for $(X_0)_1$ near unity, it is necessary to project horizontally from the first-order to the second-order spiral *near resonance and antiresonance*. For example, the point $\beta_0 h = (3.0)$ is projected to $\beta_0 h = 2.888$.

The method used to convert first-order self- and mutual impedances into approximate second-order values is to assume that the same shift in $\beta_0 h$ which brings $(Z_0)_1$ near to $(Z_0)_2$ in the complex plane may be used to bring first-order self- and mutual impedances near to second-order values. The scale factor which then converts $(R_0)_1$ *exactly* into $(R_0)_2$ is used to convert first-order self- and mutual resistances into approximate second-order values. Similarly, the scale factor used to convert $(X_0)_1$ *exactly* into $(X_0)_2$ is used to convert first-order self- and mutual reactances into approximate second-order values. This procedure may be illustrated graphically for the point $\beta_0 h = (3.0)$ on the first-order curve in Fig. 1. This point in the complex RX-plane is the first-order impedance of an antenna of electrical half-length 3.0 radians when infinitely far ($b = \infty$) from a second identical and parallel antenna. As the second antenna is brought nearer the point representing the self-impedance, Z_{s1} traces a spiral about the point $(R_0)_1$, $(X_0)_1$ for infinite separation. For great electrical distances, $\beta_0 b$, this spiral is extremely small, but grows rapidly as $\beta_0 b$ becomes smaller. Beginning with the point $(R_{s1})_1$, $(X_{s1})_1$ for $\beta_0 b = 6.3$, the spiral is shown by a broken line in the insert in Fig. 2. As $\beta_0 b$ becomes smaller the spiral increases until from $\beta_0 b = 1.5$ to $\beta_0 b = 0.3$ it is, in effect, a great circle, as shown in Fig. 2. By multiplying $(R_{s1})_1$ by 1.062 and $(X_{s1})_1$ by 1.0, this first-order spiral is converted into the spiral in solid line. This latter spiral occupies the same position relative to the true *second-order* impedance $(Z_0)_2$ with infinite separation for $\beta_0 h = 2.888$ as does the spiral in broken line to the first-order impedance $(Z_0)_1$ for $\beta_0 h = (3.0)$. It appears reasonable to assume that the self-impedance given by the solid-line spiral in Fig. 2 is a good approximation of the second-order self-impedance. It is therefore called the *approximate second-order self-impedance.*

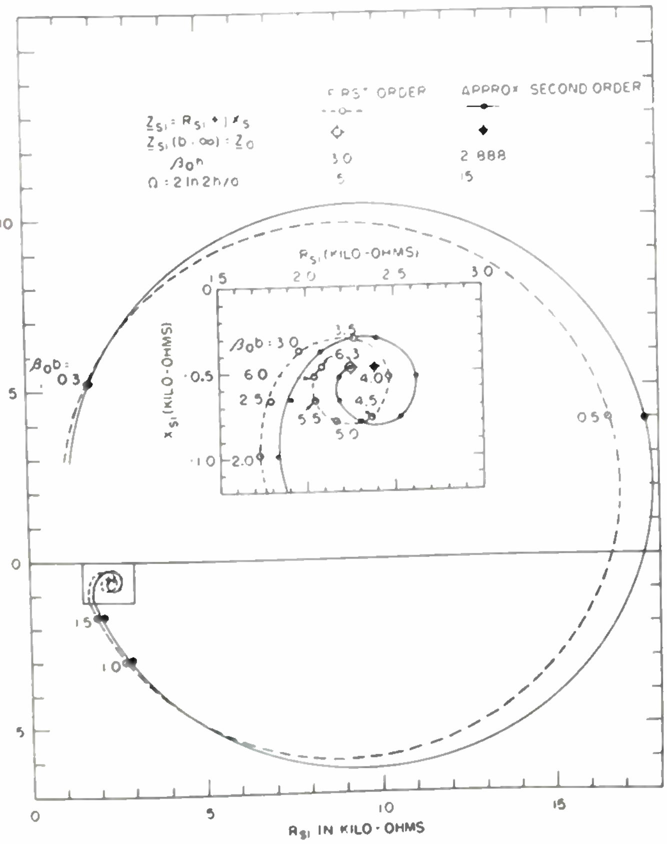


Fig. 2—Self-impedance spiral of parallel antennas.

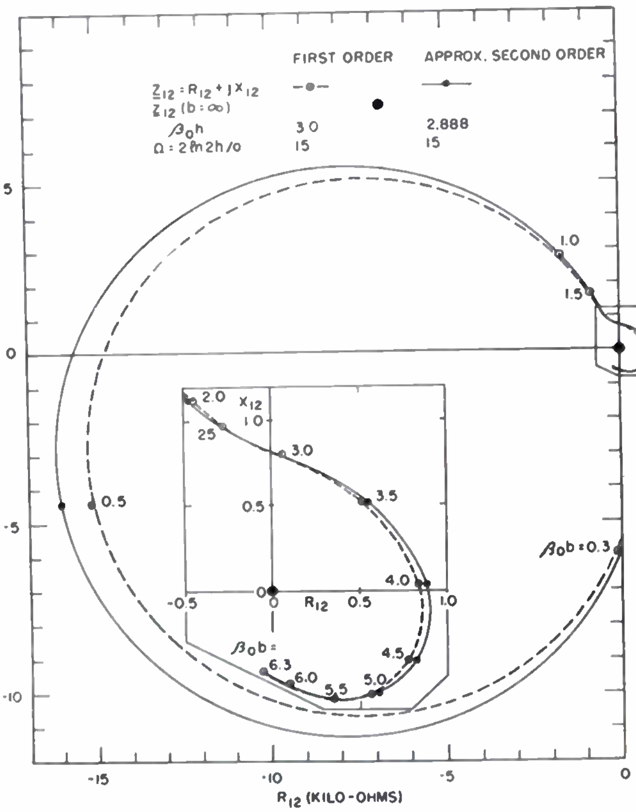
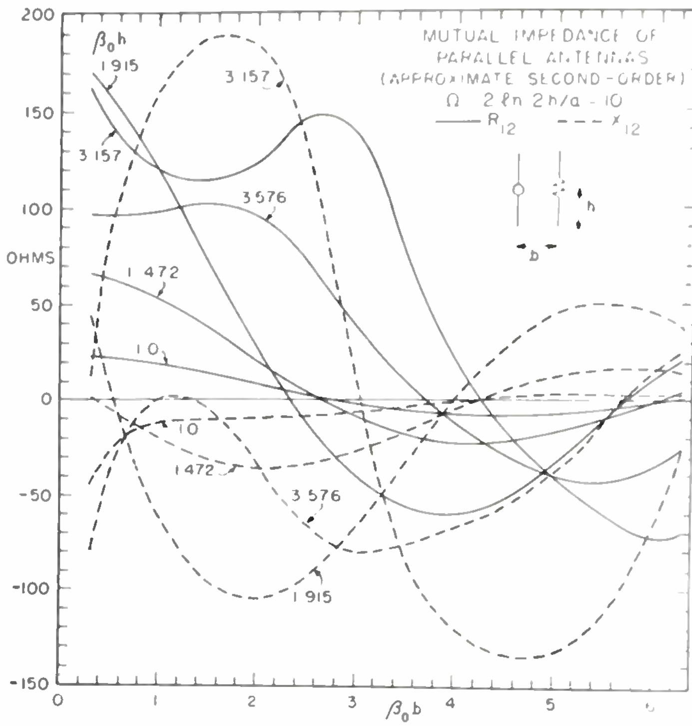


Fig. 3—Mutual impedance spiral of parallel antennas.

At infinite separation the mutual impedance, Z_{12}, is zero. As the two identical antennas are brought together the mutual impedance traces a small spiral about the zero point in the RX-plane. Beginning with $\beta_0 b = 6.3$, the first-order spiral for $\beta_0 h = (3.0)$ is shown by a broken line in Fig. 3 and enlarged in the insert. Beginning with $\beta_0 b = 1.5$ this spiral also becomes a great circle. By multiplying $(R_{12})_1$ by 1.062 and $(X_{12})_1$ by 1.0, the spiral is converted into the spiral in solid line. This is assumed to be a good approximation of the second-order mutual impedance for $\beta_0 h = 2.888$, and the impedance it describes is called the *approximate second-order mutual impedance*. Note that it is necessary to use the same scale factors for R_{12} and X_{12} as for R_{s1} and X_{s1} in order that $Z_0^s = Z_{s1} + Z_{12}$ and $Z_0^a = Z_{s1} - Z_{12}$ are transformed properly, in particular that Z_0^s may be equal to $2Z_0$ when b becomes sufficiently small—as must be true when $\beta_0 h$ is near $\pi/2$. The same scale factor is suggested by the fact that the maximum value of Z_{12} is of the same order of magnitude as Z_{s1} for each value of $\beta_0 h$. It is to be expected, of course, that a representation of second-order self- and mutual impedances from first-order quantities by an appropriate change in $\beta_0 h$ and by use of scale factors for resistance and reactance is quantitatively accurate *only if both scale factors are quite near unity.*

Tables of approximate second-order self- and mutual impedances were prepared using the extensive first-order computations previously referred to. For each first-order value of $\beta_0 h$ an appropriate second-order value was determined from enlarged sections of spiral diagrams such as those in Fig. 1. The true second-order impedance of an isolated antenna with this electrical half-length was obtained directly from the impedance tables[9] or by interpolation from them. With the corresponding first-order values known, the ratio factors



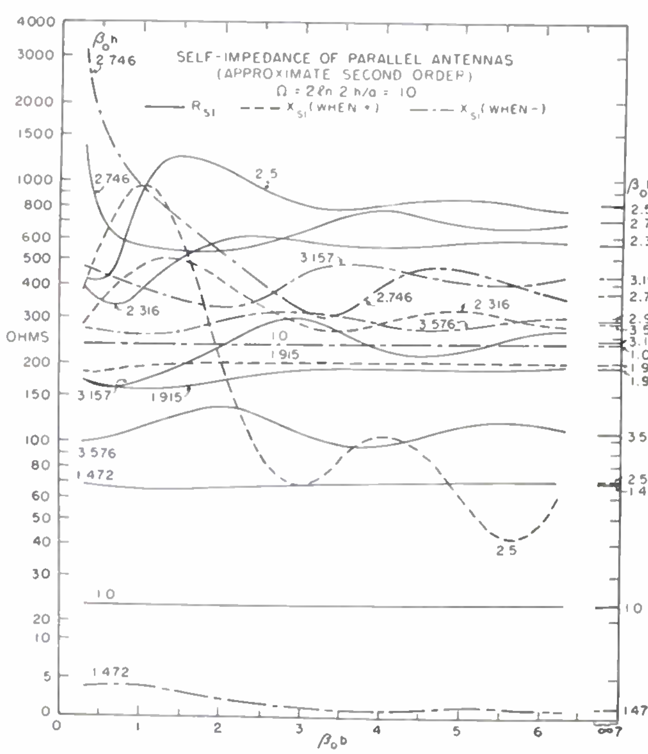


Fig. 4—Self-impedance of parallel antennas (approximate second-order); Ω = 10.

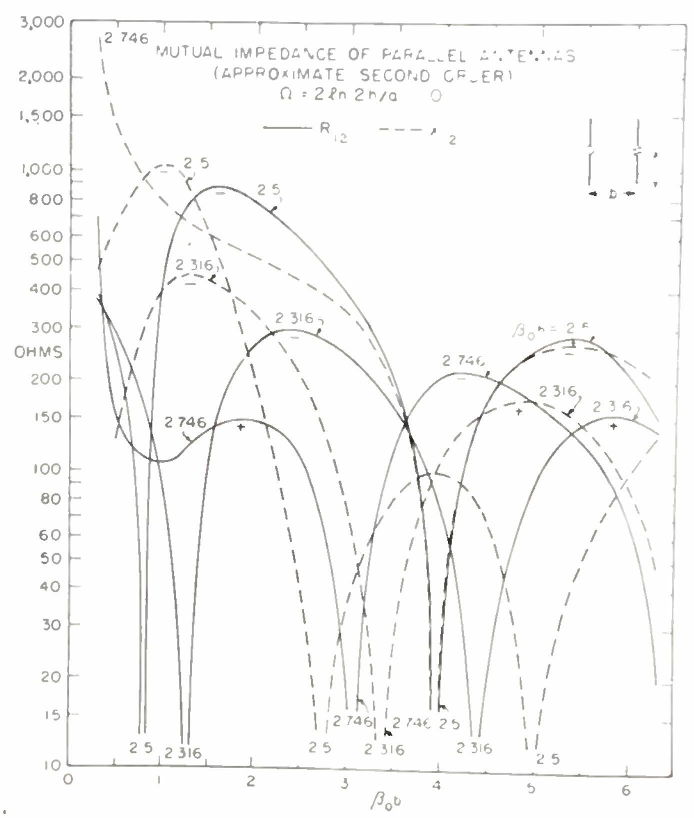
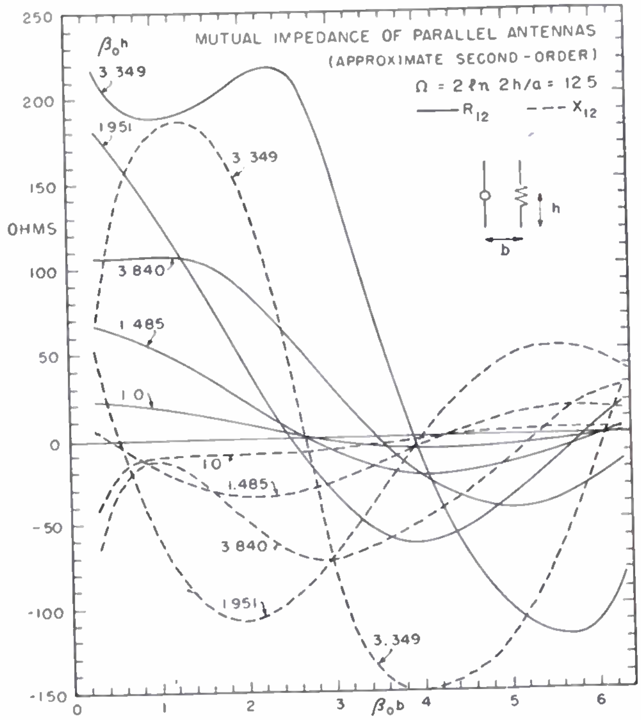


Fig. 5—Mutual impedance of parallel antennas (approximate second-order); Ω = 10.

[9] J. E. Storer, "Variational solution to the problem of the symmetrical cylindrical antenna," Cruft Laboratory Technical Report No. 101; February, 1950.

for resistance and reactance were computed and these used as factors multiplying the first-order values of self- and mutual resistance and reactance. The values obtained in this way are represented graphically in Figs. 4 through 9. In some instances, notably for $\beta_0 h=2.875$, $\Omega=12.5$; $\beta_0 h=2.888$, $\Omega=15$, the isolated antenna ($\beta_0 b=\infty$) is so near antiresonance that the change in impedance caused by moving a second antenna close produces antiresonance in the self- and mutual impedances. Since computed points were not sufficiently close-spaced to plot the curves over ranges of very rapid variation, the impedances were plotted on the RX-plane and the associated values of R and X determined from the spiral diagram. Figs. 2 and 3 are such diagrams for $\beta_0 h=2.888$, $\Omega=15$.

The self- and mutual impedances of antennas of half-length $\beta_0 h=\pi/2$ are of particular interest. Extensive first-order computations were made by Tai. To obtain approximate second-order values, however, first-order values are required for $\beta_0 h$ somewhat greater than $\pi/2$ by an amount depending on the ratio h/a of the antenna. As these were not available and the long additional computation did not seem justifiable, approximate second-order self- and mutual impedances for $\beta_0 h=\pi/2$ were obtained by interpolation from impedances already available for values of $\beta_0 h$ near 1.0, 1.5, and 2.0.



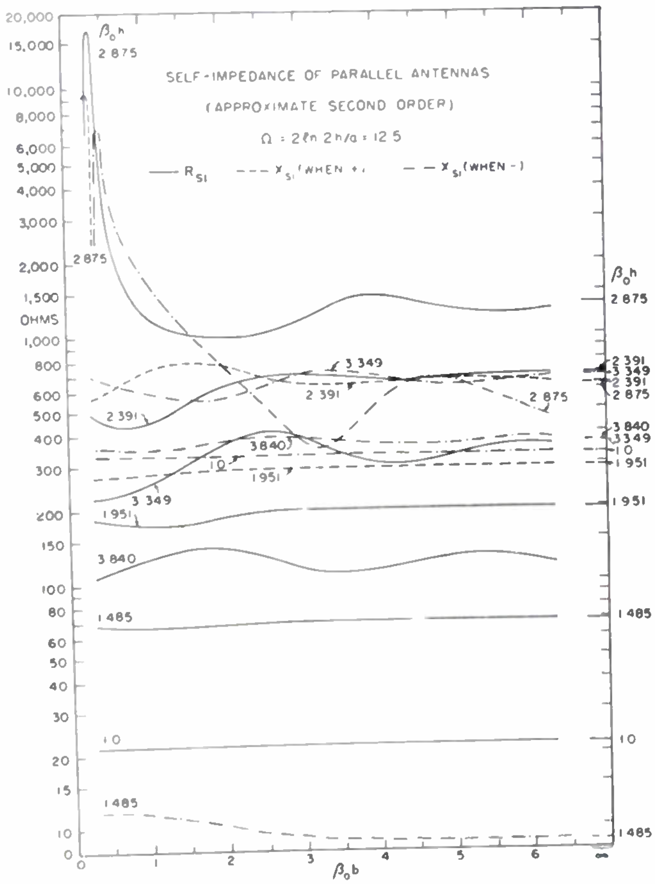


Fig. 6—Self-impedance of parallel antennas (approximate second-order); $\Omega=12.5$.

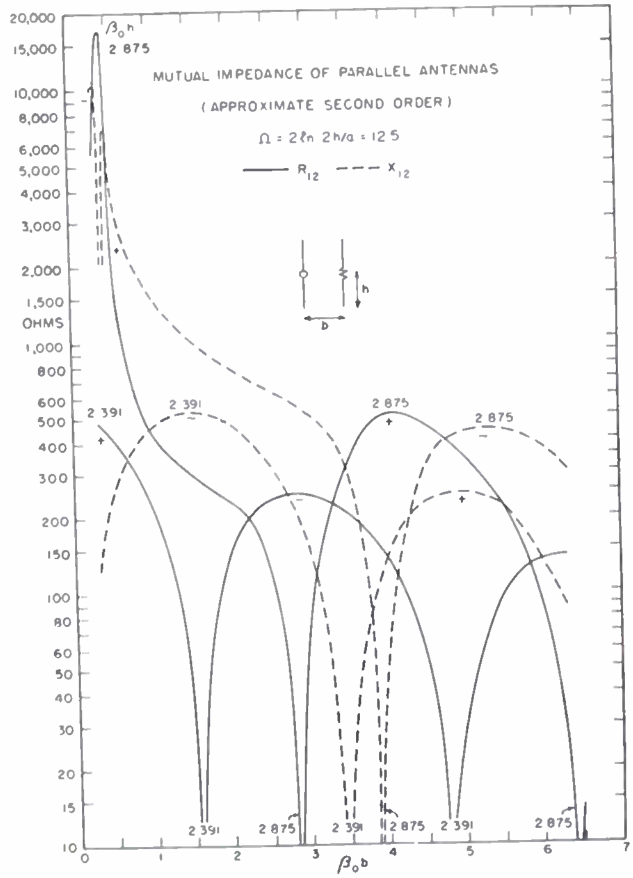


Fig. 7—Mutual impedance of parallel antennas (approximate second-order); $\Omega=12.5$.

At infinite separation the mutual impedance, Z_{12}, is zero. As the two identical antennas are brought together the mutual impedance traces a small spiral about the zero point in the RX-plane. Beginning with $\beta_0 b = 6.3$, the first-order spiral for $\beta_0 h = (3.0)$ is shown by a broken line in Fig. 3 and enlarged in the insert. Beginning with $\beta_0 b = 1.5$ this spiral also becomes a great circle. By multiplying $(R_{12})_1$ by 1.062 and $(X_{12})_1$ by 1.0, the spiral is converted into the spiral in solid line. This is assumed to be a good approximation of the second-order mutual impedance for $\beta_0 h = 2.888$, and the impedance it describes is called the *approximate second-order mutual impedance.* Note that it is necessary to use the same scale factors for R_{12} and X_{12} as for R_{s1} and X_{s1} in order that $Z_0^s = Z_{s1} + Z_{12}$ and $Z_0^a = Z_{s1} - Z_{12}$ are transformed properly, in particular that Z_0^s may be equal to $2Z_0$ when b becomes sufficiently small—as must be true when $\beta_0 h$ is near $\pi/2$. The same scale factor is suggested by the fact that the maximum value of Z_{12} is of the same order of magnitude as Z_{s1} for each value of $\beta_0 h$. It is to be expected, of course, that a representation of second-order self- and mutual impedances from first-order quantities by an appropriate change in $\beta_0 h$ and by use of scale factors for resistance and reactance is quantitatively accurate *only if both scale factors are quite near unity.*

Tables of approximate second-order self- and mutual impedances were prepared using the extensive first-

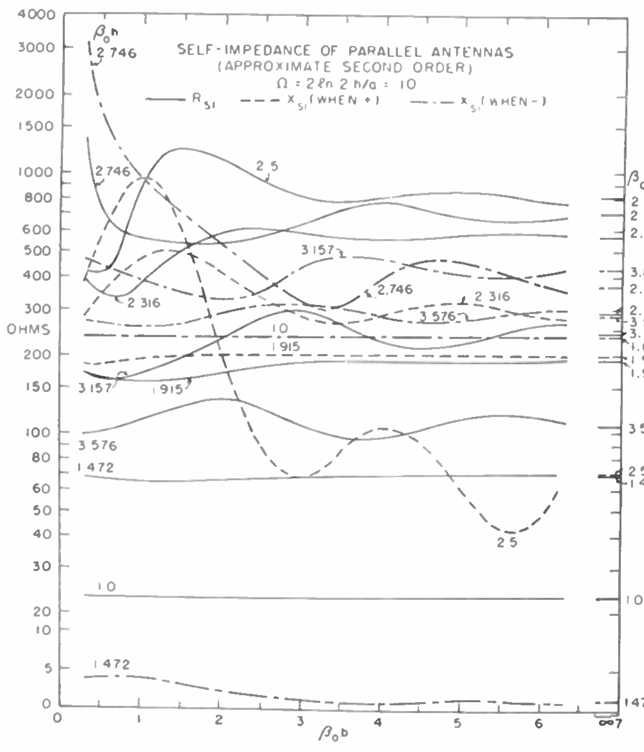
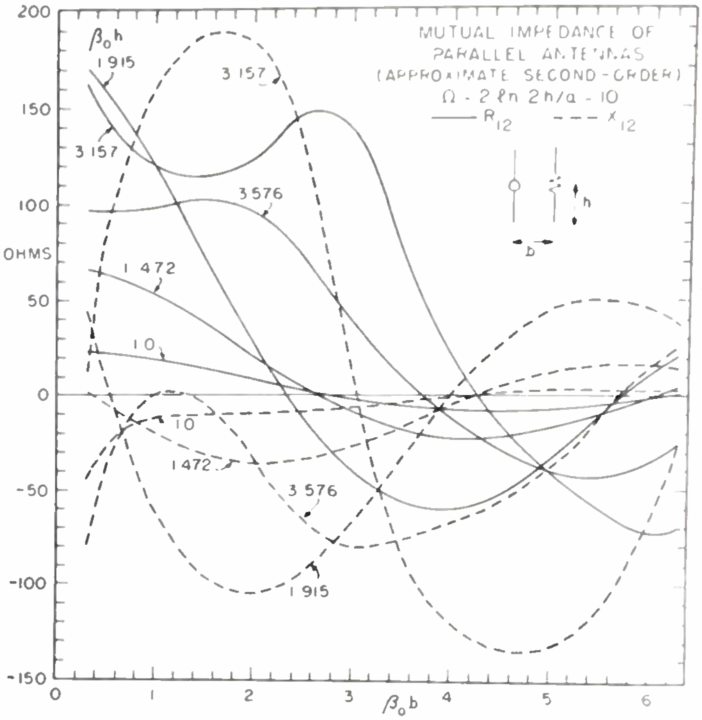


Fig. 4—Self-impedance of parallel antennas (approximate second-order); $\Omega = 10$.

order computations previously referred to. For each first-order value of $\beta_0 h$ an appropriate second-order value was determined from enlarged sections of spiral diagrams such as those in Fig. 1. The true second-order impedance of an isolated antenna with this electrical half-length was obtained directly from the impedance tables[9] or by interpolation from them. With the corresponding first-order values known, the ratio factors



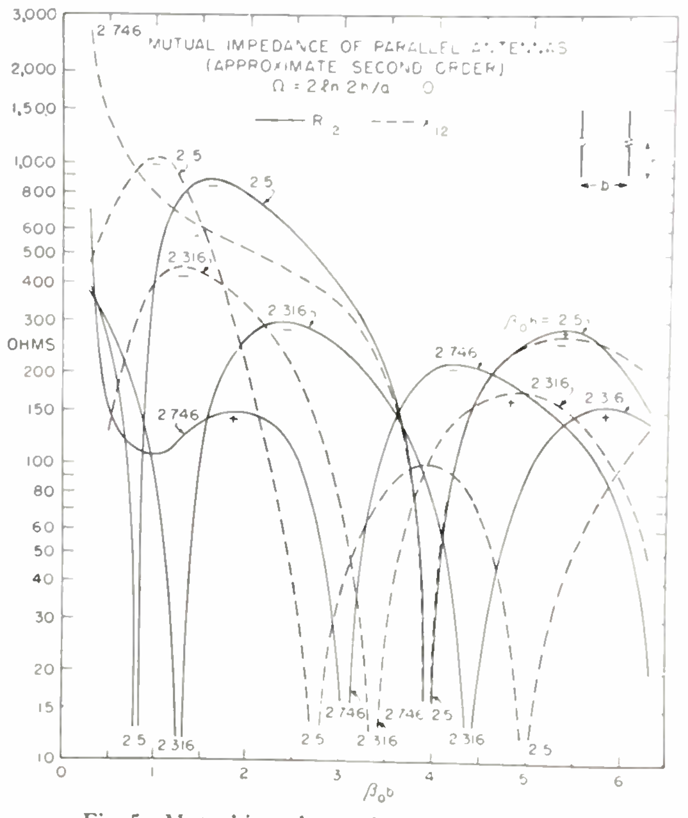
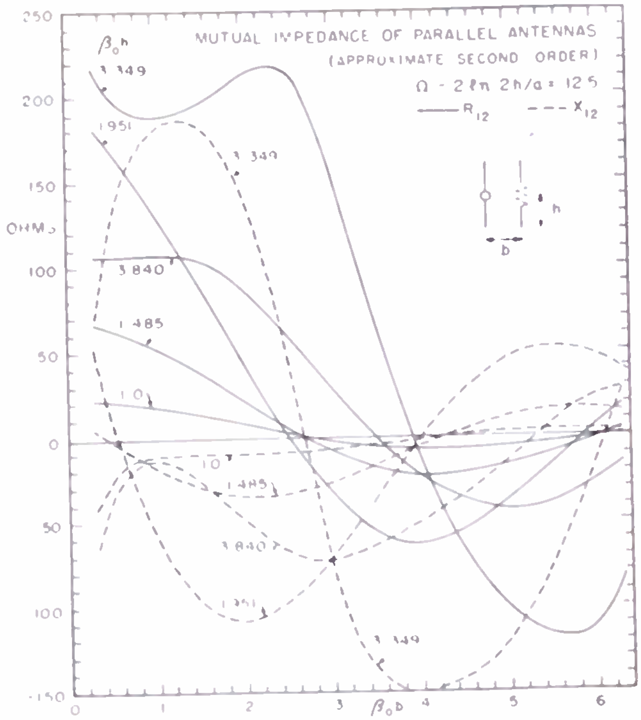


Fig. 5—Mutual impedance of parallel antennas (approximate second-order); $\Omega = 10$.

[9] J. E. Storer, "Variational solution to the problem of the symmetrical cylindrical antenna," Cruft Laboratory Technical Report No. 101; February, 1950.

for resistance and reactance were computed and these used as factors multiplying the first-order values of self- and mutual resistance and reactance. The values obtained in this way are represented graphically in Figs. 4 through 9. In some instances, notably for $\beta_0 h = 2.875$, $\Omega = 12.5$; $\beta_0 h = 2.888$, $\Omega = 15$, the isolated antenna ($\beta_0 b = \infty$) is so near antiresonance that the change in impedance caused by moving a second antenna close produces antiresonance in the self- and mutual impedances. Since computed points were not sufficiently close spaced to plot the curves over ranges of very rapid variation, the impedances were plotted on the RX-plane and the associated values of R and X determined from the spiral diagram. Figs. 2 and 3 are such diagrams for $\beta_0 h = 2.888$, $\Omega = 15$.

The self and mutual impedances of antennas of half-length $\beta_0 h = \pi/2$ are of particular interest. Extensive first order computations were made by Tai. To obtain approximate second order values, however, first-order values are required for $\beta_0 h$ somewhat greater than $\pi/2$ by an amount depending on the ratio h/a of the antenna. As these were not available and the long additional computation did not seem justifiable, approximate second-order self- and mutual impedances for $\beta_0 h = \pi/2$ were obtained by interpolation from impedances already available for values of $\beta_0 h$ near 1.0, 1.5, and 2.0.



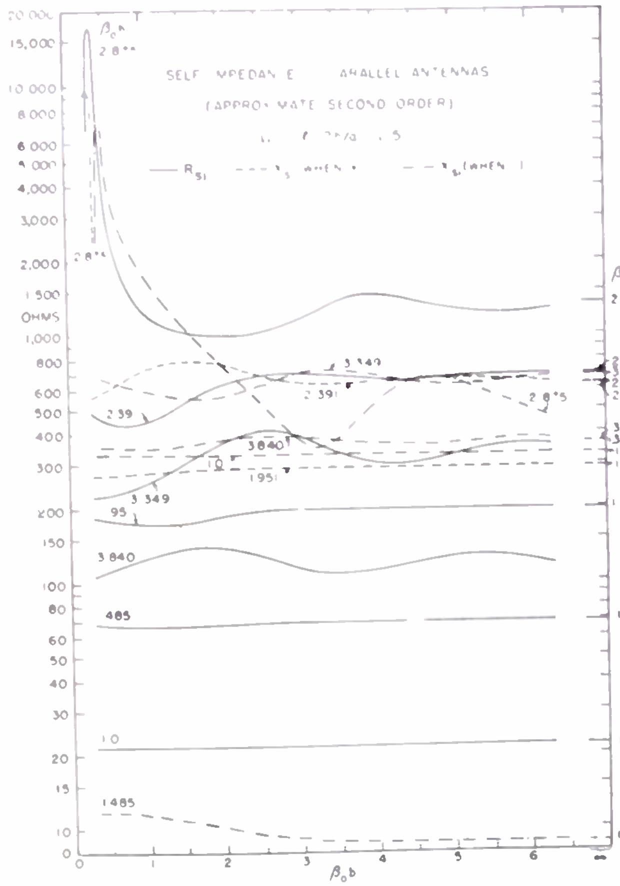


Fig. 6—Self-impedance of parallel antennas (approximate second-order); $\Omega = 12.5$.

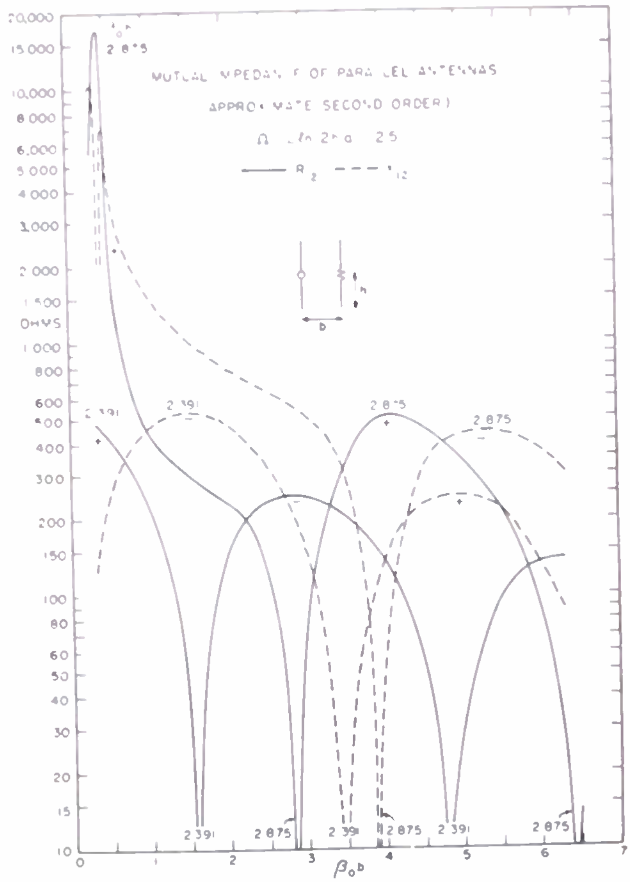


Fig. 7—Mutual impedance of parallel antennas (approximate second-order); $\Omega = 12.5$.

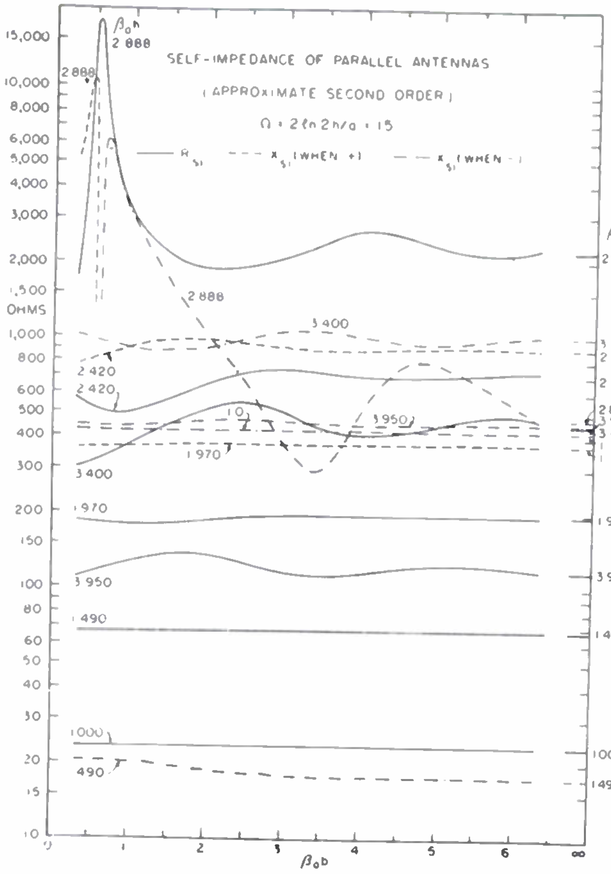


Fig. 8—Self-impedance of parallel antennas (approximate second-order); $\Omega=15$.

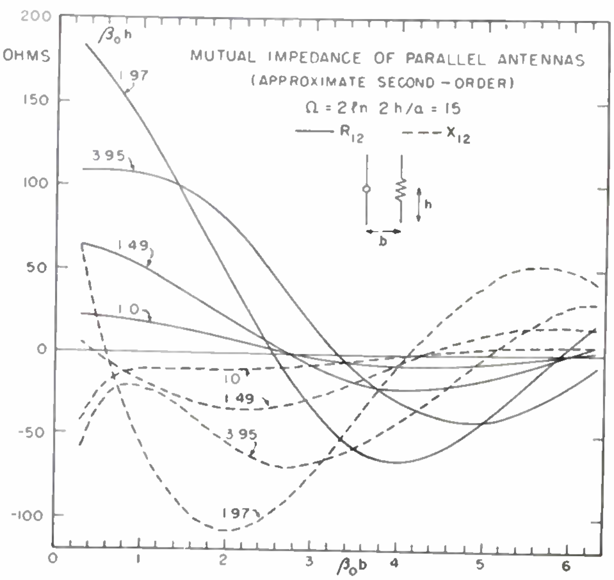
Approximate second-order self- and mutual impedances for $\beta_0 h=\pi/2$ are in Fig. 10 for $\Omega=10$, 15, 20, and ∞. Self- and mutual impedances for the important range of $\beta_0 h$ from 1.0 to 2.0, i.e., near resonance, are in Figs. 11 and 12. Zeroth-order curves have been computed by Starkey and Fitch.[10]

The symmetrical and antisymmetrical impedances Z_0^s and Z_0^a are obtained by using (3) and (4). Note that the latter is the input impedance of an antenna parallel to and at a distance $b/2$ above a perfectly conducting infinite plane. The values for $\Omega=\infty$ are the conventionally given zeroth-order values. The symmetrical and antisymmetrical impedances are represented graphically in Figs. 13 through 15.

Conclusion

In the absence of accurate second-order self- and mutual impedances the approximate second-order values obtained by superimposing the first-order effect of coupled antennas on the accurate second-order impedance of an isolated antenna should be superior in accuracy to zeroth- and first-order values. The approximate second-order self-impedance Z_{s1}, symmetrical impedance $Z_0^s=Z_{s1}+Z_{12}$, and antisymmetrical impedance $Z_0^a=Z_{s1}-Z_{12}$ all represent oscillations about or variations from the correct limiting value Z_0 as the separation is increased without limit.

[10] Starkey and Fitch, "Mutual impedance and self-impedance of coupled antennas," *Jour. IEE* (London), part III, vol. 97, pp. 129–137; May, 1950.



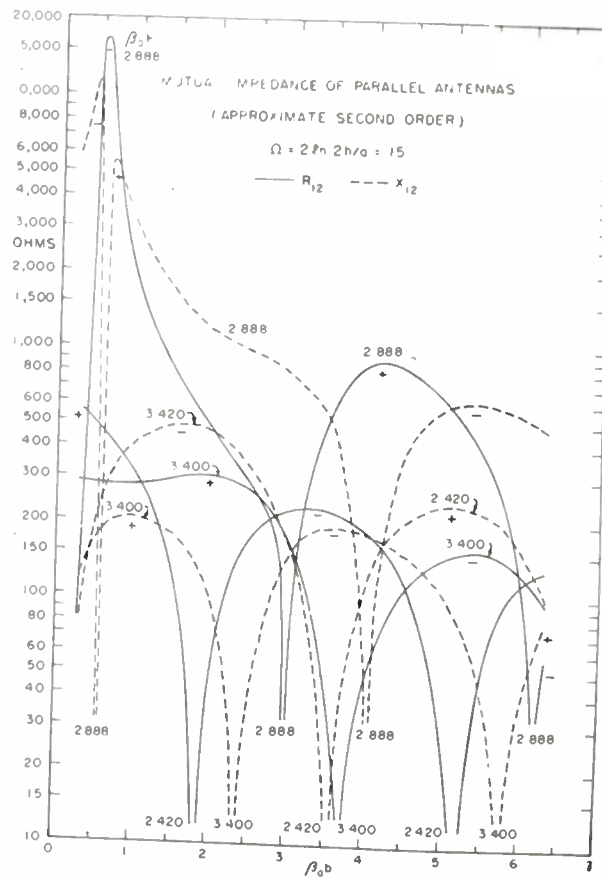


Fig. 9—Mutual impedance of parallel antennas (approximate second-order); $\Omega=15$.

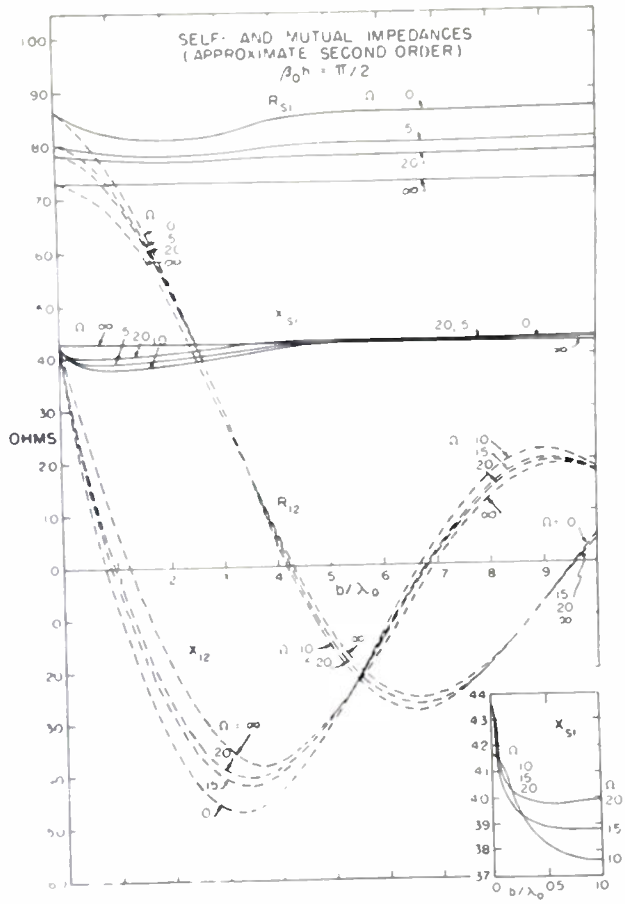


Fig. 10—Self- and mutual impedances (approximate second-order); $\beta_0 h = \pi/2$.

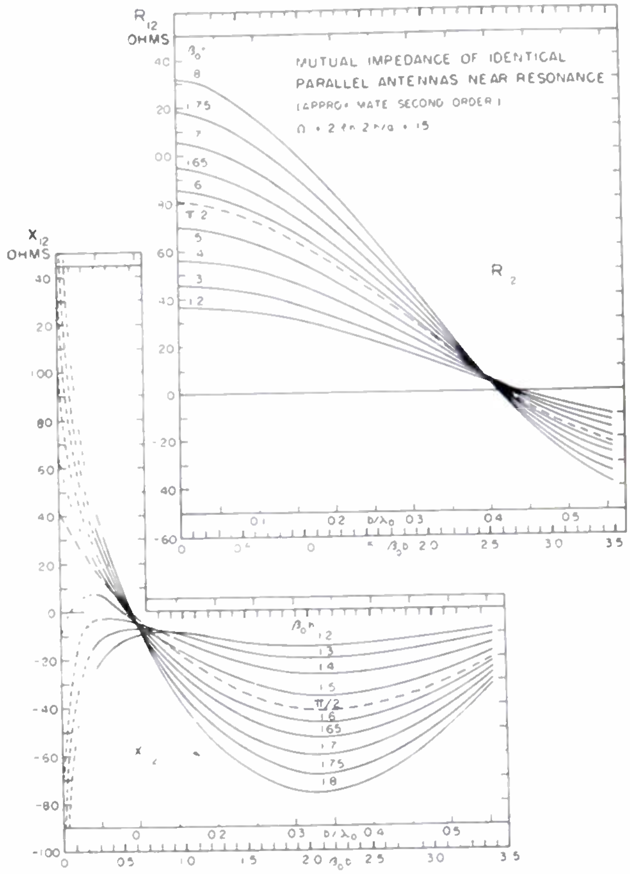


Fig. 12—Mutual impedance of identical parallel antennas near resonance (approximate second-order); $\Omega = 15$.

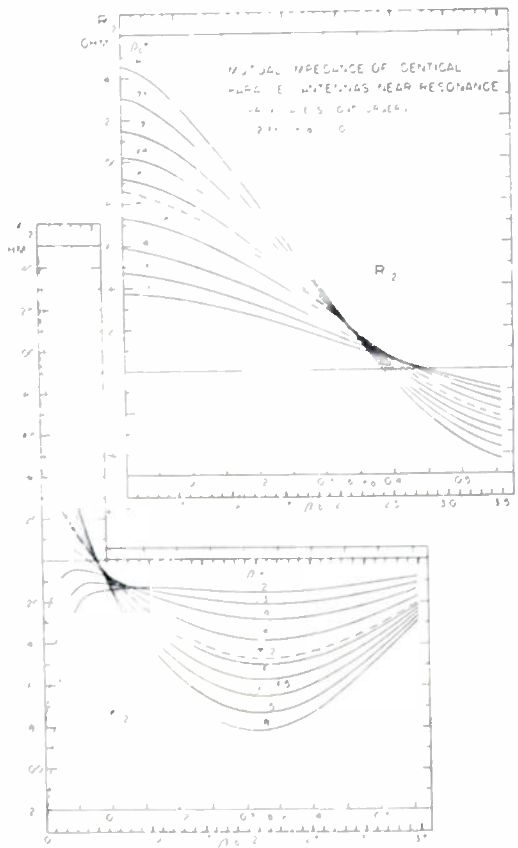


Fig. 11—Mutual impedance of identical parallel antennas near resonance (approximate second-order); $\Omega = 10$.

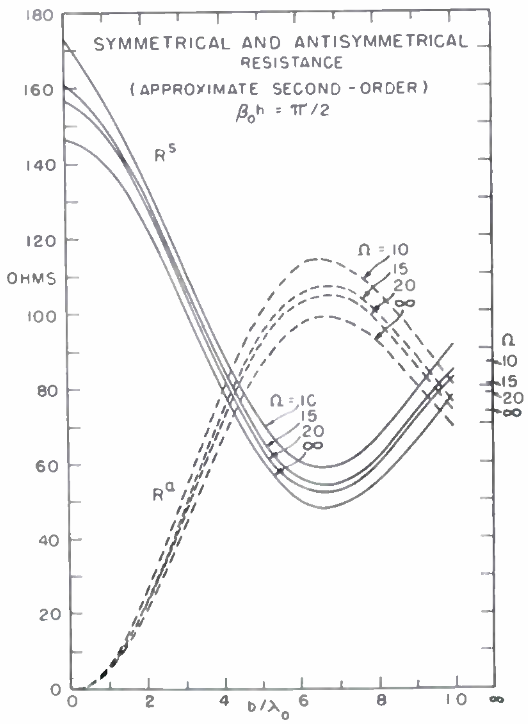


Fig. 13—Symmetrical and antisymmetrical resistance (approximate second-order); $\beta_0 h = \pi/2$.

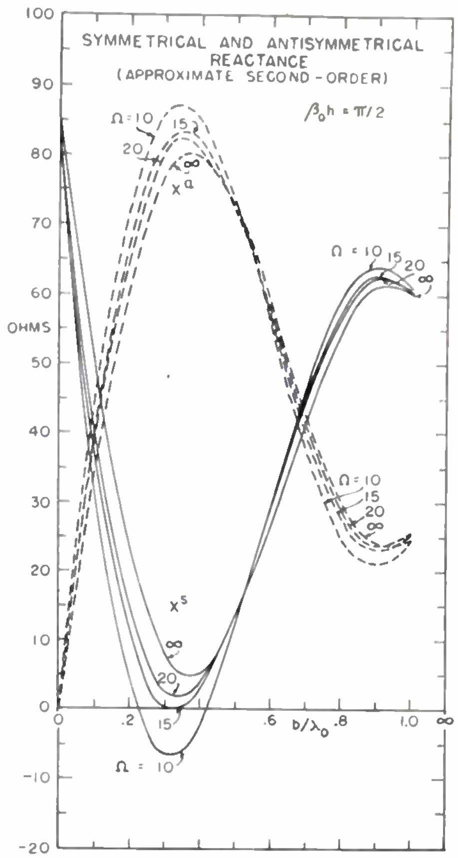


Fig. 14—Symmetrical and antisymmetrical reactance (approximate second-order); $\beta_0 h = \pi/2$.

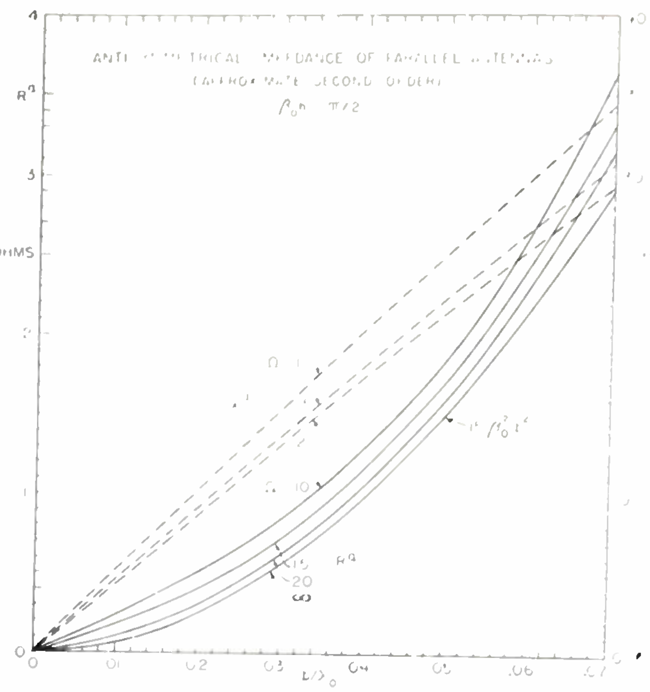


Fig. 15—Antisymmetrical impedance of parallel antennas (approximate second-order); $\beta_0 h = \pi/2$.

It is essential to bear in mind that the impedances represented apply when antennas are given by so-called slice generators or their equivalent. This means transmission-line end and coupling effects are negligible, as when the spacings of the driving lines approach values that are small compared with the wavelength. In most practical cases, significant end and coupling effects exist of which account must be taken in the form of equivalent lumped networks. Determination of the elements of such networks is more intricate with coupled antennas than with isolated antennas as analyzed by King,[11] King and Tomiyasu,[12] and by Hartig,[13] since it is not sufficient to take account of the coupling between each antenna and its feeding line but account also must be taken of coupling between the two lines and between each antenna and the feeding line of the other antenna. In comparing theoretical predictions obtained using the impedances of this paper with experimental observations that *apparently* correspond to quantities determined theoretically, the significance of these end and coupling effects must not be overlooked. Comparisons of impedances and front-to-back ratios determined theoretically using the approximate second-order self- and mutual impedances obtained in this paper with experimental measurements are described by King[14] and by Morita and Faflick.[15] The agreements are good.

Acknowledgments

Mr. Seymour Stein and Miss Phyllis Kennedy assisted with phases of the work.

[11] R. King, "Antennas and open-wire lines I," *Jour. Appl. Phys.* vol. 20, pp. 832–850; September, 1949; Cruft Laboratory Technical Report No. 41.

[12] R. King and K. Tomiyasu, "Terminal impedance and generalized two-wire line theory," Proc. I.R.E., vol. 37, pp. 1134–1139; October, 1949; Cruft Laboratory Technical Report No. 74.

[13] E. O. Hartig, "Circular apertures and their effects on half-dipole impedances," Cruft Laboratory Technical Report No. 107; June, 1950.

[14] R. King, "A dipole with a tuned parasitic Radiator," *Jour IEE* (London), vol. 99, part III, pp. 6–14; January, 1952.

[15] T. Morita and C. Faflick, "Measurement of current distributions along coupled antennas and folded dipoles," Proc. I.R.E., vol. 39, pp. 1561–1565; December, 1951.

Radiation Characteristics of Helical Antennas of Few Turns*

OBED C. HAYCOCK†, SENIOR MEMBER, IRE AND JAMES S. AJIOKA‡, STUDENT, IRE

Summary—**A one-turn helical antenna designed to produce axial radiation is linearly polarized. A similar antenna can be made circularly polarized by suitable resistance loading. This paper presents a method of producing circular polarization by resistance loading a $1\frac{1}{4}$-turn helix.**

Introduction

THE NEED for suitable antennas for ionosphere measurements at frequencies near the critical, approximately 3–8 mc, prompted the authors to investigate helical antennas of few turns. The purpose of this paper is to report results of this investigation. Desirable characteristics are circular polarization, broadband, vertical radiation, and reasonable size. A helix to operate in this frequency band would be limited to approximately one turn, and in this study was to be supported by four poles placed at corners of a square.

Description of Antennas Tested

A. Single-Turn Square Helix

Scale models were constructed, designed to operate at 600 mc, using 14° pitch[1,2] as required for transmission in the axial mode. Both corner-fed and center-fed antennas were constructed, as shown in Fig. 1. These use dowel rods for poles and copper screen for a ground plane. The rods were placed at corners of a 12-cm square; the helix used a 14° pitch and the ground plane was approximately two wavelengths square.

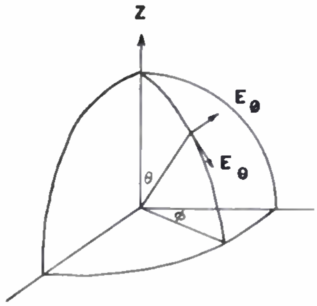
B. Resistive Loaded Antennas

Square helical antennas of both one turn and $1\frac{1}{4}$ turns of 14° pitch were constructed in the same manner as in part A, except that a 300-ohm resistor was placed near the open end. Beyond this resistor, a length of wire approximately a quarter wavelength was added, going to center of the helix and down the axis. The end of the wire was unterminated. (See Fig. 1.)

Experimental Procedure

The pattern measurements were taken in the usual manner with about 10 wavelengths between the transmitting and receiving antenna. The test antenna was placed horizontally against the vertical ground plane of copper screen. The ground plane could rotate about a vertical axis, and the antenna could rotate about the helix axis.

The field-intensity detecting device consisted of a balanced dipole and a 1N21B crystal. The crystal was operated at a level that insured its operation as a square law detector.[3]



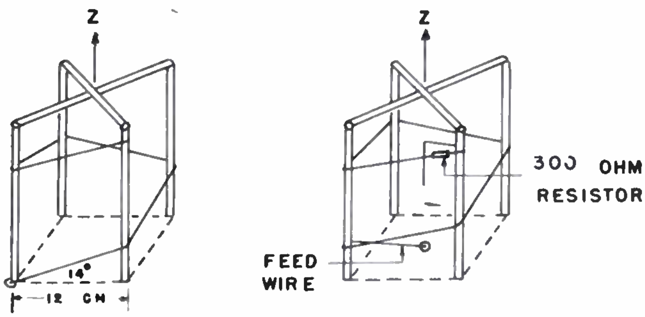


Fig. 1—Antenna construction. (Left) Corner-fed, nonloaded. (Right) Center-fed, resistance loaded.

Tests for circularity of polarization were made by rotating the antenna about its own helix axis with the ground plane stationary. The polarization was obtained by taking readings versus antenna rotation.

Impedance measurements were made using a slotted line. A horizontal ground plane was used with the antenna pointing upward. The observer and all apparatus were kept below the ground plane.

Results

The $1\frac{1}{4}$-turn helix gave the radiation patterns shown in Fig. 2(a). As indicated, these are for 500, 600, and 700 mc. The radiation was linearly polarized with the patterns in the plane of polarization as shown.

The pattern for 500 mc is sharper than the one for 600 mc, and at 700 mc, the radiation is beginning to broaden out into the normal mode instead of the axial mode.

Changing the frequency from 500 to 700 mc caused the plane of polarization to change by approximately 90°.

* Decimal classification: R120.2×R326.6. Original manuscript received by the Institute, June 13, 1951; revised manuscript received May 5, 1952.

† Electrical Engineering Dep't., University of Utah, Salt Lake City 1, Utah.

‡ U. S. Navy Electronics Laboratory, San Diego, Calif.

[1] J. D. Kraus, "Helical beam antennas," *Electronics*, vol. 20, pp. 109–111; April, 1947.

[2] J. D. Kraus, "Helical beam antennas for wide-band applications," Proc. I.R.E., vol. 36, pp. 1236–1242; October, 1948.

[3] "The Preliminary Operating Instructions for the GR Type 874-LB Slotted Line," Form 727A, General Radio Company, Cambridge, Mass.

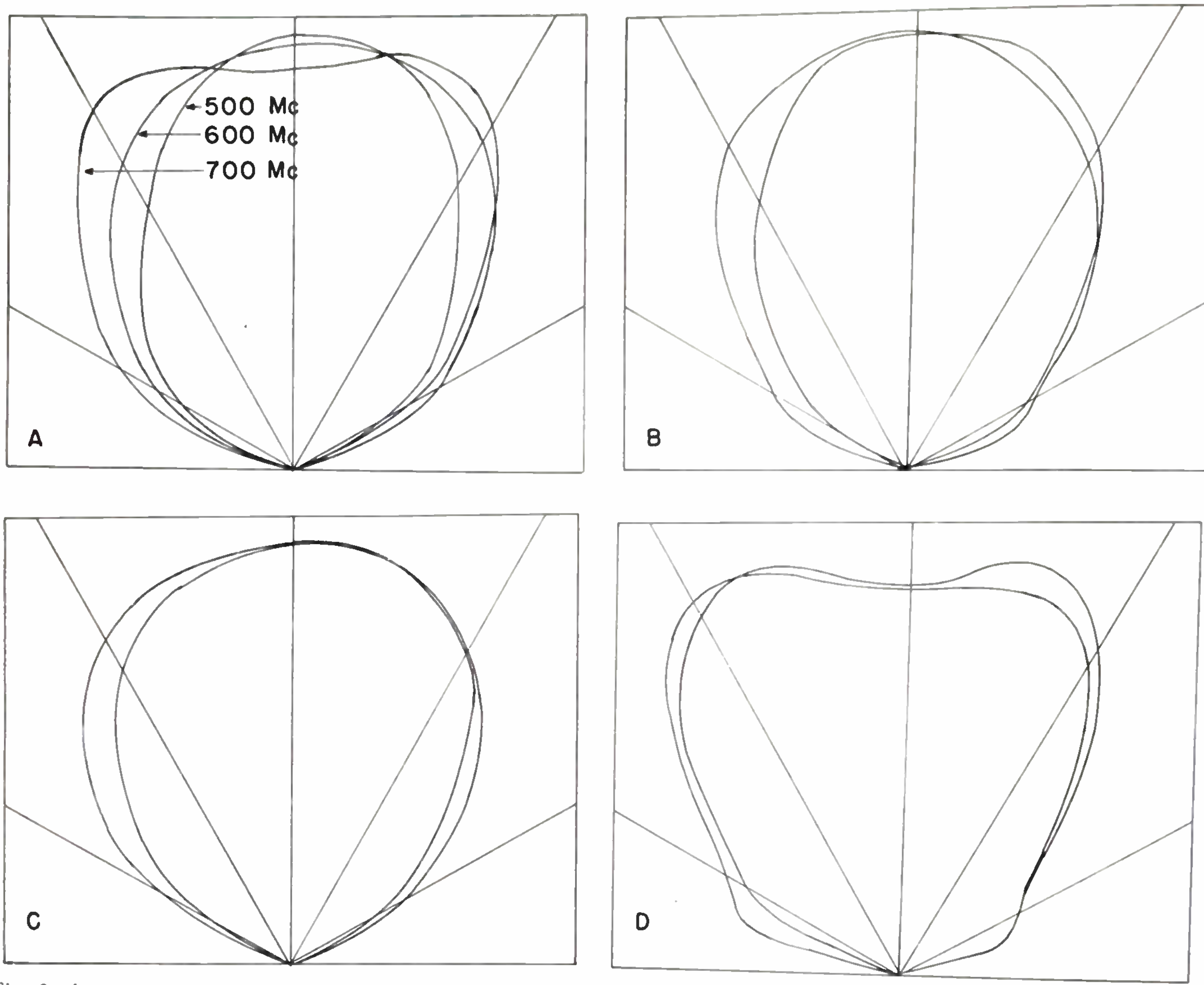


Fig. 2—Antenna patterns showing E_ϕ for $1\frac{1}{4}$ turn square helix. (A) Nonloaded helix. (B) Loaded helix, pattern in two right-angle planes at 500 mc. (C) Loaded helix, pattern in two right-angle planes at 600 mc. (D) Loaded helix, pattern in two right-angle planes at 700 mc.

At 600 mc the plane of polarization was at right angles to the lead going from the center of the ground plane at the feed point to the helix. (See Fig. 1.) It advanced about 45° in the direction of the helix at 500 mc, and was retarded about 45° at 700 mc.

The plane of polarization could also be altered by changing the distance from the ground plane to the helix. No signal above noise level could be detected when the receiving dipole was normal to the plane of polarization mentioned above.

Further study indicated that the linear instead of circular polarization resulted from standing waves on the helix. A small exploring coil was used to measure the standing current wave on the helix, and it was determined that these standing waves fixed the plane of polarization.

When the helix was increased to three turns, the standing wave decreased and essentially circular polarization resulted.

Methods of reducing the standing waves were considered; each, of course, required the elimination or reduction in the wave reflected from the open end of the helix, without increasing the physical length of the helix. Included in the methods were reactive loading and two-layer helix. These gave negligible improvement as far as circular polarization was concerned. The solution seemed to be in dissipative end loading at, of course, some loss in efficiency. It was reasoned that if the helix could be terminated in its characteristic impedance, the reflected wave would be eliminated.

A resistance placed at the open end of the line could not serve because the current would be zero. However, if an additional length of approximately $\frac{1}{4}$ wavelength were added beyond the resistance, then it would be placed at a current maximum and could reduce the reflected wave. This was done and the resistance value found not critical. Values between 250 and 300 ohms gave essentially the same results. The resulting antenna is shown in Fig. 1 with a 300-ohm loading resistance.

This antenna gave essentially circular polarization, with the pattern in the two planes shown in Fig. 2(b), (c), and (d). These are for E_θ in two planes at right angles. Both planes contain the axis of the helix. As the principal interest was in axial radiation, E_ϕ patterns were not obtained.

Twice as much power is necessary for a circularly polarized field as for a linearly polarized field to give the same field-intensity reading. For this reason, the readings obtained for the loaded antenna were added[4] for comparison with the reading for the unloaded antenna.

The ratio of the sum of the loaded readings to the unloaded reading gives the approximate ratio of the radiated powers in the forward direction.

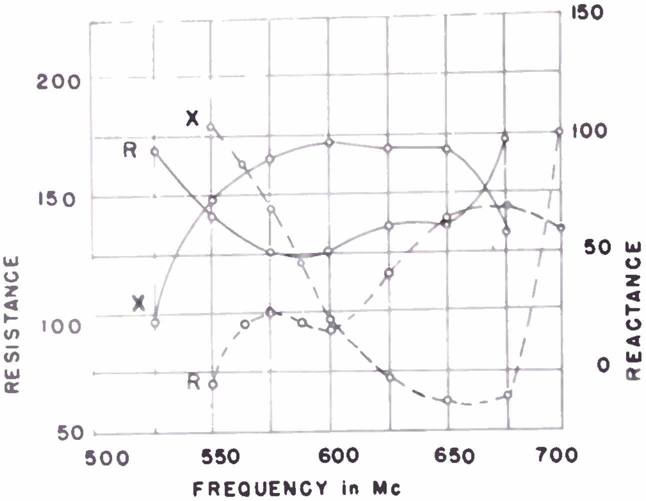


Fig. 3—Input impedance to 1¼ turn helix. Unloaded antenna shown with dashed line. Loaded antenna, with solid line.

The radiation patterns of the loaded and unloaded antennas were found to be nearly the same. Therefore, the above ratio could be considered as the approximate ratio of the radiation efficiency in the axial direction of the loaded antenna as compared to the unloaded antenna. This was found to be approximately 80 per cent.

[4] A square law detector will give reading proportional to power.

The input impedance of 1¼-turn helix is fairly broad-band with a resistive component varying between 90 and 150 ohm. (See Fig. 3.) The 1¼-turn loaded antenna is considerably more broad-band. This can be seen by comparing curves of Fig. 3. The resistive component remains between 120 and 145 ohms over a frequency range from 550 to 650 mc. The reactive component is also relatively constant over this range.

Conclusion

From the results of this investigation, the following is concluded:

(1) A simple single-turn helical antenna gives essentially linear polarization. Circular polarization is approached as the number of turns is increased, becoming essentially circular with three turns.

(2) The plane of polarization can be changed by adjusting the distance in wavelength of the helix above the ground plane.

(3) A single-turn helical antenna can be made to give circular polarization by proper resistive loading. The resistor is placed where there would normally be a current maximum. The resistance required for this configuration was experimentally determined to be approximately 300 ohms.

(4) The input impedance to the loaded antenna is approximately the same in resistance, and is more constant with frequency than an unloaded antenna. It has a resistive component in the order of 130 ohms. The reactance is primarily inductive.

(5) The radiation patterns of the loaded and unloaded antennas are very similar, with a half-power beam width from 60°–80°.

Acknowledgments

The authors wish to express thanks to the Research Group, Physics of the Upper Atmosphere, University of Utah, for their aid in making this project possible.

Correspondence

More on "A Method for Calculating the Current Distribution of Tschebyscheff Arrays"*

It has been pointed out to me by R. L. Pritchard of the General Electric Communications Research Laboratory that if a table is constructed for an odd number of elements (design (20′)) the sum down each column does not equal the corresponding Tschebyscheff coefficient.[1] He shows that this is due to an error in (13′) whose left-hand member should be $(2\pi/\epsilon_k)I_K$ where ϵ_k is the Newmann factor, $\epsilon_k \equiv 2$ for $k>0$, $\epsilon_0 \equiv 1$. The right-hand members of (15′), (16′), and (20′) should therefore be multiplied by $\epsilon k/2$. The only change in a table based on (20′) is that the first row ($k=0$) is halved. I also wish to point out that the factor 2^{2n-1} in (17) and (18) should be 2^{n-1}.

Domenick Barbiere
The Johns Hopkins University
Radiation Laboratory
1315 St. Paul St.
Baltimore 2, Md.

* Received by the Institute, April 21, 1952.

[1] D. Barbière, "A Method for Calculating the Current Distribution of Tschebyscheff Arrays," Proc. I.R.E., vol. 40, pp. 78–82; January, 1952.

Correspondence

Proposal of Piezomagnetic Nomenclature for Magnetostrictive Materials*

The writer wishes to propose a system of nomenclature for piezomagnetic equations of state which would be comparable to piezoelectric terminology. According to Cady's[1] definition, piezomagnetism is a linear magnetomechanical effect analogous to piezoelectricity. Similarly, magnetostriction and electrostriction are analogous second-order effects. The second-order and higher-order effects can be represented as effectively first-order when variations in the system parameters are small compared with the initial values of the parameters. Comparable terminology and nomenclature should therefore be useful.

The IRE has already published standards for piezoelectric crystals, and Mason has treated the subject of electrostriction in polarized barium titanate ceramics using piezoelectric terminology.[2,3] Table I compares the writer's proposed mks rationalized piezomagnetic notation in matrix form with equivalent equations from "Magnetostriction Transducers"[4] in Gaussian units and the analogous equations from the IRE Standard. Simple substitution of magnetic for electric variables is necessary. A glossary of the magnetomechanical parameters is presented in Table II.

The proposed system of notation should facilitate any comparison of magnetostrictive, piezoelectric, and electrostrictive materials and devices from a basic point of view. A specific field in which this system would be useful is that of underwater sound transducers.

In addition, the proposed notation is intended as a step in standardizing terminology for other transducing mechanisms.

Stanley L. Ehrlich
U. S. Navy Underwater Sound Laboratory
Fort Trumbull
New London, Conn.

TABLE I

Piezomagnetic Equations of State		Piezoelectric Equations of State
Proposed Notation	"Magnetostriction Transducers"	IRE Standard
$S=s^HT+d_tH$	$s=(1/E')P+\Lambda(H/4\pi)$	$S=s^ET+d_tE$
*$B=dT+\mu^TH$	*$B'=\Lambda P+[4\pi(\mu'-1)](H/4\pi)$	$D=dT+\epsilon^TE$
$T=c^HS-g_tH$		$T=c^ES-g_tE$
$B=eS+\mu^SH$		$D=eS+\epsilon^SE$
$S=s^BT+g_tB$		$S=s^DT+g_tD$
$H=-gT+\gamma^TB$		$E=-gT+\beta^TD$
$T=c^BS-h_tB$	$P=Es-\lambda B'$	$T=c^DS-h_tD$
$H=-hS+\gamma^SB$	$H/4\pi=-\lambda s+B'/[4\pi(\mu-1)]$	$E=-hS+\beta^SD$

* Note: The proposed equations use magnetic flux density, while the older forms contain intrinsic flux density as a primary variable. A similar choice between electric displacement (electric flux density) and electric polarization exists in piezoelectricity.

TABLE II

GLOSSARY OF SYMBOLS

Parameter	mks Rationalized Units		Conversion Factor	Gaussian Units	
	Symbol	Unit		Unit	Symbol
Stress	T	newton/meter²	=10	dyne/cm²	P
Strain	S	meter/meter	=1	cm/cm	s
Elastic compliance	s	meter²/newton	=0.1	cm²/dyne	$1/E'$
Elastic stiffness	c	newton/meter²	=10	dyne/cm²	E
Magnetic field intensity	H	ampere/meter	$=4\pi\times10^{-3}$	oersted	H
Magnetic flux density	B	weber/meter²	$=10^4$	gauss	B
Intrinsic flux density*	B'	weber/meter²	$=10^4$	gauss	B'
Intensity of magnetization*	M	weber/meter²		(no name)	I
Magnetic permeability	μ	henry/meter	$=4\pi\times10^{-7}$	dimensionless	μ'
Magnetic impermeability	γ	meter/henry	$=10^7/4\pi$	dimensionless	$1/\mu$
Permeability of free space	$\mu_0=4\pi\times10^{-7}$	henry/meter		$\mu_0=1$	
Effective piezomagnetic constant	d	weber/newton	$=10^3$	gauss cm²/dyne	Λ
Effective piezomagnetic constant	e	weber/meter²			
Effective piezomagnetic constant	g	meter²/weber			
Effective piezomagnetic constant	h	newton/weber	$=10^{-3}$	dyne/gauss cm²	λ

* $B'=B-\mu_0H$; $M=B'$; $I=B'/4\pi$.

* Received by the Institute, January 4, 1952.

[1] W. G. Cady, "Piezoelectricity," McGraw-Hill Book Co., New York, N. Y., p. 754; 1946.

[2] "Standards on piezoelectric crystals, 1949," Proc. I.R.E., vol. 37, pp. 1378–1395; December, 1949.

[3] W. P. Mason, "Electrostrictive effect in barium titanate ceramics," *Phys. Rev.*, vol. 74, 1134–1147; November 1, 1948.

[4] D.R.C. Report, div. 6, vol. 13, chapt. IV.

A Method for Studying Sporadic-*E* Clouds at a Distance*

It seems worthwhile at this time to call attention to the possibilities of an observing technique which enables continuous study at a distance of the growth, motion, and disappearance of sporadic-*E* clouds of ionization.

The technique depends on the fact that radio waves incident obliquely on the surface of the earth produce appreciable backscattering, whereas those incident obliquely on the ionosphere do not. It has now been conclusively shown[1,2] that when propagation is via the *F* layer the majority of backscattered energy returns from the earth's surface, and not from the *E* region of the ionosphere as had previously been believed. It follows that when propagation is via clouds of sporadic-*E* ionization, similar backscattering occurs.

Ample experimental evidence exists to show that the amount of backscattering from the surface of the sea is not noticeably different from that obtained from land.

To track sporadic-*E* clouds by means of backscatter, a pulsed high-frequency transmitter and receiver are connected to a rotatable directive antenna. Received echoes are displayed on a plan position indicator. For best sensitivity, the longest pulse length consistent with the desired minimum range discrimination should be used.[2]

The frequency of operation may be made high enough so that regular-layer transmission to any point on the earth's surface is not supported. Under these conditions, in the absence of sporadic-*E* clouds within a radius of roughly 1,200 km about the station, no echoes will appear on the indicator. A cloud of adequate size and ion density appearing within this radius will make possible transmission to and from some point on the earth beyond the cloud, and the backscattered energy will cause an echo to appear. Echo direction is the direction to the cloud, and echo time delay is roughly proportional to twice the slant range to the cloud.

An example of the possibilities of the method is shown in Fig. 1. On the morning of December 15, 1951, a sporadic-*E* cloud drifting roughly from north to south was tracked as it passed over Stanford University. The PPI photographs were made with a scatter-sounding apparatus operating at approximately 14 mc. The antenna was a standard three-element Yagi, to which was fed a radio-frequency peak power of the order of 500 watts. The h'f records are those made by a model C-3 ionosphere recorder operated at Stanford University for the Central Radio Propagation Laboratory of the National Bureau of Standards.

* Received by the Institute, February 14, 1952.

[1] W. Dieminger, "The scattering of radio waves," *Proc. Phys. Soc.* (London), vol. 64, pp. 142–158; February, 1951.

[2] A. M. Peterson, "The mechanism of *F*-layer propagated back-scatter echoes," *Jour. Geophys. Res.*, vol. 56, pp. 221–237; June, 1951.

Correspondence

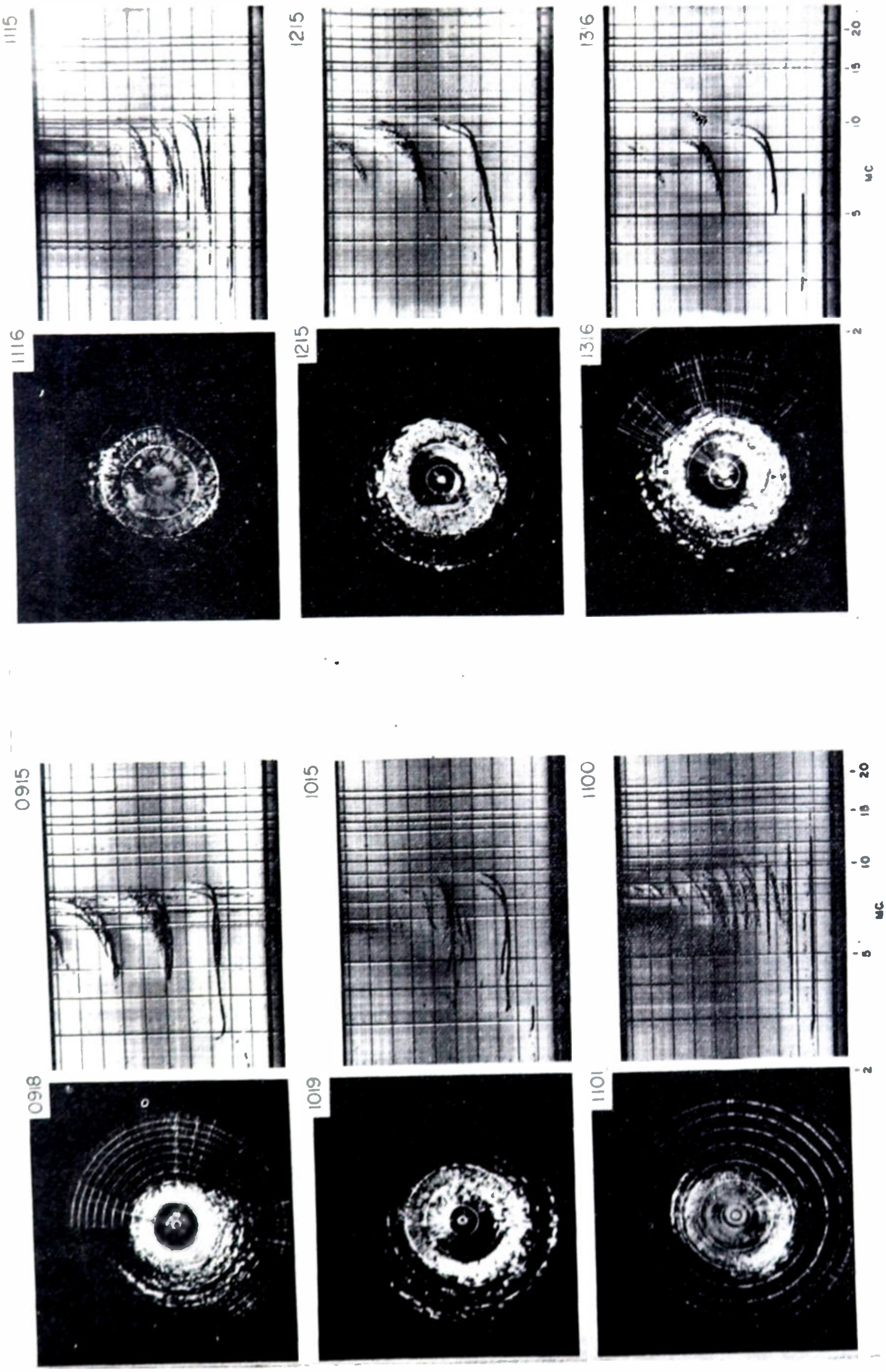


Fig. 1—Example of apparently moving sporadic-*E* observed simultaneously on scatter sounding and virtual-height-versus-frequency records at Stanford University, Calif., December 15, 1951. (Times are local.) Horizontal lines in the h'q records are 100-km Markers.

Correspondence

The first PPI record, at 0918, shows normal conditions for the frequency and time of day. The circles represent 500-km range marks. The ring of echoes just outside the 1,000-km range circle is formed by backscatter propagated by the *F*-layer. Between 500- and 1,000-km range circles in the 1019 record (whose scale has been expanded), the approaching sporadic-*E* cloud has produced echoes whose heaviest concentration is to the north and northeast. A small amount of sporadic-*E* is beginning to appear on the h'f record. By 1101 the ion cloud appears to be directly overhead, and the h'f record shows weak vertical-incidence reflections all the way up to 19 mc. By 1611 the scatter record shows a noticeable hole to the north, and by 1613 it is seen that the cloud has moved on to the south.

If the assumption is made that the ion cloud had the same intensity during the entire time it was tracked, the average rate of drift from north to south is estimated to be 175 km per hour. This drift velocity is on the high side, but is not inconsistent with values obtained at Stanford University by the meteor trail-drift method.[3] The north-south direction, it may be added, was one of those found to be favored in the meteor studies.

This particular series of records is shown because passage of the cloud overhead enabled direct comparison with the h'f measurements. Many sporadic-*E* clouds have been tracked when overlapping *F*-layer echoes were absent.

No novelty is claimed for the employment of rotating antennas and range-azimuth display in locating sporadic *E* since the technique appears to have been used first in Japan during the war.[4] Later research has, however, both confirmed the fundamental soundness of the method and demonstrated the simplicity of the equipment required.[5]

Although much more remains to be learned, it is clear that the scatter-sounding technique has important possibilities for study of the temporal and geographical characteristics of sporadic *E*.

ACKNOWLEDGMENT

These remarks are based on work supported by the Office of Naval Research, the Army Signal Corps, and the Air Force.

O. G. VILLARD, JR.
A. M. PETERSON
L. A. MANNING
Radio Propagation Laboratory
Stanford University
Stanford, Calif.

[3] L. A. Manning, O. G. Villard, Jr. and A. M. Peterson, "Meteoric echo study of upper atmosphere winds," PROC. I.R.E., vol. 38, pp. 877–883; August, 1950.

[4] T. Kono, "Experimental Study of Scattered Echoes," Report of ionosphere research in Japan, Maruzen Publishing Co., Tokyo, vol. IV; 1950.

[5] O. G. Villard, Jr. and A. M. Peterson, "Instantaneous prediction of radio transmission paths," *QST*, vol. 36, pp. 11–20; March, 1952.

The Spreading of an Electron Beam*

The effect of conducting walls on the spreading of an electron beam has received controversial treatment in the literature. In general, two approaches are used, which lead to opposite conclusions. To state the problem: Does an electron beam spread more when traveling "in free space," or when traveling in a region symmetrically bounded by conducting walls?

METHOD I

It is assumed that an electron beam of initially parallel electrons is traveling in "free space" with a velocity corresponding to beam voltage V_0 and with charge density ρ_0. The acceleration and displacement of an edge electron are easily calculated by applying Gauss's Law to the beam.[1,2] In extending the thinking to the bounded case, it is customary to perform a classical calculation for an electron beam traveling close to a single conducting surface and show that the image charges cause extra deflection.[3] From this the inference is that in the general case, image charges in walls increase beam spread.

METHOD II

An electron beam is injected into a region the bounding walls of which are maintained at V_0. Because of space-charge depression of potential, the edge electrons have a smaller potential V_m. The electric field thus set up causes beam spread which can be calculated.[4] It is found that beam spread is decreased by bringing the walls closer to the beam. This is due both to decreased transverse acceleration and increased forward velocity.

DISCUSSION

In view of these conflicting results, what answer may be given to the initial question? The methods are correct only insofar as their assumptions are consistent and approach the practical case. For instance, electron beams in most tubes flow through symmetrically placed conducting cylinders or plates. In such cases, image charges are induced symmetrically, and hence exert no integrated force on electrons in the beam. Hence, the argument concluding Method I must be discarded.

Another tacit assumption of Method I should be brought out. The electron beam could not be in free space and still maintain a potential V_0. Such an assumption holds true only if walls of potential V_0 hug the beam. Indeed, it is found that the result of Method I agrees with the result of Method II for the case in which the beam completely fills the bounded region.

Hence it is correct to conclude that the presence of symmetrically placed walls decreases beam spread. It is interesting to note that the same conclusion is valid for rf debunching, both transverse and longitudinal.

J. S. HICKEY, JR.
T. G. MIHRAN
Electron-Tube Section
General Electric Company
Schenectady, N. Y.

* Received by the Institute, November 9, 1951.

[1] K. R. Spangenberg, "Vacuum Tubes," McGraw-Hill Book Co., New York, N. Y., p. 440; 1948.

[2] D. R. Hamilton, *et al.*, "Klystrons and Microwave Triodes," McGraw-Hill Book Co., New York, N. Y., p. 209; 1948.

[3] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, New York, N. Y., p. 145; 1945.

[4] D. P. R. Petrie, "The effect of space charge on the potential and electron paths of electron beams," *Elec. Commun.*, vol. 20, pp. 100–111; 1941.

Super Directivity with Directional Coupler Arrays*

Recently Bolinder,[1] in two letters to the PROCEEDINGS, clearly stated the basis for the similarity between the theory of antennas and directional coupler arrays. The wording of these letters, however, is such as to give the impression that there exists a fundamental relationship between bandwidth and directivity which ultimately limits the directivity possible for a given bandwidth. Although such a condition may hold for a sufficiently generalized definition of bandwidth, it can be shown, by example, that it is not true for many cases of practical interest. In this respect, the situation is similar to that encountered with antenna arrays.[2]

If one considers an admittance diagram in the vicinity of unity admittance, it will be observed that the lines of constant conductance and constant susceptance form a square grid similar to a Cartesian co-ordinate system. It is an immediate consequence that the admittance transformation introduced by a length, l, of line is approximately $Y_s - 1 = (Y_0 - 1)e^{2j\theta}$, where Y_0 is the terminating admittance, Y_s is the input admittance, and $\theta = 2\pi l/\lambda g$. λg is, of course, the guide wavelength. Since $Y_s - 1$ and $Y_0 - 1$ are essentially reflection coefficients, the accuracy of the transformation is apparent and one sees how, by integration, an integral representation of the input reflection coefficient might be obtained.

The input reflection for an array of n symmetrical, equally spaced coupling elements, each one of shunt susceptance Y_n, is

$$Y_s - 1 = jY_n x^{n-1} + \cdots + jY_1,$$

where $x = e^{+2j\theta}$, n is odd, and $Y_n = Y_1$, and so on. Putting $u = \cos 2\theta$, one may follow Dolph[3] and write

$$\begin{aligned} Y_s - 1 &= P(u) \\ &= 2Y_1 \cos\left(\frac{n-1}{2}\right)\theta + \cdots, \end{aligned}$$

where the variable u ranges from u_{min} to

* Received by the Institute, January 24, 1952.

[1] F. Bolinder, "Fourier transforms in the theory of inhomogeneous transmission lines," PROC. I.R.E., p. 1354; November, 1950; "Approximate theory of the directional coupler," PROC. I.R.E., p. 291; March, 1951.

[2] H. J. Riblet, "Discussion on a current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," PROC. I.R.E., vol. 35, pp. 489–492; May 1947.

[3] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," PROC. I.R.E., pp. 335–348; June, 1946

Correspondence

u_{max} as λg and θ cover the values prescribed by the frequency range. The voltage directivity is $|P(1)|/|P(u)|$. For fixed bandwidth, n and l, the optimum directivity is obtained by selecting Y_i so that

$$P(u) = T_{\frac{n-1}{2}}\left\{K\left(u - \frac{u_{max}+u_{min}}{2}\right)\right\},$$

where $T_{\frac{n-1}{2}}$ is the Tchebyscheff polynomial of $n-1/2$ degree and K satisfies the relation $K(u_{max}-u_{min})=2$. For a fixed length array, nl=constant and the directivity, over the given frequency band, can be increased without limit by increasing n. This follows from the fact that in evaluating $P(1)$ the order of the Tchebyscheff polynomial increases without limit while its argument approaches a finite value depending only on the guide wavelengths at the edge of the frequency band.

The simplifying assumption, which is the basis of this letter, does not appear to limit the available directivity. Essentially, the Tchebyscheff performance depends on the number and spacing of the zeros of Y_i-1 in the frequency band of interest. From the Hurwitz theorem,[4] it may be expected that the well-behaved limiting process involved in the simplifying assumption does not affect the roots of the function Y_i-1 in a discontinuous or unusual manner.

It is interesting to observe that the designer of super directive arrays is not faced with the formidable problems posed by high current densities and high Q circuits[5] which confront the ambitious antenna designer. Fabrication tolerances,[6] the frequency behavior of the coupling apertures, and interaction would appear as the principal factors limiting the directivity of this type of directional coupler.

HENRY J. RIBLET
Microwave Development Laboratories
220 Grove St.
Waltham, Mass.

[4] M. Marden, "The Geometry of the Zeros," Mathematical Surveys No. III, American Mathematical Society, New York, N. Y., p. 4; 1949.
[5] L. J. Chu, "The physical limitations of directive radiating systems," *Jour. Appl. Phys.*, p. 1163; December, 1948.
[6] N. Yaru, "A note on super-gain arrays," PROC. I.R.E., pp. 1081–1085; September, 1951.

Telepathic Communication*

I am very much impressed by the ever-increasing correspondence concerning telepathic communication. This points to the urgency for extensive research in this new field opened by Professor Rhine and his school. The parapsychologists have pointed the way to the connection between physics and human psychology, in particular, extrasensory perception and teleradiasthesia. As a basis for any constructive work and further progress, we must be above any prejudice toward a study that is considered nonscientific or even occult. Before too much concern be shown over the financing of such a project, it is of great importance that the objectives be clearly delineated. In order to avoid confusion, I would like to make some corrective remarks in reply to Harry Stockman's letter.[1]

As a long-standing friend of Manfred Baron von Ardenne, I had the pleasure of being familiar with his studies and findings; I also collaborated with him on many significant projects. In all our discussions my friend never claimed any success in proving the presence of brain waves, in spite of extensive and astute experiments, as he pointed out in his paper published in 1928.[2] In his new book "Psychical Physics,"[3] Dr. Tromp confirms this statement. We are dealing here exclusively with electromagnetic radiation and not with skin potentials and bioelectric action currents which are the basis not only of electrocardiography and encephalography, but also of the Heydweiller-Schumann effect.[4]

In this light, the unique claims of Cazzamalli must be restated. On the one hand we have the negative results of von Ardenne, and on the other, Cazzamalli's oscillograms. The only possibility of getting around this contradiction is to attribute Cazzamalli's oscillograms not to brain waves, but to some other rf phenomenon: My own studies of numerous Cazzamalli papers lead me to the following hypothesis:

The most decisive element of Cazzamalli's apparatus seems to be the Faraday cage. Cazzamalli himself claims to use it merely as a shielding, whereas I consider its function as a cubical cavity resonator. Cazzamalli's oscillator excites the Faraday cage into an unknown mode of resonance which is modulated by the biochemical action of the brain under emotional excitation. Absorption or detuning causes AM or FM-AM, which controls the output of a receiver, and thus produces the characteristic oscillograms.

At a glance, this hypothesis would seem to be somewhat strange because I can neither verify nor corroborate it on the basis of Cazzamalli's nonprofessional description. We must keep in mind that Cazzamalli wrote his papers for electromagnetic radiation, which does not apply to this new conception. In any case, the question arises as to whether Cazzamalli needs an oscillator at all, because true brain waves should be powerful enough to actuate a sensitive receiver.

Moreover, I want to point out that the "bioelectric modulation" is not new. A prototype is the "Dielectrograph"[5,6] which, in the field of electrocardiography, operates as follows: The chest of the patient forms the dielectric of a capacitor which, in turn, is part of the tank circuit of an uhf oscillator. The rhythmic contractions of the heart vary the capacity, and the resulting FM is converted into output currents feeding an oscilloscope. The bioelectric action potentials of the heart are accompanied by dielectric displacements and fluctuations in the concentration of ions in the heart cells. If the analogy between electrocardiography and encephalography is kept in mind, the principle of the dielectrograph may easily be applied to Cazzamalli's apparatus so that it may be called an "Encephalic Dielectrograph" or a "Dielectric Encephalograph."

As long as the meager description of Cazzamalli's circuitry does not permit a verification of the new hypothesis and as long as no other corroboration of actual brain waves has been published, any further discussion is of little avail. On the other hand, a thorough repetition of Cazzamalli's tests would be most valuable for scientific clarification, regardless of whether brain waves would be proved, or whether the dielectric or any similar effect may be found.

HANS E. HOLLMANN
105 North "G" Street
Oxnard, Calif.

* Received by the Institute, February 28, 1952.

[1] H. Stockmann, "More on telepathic communication," PROC. I.R.E., vol. 39, p. 1571; December, 1951.
[2] M. von Ardenne, "Über elektrische Felder in der Umgebung lebender Wesen" (Electric Fields around Living Beings), *Zeit. für tech. Phys.*, vol. 9, p. 288; 1928.
[3] S. W. Tromp, "Psychical Physics," Elsevier Publishing Co., New York, N. Y. and Holland, p. 179; 1949.
[4] *Ibid.*, p. 178.
[5] E. Atzler and G. Lehmann, "Zs. für Arbeitsphysiology," Junk, vol. 5, p. 636; 1932.
[6] H. E. Hollmann, "Physik and Technik der ultrakurzen Wellen" (Physics and Technique of VHF), J. Springer, Berlin, vol. II, p. 181; 1936.

The Sweep-Frequency Response of RG-6/U*

In a recent paper[1] Alsberg reproduces curves showing the input impedance of a length of 158 feet of RG-6/U cable measured over a frequency range 0.05–20 mc. These curves show marked fluctuations which the author attributes to "the frequency dependence and irregularity of the characteristic impedance of RG-6/U cable, a common fault of all flexible coaxial cables."

While it is agreed that many cables exhibit these faults, especially when tested at 3,000 mc,[2] it is suggested that the main fluctuations in this cable are independent of irregularities in the cable construction, often referred to by the term "periodicity."

The main features to be accounted for are a succession of peaks and troughs in both resistance and reactance curves, and a general rise of resistance with decrease in frequency.

First considering the peaks and troughs, it will be seen that they are roughly equally spaced in the frequency scale. The average spacing of the maxima and minima for resistance and reactance curves have been extracted from the resonant frequencies by the Gauss method. The four average spacings differ by less than 1 per cent, the mean value being 2.10 mc. The most probable cause of a series of equispaced peaks would be a mismatch at the far end of the cable. Since the cable is 158 feet long and can be assumed to have a velocity ratio of about 0.66, such resonances would be expected at intervals of 2.08 mc. The agreement be-

* Original manuscript received by the Institute, April 10, 1952.

[1] D. A. Alsberg, "A precise sweep-frequency method of vector impedance measurement," PROC. I.R.E., vol. 49, pp. 1393–1400; November, 1951.
[2] W. T. Blackband and D. R. Brown, "The two-point method of measuring characteristic impedance and attenuation of cables at 3,000 mc," *Jour. IEE* (London), vol. 93, pt. IIIA, pp. 1383–1386; September, 1946.

Correspondence

tween this and the measured spacing is such as to leave little doubt that the peaks are due to a slight mismatch at the receiving end of the cable. It is not easy to estimate the amplitude of the fluctuations, but the mismatch would be of the order of 2–3 ohms.

The second feature, the rise in resistance with decrease in frequency, is a result of the rise in characteristic impedance at the lower frequencies. This rise is due to the increase in the "internal inductance" of the cable as the flux linkage within the conductors increases as the depth of penetration becomes greater. For a cable of the size of RG-6/U the characteristic impedance at 1 mc would be 2–3 ohms greater than the nominal 75 ohms at higher frequencies. This increase would account for the observed slope of the resistance curve.

The residual irregularities, such as the nonuniformity of peak-trough amplitude, may be attributed to irregularities in the cable, to experimental error, and so on.

The foregoing interpretation of the curves is a further illustration of the wealth of data provided by sweep-frequency impedance measurements.

ACKNOWLEDGMENT

This letter is published by permission of the Chief Scientist, British Ministry of Supply and reproduced by permission of the Controller, H.B.M. Stationery Office, England. The British Crown copyright is reserved.

W. T. BLACKBAND
Royal Aircraft Establishment
Farnborough, Hants.
England

Automatic Frequency Control*

Weaver and his coworkers have recently made a mathematical study of the servomechanism needed for automatic frequency control.[1] The formulation of the problem will be reviewed briefly:

The control frequency, f, is mixed in a converter with the slaved frequency, f_s. The output of the converter is passed by an RC filter with a time constant T_1 to the AFC circuit which generates f_s. Let

$$A \cos 2\pi z = A \cos 2\pi \int^{t} (f - f_s) dt \quad (1)$$

designate this converter output. It is related to the RC filter output, x, by

$$A \cos 2\pi z = T_1 x' + x. \quad (2)$$

If we neglect the reference frequency about which both f and f_s must vary in a practical system, the servo action of the AFC circuit can be defined by the simple relation

$$f_s = \frac{x}{AT_2}. \quad (3)$$

Relations (1) to (3) can be combined to yield the servo equation

$$T_1 z'' + z' + \frac{\cos 2\pi z}{T_2} = T_1 f' + f. \quad (4)$$

* Received by the Institute, November 5, 1951.
[1] Final report of SRI Proj. 257, Air Materiel Command, the part pertinent to this subject to be published shortly in the PROC. I.R.E.

The introduction of the dimensionless parameter, $u = t/\sqrt{T_1 T_2}$, normalizes (4)

$$\ddot{z} + 2a\dot{z} + \cos 2\pi z = b, \quad (5)$$

where

$$a = \frac{1}{2}\sqrt{\frac{T_2}{T_1}}$$

and

$$b = T_1 T_2 f' + T_2 f.$$

In the study referred to above, f was assumed to be constant, as it will be in what follows, equation (5) was visualized as that of a viscously restrained pendulum submitted to a constant torque, and an approximate relation between a and b was derived to mark the boundary between periodic and aperiodic (i.e., frequency-capturing) solutions, when $b < 1$.

Equation (5) can also be visualized as that of a viscously restrained steel ball traveling across a sinusoidally corrugated sheet, or coming to rest in one of the troughs, depending upon the inclination of the sheet and viscous resistance encountered.

It will be noted that the damping coefficient a is proportional to the converter output, which is proportional to the product of the amplitudes of the control and slaved signals of frequencies f and f_s respectively. If the control frequency is received after transmission over a variable path, an elaborate AVC circuit will be required to insure constancy of a. On the other hand, little or no AVC will be required if the control signal is clipped and fed to the converter as a square wave. If the slaved signal is also clipped, the output of the converter as a function of the phase will be a series of ascending and descending straight lines with a rigidly determined slope. This slope can be assigned the value unity, and the function itself can be limited between $-\frac{1}{4}$ and $+\frac{1}{4}$, without loss of generality. The formal expression for this new function of z is

$$C(z) = \frac{2}{\pi^2}\left(\cos 2\pi z + \frac{1}{9}\cos 6\pi z + \frac{1}{25}\cos 10\pi z + \cdots\right).$$

The servo equation of this modified system is

$$\ddot{z} + 2a\dot{z} + C(z) = b. \quad (6)$$

The validity of the corrugated sheet model can be extended to this case by substituting a series of parabolic segments for the former sine curve.

A rigorous mathematical treatment can be had for (6), which is more complicated in appearance than (5), but which corresponds to a system for which a plea of circuit realism can be made. The calculations needed are straightforward and will be sketched.

When $b > \frac{1}{4}$ no rest position exists for the system, whereas when $b < \frac{1}{4}$, that is, when the capturing frequency is a fraction $4b$ of the maximum holding frequency, the corresponding damping coefficient a required for bare capture under all initial conditions can be calculated as follows: Let the system start with a bare forward speed from a labile position where $C(z) = b$ and $C'(z) = -1$. Its velocity will be $\dot{z} = \epsilon \exp(-a + \sqrt{1 + a^2})u$, and after traveling a distance $\frac{1}{4} + b$ to a minimum of the $C(z)$ curve its velocity will be $(\frac{1}{4} + b)(-a + \sqrt{1 + a^2})$.

Setting now $u = 0$ and $z = -\frac{1}{4}$ at this event (purely for mathematical convenience), the kinematics of the system along the just reached ascending slope of the $C(z)$ curve will be given by

$$z = -(\tfrac{1}{4} + b)e^{-au}\cos\sqrt{1 - a^2}\,u + \frac{\frac{1}{4} + b}{\sqrt{1 - a^2}}(-2a + \sqrt{1 + a^2}) \cdot e^{-au}\sin\sqrt{1 - a^2}\,u + b,$$

and

$$\dot{z} = (\tfrac{1}{4} + b)(-a + \sqrt{1 + a^2})e^{-au}\cos\sqrt{1 - a^2}\,u + \frac{\frac{1}{4} + b}{\sqrt{1 - a^2}}(a^2 + 1 - \sqrt{1 + a^2}) \cdot e^{-au}\sin\sqrt{1 - a^2}\,u,$$

which can be verified to satisfy the conditions $z = -\frac{1}{4}$ and $\dot{z} = (\frac{1}{4} + b)(-a + \sqrt{1 + a^2})$ at $u = 0$.

If the stipulation is made that after traveling a certain time the system should reach the point $z = \frac{1}{4}$ with a velocity $\dot{z} = (\frac{1}{4} - b)(a + \sqrt{1 + a^2})$, i.e., barely sufficient to reach the next point of labile equilibrium, two relations are obtained between u, a and b, from which two explicit expressions for u can be derived:

$$u = -\frac{1}{\sqrt{1 - a^2}}\arctan\frac{\sqrt{1 - a^4}}{a^2} = \frac{1}{a}\lg\left[\frac{1 + 4b}{1 - 4b}(-a + \sqrt{1 + a^2})\right]. \quad (7)$$

This relation yields in turn an explicit expression for b in terms of a. The observation that u must be positive and that the first arrival of the system at $z = \frac{1}{4}$ must be considered indicates that the arc tan must be comprised between $-\pi$ and $-\pi/2$.

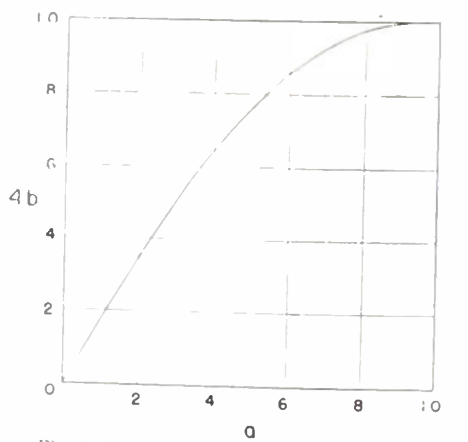


Fig. 1—Capturing to holding frequency ratio versus damping coefficient.

Figure 1 is a plot of the damping coefficient a needed for capture of the system by a frequency which is a fraction $4b$ of the maximum holding frequency. For small values of a and b we have, approximately

$$4b = \left(1 + \frac{\pi}{4}\right)a,$$

whereas for values of b which approach $\frac{1}{4}$ the system must approach critical damping.

MARCEL J. E. GOLAY
Signal Corps Engineering Laboratories
Fort Monmouth, N. J.

Correspondence

Comment on Fink's Article in the Television Symposium*

Of particular interest to me was Fink's article in the television symposium published in the PROCEEDINGS,[1] which called attention to the credit due Gray for his invention of frequency interlace systems many years ago. Of course, Gray's invention had not been forgotten in the Bell System. In fact, I am informed that this invention has been used for some years in the transmission of pilot signals along with television signals for regulating amplifiers interposed in the coaxial-cable transmission systems.

For future reference I would like to take this occasion to mention another expired patent on a Bell Telephone Laboratories invention, which is also not forgotten, and which also covers an invention of interest in recent years. I refer to the Hartley patent No. 1,666,206, which covers the so-called "phase-rotation" system for generating a single-sideband communication signal. Indeed, the various names given to that system are so imperfectly descriptive that perhaps it would be less ambiguous to refer to it as "the Hartley single-sideband modulation system." Hartley's patent was applied for January 15, 1925 and was issued April 17, 1928.

WILLIAM R. WOODWARD
Counselor at Law
195 Broadway
New York, N. Y.

* Received by the Institute, March 31, 1952.

[1] D. G. Fink, "Alternative approaches to color television," PROC. I.R.E., vol. 39, pp. 1124–1134; October, 1951.

Contributors to Proceedings of the I.R.E.

JAMES S. AJIOKA

James S. Ajioka (S'49) was born at Rexburg, Idaho, on August 9, 1923. He graduated from the Roosevelt High School in Los Angeles and received his early college education at the University of California at Los Angeles. His education was then interrupted by World War II, and he served with the 442nd Army Division in southern Europe.

At the close of the war, Mr. Ajioka entered the University of Utah, where he received the B.S. degree in 1949 and the M.S. degree in 1950. Since graduation, he has been employed in the U. S. Navy Electronics Laboratory at San Diego, working on microwave antenna design and development.

Mr. Ajioka is a member of Tau Beta Pi.

❖

HAROLD S. BLACK

Harold S. Black (M'42–SM'43–F'48) was born in Leominster, Mass. on April 14, 1898. He received the B.S. degree in electrical engineering from Worcester Polytechnic Institute in 1921, and at once joined what is now Bell Telephone Laboratories. In 1925 he was placed in charge of a group developing repeaters, regulators, filters, and other circuits for carrier telephone systems. In connection with his carrier research, Mr. Black invented the stabilized feedback amplifier, which has come into general use not only with carrier systems but with radio broadcasting and other electronic and communication fields both here and abroad.

In 1941 the Franklin Institute awarded the John Price Wetherill medal to Mr Black for his technical contribution to the modern efficiency of long-distance telephony, particularly for his development of the negative feedback amplifier.

From 1942 to 1947 Mr. Black was concerned almost exclusively with war developments, and in 1946 was awarded a Certificate of Appreciation from the War Department in official recognition of assistance. He also initiated and made important contributions to the early development of pulse-code-modulation and the application of pulse techniques to radio-communication systems. He has recently been investigating the properties of laminated transmission lines.

Mr. Black is a Fellow of A.I.E.E., and a member of Tau Beta Pi, Sigma Xi and the A.A.A.S.

❖

R. C. BOOTON, JR.

R. C. Booton, Jr. (S'48–A'49) was born on July 26, 1926, in Dallas, Texas. After attending public schools in Dallas and San Antonio, he completed in 1943–1944 the freshman and sophomore years at Texas Agricultural and Mechanical College.

From January, 1945 to August, 1946, Mr. Booton served in the United States Navy. This time was spent first as a student and then as an instructor in the airborne radar maintenance training schools. In September, 1945 he returned to Texas A. and M., where he completed the requirements for the B.S. degree in electrical engineering in 1948. He received the M.S. degree in mathematics from the same institution in 1948, and then accepted an appointment as a research assistant at M.I.T.

Since that time Mr. Booton has continued his graduate studies at M.I.T. and served as a staff member at the Dynamic Analysis and Control Laboratory, where he now heads the Analysis and Evaluation Division. His work at the D.A.C.L. has been concerned both with the operation and evaluation of the M.I.T. flight simulator and with analytic and simulator studies of various control-system problems.

Mr. Booton is a member of both the Sigma Xi and the Tau Beta Pi societies.

❖

T. J. BUCHANAN

Thomas J. Buchanan was born in 1913 in Clydebank, Scotland. He received the M.A. degree in physics and mathematics from Glasgow University in 1936 and remained there to do post-graduate research in electron diffraction.

In 1939 he joined the staff of H. M. Signal School, now known as the Admiralty Signal and Radar Establishment, where he worked on radar development until 1947. Mr. Buchanan then joined the staff of the physics department of the Middlesex Hospital Medical School to investigate possible medical and biological applications of microwaves.

❖

Ernest Buehler was born on April 12, 1913 in East Rutherford, N. J. He attended Brooklyn Polytechnic Institute and Newark College of Engineering. Before joining Bell Telephone Laboratories, he was associated with David Kahn and Co., N. Bergen, N. J.

Mr. Buehler joined the Laboratories in 1930 as a shop apprentice. In 1934 he became an instrument maker, and in 1941 a shop supervisor, making cost analysis,

Contributors to Proceedings of the I.R.E.

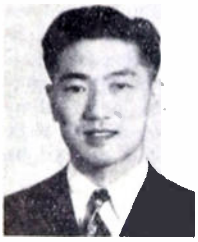
planning and proportioning the work, and getting experimental models of equipment made.

Ernest Buehler

Mr. Buehler is a member of the chemical research department at Bell Telephone Laboratories. For the last seven years he and his associates have been growing experimental crystals and studying the effects of heat treatment, growth rate, and mixing upon their physical and electrical properties, as well as the segregation characteristics of various impurities within the crystals.

❖

David K. Cheng (S'44–A'48–M'48–SM'50) was born on January 10, 1918, in Kiangsu, China. He received a B.S. degree in 1938 from the Chiao-Tung University, China, where he majored in electrical engineering. From 1938 to 1942 he was an engineer with the Central Radio Corporation, in China. In 1943 he came to this country to do graduate study at Harvard University, receiving the S.M. degree in communications engineering in 1944 and the Sc.D. degree in 1946. While at Harvard he was a Charles Storrow Scholar and Gordon McKay Scholar.



David K. Cheng

From 1946 to 1948 Dr. Cheng was an electronics engineer at the Cambridge Field Station of the USAF, following which he became assistant professor of electrical engineering at the L. C. Smith College of Applied Science of Syracuse University. At present he is an associate professor at Syracuse.

Dr. Cheng is a member of the Society of the Sigma Xi and Eta Kappa Nu.

❖

L. A. DuBridge was born on September 21, 1901 in Terre Haute, Ind. He received the A.B. degree from Cornell College in 1922 and the Ph.D. degree from the University of Wisconsin in 1926, both in the field of physics.

From 1925 to 1926 Dr. DuBridge was an instructor in physics at the University of Wisconsin and from 1926 to 1928 he was a Fellow of the National Research Council at the California Institute of Technology. During the next six years he was an assistant professor and an associate professor of physics at Washington University. From 1934 to 1946 he was affiliated with the University of Rochester as professor of physics and chairman of the department. On leave of absence from 1940 to 1945, he served as Director of the Radiation Laboratory at M.I.T., under the National Defense Research Committee. Since 1946 Dr. DuBridge has been President of the California Institute of Technology.

L. A. DuBridge

Dr. BuBridge is a fellow of the American Physical Society and a member of the American Philosophical Society and the National Academy of Sciences. He is also a member of the General Advisory Committee of the Atomic Energy Commission, the National Science Foundation Board, the President's Communications Policy Board, the Board of Trustees of the Carnegie Endowment for International Peace, and the Board of Trustees of the Rand Corporation, and is chairman of the Science Advisory Committee of the Office of Defense Mobilization.

❖

Robert C. Fletcher was born in New York, N. Y. on May 27, 1921. He received the B.S. degree in physics from the Massachusetts Institute of Technology in 1943. During the next three years he was engaged in electronic research at the MIT Radiation Laboratory, principally on magnetron oscillators. He then attended the graduate school in physics at MIT, having been awarded a National Research Council Predoctoral Fellowship.

R. C. Fletcher

In 1949, he was awarded the Ph.D. degree. Since then he has been doing research in the electron dynamics group at the Bell Telephone Laboratories Inc., Murray Hill, N. J.

❖

William F. Gabriel (S'45–A'46) was born in Sault Ste Marie, Mich., on October 17, 1925. He received the B.S. degree in electrical engineering from the University of Wisconsin in 1945. After one year of active duty in the U. S. Naval Reserve, he returned to the University of Wisconsin for graduate work in electrical engineering in the communications department, receiving the M.S. degree in 1948 and the Ph.D. degree in 1950.

W. F. Gabriel

He is now engaged in research and development work in the Antenna Research Branch at the Naval Research Laboratory in Washington, D. C.

Dr. Gabriel is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

Obed C. Haycock (A'40–SM'47) was born in Panguitch, Utah, on October 5, 1901. He received the B.S. degree in electrical engineering from the University of Utah in 1925, and the M.S. degree in electrical engineering from Purdue University in 1931.

Obed C. Haycock

After graduation, Professor Haycock joined the Westinghouse Electric and Manufacturing Company, and later returned to the University of Utah as instructor in electrical engineering. He has remained more or less permanently with the University since that time, and was promoted to the rank of professor in 1947. During the war he was given a leave of absence to do government research in Panama.

In addition to being professor of electrical engineering, he is now associate director of "Physics of the Upper Atmosphere," an Air Corps sponsored program connected with White Sands Proving Ground.

Professor Haycock is a member of A.I.E.E., Tau Beta Pi, Sigma Xi, and A.S.E.E.

❖

Archie P. King (A'27–SM'45) was born in Paris, France, on May 4, 1901. He received the B.S. degree from the California Institute of Technology in 1927. From 1927 to 1930 he was in the seismological research department of the Carnegie Institution of Washington, at the end of which time he joined the research department of the Bell Telephone Laboratories, Inc.

Archie P. King

During the interval to 1941, Mr. King was engaged in research on short-wave radio systems, waveguide techniques, and waveguide antennas. From 1941 to 1945 he was occupied with the development and performance studies of radar systems.

Since 1945 Mr. King has been engaged in

Contributors to Proceedings of the I.R.E.

microwave research at the Holmdel Laboratory of the Bell Telephone Laboratories.

He is a member of the American Physical Society.

❖

For a photograph and biography of Ronald W. P. King, see page 1462 of the November, 1951 issue of the PROCEEDINGS OF THE I.R.E.

❖

Charles O. Mallinckrodt (SM'51) was born in St. Louis, Mo. on May 28, 1907. In 1930 he was graduated from Washington University in St. Louis with a B.S. degree, and later that year joined Bell Telephone Laboratories, where in recent years he conducted research on high-frequency transmission methods. His previous work in the Laboratories included development of transmission regulators and feedback amplifiers for carrier telephony, pulse-modulation telephone systems which led to the development of the theory of instantaneous compandors; and transistors. He is now associated with Hughes Aircraft Company.


C. MALLINCKRODT

Mr. Mallinckrodt is a member of A.I.E.E.

❖

Jean Marique (M'46) was born in Brussels, Belgium, in 1900. He received the civil engineer degree from the University of Brussels in 1923 and that of radio engineer from the Ecole supérieure d'Electricité (Paris) in 1928.


JEAN MARIQUE

From 1925 to 1932, Mr. Marique was with the Belgian Civil Aviation Administration, and from 1932 to 1948 with Messrs. S.A.I.T. (Brussels), which is an exploiting company in the radio marine field. During the same period, he acted as chief of the scientific center of the Comité International Radio maritime (C.I.R.M.).

During 1946 he joined the Monitoring Centre for Marine and Aviation (C.C.R.M.) Brussels, with the function of Secretary General.

From 1930 until 1948, Mr. Marique has been in charge of the course on electron tubes at the University of Brussels and since 1948 on that of exploitation of radiocommunications.

Mr. Marique is vice-chairman of the Belgian Society of Engineers in Telecommunications and Electronics (SITEL). He attended a number of International Telecommunications conferences (Madrid, 1932; Cairo, 1938; Atlantic City, 1947), the Safety of Life at Sea Convention (1948), C.C.I.R. Meetings, and the like.

❖

Knox McIlwain (A'31–M'40–SM'43–F'48) was born on September 4, 1897, in Philadelphia, Pa. He received the B.S. degree in 1918 from Princeton University and the B.S.E.E. degree in 1921 and the E.E. degree in 1930, both from the University of Pennsylvania.


KNOX MCILWAIN

From 1921 to 1924 Mr. McIlwain was with the Bell Telephone Company of Pennsylvania Engineering Department. For the next sixteen years he was a professor at the Moore School of Electrical Engineering of the University of Pennsylvania. From 1940 to date he has been the Chief Consulting Engineer of the Hazeltine Electronics Corporation.

❖

S. P. Morgan, Jr. was born on July 14, 1923 in San Diego, Calif. He attended the California Institute of Technology, where he received the B.S. degree in 1943, the M.S. degree in 1944, and the Ph.D. degree in 1947.


S. P. MORGAN, JR.

Since 1947 Dr. Morgan has been a research mathematician with Bell Telephone Laboratories, specializing in electromagnetic theory. He has been particularly concerned with problems of waveguide and coaxial-cable transmission.

Dr. Morgan is a member of the American Physical Society, Tau Beta Pi, and is an associate member of Sigma Xi.

❖

P. B. Patton (A'46) was born on September 25, 1914, in Santa Rosa, Calif. After graduating from Polytechnic High School in San Francisco, he attended San Francisco State Teacher's College and the University of Maryland.


P. B. PATTON

Mr. Patton began his communication career with Western Union and remained with that company eleven years, after which he joined the engineering staff of the Federal Communications Commission in the common carrier branch. In 1946 he joined the Farnsworth Television and Radio Company and became Technical Co-ordinator of sales and engineering functions of the Mobile Communications Division. Associated with Lenkurt Electric since 1948, Mr. Patton now holds the position of Vice-President and Commercial Manager, and serves on the board of directors of that company.

Mr. Patton is a member of the Armed Forces Communication Association.

❖

E. G. Ramberg was born in Florence, Italy, on June 14, 1907. He received the B.A. degree with honors in physics at Cornell University in 1928 and the Ph.D. degree in theoretical physics at the University of Munich in 1932. From 1932 to 1935 he worked on the theory of X-ray spectra as research assistant at Cornell University.


E. G. RAMBERG

Since 1935, except for three years spent in Civilian Public Service (1943–45) and a summer term as visiting professor at the University of Munich (1949), Dr. Ramberg has been a staff member of the Electronic Research Laboratory, first of the RCA Manufacturing Company in Camden, N. J., then of the RCA Laboratories Division of the Radio Corporation of America in Princeton, N. J. Here he has been concerned principally with electron optics as applied in electron microscopy and with television, physical electronics, and light optics.

Dr. Ramberg is a member of the American Physical Society, the Electron Microscope Society of America, and Sigma Xi.

❖

Ronald E. Scott (S'43–A'50) was born in Leslie, Saskatchewan in 1921. In 1943 he received the B.A.Sc. degree from the University of Toronto, where he was president of the student council in 1943. From 1943 to 1945 he served as a radar officer in the Royal Canadian Naval Reserve. In June, 1946 he received an M.A.Sc. from the University of Toronto, and in February, 1950 he received an Sc.D. degree from the Massachusetts Institute of Technology.


RONALD E. SCOTT

Contributors to Proceedings of the I.R.E.

From 1946 until the present Dr. Scott has been on the staff at the Massachusetts Institute of Technology as a research assistant and research associate in the Research Laboratory of Electronics, and finally as an assistant professor of electrical engineering. He has worked in the fields of network theory and analog computers. Dr. Scott is a member of Sigma Xi.

❖

Richard Silberstein (A'30–SM'47) was born in New York City on September 18, 1906. He was a shipboard radio operator in 1927, joining R.C.A. at Riverhead in 1929. In 1930 he received the E.E. degree from Columbia University, following which he returned briefly to R.C.A. In 1936 he formed the precision Inductance Corp., remaining as president until 1938, when he moved into the presidency of Microphone, Inc.

R. Silberstein

Mr. Silberstein became a radio engineer at the National Bureau of Standards in 1941 and is there at present. In 1945 and 1946 he edited *Basic Radio Propagation Predictions;* since then he has been assistant chief of the Ionospheric Research Section. Beginning in 1944 he performed research in radio propagation effects on high-frequency direction finders, and in 1947, high-frequency oblique-incidence pulse and backscatter reception.

Mr. Silberstein has been alternate secretary of the Committee on Standard Direction Finder Measurements, a subcommittee of the Committee on Electronic Navigation, since 1946. Beginning in 1950 he has also been active in various phases of CCIR work.

❖

Morgan Sparks was born on July 6, 1916 in Pagosa Springs, Col. He received a B.A. degree from Rice Institute in 1938 and an M.A. degree from the same institution two years later. He won a University of Illinois' Rockefeller Foundation Fellowship for 1940–1942, and in 1943 was awarded a doctorate by that university. While at Illinois, he worked for the National Defense Research Committee.

Morgan Sparks

Dr. Sparks joined the Bell Telephone Laboratories in 1943, where he worked on military projects in the electrochemical group. After World War II he conducted research on primary batteries, electrolytic capacitors, and rectifiers. For the past four years he has been engaged in research on solid-state physics problems related to the transistor.

Dr. Sparks is a member of A.C.S., A.P.S., Phi Beta Kappa, Sigma Xi, and Phi Lambda Upsilon.

❖

Constantin S. Szegho (A'41–SM'51–F'52) was born on March 15, 1905, in Hungary. In 1927 he received his diploma of electrical engineering from the Institute of Technology in Munich. He served as an assistant in the Electrical Engineering Department of the Institute of Technology in Aix-la-Chapelle, where he received the Doctor of Engineering degree in 1931.

C. S. Szegho

In 1933 Dr. Szegho joined Baird Television, Ltd. in London and was in charge of the Cathode-Ray Tube Research Department of that firm, which later became Cinema-Television, Ltd. In 1942 he joined the Rauland Corporation in Chicago, of which he is director of research.

During World War II Dr. Szegho served as a member of the vacuum tube development committee of O.S.R.D. He is a member of the Society of Motion Picture and Television Engineers and of the American Physical Society, and serves on the Cathode-Ray Tube Committee of the Joint Electron Tube Engineering Council.

❖

Gordon K. Teal was born in Dallas, Texas, on January 10, 1907. After receiving the A.B. degree from Baylor University in 1927, he went to Brown University as a Marston Scholar. From 1928 to 1929 he was a University Fellow and the following year a Metcalf Fellow. Brown University awarded him the Sc.M. degree in 1928 and a doctorate in 1931. From 1933 to 1935 he was an honorary research associate at Columbia University, where he studied heavy hydrogen.

Gordon K. Teal

Dr. Teal joined Bell Telephone Laboratories in 1930. For two year he was in the chemical department, where he studied the chemical properties of photoelectric materials. In 1932 he became a member of the television group, and was engaged in the research and development of photocells, electron multipliers, and camera tubes. From 1942 until the present time his major interest has been in microwave attenuators and semiconductors. He is also in charge of a group responsible for research and development of transistor and varistor materials.

Dr. Teal is a member of A.P.S. and Sigma Xi.

❖

B. R. Tupper (A'36–SM'46) was born on April 15, 1906, in Vancouver, B. C., Canada. He received the B.A.Sc. degree in 1928 at the University of British Columbia.

B. R. Tupper

After graduation, Mr. Tupper joined the Transmission Engineering Department of the British Columbia Telephone Company. In 1935 he was appointed radio engineer of the Northwest Telephone Company. From 1941 to 1946, he was programme engineer for the Pacific Communication Programme, which provided voice and teletype circuits by means of radio and wire lines in the remote areas of British Columbia for the Canadian Military Services. In 1951, Mr. Tupper became Manager and Chief Engineer of the Northwest Telephone Company.

Mr. Tupper is a member of the Association of Professional Engineers of British Columbia and of the Canadian Radio Technical Planning Board, and is past Chairman of the Vancouver Section of the IRE.

❖

An Wang (S'48–A'50) was born in Shanghai, China, on February 7, 1920. He received the B.S. degree in electrical engineering from Chiao-tung University, China, in 1940. He attended Harvard University from 1945 to 1948, receiving the M.S. degree in 1946 and the Ph.D. degree in 1948.

An Wang

From 1940 to 1941 Dr. Wang was an assistant of Chiao-tung University. The following four years he was an engineer of the Central Radio Corporation in China. From 1948 to 1951 he was a research fellow at Harvard University, where he worked on the development of basic components and systems of digital computing machines.

Since 1951 Dr. Wang has been active in the research and development of electronic systems and components with the Wang Laboratories, which he organized.

Dr. Wang is a member of Sigma Xi.

Institute News and Radio Notes

Calendar of COMING EVENTS

AIEE-IRE Telemetering Conference, Lafayette Hotel, Long Beach, Calif., August 26–27

1952 IRE Western Convention, Municipal Auditorium, Long Beach, Calif., August 26–29

Radar Weather Conference, McGill University, Montreal, Canada, September 15–17

Cedar Rapids IRE Technical Conference, Roosevelt Hotel, Cedar Rapids, Iowa, September 20

National Electronics Conference, Sherman Hotel, Chicago, Ill., September 29–October 1

57th Annual Convention, International Municipal Signal Assoc., Inc., Hotel Statler, Boston, Mass., September 29–October 2

IRE-RTMA Radio Fall Meeting, Syracuse, N. Y., October 20–22

Symposium on Microwave Circuitry, New York, N. Y., November 7

7th Midwest Conference, American Society for Quality Control, Claypool Hotel, Indianapolis, Ind., November 20–21

IRE-AIEE Computers Conference, Park Sheraton Hotel, New York, N. Y., December 10–12

IRE-AIEE Meeting on High Frequency Measurements, Washington, D. C., January 14–16

IRE Southwestern Conference and Electronics Show, Plaza Hotel, San Antonio, Tex. February 5–7

1953 IRE National Convention, Waldorf-Astoria Hotel, and Grand Central Palace, New York, N. Y., March 23–26

1953 National Conference on Airborne Electronics, Dayton, Ohio, May 11–13

FIRST CALL! AUTHORS FOR IRE NATIONAL CONVENTION!!

Lloyd T. DeVore, Chairman of the Technical Program Committee for the 1953 IRE National Convention, to be held March 23–26, requests that prospective authors submit the following information: (1) Name and address of author, (2) Title of paper, (3) A 100-word abstract and additional information up to 500 words (both in triplicate) to permit an accurate evaluation of the paper for inclusion in the Technical Program.

Please address all material to: Lloyd T. DeVore, c/o IRE Headquarters, 1 E. 79 Street, New York 21, N. Y.

The deadline for acceptance is November 17, 1952. Your prompt submission will be appreciated.

IRE Fort Wayne Section Honored

IRE Fort Wayne Section was guest of the radio engineering department at a recent banquet at Tri-State College, Angola, Ind. Speakers are: (left to right) L. F. Mayle, Magnavox; J. F. Conway, Chairman of IRE Fort Wayne Section; Gilbert Lawson, Westinghouse; Professor Leland Ax, Tri-State College; Dr. Warfield, Fort Wayne; Bruce Ratts, Station WOWO; and, R. B. Jones, Signal Department, Fort Wayne.

Cedar Rapids Conference Promises Notable Speakers

A registered attendance of 400 is expected at the two-day communications conference, to be sponsored by the Cedar Rapids IRE Section, September 19–20, Cedar Rapids, Iowa.

Outstanding authorities on the various phases of communication will be heard on the two-day program which will include exhibits and plant tours. Listed below are the subjects and the speakers to be presented during the meeting.

"Keynote Address"
A. A. Collins, Collins Radio Company,

"The Transmission of Intelligence in Typescript"
I. S. Coggeshall, Western Union Telegraph Company

"Long-Range Communication Trends"
M. G. Crosby, Crosby Laboratories

"Comparative Study of Modulation Methods"
R. M. Page, Naval Research Laboratory

"Design Trends in Communication Equipment"
L. Morgan Craft, Collins Radio Company

"The Voice of America in the Electronic War"
G. Q. Herrick, Broadcast Service, United States Department of State

"The Evolution of Communications"
L. V. Berkner, Associated Universities, Inc.

"IRE Region 5 Activities"
A. W. Graf, Director of Region 5

Southwestern IRE Conference Set for San Antonio

The 1953 Southwestern IRE Conference and Electronics Show will be held on February 5, 6, and 7, at the Plaza Hotel, San Antonio, Tex.

Authors are invited to submit papers in the general fields of television, microwave communications, new components, instrumentation, servomechanisms, audio, and electronic applications in medicine and geophysics. Two-hundred-word abstracts are solicited. They are to be sent to A. W. Straiton, Box F, Univ. Station, Austin, Tex., and must be received at that address by December 1, 1952.

Western Convention Starts One Day Earlier

The IRE Western Convention, originally scheduled to be held on August 27–29 at the Municipal Auditorium, Long Beach, California, will now start on Tuesday, August 26 and will continue through August 29. The added Tuesday sessions will be held at the Pacific Coast Club, Long Beach. The program for August 27–29 remains unchanged and will be held at the Municipal Auditorium, Long Beach, as previously announced.

The added program for Tuesday, August 26 will consist of a morning session, starting at 9:00 A.M., on Scanners and Antennas, and an afternoon session, starting at 2:00 P.M., on Electron Devices.

Institute News and Radio Notes

The proceedings of the joint IRE-AIEE Computer Conference, held in Philadelphia, December, 1951, are now available in sufficient quantity at IRE Headquarters, 1 E. 79 St., New York 21, N. Y. The price is $3.50 per copy.

Technical Committee Notes

The **Standards Committee** met May 8, M. W. Baldwin took the Chair in the absence of A. G. Jensen. After a further consideration of the Proposed Standards on Receivers: Definitions of Terms, The Committee then turned its attention to a report by W. J. Pease on "Present Status of ASA Committee Y10.14 Activities," with particular reference to the AIEE committee report on Proposed Symbols and Terms for Feedback-Control Systems. It was noted that the committee in its report had set forth the principles upon which the AIEE-ASA symbols were based and concluded that the principles are reasonable although not necessarily the only basis on which work could be done. Professor Pease added that ASA Y10.14 is now considering to turn its attention to terminology. The committee members would feel stronger in their objections if they could cite alternative symbols, which would be a function primarily of the Symbols Committee and, to a secondary degree, a function of the Servo-Systems Committee. After further discussion, it was resolved that formal request be made to ASA Committees Y10 and Y10.14, and to the AIEE Standards Committee that no action be taken in the form of approval of this report until the IRE Standards Committee has had time to review the matter in detail. The Committee then turned to a letter from R. R. Batcher, Secretary of ASA C16 on Radio, enclosing a letter ballot covering eleven RTMA standards. These RTMA standards have been circulated to interested committee chairmen and will be reviewed at the next Standards Committee meeting. The Committee also considered a request from ASA, for revision or reaffirmation of eight American standards. A list of these standards has been circulated to committee chairmen with a request that they indicate which standard or standards they would comment upon before the next meeting of the Standards Committee. The list of IRE Representatives on ASA Sectional Committees for the year 1952–1953 was approved with one change.

Under the Chairmanship of D. C. Ports, the **Antennas and Waveguides** Committee met on May 13. A resolution was passed that this Committee extend to A. G. Fox its fullest appreciation for his guidance and the excellent record the Committee has achieved during his two years as Chairman. The Committee turned to the reviewing of proposed definitions recommended to the Standards Committee by three technical committees. Consideration was given to the immediate work confronting the Committee which will be reviewing criticisms to be submitted in response to the circulation of the proposed definitions of Waveguide Terms among the various technical committees and finally to the Standards Committee. Future work of the Committee will include: (a) Revision of the 1948 Antenna Terms, and broadening their scope to include the recent developments in the field; (b) Definitions of terms relating to waveguide components; (c) Formulation of methods of testing waveguides and microwave components; (d) Preparation of a selected bibliography of microwaves and waveguides. It was decided to explore the possibility of establishing a subcommittee whose members would be drawn from the West Coast region. This subcommittee will have a large measure of autonomy and will be given as its initial charge, preparation of a selected bibliography of microwaves and waveguides. The existing Subcommittee (2.2), under the Chairmanship of P. H. Smith, will be charged with the task of preparing a list of waveguide components to be defined for consideration at an early meeting of this Committee; and, eventually, preparing preliminary definitions of these terms.

On May 22, the **Audio Techniques** Committee met under the Chairmanship of C. A. Cady. It was noted that this Committee has been inactive for many months and that a concerted effort will be necessary to restore the Committee to its former active state. Mr. Cady who is also Chairman of Subcommittee 3.5 submitted a report on the work of this group. Discussion by the Committee followed. H. W. Augustadt reported on the present status of Subcommittee 3.1 and agreed to continue as Chairman to implement the considerable volume of work awaiting action. A report dated September 24, 1952, "Audio Techniques Definitions," was submitted and members were requested to review this material and be prepared to discuss this document at the next meeting. Definitions on Noise, 51 IRE 7.6PS were reviewed. The objectives of Subcommittee 3.1, Definitions of Audio Systems and Components, was discussed in relation to other committees performing related functions. D. E. Maxwell reported on the status of his Subcommittee 3.2, and consented to remain as Chairman to expedite the completion of the work now in progress. The "Methods of Measurement for Single-Frequency Harmonic Distortion," may be completed and submitted to the Committee by the next meeting. The Committee membership was discussed, and it was recognized that further additions should be made with a view toward greater representation of active members in the Professional Group on Audio, and other Groups whose work paralleled the interests of the Committee.

The **Electron Devices** Committee met on May 23, under the Chairmanship of G. D. O'Neill. Chairman H. L. Thorson of the Subcommittee on Gas Tubes reported that the material on Radiation Counter Tubes is in the hands of the Editorial Department. Chairman Thorson also reported that a task group to devise standard methods of test for TR and ATR tubes was being formed. Chairman G. A. Espersen of Subcommittee 7.4.1 reported that methods of test for magnetrons were now being formulated and that he wished to renew a long-standing invitation to the task groups on klystrons and traveling-wave tubes to attend a joint meeting for the purpose of considering test methods common to the various types of microwave tubes. Chairman T. J. Henry of the Subcommittee on Small High-Vacuum Tubes reported in a general review of the Subcommittee's activity that the ad hoc committee for co-ordination of microwave-tube definitions had completed its work, and that the noise standards were ready for final review by the Standards Committee. E. M. Boone of Subcommittee 7.6.4 submitted the revised definitions for klystrons, traveling-wave tubes, and magnetrons with the recommendation that the ad hoc committee be dissolved. Discussion followed and a proposal by L. S. Nergaard was submitted proposing the formation of a Joint Subcommittee on High-Frequency Tubes, to draft Methods of Test for High-Frequency Tubes; the Subcommittee will report directly to the Chairman of the Electron Devices Committee. The proposed standards will be subject to review of interested subcommittees prior to submission to the Committee. This proposal was passed unanimously. G. A. Espersen was appointed Joint Subcommittee Chairman. Chairman Thorson of the Klystron Task Group reported that he is reactivating this group in preparation for considering methods of test. Most of the definitions have been submitted to the ad hoc co-ordinating group, but the remainder will be submitted for Committee approval. Chairman O'Neill appointed M. E. Hines, chairman of an ad hoc committee to make a critical review of the organization of the Electron Devices Committee. It was reported that R. M. Ryder was appointed Committee Vice Chairman.

On April 29, the **Information Theory and Modulation Systems** Committee met under the Chairmanship of W. G. Tuller. The meeting was called to consider comments on modulation systems definitions circulated.

Under the Chairmanship of R. A. Sykes the **Piezoelectric Crystals** Committee met on May 5. W. P. Mason's paper on, "Methods for Measuring Properties of Piezoelectric Crystals and Electrostrictive Ceramics," was reviewed in draft form with a view to early publication. New work of Hans Jaffe in this field was discussed. After consideration, the Committee recommended that Dr. Jaffe and Dr. Mason work toward the early publication of a joint paper incorporating the major results of both. Edward Gerber's paper on "Methods of Measuring the Constants of Piezoelectric Vibrators" was also discussed in some detail. Several minor comments and corrections were offered for inclusion in the final draft. With corrections made, the Committee endorsed the paper for publication at the earliest practical date. The brief paper of Dr. Ehrlich was discussed and should be published with the endorsement of the Committee at the earliest convenient time. The questions that arose concerning this paper will be resolved between Drs. Mason, Smith, Jaffe, Ehrlich, Baervald.

Professional Group News

Airborne Electronics

The Administrative Committee of the Airborne Electronics Group has approved the inclusion of all phases of navigation within the scope of the Group.

The Los Angeles Chapter of the Airborne Electricons Group has appointed G. M. Greene, Chairman of the Group, and the Baltimore Chapter has appointed C. E. McClellan, Chairman.

The Group will co-sponsor the 1953 IRE-IAS-RTCA-ION Symposium on Electronics in Aviation.

Antennas and Propagation

Officers and Administrative Committee Members for the Antennas and Propagation Group have been appointed as follows: A. H. Waynick, Chairman; D. C. Ports, Secretary-Treasurer; and H. G. Booker, J. B. Smythe, D. C. Ports, Administrative Committee.

Broadcast and Television Receivers

The Group on Broadcast and Television Receivers will hold sessions on uhf and color TV at the Syracuse Fall Meeting. The Administrative Committee of the Group also will meet at this time.

The first TRANSACTIONS of this Group have been mailed to all members who have paid their dues. The TRANSACTIONS consist of the round table discussion held at the IRE National Convention, and the following papers presented at the Convention: "Amplifiers for UHF Distribution Systems," by T. Murakami; "Practical TV Antennas for UHF," by E. O. Johnson and R. F. Kolar; "UHF Hybrid Ring Mixers," by Walter V. Tyminski and A. E. Hylas; "A VHF-UHF Television Turret Tuner," by M. G. Beier, J. F. Bell, A. Cotsworth, and J. F. White; "82 Channel Turret Tuner," by A. M. Scandurra; "The Design and Performance of a Compact UHF Tuner," by H. F. Rieth; "Comparison of Present-Day UHF and VHF Television Receivers," R. A. Varone.

Broadcast Transmission Systems

The Los Angeles Chapter of the Broadcast Transmission Systems Group will conduct a one-day symposium at the IRE West Coast Convention, Long Beach, Calif. Four papers and a panel discussion will be featured.

Papers to be heard are "Super Power AM Standard Broadcast Transmitters for the Voice of America," by J. O. Weldon, Continental Electronics Mfg. Co.; "Control Systems for Automatic Switching of Microwave Radio Channels," by Harold Pruden, Bell Telephone Labs.; "Compatible Color Television and the Broadcaster," by R. E. Shelby, NBC; "The Economics of TV Broadcasting," by Joseph Herold, RCA.

The subject of the panel discussion will be "What the End of the TV Freeze Means to the West." Serving on the panel will be A. E. Cullum, consulting engineer; Fred Albertson, radio-TV attorney; J. W. Kingsbury, A T & T; H. L. Hoffman, Hoffman Radio Corp; Rosel Hyde, FCC Commissioner.

Circuit Theory

The officers of the Los Angeles Chapter of the Circuit Theory Group have been elected as follows: Louis Weinberg, Chairman; J. J. Burke and N. H. Enenstein, Vice Chairman; J. A. Aseltine, Secretary-Treasurer; and W. R. Abbott, Convention Papers Chairman.

Communications

The recently formed Communications Group has chosen Professor C. A. Hachemeister to serve on the Group's Administrative Committee.

Electron Devices

Papers are being solicited for the Electron Devices Group's technical program at the Syracuse Fall Meeting. A drive also is now under way for new Group members.

Electronic Computers

The dates for the Joint Computer Conference on Input-Output Systems, sponsored by the Electronic Computers Group, is set for December 12–14, Park Sheraton Hotel, New York, N. Y.

Engineering Management

The Engineering Management Group is formulating plans for a symposium to be held during the National Electronics Conference, September 29–October 1, Sherman Hotel, Chicago, Ill.

Industrial Electronics

The Administrative Committee of the Industrial Electronics Group has voted to assess the membership $2.00. The assessment is for the publication of a TRANSACTIONS consisting of papers presented at the recent conference held in Chicago, Ill.

Information Theory

A symposium on information theory, with five technical sessions, will be held in New York City during the latter part of October. Papers are being solicited. Abstracts of papers should be sent to: M. J. DiToro, Federal Telecommunications Labs., 500 Washington Ave., Nutley 1, N. J. The subjects are as follows: Tutorial review of information theory; Tutorial review of statistics; Advances; Miscellaneous applications of information theory; and, Applications to communications systems. The deadline for the abstracts is set for the end of August.

Instrumentation

Officers and the Administrative Committee for the Instrumentation Group have been appointed for the coming year. They are: I. G. Easton, Chairman; R. L. Sink, Vice Chairman; S. N. Alexander, H. L. Byerlay, I. G. Easton, E. P. Felch, Rudolf Feldt, W. D. Hershberger, S. C. Lawrence, L. E. Packard, Walther Richter, R. L. Sink, Ernst Weber, Administrative Committee members.

Nuclear Science

Local chapters in Dallas, Tex., and Oak Ridge, Tenn., are being promoted by the Nuclear Science Group.

Microwave Theory and Techniques

The IRE Professional Group on Microwave Electronics has officially changed its name to the IRE Professional Group on Microwave Theory and Techniques.

A. C. Beck of Bell Telephone Laboratories has been named chairman of the technical program for the one-day Symposium on Microwave Circuitry, to be held November 7, 1952, New York, N. Y. Further details of the symposium will be announced in the September PROCEEDINGS.

Medical Electronics

A NEWSLETTER has been sent to all members requesting information for a bibliography which will be compiled of available material on medical electronics.

Quality Control

The first TRANSACTIONS of the Quality Control Group, containing six papers from the last two symposia of the Group, has been sent to the members.

The officers appointed for the Group for the coming year are: Leon Bass, Chairman; Harold May, Vice Chairman; and Victor Wouk, Secretary-Treasurer.

Vehicular Communications

Officers and the Administrative Committee of the Vehicular Communications Group have been appointed for the coming year. They are: F. T. Budelman, Chairman; Waldo Shipman, Vice Chairman; G. M. Brown, Treasurer; and, Austin Bailey, G. M. Brown, A. B. Buchanan, F. T. Budelman, E. C. Denstaedt, R. V. Dondanville, C. M. Heiden, E. H. I. Lee, Newton Monk, W. M. Rust, Jr., Waldo Shipman, R. C. Stinson, Edwin White, Administrative Committee Members.

TRANSACTIONS NOW AVAILABLE BY SUBSCRIPTION

The TRANSACTIONS of the IRE Professional Groups are now available on a subscription basis. All TRANSACTIONS published during the 12-month period beginning July 1, 1952 may be ordered for $70. University libraries, public libraries, and subscription agencies may subscribe at the special rate of $50.

It is anticipated that at least 25 issues of TRANSACTIONS will be published during the 12-month period.

Donald B. Sinclair (below) IRE President, made opening address at Conference.

Gilbert H. Arenstein, President of the Conference, is shown addressing the banquet guests.

Airborne Electronics Featured at Dayton Conference

The fourth annual National Conference on Airborne Electronics attracted, for the year 1952, more than 1300 scientists, engineers, and technicians from all parts of the United States and Canada. The majority of the 81 technical papers and all the 53 engineering exhibits were located at the Biltmore Hotel, Dayton, Ohio. However, two outstanding technical sessions, which included nine technical papers, were held in the auditorium of the Dayton Engineer's Club in order to accommodate the large audiences which attended these sessions. The keynote of last year's conference, "Electronics—Key to Air Supremacy," was used again this year and extended to specifically include commercial aviation by the presentation of two symposia devoted entirely to this field of airborne electronics.

On Monday noon, May 12, 1952, the convention was officially opened by Gilbert Arenstein, president of the 1952 National Conference on Airborne Electronics, at a luncheon given in the Grand Ballroom of the Biltmore Hotel.

Airborne Electronics Conference Symposium, "Electronics and the Air Lanes," (left to right) S. P. Saint, Air Transport Association of America; F. B. Lee, Department of Commerce; L. M. Sherer, Radio Technical Commission for Aeronautics; C. Morrow, Corporation Aircraft Owners Association; M. J. Riddle, Aircraft Owners and Pilots Association; L. B. Hallman, Wright Air Development Center.

Technical Sessions

The technical papers presented at the Biltmore Hotel and the nearby Engineer's Club were of outstanding quality and constituted the major activity of the conference.

In addition to the 18 technical sessions two symposia were presented at the National Conference on Airborne Electronics dealing exclusively with the electronic problems of commercial aviation. The highlight of the Monday sessions was the evening symposium "Electronics and the Air Lanes." The 800 persons attending the symposium heard veteran pilots request more simplicity in the operation of electronic navigation devices.

On Tuesday morning the transistor session filled the 500 seat auditorium of the Dayton Engineer's Club to capacity. A similar capacity audience at the Engineer's Club heard the magnetic amplifier session papers in the afternoon.

Commercial Aviation again held the spot light Wednesday morning when a second symposium, "Six Years of Progress in the Common System of Air Navigation and Traffic Control," was attended by a capacity audience.

In addition, to the excellent material on electronics for commercial aviation, papers were presented on measurements, human engineering, vacuum tubes, microwaves.

Exhibits

The number of exhibitors at the National Conference on Airborne Electronics was increased this year to a total of 53. The main lobby of the Biltmore Hotel was devoted this year for the first time to exhibit space in addition to the exhibits covering the fourth and fifth floors.

While electronic components and measuring equipments were featured by many exhibitors, more spectacular displays such as the instrument landing system of the Collins Radio Corps., the industrial television display of RCA, and the radar range of the Raytheon Manufacturing Company, also figured in convention exhibits.

Social Events

The social events of the conference were initiated by the President's luncheon given on Monday noon, May 12, 1952. Gilbert H. Arenstein welcomed to the convention, members of the IRE, members of the IRE Professional Group of Airborne Electronics,

Among the highlights of the exhibits in the lobby of the Biltmore Hotel were: (left) an instrument landing system, and (right) a television display.

their wives, friends, and guests. I. B. Johns, assistant research director of the central research department, Monsanto Chemical Company, was the guest speaker of the luncheon. His topic, "The Electron in Modern Science," proved to be most interesting and was especially suited as an introduction to the many technical sessions presented throughout the convention.

On Tuesday evening preceding the annual banquet, a cocktail party was given, sponsored by the National Conference on Airborne Electronics. This function, the first of its kind at the convention, was attended by more than 300 guests.

Immediately following the cocktail party was the annual banquet attended by more than 500 members and guests. Charles B. Jolliffe, vice president and technical director of RCA, spoke to a capacity audience in the grand ballroom of the Biltmore Hotel. Dr. Jolliffe spoke on the subject, "Pioneering in Science," in which he emphasized the urgency of replenishing science's storehouse of fundamental information. George L. Haller, dean of chemistry and physics, Pennsylvania State College, again this year notably filled the role of toastmaster for the banquet. A. Hoyt Taylor, former chief scientist of the Naval Research Laboratory, received in absentia, the annual award as a "Pioneer in Airborne Electronics." Leo C. Young, Dr. Taylor's long time associate and fellow scientist at NRL, accepted the award and recounted many of his interesting experiences with Dr. Taylor in the discovery and evolution of radar.

The IRE Professional Group on Airborne Electronics sponsored the final luncheon on Wednesday, May 14, 1952. George Rappaport, Chairman of the Group, made a progress report on group activities and invited all interested in the field of airborne electronics to join the Professional Group on Airborne Electronics. K. C. Black, technical advisor to the commander, Naval Air Development Center, Johnsville, Pa., was the guest speaker.

A ladies luncheon, lecture, and sewing demonstration was given on Monday as a feature of the ladies program.

The De Forest Pioneers Testimonial Dinner

April 18, 1952—New York, N. Y.

A testimonial dinner to Dr. Lee de Forest, inventor of the three-element tube which underlies all present-day radio, television, and electronic equipment was held at the Waldorf-Astoria Hotel, April 18, 1952, New York, N. Y. Former President Herbert Hoover made the principal address, and leading industry figures, including members of the De Forest Pioneers, took part.

A recent informal estimate places at $60 billions the total value of the equipment based on Dr. de Forest's primary invention, in the fields of radio, TV, sound pictures, long-distance telephony, communications, medicine, instrumentation, and miscellaneous electronic applications.

Admiral Ellery W. Stone, president of American Cable and Radio Corporation, presided at the 50th anniversary of "Doc's" entry into "wireless" in 1902. Others who joined in the de Forest tribute were Brigadier General David Sarnoff, Dr. Allen B. DuMont, Colonel Sosthenes Behn, William J. Barkley, John V. L. Hogan, Frank Andrae, Haraden Pratt, Charles A. Rice, Walter Marshall, Admiral S. C. Hooper, ex-Governor Charles A. Edison, Dr. Marvin J. Kelly, and Dr. Orstes H. Caldwell, Editor of *Tele-Tech*, and IRE Executive Secretary George W. Bailey.

The De Forest pioneers sponsored the dinner in co-operation with AIEE, IRE, NARTB, RTMA, ARRL, SMPTE, VWOA, and the Radio Pioneers.

Following is the text of the address given by Dr. De Forest, during the events of the dinner.

"There seems to be a lasting cement, intangible yet strong, that binds together all pioneers in any exploration project, whether that was a daring expedition into an untrod, unknown wilderness, or into some strange unknown realm of technological exploration where they, as pioneers, have long striven, despite general disbelief, personal sacrifice, ridicule, and discouragement.

If the sacrifices such early work demanded were heavy and long endured, if recompense was small and interminedly postponed, calling for unyielding faith in the future of that cause, and staunch reliance on the leader of the strange adventure, then so much the more strongly do we find those pioneers united, in a sense of fellowship and affection which is rarely realized in more modern times, when the element of personal risk is almost wholly lacking.

Such has been the romantic and "hard-way" history of the Wireless Telegraph—the Radio Telephone and Broadcasting, and the Talking Motion Picture. It was my good fortune to have been born early enough to have played a pioneer role in each of those developments. And in every case, I have been loyally assisted by stalwart young men, gifted with *vision to see, faith to believe, courage to do.*

Those few here tonight from my wireless telegraph beginnings include: Elmo Pickerill, dating from 1904; Charley Cooper, the oldest here, whose work dates from 1903; Charley Pannill; Elmer Bucher, also 1903; and William Medd, who helped me install wireless at Key West Navy Yard in 1905, and later operated the first radio phone for the Navy, on the Battleship *Maine's* historic voyage around the globe in 1908, under Admiral Bob Evans.

Those here from the radio telephone and broadcast era include: Jack Hogan, from 1906 (the youngest of them all, who has since pioneered brilliantly in facsimile, and who, like a faithful disciple of mine, has convincingly demonstrated that radio broadcasting of only fine music, almost purged of sponsors, can be self sustaining); here, too, is E. J. Simon, who started with me in 1909, and who today has lost no whit of his devoted zeal; Lloyd Espenschied, who with Dr. Llewellyn, represents, (as sponsors) the Bell Telephone Laboratories, typically epitomizing the pioneer spirit of that great research institution which was the first to recognize the value of the 3-electrode tube and, with the American Telegraph and Telephone Company, early adapted it to transcontinental telephone service, and now to microwave linkage for nation-wide television; Bill Barkley, who pioneered Collins Radio to a quarter-billion war orders; Hugo Gernback, pioneer in radio journalism and invention; William Dubilier, pioneer in condensers, himself a condenser highly charged with electric energy; Joel Michaels (to whose vision, in the early thirties, we owe the origin of the De Forest Pioneers); Jack Poppele, who built up one of the earliest national networks; Frank Hinners; Frank Andrea, who at old Highbridge learned how to manufacture a thousand TV sets a day; Bob Gowan, who broadcast from Ossining to Chicago in 1916; Admiral Ellery Stone, whose devoted, tireless energies, together with those of Charles Rice, have made this great meeting possible; I heartily thank them and all of you individually who have aided in preparation of this grand meeting. Here, too, is the pioneer radio's historian, George Clark; here, Dr. Allen Dumont, that outstanding pioneer of *television*, surely the most miraculous of all the electron's marvelous achievements. An even earlier video pioneer is with us tonight, U. A. Sanabria of Chicago, who even before the Englishman Baird, demonstrated his American television system, as early as 1926. He was the original inventor of interlaced scanning and he and I have worked closely together in recent years. From Chicago, also, is Lyndon A. Durant, who last year formed the Lee de Forest Foundation, a charitable society.

Of talking picture pioneers; we have with us Earl Sponable, Earle Hines, and many others seated at the Society of Motion Picture Engineers table.

I am especially pleased to see here so many of the old patent attorneys: George Folk, Cornelius Ehret, George Shirer, and

(*Cont. on page 1006.*)

(Cont. from page 1005.)

others. It is indeed sad that both Captain Darby and his brilliant son, Sam Darby, Jr., have passed away. Regardless of how those fierce old patent battles were won, or lost, they, the attorneys, were the first to find a profit in radio! They always won!

Thus, we find many of those early pioneers here tonight, united in a common spirit of youthful adventure, eager to renew the old affections, to form a lasting organization that will continue to bind them together, and so provide that their children, proud of what their fathers have accomplished in past years, shall continue to relate the exploits of the fathers, and to perpetuate the annals of brave pioneering days.

It gives me, therefore, a more profound sense of gratitude to those early associates, employees, and friends than I have ever before felt, to see so many of you gathered here tonight about this festive board, commemorative of so many devoted lives, your own and those of beloved associates who have already crossed the great divide. Sad indeed, is it that only this morning, Louis Pacent, beloved by all who knew him, passed away. He had been looking forward so eagerly to this meeting. His spirit is with us, nevertheless. Louis Pacent shall never be forgotten.

It is futile to try to express in words what this occasion means to me; so much of solemn recollection, such thoughts of jovial hours, as one by one our crude efforts pushed back the age-long barriers, as we rejoiced in our small, dearly bought triumphs which, in the aggregate, have remade the world we labored to improve.

Saddened though we are tonight that so many beloved comrades are no longer with us, let us cheerfully look forward to the *New World*, to that tomorrow which would be but an idle dream had we not banded together, as plodding pioneers, in the years gone by.

You know, each one of you, the love and affection with which I salute you all tonight.

Throughout my wireless-telegraph and radio-telephone work, the United States Navy and the Signal Corps were ever eager to encourage such pioneering efforts. Their assistance could always be counted upon; their encouragement was most essential to my progress. So I am happy to see the Navy and Army represented here tonight.

As then Secretary of Commerce, you, Mr. Hoover, brought radio out of the chaos that had so engulfed it from its beginning, and laid down those wise regulations under which it has grown to be one of the bulwarks of America.

A most worthy son of the world's greatest inventor, you, Mr. Edison, have nobly carried on the tradition that will stand forever as a goal which we subsequent inventors shall ever strive, hopelessly, to attain. By his discovery of the "Edison Effect," Thomas A. Edison may be justly called the first electronic pioneer.

Perhaps the most outstanding radio pioneer of this present age (when measured by the tangible resources he has created), but a pioneer regrettably not of our organization, is General David Sarnoff, chairman of the most powerful organization in radio history. The entire framework of today's mighty structure of radio communication, including the wireless telegraph and broadcasting by sound and sight, owes more than can be computed to this man's genius. We are proud to have David Sarnoff as our honored guest.

No cultural institution owes more to the radio broadcast than does grand opera. I am delighted to have as our guest tonight, Mr. George Sloan, chairman of the Metropolitan Opera Association, mindful of that first broadcast of grand opera in January, 1901, when Caruso's voice graced ether waves.

Few realize as does Mr. Sloan how much radio has achieved in acquainting the public with the finest of great music. And, when it is dressed in polychrome video, opera's appeal will indeed be irresistible!

The significance of this great meeting is to me completely overwhelming. It, indeed, gives me pause—I thank God that my life has been so extended that I am privileged to witness this gathering, representing as it does so many of the vast industries and the far-reaching instrumentalities, associations, and institutes, nation-wide, world-wide, which have grown up in the past forty-five years, thanks to all their united and sustained efforts, starting from this tiny acorn, a replica of the first Audion of 1906.

Emotions which this magnificent tribute to this tiny tube and its originator arouses within my heart are far beyond my power to express. I can only say that my heart is very full, and deeply happy."

IRE People

John H. Ganzenhuber (A'42), has been appointed manager of the government contracts department of the Hoffman Laboratories, Inc., Los Angeles, Calif.

J. H. Ganzenhuber

Mr. Ganzenhuber was born on September 19, 1909, and received the B.S. degree in electrical engineering, in 1933, from the University of Southern California.

Prior to Mr. Ganzenhuber's recent appointment, he was field sales engineer and district manager of broadcasting sales for Graybar Electric Company for 9 years, manager of broadcast sales for Western Electric Company, and vice president in charge of sales and product development of Standard Electronics Corporation where his work was closely associated with the Television broadcast field. During World War II, Mr. Ganzenhuber took active part in Western Electric's radar program.

J. T. Cataldo (M'50) has been appointed general manager of the International Rectifier Corporation, El Segundo, Calif.

J. T. Cataldo

Mr. Cataldo received his B.M.E. degree from Clarkson College of Technology and the B.E.E. degree from Brooklyn Polytechnic Institute, with supplementary graduate studies at Rutgers and Brooklyn Polytechnic. Mr. Cataldo served with the Government from 1939, and was assistant chief of the instrument section at the Material Laboratory, New York Naval Shipyard, until 1948. He then joined the Signal Corps Engineering Laboratories, at Fort Monmouth, N. J., as a research and development engineer.

Mr. Cataldo has published articles in engineering, radio, and electronic periodicals and has presented a number of papers on the theory and application of selenium rectifiers.

Philips B. Patton (A'46) has been elected vice president of the Lenkurt Electric Company, Inc., in San Carlos, California. Prior to this appointment, he had been sales engineering manager, commercial manager, and a member of the board of directors of the Lenkurt Company.

P. B. Patton

Mr. Patton was born in California, in 1914, and studied in the San Mateo Junior College, Calif., and the University of Maryland. He has held positions with the radiotelephone and telegraph section of the Federal Communication Commission, Pan American Airways, and Western Union. Before joining the Lenkurt Company in 1948, Mr. Patton had been technical co-ordinator of sales and engineering functions of the mobile communications, Farnsworth Television and Radio Company. He continues as commercial manager of Lenkurt.

Walter Evans (M'26–SM'43–F'45) president of Westinghouse Radio Stations, Inc., died recently in Baltimore, Md., after a long illness. He was in charge of fourteen broadcasting stations for Westinghouse.

WALTER EVANS

Born in 1898 in Columbus, Ohio, Mr. Evans studied electrical engineering at the University of Illinois, where his courses were interrupted for service with the Navy during World War I. He was a ship's radio operator aboard Great Lake steamers at the age of 16, which later brought him an assignment as an instructor at the Naval Radio School at Harvard, and as a radio operator aboard a submarine chaser.

He completed his courses after the war at the University of Illinois, where he was an undergraduate instructor in radio engineering.

Mr. Evans then served as wireless operator for the Marconi Company, Radio Corporation of Mexico, and the United Fruit Company. While with the latter organization, he was credited with supervising the first radio telephone installation on an American merchant ship.

Joining Westinghouse in 1921 as a radio operator at the then-new station KYW in Chicago, he was named chief engineer the next year, became general manager four years later, and in 1932, was placed in complete charge of all Westinghouse stations.

In 1933 his duties were expanded, in 1936 Mr. Evans was elected a director of Westinghouse Radio Stations, Inc., in 1939 vice president, and in 1947, president of that subsidiary. He was vice president of the parent Westinghouse Electric Corporation from 1942 until his death.

Mr. Evans was named chairman of the International Broadcasting Committee of the Defense Communications Board in 1941. Recognition of this service and "his contributions in connection with the development and production of radio and radar equipment during World War II" came in Certificates of Appreciation from both the Army and Navy.

On leave of absence from Westinghouse, Mr. Evans served, at the request of the Government, as industry technical adviser to the State Department at the International Telecommunication Conference at Madrid in 1932 and at Cairo in 1938.

Mr. Evans was a director of Broadcast Music, Inc., and the Radio Manufacturers Association. He served on the IRE Awards Committee in 1948, and IRE Public Relations Committee in 1947.

B. Richard Teare, Jr. (A'41–SM'45–F'51) has been appointed associate dean of the Carnegie Institute of Technology College of Engineering and Science. He will work closely with the dean in general administration and policy planning of the College of Engineering and Science, including financial and educational development.

B. R. TEARE, JR.

A native of Menomonie, Wis., Dr. Teare received his B.S. and M.S. degrees in 1927 and 1928, respectively, at the University of Wisconsin.

In 1929 Dr. Teare joined the General Electric Company at Schenectady as a student in the advanced course in engineering, and continued for four years with the educational program. In 1931 he became supervisor of the advanced course until two years later, when he became an instructor and assistant professor of electrical engineering at Yale University. He received his Ph.D. degree in 1937, while teaching at Yale.

Dr. Teare joined the Carnegie faculty in 1939, and was named Buhl Professor there in 1943, and dean of graduate studies in 1950. During World War II, he served as an electrical engineer in the Naval Ordnance Laboratory and worked on research problems at Carnegie Tech for the National Defense Research Committee, the Office of Scientific Research and Development, Army Air Forces, Office of Naval Research.

Dr. Teare is the author of several papers on engineering education and was voted the George Westinghouse Award in 1947 for his distinguished contributions in the field of engineering education.

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

❖

James H. Mulligan (S'41–A'45–M'45) has been appointed chairman of the department of electrical engineering, New York University. Prior to this appointment, Dr. Mulligan was a joint co-ordinator of electrical engineering research for the university, directing sponsored research projects.

J. H. MULLIGAN

Dr. Mulligan was born on October 29, 1920, in Jersey City, N. J. He received the B.E.E. and E.E. degrees, in 1943 and 1947, from the Cooper Union School of Engineering, the M.S. degree from Stevens Institute of Technology in 1945, and the Ph.D. degree from Columbia University in 1948.

First associated with the Bell Telephone Laboratories, he later became a member of the Combined Research Group of the Naval Research Laboratory, engaged in the development of radar identification equipment. In 1945 he joined the Allen B. DuMont Laboratories where he became the chief engineer of the television transmitter division, and was active in the development of the first DuMont Image Orthicon field pickup equipment. In 1947 he was awarded the Watson Scientific Computing Laboratory Fellowship, and in 1949, he joined the faculty of New York University.

Dr. Mulligan is the Chairman of the IRE New York Section, a member of the American Institute of Electrical Engineers, the American Physical Society, American Mathematical Society, Tau Beta Pi, Sigma Xi, Cooper Union Alumni Association.

❖

Francis S. Benson (A'28), an assistant engineer of the Pacific Gas and Electric Company, died recently in San Francisco, Calif.

Mr. Benson, a native of Illinois, was born in 1895, and studied engineering at the University of California. He was the author of numerous electrical engineering textbooks and technical articles. Mr. Benson was a member of the American Institute of Electrical Engineers and the American Association for the Advancement of Science.

❖

Frederick G. Diver (J'24-A'26-SM'46) has been appointed a Member of the Order of the British Empire. His name appeared in the Queen's Birthday Honors List, published recently.

F. G. DIVER

Mr. Diver was born at Holmwood, Surrey, Eng., in 1904, and received technical training at the London Telegraph Training College from 1920–1921.

From 1921–1922 Mr. Diver was associated with the Low Engineering Company Ltd. at Feltham, Middlesex, as a radio engineer in charge of experimental radio reception. He continued there as a technical assistant until 1924, when he entered the sea-going staff of the Marconi International Marine Communications Company, Ltd.

In 1928, Mr. Diver joined McMichael Radio Ltd., as chief of test and in that capacity was responsible for test, inspection, and test-gear design activities until 1939. He was then appointed chief engineer in charge of all design, development, and test engineering. He has held that position in the equipment division of the company since 1946 to the present.

Mr. Diver is a member of the British Institution of Radio Engineers and has been serving on its technical committee.

Books

Short-Wave Radiation Phenomena, Volume I and II by August Hund (4297)

Published (1952) by McGraw-Hill Book Company, Inc., 330 West 42 St., New York 18, N. Y. 1243 pages+45-page appendix+14-page list of references +80-page index. 388 figures. Two volumes. 6×9. $20.00. Not sold separately.

August Hund was a scientific and technical radio consultant, Santa Monica, Calif. He is deceased.

This latest work from the "pen" of August Hund will, in this reviewer's opinion, be subject to the same varied reception as his previous works. The style is generally the same, and the controversy will probably also be the same.

As indicated on the jacket of the book (it is a single work of 1,382 pages bound in two parts), it contains a wealth of data, formulas and theory on short-wave radiation effects. This is precisely the major weakness of the work; it is too much a collection of data and formulas, and not enough an integrated development. The average reader may be critical; the writing level of this book is high.

The text consists of nine chapters and an appendix. This means, of course, that each chapter is almost of normal book length. The first chapter deals with fundamental concepts and relations between currents and electromagnetic fields. The second treats space-electromagnetic fields, considering both elementary electric and magnetic dipoles from a comparative viewpoint. Fundamental methods used in electromagnetic theory are viewed in the third chapter. A fourth chapter examines certain features of radio wave propagation and the influence of the medium through which the propagation takes place. A rather extensive treatment is contained in the fifth chapter of transmission on lines and the problems of matching lines to radiators. Chapter six, which is 261 pages in length, discusses unobstructed radiation in space, a fairly broad treatment of radiation of antennas all sizes, shapes, complexity.

Volume II begins with chapter seven which is a 238-page discussion on space radiation in the presence of electromagnetic obstructions, and includes considerations of the influence of the earth, the troposphere and ionosphere, and of metallic reflectors. Chapter eight describes obstructions and electromagnetic diffraction effects, giving details of wave-front radiators and the field effects in the obstructed as well as in the unobstructed region. Extensive use is made of the Huygens wavelets, and the use of the Fresnel integrals and the Cornu spiral interpretation in the solution of diffraction problems. It deals with diffraction radiators which are driven by radio-frequency potentials across slits, by circular-ring slit radiators, and narrow slits in the axial direction of metal cylinders. The final chapter discusses aspects of waveguides and cavities.

There is much to recommend this work. It contains an enormous amount of material relating to all aspects of electromagnetic theory and its practical applications. The explanatory notes which are included with mathematical equations do much to add physical significance to the equations. Many numerical examples are included throughout the text. Considerable discussions are included on (a) Hertzian vector, chapter 3; (b) phase and group velocities, chapter 4; (c) antenna theory for arrays of all types and complexity, chapter 6; and, (d) diffraction problems, and the use of Cornu's spiral for interpreting the Fresnel integrals, chapter 8. A complete reference and index is included.

There are likewise many criticisms that can be levelled against this work. While explanatory notes regarding vectors and vector operations are given on page 119 and page 1268, there has been a careless disregard in distinguishing between scalars and vectors. Often, too, dots and crosses in vector operations have been ommitted, and the complex numbers of ac circuit theory and space vectors seem to get thoroughly confused. A careful avoidance of transmission line charts in the solution of impedance matching problems is difficult to understand. A number of very important features of waveguides and cavities are neglected in the brevity of chapter 6. Perhaps the major criticism here is that the term "cavities" appears in the chapter title, and the consideration of these is quite limited. The book also contains many typographical errors.

Because of the book's length, content, and manner of presentation, it is unlikely that it will serve as a textbook. On the other hand, it is an excellent reference book for students and experienced practicing engineers, and will undoubtedly be warmly received by these persons.

SAMUEL SEELY
Syracuse University
Syracuse, N. Y.

Microphones by the Staff of the Engineering Training Department British Broadcasting Corporation (4298)

Published (1952) by Iliffe & Sons, Ltd., Dorset House, Stamford St., London S.E. 1, Eng. 93 pages +2-page index+19-page appendix. 78 figures. 8¾×5½. Price, 15s.

The title of this book would lead one to believe that all currently used microphones systems would be discussed; however, there are two major omissions: The Western Electric-Altec capacitor microphone with its rather unique head design and equalizer system, and the Stephens microphone system with its RF modulation. The German Telefunken microphone system is also omitted. Since these three microphones represent the highest standards in quality transducers, it is presumed that they are either of no direct interest to the broadcasting engineer or that their omission was an oversight by the authors. Other microphone systems of past and current usage are adequately discussed and are of value to the engineer who has a reasonably good basic knowledge of the mathematics involved.

This volume could not be classified as "new information" since all of the subject matter is perhaps better covered in volumes published by the Society of Motion Picture Engineers. It is a little surprising that, because of the title "Microphones," the authors did not consider the recording field as well as the broadcast field. The two are closely allied in many respects. It is common knowledge that the quality of British recording is of the highest order and surpasses the quality of recording in this country. Inclusion of information of the latest British and German developments, therefore, would have been in order.

From an academic point of view the approach to various microphones and systems is satisfactory but seems too theoretical and too technical for the average broadcast engineer whose value to his station lies in his ability to use the microphone effectively with a thorough knowledge of its operation a secondary consideration.

The reviewer would consider this book to be best suited as an elementary manual for classes of broadcast engineering students. It seems to meet this requirement most satisfactorily. The information is clearly presented and no serious repetition or inconsistency is noted.

E. D. NUNN
Northern Signal Company, Inc.
Saukville, Wis.

The Measurement of Radio Isotopes by Denis Taylor (4299)

Published (1951) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 116 pages+2-page index+viii pages. 40 figures. 4×6½. $1.50.

Denis Taylor is head of the electronics division, Atomic Energy Research Establishment, Harwell, Eng.

This slender volume is written for the nonspecialist planning to use radioisotopes as a research tool, to acquaint him with the various experimental techniques available, and to provide him with the necessary knowledge of the chief characteristics of radioactive substances. Emphasis is placed upon the physical principles involved and their limitations, as well as the advantages of the methods described. The greatest amount of discussion is devoted to counting systems employing Geiger-Müller tubes, however; proportional and scintillation counters are also included.

A brief introduction on the principal uses of radioisotopes is followed by a chapter on the fundamentals of radioactive decay and units. Chapter three covers apparatus for measuring radioactivity, such as ionization chambers, vacuum tubes, and vibrating-reed electrometers, and quartz-fiber electrometers, while chapter four is devoted to counting systems.

Three chapters are concerned with reduction and correction of the data. One is devoted to statistics, the next is a discussion of source geometry and self-absorption, and another includes methods of measurement and correction factors. Some material on diagnosis of equipment faults is included. Chapter eight deals with further counting systems, followed by a final chapter on health hazards and radiation monitors.

Conciseness is at once a selling point of the book if one is not concerned with technical details. This, however, is also a drawback in the book. In his anxiety to spare the reader the intricacies of electronics, the author almost fails to mention the matter of speed of response, pertaining to the topic of ion chambers with dc amplifiers. In the reviewer's opinion, the chapters on reduction of data are likely to prove most valuable, especially in the methods of correcting for resolving time and absorption. As a quick reference book, the value would have been enhanced if the author had not chosen t to mean "sample thickness" (in chapter six), as well as "time."

There are only a few errors which should not be any trouble to an alert reader; the most serious is noted on page 91, where the reader is told to multiply instead of divide by the geometrical factor in order to obtain the sample activity. This book is well written and recommended for every beginner in radioisotopes.

J. B. H. KUPER
Brookhaven National Laboratory
Upton, L. I., N. Y.

Transmitting Valves by J. P. Heyboer and P. Zijlstra (4300)

Published (1951) by the Phillips Technical Library Eindhoven, Holland. 281 figures. 6×9.

This is the seventh of a series of books, published by the Phillips Technical Library, dealing with the construction and use of vacuum tubes. The book describes methods of determining the operating characteristics of vacuum tubes in transmitting or other equivalent circuits. Since a straight-line static vacuum-tube characteristic is assumed, the treatment of the material is such that an extensive mathematical background is not required. This straight-line approximation is not invalid in the region in which these tubes are generally operated.

A large section of the book is concerned with the performance of triodes, tetrodes, and pentodes as Class-C power amplifiers. Described in detail are the methods of calculating input and output power, efficiency. driving power, angle of current flow, load impedance, and limitations of the various tube types. Comparisons as to accuracy and work involved are also made of analytic and graphic methods, determining performance characteristics of transmitting tubes.

Another large section covers methods of amplitude modulation of Class-C power amplifiers. Conventional methods of modulation (for example, control grid, screen grid, and anode) and combined methods of modulation (control grid-anode) are described. The various methods of calculating modulation power, efficiency, and the relative advantages of each are presented in detail. The sections on power amplifiers and modulation methods cover approximately 50 per cent of the book.

The chapter on oscillators discusses problems of stabilization (neutralization) of oscillators, calculation of frequency and amplitude of oscillation, and application of power amplifiers as oscillators.

The last three chapters are concerned with the transmitting tube as a frequency multiplier, some special items encountered in transmitting tubes (gas discharges), and methods of increasing the frequency range of conventional tubes to the vhf region by special circuits and construction techniques.

The Appendix describes a method used to calculate the operating characteristics of tubes as Class-A and Class-B amplifiers, and shows how constant-current characteristics can be used to advantage in computing Class-C amplifier operation.

Unfortunately, there are several defects in the book. In some cases the translation from Dutch to English results in rather stilted reading. The absence of a list of symbols causes some confusion since some symbols do not correspond to those used in this country and, in some instances, no meaning of the symbol has been given. A few errors in the symbols are also present.

Despite these limitations, this book should be of value to those concerned with the use of vacuum tubes in transmitters or other equivalent circuits.

MAX KRAWITZ
Sylvania Electric Products Inc.
Bayside, N. Y.

Advanced Theory of Waveguides by L. Lewin (4301)

Published (1951) by Iliffe & Sons, Ltd., Dorset House, Stamford St., London, S.E. 1, Eng. 165 pages+23-page bibliography+2-page index. 54 figures. 8¾×5½. $6.75 at The British Book Centre, 122 E. 55 St., New York, N. Y.

This book will be welcomed by mathematically equipped microwave engineers and physicists who are interested in the actual analysis of the effects of discontinuities in waveguides. Fairly complete detailed analyses of the various problems makes it a book for serious study, rather than for browsing. It is unfortunate that the outlined mode of attack on each type of problem follows the detailed algebra rather than precedes it; it would be better pedagogy to let the reader know the campaign plans in advance, and it would facilitate the reading.

For uniformity and brevity, only rectangular guides are discussed; however, the methods are general. The book opens with a useful summary of the use of electric and magnetic Hertz vectors for the generation of propagation modes. Inductive and capacitive posts are then discussed by mode matching a multipole current distribution to the incident field. The equivalent circuit and reflection coefficient are deduced. Diaphragms are treated by mode matching the fields on opposite sides of the aperture; the resulting integral equation for the aperture field is solved to the quasi-static and third-order approximations. Single strip obstructions are treated with an integral equation for strip current; quasistatic and second order approximations are derived.

Following, is a treatment of the tuned post with finite conductivity by an integral equation for the post current distribution. Impedance is expressed as a stationary quadratic form associated with the integral equation, and an approximate result is obtained by assuming a sinusoidal current distribution. The tuned window is handled by an integral equation for the aperture field, and the impedance is approximated by the variational method already introduced.

Also discussed is the H-plane step by the integral equation variational procedure, as is the E-plane T-junction. The E-plane step is handled to the quasi-static approximation by conformal mapping. The small angle E and H-plane tapers are treated by dominant mode matching, and a matching scheme using approximate modes is applied to the double taper.

Radiation from a guide with an infinite flange is again presented by the mode matching, integral equation for aperture field, and variational approximation technique. Radiation from parallel planes with no flange introduces a Wiener-Hopf integral equation with the factored Fourier-Transform solving technique.

In the final chapter, guides with homogeneous and inhomogeneous loading are considered, and the corrugated guide is illustrated by the detailed analysis developed for a linear accelerator.

This text is well annotated with references to other ways of dealing with the same problems. Following the text is a 23-page classified bibliography covering the last decade. There are a few places with careless wording, and a few places where the reader loses the thread, but on the whole the book is quite readable.

CHESTER H. PAGE
National Bureau of Standards
Washington, D. C.

Fundamentals of Electronics and Control by Milton G. Young and Harry S. Bueche (4302)

Published (1952) by Harpers and Brothers, 49 E. 33 St., New York 16, N. Y. 498 pages+15-page index +12-page appendix+x pages. 318 figures. 6×9¼ $6.00.

Milton G. Young is a professor and Harry S. Bueche is an associate professor in the electrical engineering department, University of Delaware, Newark, Del.

In the authors' words, this book "has been written as a fundamental text to serve the needs of the nonelectrical as well as the electrical engineering student and the practicing engineer." In line with their stated purpose, the authors have produced a well-balanced descriptive text which covers a wide range of topics in the field of electronics and control.

In the main, the material is ably organized, clearly presented, and well illustrated with numerous photographs, circuit diagrams, and graphs. Unfortunately, because of space limitations, some rather important topics are discussed so briefly that the reader can achieve only a superficial understanding of the subject. However, ample references are included to facilitate further study if desired. A fault to be found with the references is that many of them give neither the name of the author nor the title of the paper. On the positive side, a fairly large selection of problems, most of them with answers, is at the end of each chapter.

The scope of the book is indicated by the chapter headings as follows: electron theory; basic circuit components and control; principles of electron emission; high vacuum tubes; tubes utilizing gas; mercury pool tubes; amplifiers; oscillators; modulation and demodulation; and, rectifiers. Included in these chapters are brief discussions of such modern devices as the transistor, dynamo-electric amplifier, and magnetic amplifier. In general, this is a carefully written book which should serve well the needs of the nonelectrical engineering student and the practicing engineer. In regard to the electrical engineering student, this reviewer is inclined to feel that the book does not cover the fundamentals of electronics and control in a sufficiently thorough and quantitative manner to be suitable as a textbook for students of electronics or communications.

L. A. ZADEH
Columbia University
New York, N. Y.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with That Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published of the PROC. I.R.E. from February 1951, through January 1952, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London, S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

ACOUSTICS AND AUDIO FREQUENCIES

016:534 1797

References to Contemporary Papers on Acoustics—R. T. Beyer. (*Jour. Acous Soc. Amer.*, vol. 24, pp. 92–97; January, 1952.) Continuation of 1487 of July.

534.213:621.395.623.75 1798

Exact and Approximate Equations for Wave Propagation in Acoustic Horns—A. F. Stevenson (*Jour. Appl. Phys.*, vol. 22, pp. 1461–1463; December, 1951.) "Exact equations are given for the propagation of acoustic waves in horns of arbitrary shape. These equations are similar to, though simpler than, the equations previously found for electromagnetic horns [1841 below], and can be regarded as giving rise to an infinite number of coupled modes of propagation. If the coupling is neglected, the equation for the fundamental mode is the familiar one, but the theory also furnishes equations for the higher modes. The error involved in neglecting coupling is discussed."

534.232 1799

The Design of Optimum Directional Acoustic Arrays—N. Davids, E. G. Thurston and R. E. Mueser (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 50–56; January, 1952.) Dolph's theory (2187 of 1946) for broadside antenna arrays is applied to the design of acoustic arrays with optimum directive properties. Tchebycheff polynomials are used to obtain the best possible relation between side-lobe level and main-beam width. Experimental results are shown in diagrams and confirm the theory.

534.232 1800

Directionality Patterns for Acoustic Radiation from a Source on a Rigid Cylinder—D. T. Laird and H. Cohen. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 46–49; January, 1952.) Morse's theory (1575 of 1949) of radiation from an infinitely long strip vibrating on the side of a rigid cylinder is extended to apply to a finite source such as a rectangular strip.

534.232:538.652 1801

Ferrite [Magnetostrictive] Oscillators—K. Sixtus. (*Frequenz*, vol. 5, pp. 335–339; November/December, 1951.) The temperature coefficient for the resonance frequency and the Q value of ferrite-cored magnetostriction oscillators are determined from results of impedance measurements. Equivalent-circuit elements are evaluated. Properties and applications are discussed.

534.232:538.652:621.3.012.8 1802

The Equivalent Circuit of the Ferro-Magnetostrictive Transducer—H. Schonfeld. (*Frequenz*, vol. 5, pp. 331–334; November December, 1951.) The equivalent circuit is derived by analysis based on the change in length with magnetization, and the variation of the magnetization curve with tensile stress. At frequencies near natural resonance it comprises the coil inductance in parallel with a series resonant circuit.

534.26:535.43 1803

Multiple Scattering of Radiation by an Arbitrary Configuration of Parallel Cylinders—V. Twersky. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 42–46; January, 1952.) "A formal solution in terms of cylindrical wave functions is obtained for the scattering of a plane acoustic or electromagnetic wave by an arbitrary configuration of parallel cylinders, which takes into account all possible contributions to the excitation of a particular cylinder by the radiation scattered by the remaining cylinders."

534.321.9:532.528 1804

Cavitation produced by Ultrasonics: Theoretical Conditions for the Onset of Cavitation—E. A. Neppiras and B. E. Noltingk. (*Proc. Phys. Soc.* (London), vol. 64, pp. 1032–1038; December 1, 1951.) Theoretical investigation indicates that cavitation is restricted to a definite range of variations of alternating-pressure amplitude, pressure-wave frequency, radius of bubble nucleus, and hydrostatic pressure.

534.321.9:534.22:538.69 1805

Effect of a Magnetic Field on the Propagation of Sound Waves in a Ferromagnetic Material—J. de Klerk. (*Nature* (London), vol. 168, pp. 963–964; December 1, 1951.) Transverse or longitudinal magnetic fields applied to a rod of ferromagnetic material cause a decrease in the attenuation of longitudinal ultrasonic waves, a minimum value being attained at magnetic saturation. Graphs of attenuation against magnetic-field strength are given for Ni and Fe-Ni using a frequency of 2.5 mc.

534.321.9:534.373-14 1806

An Anomalous Effect in the Ultrasonic Absorption of Electrolytic Solution—R. E. Barrett and R. T. Beyer. (*Phys. Rev.*, vol. 84, pp. 1060–1061; December 1, 1951.) Absorption measurements were made on aqueous solutions of sodium acetate in the frequency range 9–45 mc. The results indicate that the absorption coefficient cannot be assumed to be the sum of the coefficients for solvent and solute considered separately.

534.321.9:534.511.1 1807

A Precise Recording Ultrasonic Interferometer and its Application to Dispersion Tests in Liquids—R. Barthel and A. W. Nolle. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 8–15; January, 1952.)

534.321.9:534.511.1 1808

On the Theory of the Fixed-Path Acoustic Interferometer—F. E. Borgnis. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 19–21; January, 1952.) A general expression is given for the electrical input impedance of the acoustic interferometer. From this expression formulas are derived for determining the velocity of sound by a frequency-variation method, or for determining changes in velocity due to variations of pressure, temperature, etc.

534.321.9:534.511.1 1809

A Recording Ultrasonic Interferometer and its Alignment—J. L. Stewart and E. S. Stewart. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 22–26; January, 1952.)

534.321.9:535.37 1810

The Application of Phosphorescent Materials in the Detection of Ultrasonic Waves—L. Petermann (*Helv. Phys. Acta*, vol. 24, pp. 596–599; December 31, 1951. In French.) 1951 Société Suisse de Physique Lucerne Meeting paper.

534.321.9-14:534.22 1811

Ultrasonic Propagation in Liquids under High Pressures: Velocity Measurements on Water—G. Holton. (*Jour. Appl. Phys.*, vol. 22, pp. 1407–1413; December, 1951.) Values are given for the velocity of 15 waves in water at 30° and 50°C as a function of pressure to about 6,000 atm. Information on the temperature coefficient of the velocity and on the ratio of specific heats at increasing pressures is derived.

534.373-14 1812

Attenuation of Sound in Water Containing Air Bubbles—D. T. Laird and P. M. Kendig. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 29–32; January, 1952.) The bubbles were produced by forcing air through the cloth covering of four metal trays lying on the bottom of a lake. The attenuation of sound was found to be very large at frequencies coincident with bubble resonance frequencies, and much less at other frequencies, indicating that bubble resonance is the principal phenomenon concerned.

534.43:534.372 1813

The Application of Damping to Phonograph Reproducer Arms—W. S. Bachman. (PROC. I.R.E., vol. 40, pp. 133–137; February, 1952.) 1951 I.R.E. National Convention paper.

534.6:621.395.623 1814

Experimental Comparison between the Average Human Ear and some Artificial Ears—P. Schiaffino. (*Poste e Telecomunicazioni*, vol. 19, pp. 567–570; December, 1951.) The results of tests on British, Swiss, Italian and American artificial ears and on the average human ear, using six different types of telephone receiver, are presented in graphs of acoustic pressure, referred to a level of 1 dyne/cm² for 0.1 ma current, against frequency.

534.76:534.861 1815

Information Theory. Elements of Space Information in Microphone-Transmission Systems—A. Moles. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1583–1585; December 19, 1951.) Stereophonic transmission is considered; the element of space information is defined as the smallest displacement of the sound source perceptible by the listener. Experiments using three distinct microphone channels indicate that this element is of the order 1–2 m for an ordinary broadcasting studio. The number of elements of space information to be transmitted is thus small compared with the number of elements of sound information proper; hence instead of using separate channels, the space information can conveniently be transmitted as an infrasonic signal via the main sound channel.

534.833 1816

The Problem of Noise Abatement in the Offices of the Administration française des P.T.T.—P. Chavasse and R. Lehmann. (*Ann. Télécommun.*, vol. 6, pp. 381–396; December, 1951.) Causes of noise, and methods of reducing it, are considered from both the theoretical and the practical points of view, with details of work actually carried out on walls, floors, etc.

534.844.1 1817

Equipment for Acoustic Measurements: Part 4—The Direct Measurement of Reverberation Time—C. G. Mayo and D. G. Beadle. (*Electronic Eng.*, vol. 23, pp. 462–465; December, 1951.) A description is given, with detailed circuit diagrams, of (*a*) a logarithmic amplifier, (*b*) a decay calibrator and noise generator. A special scale enables reverberation times to be read directly from the decay curve displayed on a cro. Part 1: 946 of May. Part 2: 1382 of June. Part 3: 1681 of July.

534.845 1818

The Theory of Sound Absorptive Materials—C. M. Harris and C. T. Molloy. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 1–7; January, 1952.) Review of modern theories and discussion of the relation between the absorption coefficient and acoustic impedance.

621.395.61:546.431.824 31 1819

The Frequency Response of Barium Titanate Transducers—T. F. Hueter and E. Dozois. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 85–86; January, 1952.) The frequency response curves of a large number of $BaTiO_3$ ceramic transducers resonant at frequencies from 0.5 to 3 mc, operated in the thickness mode into a water load, consistently show one subsidiary peak (in some cases two) at a frequency slightly higher than the main resonance, the frequency ratio being near 1.05. All the available evidence indicates that the side peak is excited because of incomplete alignment of the domains in polarized $BaTiO_3$. Domains inclined to the direction of the driving field excite a symmetrical shear mode that has a thickness component.

621.395.623.7:621.395.42 1820

Acoustic Problems in Intercommunication Systems—H. Gemperle. (*Radio Tech.* (Vienna), vol. 27, pp. 522–524; December, 1951.) The response characteristics of a 9-cm loudspeaker (*a*) without baffle, (*b*) fitted in a cabinet with cotton-wool damping, are shown when used for sound output and when used as a microphone. The overall characteristic of an intercommunication system obtained by combination of the two curves shows for (*a*) a resonance peak near 200 cps, and for (*b*) undesirable accentuation of the middle frequencies.

621.395.623.8 1821

Sound Reinforcement and Production for Royal Festival Hall—J. L. Goodwin. (*Elec. Commun.*, vol. 28, pp. 243–250; December, 1951.) An outline description of the system installed for reinforcing the voice of a speaker, for distributing programs relayed from outside the hall and for similar purposes. The ancillary arrangements for deaf-aids and announcing are also reviewed. Two pairs of loudspeakers mounted on the orchestra canopy and two supplementary ones behind the platform are fed by two separate amplifier chains, thus permitting stereophonic reproduction when required. At the inputs to the amplifiers are two six-channel fader units which in turn are fed from any desired jacks in the 20-position field to which the incoming circuits are brought. Suitable low-level switching arrangements permit considerable flexibility of operation: the control console is situated within the auditorium. The deaf-aid jacks, which are associated with 15 per cent of the seats, are supplied through amplifiers from the output circuit of the system described above and also from a separate microphone suspended over the orchestra. An independent announcing system includes arrangements for playing gramophone records and for radiating a musical interval tone, the latter replacing the usual bells in refreshment rooms and foyers.

621.395.623.8 1822

Note on [sound-] Radiator-Array Technique—S. Sawade. (*Elektrotech Z.*, vol. 72, p. 720; December 15, 1951.) Radiation patterns for 1 kc and 6 kc are given to illustrate how the characteristics of modern cone loudspeakers have made the use of loudspeaker arrays practicable for public address work.

621.395.625.3 1823

A Tape Editing and Duplicating Machine—R. P. Ledbetter. (*Audio Eng.*, vol. 35, pp. 18–20, 45; December, 1951.) A 4-unit rack-mounted equipment with separate amplification and equalization for the play-back, duplicating and cueing operations.

621.395.625.3:621.317.35 1824

Boundary-Displacement Magnetic Recording—Daniels. (See 1963.)

ANTENNAS AND TRANSMISSION LINES

621.315.052.63:621.396.44.018.8 1825

The Attenuation of Carrier-Frequency Waves on Lines due to Hoar-Frost—A. de Quervain. (*Bull. schweiz. elektrotech. Ver.*, vol. 42, pp. 949–953; December 1, 1951. In German.) Measurements were made over a period of several years on the 1,400-m high-voltage line between Schwägalp and Säntis, using carrier frequencies of 50 and 130 kc. Results are discussed in relation to the nature and thickness of the frost layer, and are compared with calculated and measured values found previously. Attentuation is attributed to dielectric losses in the frost layer; it increases with frequency.

621.315.21 1826

Plastic-Insulated Land Communication Cables—A. L. Meyers. (*Jour. Brit. IRE*, vol. 11, pp. 556–560; December, 1951. Discussion, p. 560.) A description of multiquad cables insulated and sheathed with Telcothene and of a new air-spaced coaxial cable in which the spacer is a helical membrane of Telcothene and the outer conductor a seamless aluminum tube. An account is given of jointing methods, and features of the cable in operation are considered.

621.315.212:[621.392.43+621.314.25 1827

Balancing and Transformation with Coaxial Lines—A. Ruhrmann. (*Telefunken Ztg*, vol. 24, pp. 237–250; December, 1951.) A comprehensive review of methods and equipment suitable for short and very short waves. 29 references.

621.392:621.317.34 1828

The Measurement of Image Impedance and Image Attenuation Coefficient—Guenot. (See 1961.)

621.392.2+621.385.029.63/.64 1829

Slow Electromagnetic Waves—A. I. Akhiezer and Ya. B. Faynberg. (*Uspekhi Fiz. Nauk*, vol. 44, pp. 321–368; July, 1951.) A mathematical discussion is presented on various methods for obtaining electromagnetic waves with phase velocity lower than that of light in vacuo. In the methods reviewed use is made of (*a*) waveguides partly filled with dielectric, (*b*) periodic structures such as chains of cavity resonators, and (*c*) helical transmission lines. The interaction between the slow waves and charged particles, which is of importance in the generation and amplification of uhf oscillations, is discussed and general laws governing these processes are derived.

621.392.22:517.512.2 1830

Fourier Transforms in the Theory of Inhomogeneous Transmission Lines—F. Bolinder. (*Acta polyt.* (Stockholm), no. 88, 84 pp.; 1951. Reprint. See 2909 of 1951.

621.392.26 1831

Waves on Inhomogeneous Cylindrical Structures—R. B. Adler. (PROC. I.R.E., vol. 40, pp. 339–348; March, 1952.) Analysis of some of the basic properties of exponential modes on passive cylindrical structures whose physical and electrical properties vary over the cross section.

621.392.26 1832

The Problem of the Elimination of Reflections in Waveguides with Varying Cross-Section—B. L. Rozhdestvenski and D. N. Chetaev. (*Compt. Acad. Sci.* (U.R.S.S.), vol. 79, pp. 427–430; July 21, 1951. In Russian.) It is proved mathematically that reflections in a rectangular waveguide with varying cross section can be eliminated if the waveguide is filled with a nonuniform medium the characteristics of which are related in a definite manner to the shape. As an example, the case of a waveguide bent at an angle (Fig. 2) is discussed. The conclusions reached remain fundamentally valid for dielectrics with finite conductivity. Complete elimination of reflections is possible only with an ideal medium with continuously graded characteristics, but it can be closely approached by a medium with discrete nonuniformity.

621.392.26 1833

The Short-Slot Hybrid Junction—H. J. Riblet. (PROC. I.R.E., vol. 40, pp. 180–184; February, 1952.) Theory and description of an X-band junction with outputs in phase quadrature, suitable for use in the construction of balanced duplexers and mixers. Over the frequency range 8.5–9.6 kmc, power equality within ± 0.25 db, isolation in excess of 30 db, and a SWR <1.07 are obtainable.

621.392.26:621.392.52 1834
Tunable Waveguide Filters—W. Sichak and H. A. Augenblick (*Elec. Commun.*, vol. 29, pp. 65–70; March, 1952.) Reprint. See 353 of March.

621.392.26:621.396.677:621.317.336.1 1835
Mutual Coupling of a Slot with a Dipole Antenna—W. J. Surtees. (Proc. I.R.E., vol. 40, pp. 208–211; February, 1952.) "A parameter is defined from which may be obtained the mutual coupling between a radiating slot, cut in a plane perfectly-conducting sheet, and a dipole fed at its base on the conducting plane. Using a slot cut in a sheet of copper and fed by a waveguide, experimental values of this parameter were obtained for various positions of the dipole relative to the slot. These values are plotted and compared with the theoretical ones, very good agreement being obtained."

621.392.26.012.3 1836
Square-Wave-Guide Attenuation—(*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, p. 32; December, 1951.) An abac enabling the attenuation of the $TM_{1,1}$ mode in a waveguide with square cross section to be calculated, for various materials and waveguide dimensions.

621.392.43 1837
A Method of Matching Balanced Transmitters to Unbalanced Transmission Lines by Means of Lumped Reactances—F. Moyano Reina. (*Rev. Telecomun.* (Madrid), vol. 6, pp. 2–4; December, 1951.) Design formulas are derived for the elements of a matching unit from consideration of its equivalent T network.

621.396.67 1838
Aerials at the Langenberg/Rhld High-Power Broadcasting Station, 1926–1951—A. Wurbs. (*Fernmeldetech. Z.*, vol. 4, pp. 525–530; December, 1951.) An account of the different wooden and steel mast and tower types erected during the 25-year period of operation of this station, finishing with descriptions of two modern usw antennas.

621.396.67.011.21 1839
The Input Impedances of Slit Antennas—S. Uda and Y. Mushiake. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 46–59; 1949.) The equations giving the input impedance are derived in terms of the electric and magnetic fields existing in complementary plates and slots, without using the concept of radiated power. The method can be applied to slot antennas of any length and of arbitrary shape.

621.396.67.029.63:621.397.6 1840
Receiving Antennas for U.H.F. Television—E. O. Johnson and J. D. Callaghan. (*Tele-Tech*, vol. 10, pp. 38–41, 82; December, 1951.) A review of the results of field tests carried out in the last three years near Washington, D. C., and Bridgeport, Conn. Types considered are (*a*) single and stacked fan dipoles, (*b*) single and stacked rhombic antennas, (*c*) stacked V antennas, (*d*) sheet, parabolic, and corner reflectors, (*e*) Yagi arrays. The performance of each type is analyzed as regards gain, directivity and bandwidth, with indication of suitable field of application.

621.396.677 1841
General Theory of Electromagnetic Horns—A. F. Stevenson. (*Jour. Appl. Phys.*, vol. 22, pp. 1447–1460; December, 1951.) The propagation of electromagnetic waves in a conducting horn of arbitrary shape is described exactly by an infinite set of linear differential equations, which represent a system of coupled E and H waves. By neglecting the coupling, only one equation is required to describe each E and H wave. The equations may then be solved approximately, and lead to the distinction between transmission regions and attenuation regions found by Barrow and Chu (1446 of 1939). The error caused by neglecting the coupling is examined. A detailed study is made of the propagation characteristics of several special shapes of horn.

621.396.677 1842
Radiation Patterns and Conductance of Slotted-Cylinder Antennas—O. C. Haycock and F. L. Wiley. (Proc. I.R.E., vol. 40, pp. 349–352; March, 1952.) "A theoretical solution to the radiation patterns and conductance is obtained by solving Maxwell's equations for the fields in the far zone, and requiring them to satisfy the known boundary conditions at the surface of the cylinder."

621.392.2 1843
Einführung in die Theorie der Ausbreitung elektromagnetischer Wellen in Leitungen und Hohlkabeln. (Introduction to the Theory of the Propagation of Electromagnetic Waves in Transmission Lines and Waveguides). [Book Review]—H. Bomke and J. Gefahrt. Publishers: Wissenschaftliche Verlagsanstalt, Stuttgart, Germany, 1950, 163 pp., 21.50 DM. (*Arch. elekt. Übertragung*, vol. 5, p. 530; November, 1951.) A clear mathematical presentation of the subject, developed from Maxwell's field equations.

621.396.67:[621.397.6+621.396.619.13 1844
Television and F.M. Antenna Guide. [Book Review]—E. M. Noll and M. Mandl. Publishers: Macmillan, New York and London, 1951, 311 pp., 41s. (*Electronic Eng.*, vol. 24, p. 139; March, 1952.) "Should be of great value to anyone interested in or working with vhf antennas."

CIRCUITS AND CIRCUIT ELEMENTS

537.312.6:621.315.59 1845
Conductivity of Electronic Semiconductors and Thermistors—N'Guyen Thien-Chi and J. Suchet. (*Onde élect.*, vol. 31, pp. 473–489; December, 1951.) A description of the electronic structure and conduction mechanism of semiconductors, together with a detailed review of commercial types of thermistor, their properties and applications. See also 44 of 1951, 165 and 199 of February.

621.3.015.7:621.387.4 1846
A Pulse-Height Distribution Analyzer—W. E. Glenn. (*Nucleonics*, vol. 9, pp. 24–28; December, 1951.) This 20-channel unit has a Du Mont Type K1059 cr tube with 10 collecting electrodes for pulse sorting. It has been used with a scintillation counter and an ionization chamber, and may be adapted for coincidence pulse analysis and the measurement of very short half-lives of radioactive materials.

621.314.2 1847
R. F. Current Transformers—T. J. Douma. (*Electronics*, vol. 25, pp. 156, 174; April, 1952.) An account of experiments indicating that if precautions are taken to suppress transformer-coil resonances, a rectifier type of instrument reading to 100 ma can be constructed for the frequency range 10 kc–50 mc.

621.314.25 1848
An Analysis of the Split-Load Phase Inverter—G. E. Jones, Jr. (*Audio Eng.*, vol. 35, pp. 16, 41; December, 1951.) Mathematical theory. When the inverter has substantially equal load impedances and is driven at a relatively low level, the hf response is about equal to that of a cathode follower.

621.314.3† 1849
Some Aspects of Magnetic-Amplifier Technique—F. E. Butcher and R. Willheim. (Proc. I.R.E., vol. 40, pp. 261–270; March, 1952.) Principles are described for the design of precision dc to dc magnetic amplifiers adaptable to a wide range of input impedance and output requirements. Graphical methods for design of the circuit elements and for predicting the performance of the assembly are set out and applied to push-pull amplifiers. Performance limits imposed by zero stability and response time are reviewed and estimated.

621.316.726.078.3:538.569.4.029.64 1850
Frequency Stabilization by Microwave Absorption—H. R. L. Lamont. (*Physica*, vol. 17, pp. 446–452; March/April, 1951.) Description of the method used for stabilization at wavelengths near 1.25 cm by means of NH_3 absorption lines, and discussion of possibilities for mm waves.

621.316.86:621.316.723.2 1851
Electronically Controllable Resistors—J. N. Thurston. (Proc. I.R.E., vol. 40, p. 315; March, 1952.) The bias current, and hence the dynamic resistance of a nonlinear resistor such as SiC or Ge, is controlled by a thermionic tube. The volt/ampere characteristic can be controlled by additional series and parallel linear resistors.

621.318.563 1852
Ferroresonant Flip-Flops—C. Isborn. (*Electronics*, vol. 25, pp. 121–123; April, 1952.) A low-resistance circuit comprising a capacitor and an iron-cored inductor has, with a suitably chosen operating voltage, two possible stable states, one characterized by low current and high inductive reactance, the other by high current and low capacitive reactance. Such a circuit can be triggered by application of a dc pulse. A combination of two of these circuits is described which has two inputs and two outputs, one output always being in antiphase to the other. This is well adapted to parallel-gated binary-counting systems, since each element is capable of considerable power gain. The complete circuit is given of a two-stage arrangement for binary operation and also an outline description of a decade unit using no tubes, nonlinear thyrite resistors replacing the usual diodes.

621.319.4+621.385.032.213.2]:621.771.3 1853
The Wire Capacitor and other Composite Drawn Products—J. L. H. Jonker and P. W. Haaijman. (*Philips Tech. Rev.*, vol. 13, pp. 145–151; December, 1951.) The wire capacitor is produced from a metal tube 20 cm long with outer diameter 20 mm and wall thickness 2 mm. A wire core 8 mm thick is centered in the tube and the annular space tightly packed with insulating powder. The whole is then hammered and drawn out to a wire about 40 m in length and less than 1 mm in diameter, which is cut into lengths each having a capacitance of about 100 pf. Two methods of removing one end of the jacket and insulation prior to soldering on the core lead are described. Suitable choice of insulating material and dimensions keeps the temperature dependence low, while dielectric loss fluctuation is minimized by aging. Application of these capacitors in IF transformers has enabled transformer units to be produced with dimensions of 60 mm×27 mm diameter and 36 mm×25 mm×10 mm, respectively. The drawing process has further been applied in the manufacture of indirectly heated cathodes.

621.392.016.2 1854
A Critical Study of the Circuit Concept—J. G. Chaney. (*Jour. Appl. Phys.*, vol. 22, pp. 1429–1436; December, 1951.) From Maxwells equations an expression for the complex power associated with a wire circuit is formulated and resolved into an input power and a power into the external field. The internal and external impedances of the circuit are obtained for unspecified current distributions. This con-

cept is extended to coupled circuits and the application to circuits with lumped elements is shown.

621.392.43 1855

Shorted Stubs of High Resonant Impedance—J. M. Diamond. (PROC. I.R.E., vol. 40, pp. 188–189; February, 1952.) Curves are developed from which the parameters of a coaxial or twin-line shorted stub of maximum resonance impedance can be determined, given the operating frequency, overall dimension, and the lumped-capacitance load at the open end.

621.392.43 1856

A Method of Matching Balanced Transmitters to Unbalanced Transmission Lines by Means of Lumped Reactances—Moyano Reina. (See 1837.)

621.392.43:621.396.645 1857

How to Design R.F. Coupling Circuits—W. B. Bruene. (*Electronics*, vol. 25, pp. 134–139; May, 1952.) Charts and practical rules are given which simplify the problem of selecting a suitable circuit for coupling an amplifier to an antenna, transmission line, or other load, resistive or complex, and determining the values of the components of the coupling circuit. Examples illustrate the design of L, T, II and II L coupling units.

621.392.5 1858

Maxwell's Reciprocity Law and Thévenin's Theorem applied to the Study of the Passive Quadripole—H. Thompson. (*Rev. gén. élect.*, vol. 60, pp. 516–520; December, 1951.) From Maxwell's reciprocity law, Thevenin's theorem and the principle of superposition, three equations relating the six characteristic parameters of the passive quadripole are formulated. Starting from one of these equations the construction of the circle diagram is explained and the significance of the different vectors in the diagram pointed out. The cases of the transformer and the asynchronous motor are also examined.

621.392.5:011.21 1859

The Impedance of a Network—J. Thouzéry. (*Radio franç.*, no. 12, pp. 1–5; December, 1951.) Application of tensor analysis in determination of the impedance of a complex network gives results in a useful form. The method is based on the introduction of a number of new variables equal to the number of closed loops in the network, and transformation of the corresponding matrix into one of lower order. The procedure is applied to Wheatstone's bridge and to a double T filter. See also 2670 of 1951.

621.392.5.018.782.4† 1860

The Effect of Delay Distortion on Waveform Fidelity in the Transmission of Signals—G. Schaffstein. (*Frequenz*, vol. 5, pp. 328–331; November/December, 1951.) Extension of the frequency range towards higher frequencies to improve the transmission quality of a wideband system is only effective if the phase errors caused by delay distortion at these frequencies are <180°. A brief analysis is made of the effects of delay distortion on (*a*) the ripple amplitude of square pulses, (*b*) triangular pulses, and (*c*) step voltages.

621.392.5.092 1861

Cascade Connection of 90-Degree Phase-Shift Networks—O. G. Villard, Jr. (PROC. I.R.E., vol. 40, pp. 334–337; March, 1952.) A method applicable to af networks for use in selective-sideband transmission and reception. Three networks are needed to obtain twice the rejection (in db) of a single circuit.

621.392.52 1862

Proposed Revision of the Conventional Method of Wave-Filter Design—P. J. Selgin. (*Bur. Stand. Jour. Res.*, vol. 47, pp. 479–490; December, 1951.) The revision consists of introducing new parameters termed "frequency numbers" (n), which permit the design to be made from specification of peak and cut-off attenuation rather than the idealized cut-off frequency as in the conventional method. The ratio of the impedances Z_s and Z_p of the component half-sections is expressed in terms of appropriately chosen values of n for high-pass, low-pass and symmetrical band pass filters. The calculation of the maximum permissible dissipation factor of terminating half-sections and of the filter elements is described and illustrated.

621.392.52:621.314.2 1863

Tuned Transformers and Filters of Maximum Bandwidth—G. Rutelli. (*Alta Frequenza*, vol. 19, pp. 26–49; February, 1950.) Theory is developed for the filter properties of tuned-circuit transformers having minimum attenuation for a given bandwidth or maximum bandwidth for a given attenuation. Design curves are given, whose use is explained by numerical examples.

621.392.52.029.4 1864

The Heterodyne Filter—G. Fant. (*Kungl. tekn. Hogsk. Handl.* (Stockholm), no. 55, 78 pp.; 1952.) A heterodyne filter for the speech-frequency range is described. It can be used as a band-pass filter with continuously variable low-frequency cut-off and high-frequency cut-off chosen independently in the frequency range 40–4,000 cps. The attenuation outside the pass band is >60 db and the steepness of the cut-off slopes is 1 db per 1 cps deviation. Bandwidth and low-frequency cut-off are varied by two separate controls. A band stop filter can also be obtained, with variable mid-frequency of the suppressed band but with bandwidths restricted to two alternative values, 300 cps and 800 cps. The filter is designed for articulation and hearing tests, and also as a wave analyzer of bandwidth continuously variable from 45 to 4,000 cps in the frequency range 40–20,000 cps.

621.396.6:621.317.7 1865

Components for Instruments—R. E. Hall and E. Coop. (*Proc. IEE*, Part II, vol. 98, pp. 738–752; December, 1951. Discussion, pp. 753–759.) Tubes, resistors and capacitors (both fixed and variable), transformers and chokes, meters, and quartz-crystal units are considered. A precision variable capacitor which can be made to extremely fine limits on a production basis is described. Details are given of the shortcomings and failures of some present-day components and of the effort to overcome them by using new materials, processes, methods of manufacture and by improved design. 21 references.

621.396.611.4 1866

Mode Conversion Losses in a TE_{01}-Type Cavity Resonator with Tilted End-Plate—K. Shimoda. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 378–383; September/October, 1951.) The reduction in Q value due to the tilting of the end-plate is deduced theoretically. In experiments with a copper resonator at 3kmc, the Q value was halved by a tilt of about ¼ degree, in good agreement with a calculated value.

621.396.615 1867

Theory of RC and RL Oscillators—A. Blaquière. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1434–1436; December 3, 1951.) Expressions for the stabilized amplitude are derived from theory previously given (335 of March), and are in satisfactory agreement with results obtained by other methods.

621.396.615:621.316.729 1868

On Synchronization of LC Oscillators—J. van Slooten. (*Electronic Appl. Bull.*, vol. 12, pp. 105–110; June/July, 1951.) Analysis of this problem has previously been based on the solution of a nonlinear differential equation. A simple method of calculation is shown to be possible if the oscillations produced are assumed to be nearly sinusoidal. The treatment gives a clear explanation of the way in which synchronization is effected by a series of pulses or by a signal of arbitrary wave form. The representation is similar to that of the synchronization of a multivibrator or a blocking oscillator.

621.396.615.018.424† 1869

Seven-League Oscillator—F. B. Anderson. (PROC. I.R.E., vol. 40, p. 328; March, 1952.) Corrections to paper abstracted in 81 of February.

621.396.615.11 1870

RC Oscillators—D. J. H. Admiral. (*Electronic Appl. Bull.*, vol. 12, pp. 111–131; June/July, 1951.) General considerations include discussion of (*a*) calculation of the oscillator frequency, (*b*) phase shift in the amplifier, (*c*) the necessity of amplitude limitation, (*d*) loading of the amplifier output circuit by the filter. A detailed treatment of the RC oscillator with bridge input circuit is given. The phase-correcting action of the bridge is explained and two methods of limiting the oscillation amplitude are considered: (*a*) by means of an incandescent lamp, (*b*) by means of NTC (negative-temperature-coefficient) resistors. The relation between phase shift in the amplifier and the current through the NTC resistor is examined and the effect of the ambient temperature considered. A detailed description, with complete circuit diagram, is given of an oscillator using ganged wire wound resistors for frequency adjustment and a NTC resistor for amplitude limiting. Temperature compensation is effected by including a NTC resistor in the voltage divider used to feed the power amplifier. A push-pull output can be obtained by operation of a switch. The output voltage is constant within 1 per cent from 20 cps to 20 kc.

621.396.615.17/.18 1871

Multivibrator Frequency Divider—R. R. Rathbone and R. L. Best. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 6–7, 29; December, 1951.) A unit using two multivibrators to cover the range 60 cps–200 kc and having associated delay circuits and pulse generators. Two output pulses are available, one of 0.1-μs duration variable up to 25 v across a 93-Ω load, the other a trigger pulse variable up to −100 v and with a 5–100-μs delay time.

621.396.615.17 1872

Generation of Sawtooth Voltages by Transformation of Sinusoidal Voltages—G. Francini. (*Alta Frequenza*, vol. 19, pp. 9–25; February, 1950.) Square pulses are first derived. These are applied to a RC or LC integrating circuit, the design of which is discussed. Voltages with symmetrical or asymmetrical triangular wave form of high precision may be obtained.

621.396.615.17.015.7 1873

2-Channel Rectangular-Pulse Generator—T. D. Graybeal. (*Electronics*, vol. 25, pp. 141–143; April 1952.) Description of a stimulator useful for studies in neuromuscular physiology. Pulse recurrence frequencies are adjustable from 0.5 to 500/second, pulse voltage from 0 to 80v and duration from 25 μs to 7.5 ms.

621.396.619.27.018.78 1874

Cubic Distortion in the Ring Modulator—L. Christiansen. (*Frequenz*, vol. 5, pp. 298–303; November/December, 1951.) Discussion of distortion occurring in the conducting and the blocking arms of a modulator using Ge rectifiers. Methods discussed for reducing this include the addition of resistance in the conducting arms, insertion of resistance in front of the modulator ring, division of the ring to form a push-pull circuit, and adjustment of the load resistance.

621.396.645:536.48 1875

The Possibility for using an Amplifier at Low Temperatures—A. N. Gerritsen and F. van den Burg. (*Physica*, vol. 17, pp. 930–932; October, 1951.) The gain of a 3-tube amplifier using Type-DF65 triodes was measured at ambient temperatures of 290°, 77°, 20° and 14°K. At the two lowest temperatures the gain was only about a third of that at room temperature, the reduction being partly accounted for by capacitance variation. A suitable choice of temperature-dependent resistors and capacitors and the operating points of tubes should enable a pentode amplifier to be constructed with a much higher gain at low temperatures.

621.396.645:621.396.822 1876

Noise Suppression in Triode Amplifiers—A. van der Ziel. (*Canad. Jour. Tech.*, vol. 29, pp. 540–553; December, 1951.) An experimental verification of previous theoretical work (2751 of 1950). Measurements of the correlated and uncorrelated parts of the induced grid noise are presented; for 6J4 and 6AC7 tubes only 30 to 40 per cent of the induced grid noise is correlated with the tube noise itself. The noise factor of a grounded-grid circuit and that of a grounded-cathode circuit with tuned anode-grid capacitance and properly detuned input circuit are shown to be identical, as required by the theory.

621.396.645.35 1877

Driftless D.C. Amplifier—F. R. Bradley and R. McCoy. (*Electronics*, vol. 25, pp. 144–148; April, 1952.) Description, with detailed circuit diagrams, of the high-gain phase-inverting amplifiers used in the REAC analogue computer. Drift is counteracted by means of a chopper and auxiliary amplifier which provide continuous balancing. A 2-page table shows input and output networks for use with the amplifier to generate different transfer functions for summation in the computer.

621.396.645.35.087.6 1878

A Frequency-Compensated Direct-Coupled Amplifier for Use with a Four-Channel Pen Recorder—J. A. Tanner and B. G. V. Harrington. (*Jour. Sci. Instr.*, vol. 28, p. 384; December, 1951.) Discussion on 1603 of 1951.

621.396.645.371:621.3.015.3 1879

The Transmission of the Step Function by the Negative-Feedback Amplifier—J. Müller. (*Fernmeldetech. Z.*, vol. 4, pp. 547–551; December, 1951.) Advantages of the transient-response method over the steady-state-response method for investigating wide-band and television circuits are indicated; the transient-response method can provide direct information regarding the quality of the television picture. The effect on transient response of varying the degree of negative feedback is examined theoretically for simple circuits; results are shown graphically. For more complex feed-back arrangements the effect is investigated experimentally using an oscillographic method with a square-wave generator (209 of February).

GENERAL PHYSICS

530.145.61:535.37 1880

Attempts to apply Wave Mechanics in [the theory of] Phosphorescence—D. Curie. (*Jour. Phys. Radium*, vol. 12, pp. 920–929; December, 1951.)

534.014.5 1881

Studies in Nonlinear-Vibration Theory—S. Fifer. (*Jour. Appl. Phys.*, vol. 22, pp. 1421–1428; December, 1951.) The stability of periodic solutions of the Duffing and van der Pol equations is determined. Application is made to the forced oscillations of a triode oscillator with a fifth-order tube characteristic.

535.37:539.2 1882

Transfer and Transport of Energy by Response Processes in Luminescent Solids—T. P. J. Botden. (*Philips Res. Rep.*, vol. 6, pp. 425–473; December, 1951.)

535.42 1883

Diffraction of Light by a Semi-transparent Sheet—F. B. Pidduck. (*Quart. Jour. Math.*, vol. 2, pp. 316–320; December, 1951.) The perfected Fresnel-zone theory for the cases of a thin disk and a straight edge is developed, with discussion of the assumptions made and the effect of the approximations involved. The field in which the theory may be applied excludes large angles of diffraction and small apertures, such as those of gratings.

535.43+534.26 1884

Multiple Scattering of Radiation by an Arbitrary Configuration of Parallel Cylinders—Twersky. (See 1803.)

537.221 1885

The Volta Effect—R. Bourion. (*Jour. Phys. Radium*, vol. 12, pp. 930–941; December, 1951.) The origin of contact potential differences is discussed, various measurement techniques recently used are described, and results obtained on metals and semiconductors are summarized and discussed in relation to the experimental conditions. The influence of surface structure, connected with the anisotropy of the work function for single crystals, is also considered.

537.315:539.23 1886

Contact-Potential Variations on Freshly Condensed Metal Films at Low Pressures—L. L. Antes and N. Hackerman. (*Jour. Appl. Phys.*, vol. 22, pp. 1395–1398; December, 1951.) Curves showing the variation with time of the contact potentials of films of Al, Cu, Au, Ni, Fe, Cr are shown for different pressures, Pt serving as reference standard. Resistance measurements are correlated with the potential measurements.

537.52 1887

Theoretical and Experimental Research on the Excitation of Gases by Ultra-High Frequencies—M. Bayet. (*Rev. Sci.* (Paris), vol. 89, pp. 351–394; November/December, 1951.) A comprehensive treatment. Tests with short-duration 600-mc pulses show an increase in the recombination coefficient of positive ions and electrons when the electron velocity decreases on suppression of the excitation. 79 references.

537.523.4:535.89 1888

A Repetitive Spark Source for Shadow and Schlieren Photography—G. K. Adams. (*Jour. Sci. Instr.*, vol. 28, pp. 379–384; December, 1951.) Description of two methods of producing sparks of high brightness and short duration at high repetition frequencies. The first method gives a limited number of discharges (4–8) at individually determined time intervals of 10 μs or more with an error <1 μs. The second method, used with a rotating-drum or -mirror camera, gives a greater number of light pulses at preset repetition frequencies up to 10,000 per second. The energy of each spark is of the order of 1 joule, and the effective photographic duration <1 μs.

537.525 1889

Self-Magnetic Field in High-Current Discharges—M. Blackman. (*Proc. Phys. Soc.*, vol. 64, pp. 1039–1045; December, 1951.) Theory is developed for the case of a non-conducting envelope. Comparison with the theory of Thonemann and Cowhig (2146 of 1951) shows good agreement.

537.525.6 1890

Behavior of Gas-Discharge Plasma in High-Frequency Electromagnetic Fields—L. Goldstein and N. L. Cohen. (*Elec. Commun.*, vol. 28, pp. 305–321; December, 1951.) Studies were made of the complex conductivity of the gas-discharge plasma situated in a hf em field. Measurements were also made of the noise produced by the discharge and of the transmission characteristics when the discharge was pulsed. For the experiments in the frequency band 1,500–2,300 mc, the monatomic rare gases were used, the discharge tube being arranged as part of a coaxial-line circuit. In supplementary measurements on phase velocity, made at a frequency of 9,450 mc, the discharge was set up in a waveguide. Experimental results are given in some detail.

537.525.6:538.56 1891

Current Fluctuations in the Direct-Current Gas Discharge Plasma—P. Parzen and L. Goldstein. (*Elec. Commun.*, vol. 29, pp. 71–74; March, 1952.) Reprint. See 2974 of 1951.

537.582 1892

Some Remarks on the Equation of Thermionic Emission—Y. Watanabe. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 1–9; 1949.) Discussion of the Richardson and Nordheim emission equations, which are consistent if a different value of the apparent electron density is used in Richardson's equation.

538.312:621.318.423:513.647 1 1893

A General Theory of the Helical Line—H. Kaden. (*Arch. elekt. Übertragung*, vol. 5, pp. 534–538; December, 1951.) Formulas are derived for the em field and the transmission parameters of a helical line for any values of operating frequency and pitch of the helix. The characteristic impedance is a measure of the energy transmitted in the axial direction; when plotted against pitch, for a given wavelength it exhibits a minimum and approaches a constant value for very small values of pitch. Both delay time and attenuation increase monotonically with decreasing pitch.

538.56:537.2-7 1894

Radiation Capacity—H. Weyl. (*Proc. Nat. Acad. Sci.*, vol. 37, pp. 832–836; December, 1951.) The mathematical theory of capacitance in the electrostatic field is outlined and the extent of its applicability to the unquantized radiation field is discussed.

538.566.2 1895

Note on Cut-Off Frequency in Dielectric Plates—H. Ott. (*Z. angew. Phys.*, vol. 3, pp. 456–458; December, 1951.) Using Fresnel's optics formulas it is possible to determine whether a lower cut-off frequency exists for the propagation of em waves in a dielectric plate under given boundary conditions; a formula is derived for calculating this cut-off frequency. The question whether waves of unlimited length can be excited in the plate is discussed briefly as a separate issue.

538.569.4.029.65 1896

Microwave Spectroscopy in the Region from Two to Three Millimeters: Part 2—C. M. Johnson, R. Trambarulo and W. Gordy. (*Phys. Rev.*, vol. 84, pp. 1178–1180; December 15, 1951.) Report of results obtained by using the 4th and 5th harmonics of K-band klystrons. A list of all frequencies so far measured in the 2–3-mm range is included. See also 1668 of 1950 (Gilliam et al.) and *Phys. Rev.*, vol. 83, p. 1061; 1951. (Anderson, Johnson and Gordy.)

621.39.001.11 1897

Information Aspect of some Uncertainty Relations—R. Vallée. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1580–1581; December 19, 1951.) An extension of the relation discussed by Gabor (1057 of 1947) between the uncertainty regarding the mean time of occurrence of a signal and the uncertainty regarding its mean frequency.

621.39.001.11:517.433 1898

"Observation Operators" and Information Theory—R. Vallée. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1428–1430; December 3, 1951.) A general expression for the maximum quantity of information furnished by any experiment is derived, using the observation operators defined in 1570 of July. Among the particular cases to which this applies is Shannon's expression for maximum information content of a transmission channel.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 1899

The Velocity Distribution of Sporadic Meteors: Part 1—M. Almond, J. G. Davies and A. C. B. Lovell. (*Mon. Not. R. Astr. Soc.*, vol. 111, pp. 585–608; 1951.) Report of three sets of measurements of velocity distributions, using radio-echo diffraction technique. No evidence was found for a significant hyperbolic velocity component.

523.72:537.562 1900

Excitation of Electron Oscillations in a Shock Wave. Application to Radio Astronomy—J. F. Denisse and Y. Rocard. (*Jour. Phys. Radium*, vol. 12, pp. 893–899; December, 1951.) Analysis of the propagation of a shock wave in a strongly ionized medium. Effects of diffusion and thermal diffusion combine to concentrate electrons ahead of the wave front. This causes a polarization of the medium which finally limits the diffusion effects, and a distribution of electron velocities widely different from a Maxwellian distribution; if the energy of the shock wave is sufficiently great, the electrons may be divided into two groups with different mean velocities. Amplified plasma oscillations result; this mechanism may explain the most intense solar rf radiation.

523.72:621.396.822 1901

The Position and Movement on the Solar Disk of Sources of Radiation at a Frequency of 97 Mc/s: Part 1—Equipment—A. G. Little and R. Payne-Scott. (*Aust. Jour. Sci. Res., Ser. A*, vol. 4, pp. 489–507; December, 1951.) Description of an interferometer using spaced antennas. The interference pattern is produced by changing the phase of one antenna relative to that of the other 25 times per second, thus swinging the lobe pattern across the source. Position and polarization of a source can be determined in 1 second and the accuracy of location is within ±2 minutes of arc. The system may be used as a fixed-lobe interferometer to measure the angular size of a source. Part 2: 1902 below.

523.72:621.396.822 1902

The Position and Movement on the Solar Disk of Sources of Radiation at a Frequency of 97 Mc/s: Part 2—Noise Storms—R. Payne-Scott and A. G. Little. (*Aust. Jour. Sci. Res., Ser. A*, vol. 4, pp. 508–525; December, 1951.) Analysis of experimental results obtained with the equipment described in 1901 above. Storm radiation is associated with the largest sunspot of a group and not with the rest of the group; the spot size gives the best criterion for the occurrence of storms. The direction of rotation of the circular polarization depends on the magnetic polarity of the spot. Deviations between the apparent positions of the radio storm centers and visible spots are explained by assuming the origin of the storm radiation to be high in the corona. Part 1: 1901 above.

523.8:621.396.822 1903

An Attempt to Measure the Annual Parallax or Proper Motion of Four Radio Stars—F. G. Smith. (*Nature* (London), vol. 168, pp. 962–963; December 1, 1951.) The errors in interferometer measurements of the positions of radio stars due to phase-retardation inconstancy of the transmission lines to the antennas of the interferometer may be largely eliminated by measuring the relative positions of radio stars. From the results of two series of measurements during 1951 of the apparent right ascension of four intense radio stars, using wavelengths of 3.7 m and 1.4 m, it is concluded that their angular movements are attributable to ionospheric refraction. The distances of these stars are probably greater than ½ parsec.

523.85:621.396.822 1904

A Radio Survey of the Cygnus Region: Part 1—The Localized Source Cygnus (1)—R. H. Brown and C. Hazard. (*Mon. Not. R. Astr. Soc.*, vol. 111, pp. 576–584; 1951.) An account of the observations, with comparison of the values of the celestial coordinates and intensity of the source with the results obtained by other observers using interferometer methods.

523.854:621.396.822 1905

Galactic Radiation at Radio Frequencies: Part 4—The Distribution of Radio Stars in the Galaxy—J. G. Bolton and K. C. Westfold. (*Aust. Jour. Sci. Res., Ser. A*, vol. 4, pp. 476–488; December, 1951.) The distribution was deduced from a previous survey at 100 mc. The effects of absorption and emission within the interstellar gas are considered and found negligible at this frequency. The existence of an isotropic background radiation is discussed and estimates are made of the local number density, the flux from a typical star, and the distances of certain observed sources. Part 3: 866 of 1951.

523.854:621.396.822 1906

Observations of Galactic Radiation at Frequencies of 1210 and 3000 Mc/s—J. H. Piddington and H. C. Minnett. (*Aust. Jour. Sci. Res., Ser. A*, vol. 4, pp. 459–475; December, 1951.) The intensity was measured near the galactic center and a new discrete source discovered whose spectrum resembles that of an optically thin thermally radiating gas. Radiation was also observed from the Crab nebula, Centaurus and the moon. No radiation was observed from the nebulae M31 and NGC 7293. Details are given of the antenna systems and the experimental technique. An antenna-temperature change of 0.4°K was detectable.

550.38 1907

Indices of Geomagnetic Activity of the Observatories Abinger, Eskdalemuir and Lerwick, August to December 1951.—(*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 196–199; 1952.) *K*-indices for 3-hour intervals.

550.384 1908

World-wide Simultaneous Magnetic Fluctuations and their Relation to Sudden Commencements—W. Jackson. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 160–172; 1952.) Analysis of magnetograms obtained at a number of widely distributed stations during a period of 2½ years reveals the existence of world-wide perturbations, lasting for several hours and often dissociated from large disturbance, with properties similar to those of sudden commencements.

550.384.4:551.510.535 1909

Anomalies in the Diurnal Variation of the Geomagnetic Field and their Correlation with the Winds in the Lower Ionosphere—H. Wiese. (*Z. Met.*, vol. 5, pp. 373–377; December, 1951.) The quiet-day variations of the geomagnetic field components are examined. Anomalies observed in the month-to-month variation of the amplitude and phase of the daily fluctuations are related to seasonal inversions of steady circulating currents in the lower ionosphere.

550.385"1951.09.21" 1910

New Observations of Very Rapid Pulsations during a Magnetic Storm—G. Gibault. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1655–1656; December 19, 1951.) See also 725 of 1947.

551.510.52/.53:546.214 1911

Ozone Variations in the Troposphere and Stratosphere—E. Regener. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 173–182; 1952. In German.) In the air close to the ground, the ozone content may sink to zero owing to the proximity of oxidizable substances. In the troposphere, advection is the main cause of fluctuations. Large variations of the vertical distribution of ozone have been found recently by spectrographic observations in balloon ascents. To explain this, large-scale horizontal and vertical movements of air at great heights must be assumed.

551.510.535 1912

The Nature of the Sporadic-E Layer and Turbulence in the Upper Atmosphere—R. Gallet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1649–1650; December 19, 1951.) A theory is advanced-according to which the properties of the E_s layer are produced not by additional ionization but as a result of turbulence in the ionosphere, where fluctuations of atmospheric density are accompanied by fluctuations of electron concentration. The corresponding fluctuations of dielectric constant are relatively very much larger than the fluctuations of density at frequencies near the critical frequency. A more complete theory, in course of development, will explain the different effects at equator and polar regions by taking account of the magnetic field. The theory is relevant to the "bright ring" of the solar corona.

551.510.535 1913

The Theory of Magneto-ionic Triple Splitting—O. E. H. Rydbeck. (*Acta polyt.*, (Stockholm), no. 83, 40 pp.; 1951.) Reprint. See 2714 of 1951.

551.510.535 1914

Continental Sporadic-E Activity—N. C. Gerson. (*Trans. Amer. Geophys. Union*, vol. 32, pp. 26–30; February, 1951.) Analysis of observations by American amateurs of contacts effected on 17th–18th June 1949 on 50–54 mc by means of sporadic-E reflection. The results indicate drift of the reflecting regions at a speed of about 175 km per hour.

551.510.535 1915

Light—Inadequacy of the Ultraviolet Theory of Ionization in the E-Layer—E. F. George. (*Jour. Frank. Inst.*, vol. 252, pp. 493–500; December, 1951.) Values of E- and F_1-layer parameters for the period 1944–1949 are analyzed. The distribution of ionization is symmetrical about the magnetic rather than the geographical equator, and is periodic with respect to latitude; possible correlation of this periodicity with large-scale air movements is discussed.

551.510.535:551.55 1916

Measurement of Winds in the Ionosphere—G. J. Phillips. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 141–154; 1952.) Report of investigations extending over a period of more than two years, based on observation of the fading pattern of ionospheric echoes of 2.4-mc signals, receiving antennas being located at the corners of a right-angled triangle (shorter sides 130 m), with a fourth antenna available for checking purposes. The antennas were switched in turn to a single receiver, with gating arrangements enabling four records of echo amplitude to be obtained on a single strip of moving film. The results probably refer to motion of the air at heights of 100–120 km. Regular daily and seasonal changes in the direction of the horizontal movement were found, partly attributable to a solar tide of amplitude 16 m per second. The order of the velocities, 70 m per second, agrees with information obtained by other methods.

551.510.535:621.3.087.4 1917

A Simple Ionosphere Sounder—R. Aschen and P. Gaillard. (*Philips Tech. Rev.*, vol. 13, pp. 152–163; December, 1951.) A simple account of the Breit and Tuve method, together with details of equipment suitable for amateur use, for another account of which see 2069 of 1949 (Maguer).

551.510.535:621.3.087.4 1918

Automatic Ionosphere Recorder—J. M. Carroll. (*Electronics*, vol. 25, pp. 128–131; May, 1952.) Description, with block diagram, of the pulse generator, transmitter, receiver and recorder elements of the Model-C3 equipment used by the Central Radio Propagation Laboratory, National Bureau of Standards. Echoes of pulse transmissions, of frequency swept over the range 1–25 mc, are recorded continuously from the cro display. Some circuit details are shown of the temperature-controlled vfo, the wide-band amplifiers, and the pulse inverter and cathode-follower keyers.

551.594.5 1919

The Aurorae. [Book Review]—L. Harang. Publishers: Chapman & Hall, London, 1951, 21s. (*Jour. Atmos. Terr. Phys.*, vol. 2, pp. 199–200; 1952.) Vol. 1 of the International Astrophysics Series.

LOCATION AND AIDS TO NAVIGATION

621.396.9 1920

Organ-Pipe Radar Scanner—K. S. Kelleher and H. H. Hibbs. (*Electronics*, vol. 25, pp. 126–127; May, 1952.) Description of an experimental model using a rotating horn to feed in succession a set of 36 waveguide elements arranged with their output ends in a continuous line. When used in conjunction with a 6-ft parabolic cylindrical reflector of focal length 57.6 inch, the secondary patterns had good beam-width and side-lobe characteristics.

621.396.9 1921

Radar Signal Sampler Compresses Bandwidth—W. Otto. (*Electronics*, vol. 25, pp. 132–135; April, 1952.) Description and circuit details of equipment which enables ppi radar presentations to be relayed from aircraft to points far beyond line of sight. The video signal is sampled and synthesized into a similar wave form with a sufficiently low recurrence frequency for handling by a lf radio link.

621.396.9 1922

Radar Buoys [beacons] and their Special Circuits—J. Moline. (*Radio franc.*, no. 12, pp. 12–16; December, 1951.) Discussion of transponder systems. Discriminating circuits for eliminating interrogation pulses either too short, too long, or outside a given range, are shown. The operation of pwm and pcm responder circuits is described and the application of a multi-electrode cathode-ray tube for pcm is outlined. See also 3418 of 1948.

621.396.932/.933 1923

The German Decca Network—(*Telefunken ZTG*, vol. 24, pp. 251–252; December, 1951.) Brief details of the equipment of the master and the three associated slave transmitting stations, designed to extend the Decca system coverage over the whole of the North Sea and the West German Federal Republic.

621.396.932(083.75) 1924

British Commercial Radar—M. Hobbs. (*Electronics*, vol. 25, pp. 174–186; April, 1952.) The Ministry of Transport specifications for marine radar are outlined and the principal characteristics of radar sets for the 9.32–9.5 kmc band, manufactured by five British firms are tabulated. A description is given of the operation of magnetic modulators capable of handling high powers. These comprise cascade arrangements of "pulsactors," saturable reactive switching elements. The characteristics of a typical magnetic modulator are: peak power, 150kw; pulse voltage, 13kv; pulse current, 12.5a, and duration and rate 0.25μs and 1,500 per second. Such modulators have long and trouble-free life, simple auxiliary circuits, freedom from radiated interference and negligible maintenance costs.

621.396.933 1925

Long-Range-Navigation Instrumentation—B. Alexander. (*Elec. Commun.*, vol. 29, pp. 9–11; March, 1952.) Discussion of experience gained with the Navaglobe system [1092 of 1947 (Busignies et al.)]. The aperture of the transmitting antenna is the most characteristic feature of ground-based long-range navigational systems; a dimension of $\lambda/2$ is the best compromise for this aperture.

621.396.933 1926

Some Future Developments in Aeradio Scott-Farnie and Forsyth-Grant (See 2037.)

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 1927

Notes on the Ionization Gauge—L. Riddiford. (*Jour. Sci. Instr.*, vol. 28, pp. 375–379; December, 1951.) "The sensitivity of the hot-filament ionization gauge is in good agreement with theoretical values calculated on the basis of a physical picture of a stream of electrons passing to and fro through the ionizing space. The calibration curve is not exactly linear, the sensitivity decreasing as the pressure increases from 10^{-4} to 10^{-3} mm of mercury. The tungsten filament in such a gauge 'pumps' oxygen at a rate which is in agreement with earlier work of Langmuir. The remanent molecules which determine the ultimate pressures of diffusion pumps are strongly adsorbed by the gauge, which behaves as a pump of constant speed S. An account of the nature of this phenomenon is given."

533.5 1928

Vacuum Technique—its Application to Radio and Electronics—D. Latham and B. D. Power. (*Jour. Brit. IRE*, vol. 11, pp. 561–568; December, 1951.) "A survey of methods of producing and measuring high vacua, special emphasis being given to modern tendencies in the manufacture of electronic tubes. Methods of making vacuum joints and of leak detection are described. Various vacuum processes of interest to the electronics industry such as evaporation, impregnation and resistor manufacture are also discussed."

534.232:538.652 1929

Magnetostrictive Vibration of Prolate Spheroids. Analysis and Experimental Results—F. J. Beck, J. S. Kouvelites and L. W. McKeehan. (*Phys. Rev.*, vol. 84, pp. 957–963; December 1, 1951.) Investigation of systems in which specimens of Ni and Ni-Fe alloys are allowed to vibrate in the fundamental mode while subjected simultaneously to a steady and a hf magnetic field. From the effects induced in the hf magnetizing coil, values are computed for the incremental permeability, magnetostriction constant, modulus of elasticity and dissipation constant of the specimens. Results are consistent with domain theory. See also 1873 of 1951 (Kouvelites and McKeehan).

535.215:546.817.221 1930

The Photovoltaic Effect in Natural Lead Sulphide—R. Lawrence. (*Aust. Jour. Sci. Res., Ser. A*, vol. 4, pp. 569–578; December, 1951.) Report of investigation of the use of Australian samples of natural PbS as radiation detectors in the near infrared. A close correlation is established between the electrical properties of bulk samples and those given by other workers for thin PbS films.

535.371 1931

R.T.M.A. Screen Phosphors—(*Oscillographer*, vol. 12, pp. 3–6; October/December, 1951.) A review and tabulation of the characteristics of standard R.T.M.A. screen materials and their applications in oscillography.

537.226:537.29 1932

Field Dependence of the Dielectric Constant—J. J. O'Dwyer. (*Proc. Phys. Soc.*, vol. 64, pp. 1125–1132; December 1, 1951.) Fröhlich's calculations (1674 of 1950) for the dielectric constant of a material are extended so as to include the first term of the field dependence. The theory is applied to Kirkwood's model of a dipolar liquid.

537.228.1:621.396.611.21 1933

New Synthetic Piezoelectric Crystals in Electroacoustics and High-Frequency Technique—F. Spitzer. (*Arch. elekt. Übertragung*, vol. 5, pp. 544–554; December, 1951.) Theory of piezoelectric crystals is discussed with particular reference to the different requirements for af and hf purposes. Properties are tabulated of a large number of materials investigated as possible substitutes for Rochelle salt and quartz respectively, for the two fields of application. Consideration of the values of and relation between the piezoelectric, elastic and dielectric properties enables the best materials to be chosen.

537.311.33:621.396.822 1934

Noise in Photosensitive Semiconductors—P. Gorlich. (*Optik*, vol. 8, pp. 512–516; November, 1951.) Results of qualitative investigation of semiconductor noise suggest two sources of noise additional to thermal, shot, flicker and lattice effects [3035 of 1950 (van der Ziel)]. These are (a) boundary layers formed by the application of a direct voltage; their effect may be calculated by the formula of Mataré (2976 of 1950) or estimated by MacFarlane's theory (910 of 1951). (b) grain boundaries in crystal structures which may affect the electron flow.

538.221 1935

Ferromagnetic Properties of Semioxidized Iron and Iron-Cobalt Powders—F. Lihl. (*Acta Phys. austriaca*, vol. 4, pp. 360–379; May, 1951.) An experimental investigation was made of the influence of the degree of reduction of the initial iron salts on the coercive force, remanence, and BH product of permanent magnets moulded from powders. Highest values of coercive force and BH product respectively are obtained at different incomplete stages of reduction. When the initial material consists of Fe-Co mixed crystals, coercive force and BH product increase with Co content. The experimental results support Néel's theory.

546.289:548.55:621.396.822 1936

Shot Noise in Germanium Single Crystals—G. B. Herzog and A. van der Ziel. (*Phys. Rev.*, vol. 84, pp. 1249–1250; December 15, 1951.) The noise ratio n of a particular Ge filament carrying current I at frequencies f in the range 1–1,600 kc is represented by $n=1+AI^2/f+BI\cdot/[1+(f/f_0)^2]$, where A and B are constants and $f_0(1.5\times10^5\text{cps})$ is practically independent of I. The third term is interpreted as representing the shot noise of the holes, in accordance with the theory of semiconductors [3035 of 1950 (van der Ziel)]. For another filament, deviations from the inverse-frequency law could not be detected.

546.289:621.314.7 1937

Electric Forming of n-Germanium Transistors using Donor-Alloy Contacts—R. L. Longini. (*Phys. Rev.*, vol. 84, p. 1254; Decem-

. . . 15, 1951.) The suggestion is made that lattice vacancies diffuse from the collector probe into the Ge, this process preceding the diffusion of donor impurities, so that a *p-n* hook is set up near the probe. Values of current gain higher than can be expected from mobility considerations alone are thus made possible.

546.817.221+546.817.241 1938

The Optical Constants of Lead Sulphide and Lead Telluride in the Region 0.5–3 Microns—D. G. Avery. (*Proc. Phys. Soc.*, vol. 64, pp. 1087–1088; December 1, 1951.) Refractive indices and absorption constants are shown graphically for polished PbS and PbTe crystals; the method of measurement is outlined.

621.314.6:537.311.33 1939

On Rectifiers—W. C. van Geel. (*Physica*, vol. 17, pp. 761–776; August, 1951.) Combinations of (*a*) excess-semiconductor/deficit-semiconductor, (*b*) Al/Al_2O/semiconductor, (*c*) metal/resin-layer/semiconductor were all found to have rectification properties. It is suggested that in all three cases the contact between two layers with charge carriers of opposite sign is the source of the rectification effects.

621.315.61:621.317.7 1940

The Properties of Insulating Materials Used in Instruments—C. G. Garton. (*Proc. IEE*, Part II, vol. 98, pp. 728–737; December, 1951. Discussion, pp. 753–759.) Mechanical and electrical properties, and chemical structure of newer synthetic materials, are discussed and tabulated. The use of composite material is also discussed. 19 references.

621.315.612.4:546.831.824-31 1941

Effects of Firing Temperature on the Dielectric Properties of Barium-Titanate Ceramics—A. Kobayashi and H. Hino. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 371–373; September/October, 1951.)

621.318.13:621.317.7 1942

Some Special Characteristics of Soft Magnetic Materials used in Instrument Manufacture—G. A. V. Sowter. (*Proc. IEE*, Part II, vol 98, pp. 714–727; December, 1951. Discussion, pp. 753–759.) Merits of high-μ materials for various applications are considered. Losses and harmonic generation are calculated, especially as affected by nonuniform flux distribution. Magnetostriction and variation of magnetic properties with temperature are treated, and costs of various materials are compared. Many design curves are given.

621.396.611.21.002.2 1943

The Deposition of H.F. Crystal Electrodes by Vacuum Coating—L. Holland. (*Electronic Eng.*, vol. 24, pp. 10–13; January 1952.) Comparison of sputtering and evaporation methods for depositing thin gold films as electrodes or for frequency adjustments. Direct adjustment of crystal frequency during coating is possible with the evaporation technique. An evaporation unit is described which has been used successfully for coating and calibrating 4-mc crystals.

MATHEMATICS

517.56 1944

Epicycloidal Functions and some New Relations between Bessel Functions—C. Agostinelli. (*Rend. Accad. nac. Lincei*, vol. 11, pp. 339–344; December, 1951.)

517.93 1945

Nonlinear Systems. A Method of Solution by Graphical Analysis—(*Elec. Times*, vol. 120, pp. 1128–1129; December, 1951. Discussion, p. 1129.) Short account of a paper on "A Graphical Analysis for Nonlinear Systems" by Miss Pei-Su Hsia, describing a method of solving second-order nonlinear differential equations.

521.401.3:621.385.83 1946

Perturbation Characteristic Functions and their Application to Electron Optics—P. A. Sturrock. (*Proc. Roy. Soc. A*, vol. 210, pp. 269–289; December 20, 1951.)

681.142 1947

Some Limitations on the Accuracy of Electronic Differential Analyzers—A. B. Macnee (Proc. I.R.E., vol. 40, pp. 303–308; March, 1952.) Discussion of probable errors in the solution of differential equations with constant coefficients by means of analogue computers.

681.142 1948

Automatic Calculating Machines—M. V. Wilkes. (*Jour. Roy. Soc. A.*, vol. 100, pp. 56–90; December 14, 1951.) An account of the development of digital computers, with descriptions of their operation and use.

681.142 1949

Multi-Stable Magnetic Memory Techniques—J. D. Goodell and T. Lode. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 3–5; December, 1951.) Using suitable materials, more than two stable states in a magnetic core may be obtained by arrangements ensuring that the flux change, which is proportional to the voltage/second integral of the applied force, takes place in discrete steps. Core magnetization is then related to the number of applied pulses with an accuracy limited mainly by the slip-back to remanence from saturation. Information may be stored or read in many different ways, either by pulse train, or by single pulses, whose duration or amplitude is varied. The block diagram of an analogue-to-digital translation system is shown; other applications of magnetic structures of this type are as light-weight high-speed storage devices and in arithmetical computing units.

MEASUREMENTS AND TEST GEAR

531.761:621.318.5 1950

Time-Interval Measurements on Installations during Normal Operation—J. Schalkwijk. (*Commun. News*, vol. 12, pp. 49–56; December, 1951.) A three-range, direct-reading, mains-operated instrument for measuring closing and release times of telephone-exchange relays, or for any time intervals (up to 1 second) defined by a voltage variation at beginning and end of interval.

621.3.018.4(083.74.) 1951

Frequency Standards in the Microwave Region—R. Ferrero. (*Ricerca Sci.*, vol. 21, pp. 2142–2144; December, 1951.) The 1-mc signal from a quartz oscillator is multiplied to 200 mc, and is then passed to three Ge crystals acting as harmonic generators, which, with their associated cavity filters, provide signals of 2, 3 and 10kmc respectively. The power available is of the order of several milliwatts.

621.317:537.71(083.74) 1952

The Accuracy of Measurement of Electrical Standards—A. Felton. (*Proc. IEE*, Part II, vol. 98, pp. 694–700; December, 1951. Discussion, pp. 710–713.) "The difference between 'international' and 'absolute' electrical units is explained, and the reason for the change made in 1948 from one system to the other is given. The uncertainty of the electrical units defined in terms of length, mass and time is estimated to be about 20 parts in a million, although the accuracy of comparison of electrical quantities may be as high as one part in ten million. The loss of accuracy in the measurement of voltage, current and power when transferring from direct to alternating current is explained: it is estimated that these quantities cannot be measured at power frequencies to better than one part in 10,000." 20 references.

621.317.3:551.594.6 1953

A Statistical Approach to the Measurement of Atmospheric Noise—R. S. Hoff and R. C. Johnson. (Proc. I.R.E., vol. 40, pp. 185–187; February, 1952.) Description of a measurement method based on determination of the fraction of time that the noise envelope exceeds certain reference levels during an interval of several seconds. Results obtained are compared on a statistical basis with those obtained by other methods.

621.317.3.029.5/.6 1954

High-Frequency Measurement Technique—W. Druey. (*Bull. schweiz. elektrotech. Ver.*, vol. 42, pp. 989–1000; December 15, 1951. In German. Discussion, pp. 1000–1003, in French and German.) A survey paper dealing with measurements of current, voltage, impedance, dielectric properties, frequency and time. A description is given of a telemetry system using a hf channel, and the advantag s of distributed amplifiers for wide-band operation in cathode-ray oscillography are discussed.

621.317.326 1955

A Method for the Measurement of the Peak Voltage of Periodic Low- or High-Frequency Pulses—W. Hasselbeck. (*Funk. u. Ton*, vol. 5, pp. 617–626; December, 1951.) Positive pulses are applied so as to charge a capacitor in series with a diode and resistance; pulses produced across the resistance are amplified, with reversal of sign, and then applied so as to give a supplementary charge to the capacitor via a second diode; the charging continues until the peak voltage is attained when the first diode cuts off. Practical circuits based on this principle are described, with particular attention to the amplifier design. The measurement error is discussed; its magnitude is about 1 per cent for rectangular lf pulses with a mark/space ratio of 1:10,000.

621.317.335.3.029.62/.63 1956

Use of Coaxial Line terminated by Different Types of Impedance for Permittivity Measurements at Metre and Decimetre Wavelengths—A. Lebrun and R. Arnoult. (*Compt. Rend. Acad. Sci.* (Paris), vol. 233, pp. 1591–1593; December 19, 1951.) Description of a method and apparatus permitting measurement accurate to within 1 per cent. Three different types of dielectric cell are shown. The apparatus is suitable for measurements at controlled temperature, and can be used as a wavemeter.

621.317.335.3.029.64:546.217 1957

The Permittivity of Air at a Wavelength of 10 Centimeters—W. E. Phillips. (Proc. I.R.E., vol. 40, p. 164; February, 1952.) Discussion on paper abstracted in 2827 of 1950.

621.317.336.029.64 1958

Inductive Probe for Microwave Measurement Lines and Near-Field Meters—F. Tischer. (*Acta polyt.*, (Stockholm), no. 81, 18 pp.; 1951. In German.) Reprint. See 3048 of 1951.

621.317.336.029.64:621.396.611.4 1959

Resonance-Resistance Measurements at Centimeter Wavelengths—H. Döring and W. Klein. (*Arch. tech. Messen.*, no. 191, pp. T135–T136; December, 1951.) In the measurement of the resonance resistance of ordinary cavity resonators, the determination of attenuation and resonance reactance ["Schwingwiderstand" $=\omega_0 L = 1/\omega_0 C$] is involved, but for cavity resonators with extremely small losses a method involving insertion of high-value resistors in parallel with the resonator [see 2755 of 1943 (Borgnis)] is preferable. With a method based on wavelength measurements, errors are <5 per cent but with a method involving detector meter readings the error may be as high as 10 per cent. The results of measurements on three cylindrical cavity resonators and one coaxial-line cavity are in fairly good agreement with values calculated from loss data for the dielectric rods used as auxiliary resistors.

621.317.337:621.396.611.4 1960
The Q of a Microwave Cavity by Comparison with a Calibrated High-Frequency Circuit—H. LeCaine. (PROC. I.R.E., vol. 40, pp. 155–157; February, 1952.) Two superheterodyne channels are fed from a frequency-swept oscillator; the cavity is introduced into the rf stage of one channel and a calibrated variable-Q circuit is inserted in the IF stage of the other. The resonance curves are displayed together on a cro and when they coincide, the Q of the cavity is n times that of the comparison circuit, where n is the ratio of rf to IF Q factors between 5,000 and 15,000 can be measured at about 2.8 kmc, estimated errors being $<\pm 3$ per cent. The equivalent shunt resistance of the cavity can also be measured.

621.317.34:621.392 1961
The Measurement of Image Impedance and Image Attenuation Coefficient—L. Guenot. (*Ann. Télécommun.*, vol. 6, pp. 353–362; December, 1951.) Two methods are considered. In the first, the open-circuit and short-circuit impedances are measured, and the image impedance, image-attenuation and image-phase-change coefficients are evaluated from the results using either abacs, geometrical constructions or Kennelly's charts. In the second method, the total attenuation is measured for various lengths of nonloaded cable, the slope of the asymptote to the attenuation /length curve giving the attenuation coefficient directly.

621.317.35 1962
Ultrasonic Wave Analyzer—T. A. Benham. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 46, pp. 12–14; September, 1951.) Description of a heterodyne circuit handling inputs in the frequency range 10–300 kc; the design of the crystal filter circuit giving a pass band of 200 cps is discussed in some detail.

621.317.35:621.395.625.3 1963
Boundary-Displacement Magnetic Recording—H. L. Daniels. (*Electronics*, vol. 25, pp. 116–120; April, 1952.) Ordinary magnetic tape is used and is magnetized to saturation at all times. In the absence of modulation, one half of the tape has opposite polarity to that of the other half, with an unmagnetized boundary strip down the middle separating the two. With a modulated signal the unmagnetized boundary follows the wave form of the signal. A specially designed recording head is used, but a conventional pickup head serves for reproduction. With the tape mounted on a rotating drum, the system has been applied to the analysis of transient wave forms, frequencies ranging from 1 cps to 100 kc. Application to ab recording is under investigation.

621.317.361 1964
A Self-Interpolating Crystal Calibrator for Setting Up and Measuring Radio Frequencies—D. Cooke. (*Electronic Eng.*, vol. 24, pp. 23–25; January, 1952.) The fundamental (f_c) and harmonics of a crystal oscillator are used to modulate a carrier which is continuously variable between nf_c and $(n+1)f_c$; by using the sidebands corresponding to the harmonics, calibration frequencies of any value within a wide band can be obtained without loss of the accuracy associated with the crystal.

621.317.365+621.317.341]:621.392.26:621.315.61 1965
Measurements of Wavelengths and Attenuation in Dielectric Waveguides for Lower Modes—C. W. Horton and C. M. McKinney (PROC. I.R.E., vol. 40, pp. 177–180; February, 1952.) Measurements of the wavelength of a wave guided by a cylindrical dielectric rod give results for various modes and dielectric materials in good agreement with solutions of the characteristic equation. Measurements of attenuation due to losses and to bending of the dielectric guide are also reported.

621.317.7 1966
New Principles in Electrical Instruments—D. C. Gall. (*Proc. IEE*, Part II, vol. 98, pp. 665–670; December, 1951. Discussion, pp. 686–693.) For small dc measurements, magnetic amplifiers and tubes for amplifying the modulated dc signal are mentioned; also the Butterworth bridge comprising high μ wires whose resistance changes in a magnetic field. New stable rectifiers make the phase-sensitive bridge practical. Dc and ac stabilization and generation of standard voltages are discussed briefly. A shock-proof galvanometer whose coil is of the same density as the medium supporting it, power measurement by calorimetry, testing of materials by ultrasonics, and the NH_3-absorption-line time standard are mentioned. The cost of increasing instrumental accuracy is analyzed. 18 references.

621.317.7:621.385:621.396.822 1967
Direct-Reading Instrument Measures Tube Noise—A. van der Ziel. (*Electronics*, vol. 25, pp. 136–137; April, 1952.) Details of a 4-tube circuit enabling rapid and accurate determination of tube noise resistance by unskilled personnel. Results of tests on a few types of tube (6J4, 6AG5, 6AC7, 6J6 and 6AK5) are noted.

621.317.7.001.2 1968
Some Aspects of Electrical Instrument Design—L. Hartshorn. (*Proc. IEE*, Part II, vol. 98, pp. 657–664; December, 1951.) A discussion of requirements for accurate measurement. Short descriptions are given of the N.P.L. precision balance (error <1 in 10^9) and quartz-fibre micro-balance (sensitivity 10^{-8}g). Application of these to modern galvanometers and wattmeters is discussed. The Farmer tube-electrometer and recent measurements by Ramsey on residual leakage in high-quality dielectrics are described.

621.317.7.029.6 1969
Instruments for use in the Microwave Band—A. F. Harvey. (*Proc. IEE*, Part II, vol. 98, pp. 781–789; December, 1951. Discussion, pp. 789–792. Summary, *ibid.*, Part III, vol. 99, p. 32; January, 1952.) Description of instruments used in the microwave band for the measurement of power, frequency and impedance; the majority of them use waveguide methods of transmission, but coaxial systems at the longer wavelengths and free-space semi-optical systems at the shorter wavelengths are also described. Emphasis is laid upon more recent instruments and on those for the millimetre-wavelength region. Methods of manufacture and their influence on performance and design are also discussed. 35 references.

621.317.7.085/.087 1970
A Survey of Modern Methods of Presentation of Instrument Readings and Recordings—L. B. S. Golds. (*Proc. IEE*, Part II, vol. 98, pp. 671–685; December, 1951. Discussion, pp. 686–693.) Types of pointers and scales, lettering, etc., on electrical instruments (mostly of power-station type) are described in detail and compared with a view to assessing best readability. Other methods of presentation are also considered. Possible future developments are outlined.

621.317.7.088 1971
Performance Limits in Electrical Instruments—A. H. M. Arnold. (*Proc. IEE*, Part II, vol. 98, pp. 701–710; December, 1951. Discussion, pp. 710–713.) Potentiometers, voltmeters, ammeters, instrument transformers and wattmeters operating up to high audio frequencies are considered, defects found in instruments submitted to N.P.L. for test being described in detail. Some theoretical performance limits are analyzed.

621.317.715:621.396.645 1972
An Analysis of the Galvanometer Amplifier and its Response to Alternating Electromotive Forces and Mechanical Vibrations—R. G. Wylie and A. F. A. Harper. (*Aust. Jour. Sci. Res. Ser. A*, vol. 4, pp. 560–568; December, 1951.) A mathematical analysis of a representative system. Circuit constants may be chosen to provide critical damping such that the frequency response curve of the amplifier resembles that of the galvanometer alone. The bandwidth may be increased by the use of feedback, which also increases the effect of mechanical vibration having a rotational component about the axis of the galvanometer coil.

621.317.725 1973
Linear Diode Voltmeter—(*Radio Tech. Dig. Ed franc.*, vol. 5, pp. 327–337; 1951.) Adaptation of Burgess's analyses (2736 of 1948 and 736 of April) with supplementary data from other sources.

621.317.727:621.316.722.4 1974
Calculation of Capacitive Voltage Regulator with Wide Regulation Range—O. Schmid. *Funk u. Ton*, vol. 5, pp. 627–637; December, 1951.) Design formulas are derived for a capacitive voltage divider for hf measurement apparatus, using a variable capacitor with earthed rotor between two stators. Operation is practically independent of load and of generator impedance. A numerical example is calculated.

621.317.733 1975
The Wheatstone Bridge with Load-Dependent Resistances: Part 2—Applications—G. Nidetzky. (*Arch. tech. Messen.*, no. 190, pp. T129–T130; November, 1951.) Discussion of operating conditions whereby fluctuations of input voltage may be either suppressed or separated from the fundamental. Part 1: 446 of March.

621.317.733 1976
A Note on a Selective RC Bridge—P. G. Sulzer. (PROC. I.R.E., vol. 40, pp. 338–339; March, 1952.) Details of a bridge used as a frequency-determining element in an af oscillator and providing higher-selectivity than the Wien bridge.

621.317.755 1977
The Production of Three-Dimensional Characteristics with the Cathode-Ray Oscillograph—R. Grevel, F. W. Gundlach and H. Herklotz. (*Arch. tech. Messen*, no. 190, pp. T128; November, 1951.) An outline of principles and circuits used.

621.317.755.087:621.3.015.3 1978
Cathode-Ray Tube for Recording High-Speed Transients—S. T. Smith, R. V. Talbot and C. H. Smith, Jr. (PROC. I.R.E., vol. 40, pp. 297–302; March, 1952.) Description of a traveling-wave tube having a vertical sensitivity of 33v/cm and a writing speed of 1.5×10^7 m per second. Photographs of a 3-kmc sine wave and a pulse with a rise time of 0.5×10^{-9} second are shown

621.317.76 1979
Equipment for Accurate Comparison of Nearly Equal Frequencies—Andrieux and Dayonnet. (*Onde élect.*, vol. 31, pp. 469–472; December, 1951.) The particular method described is based on phase discrimination between the two frequencies f_1 and f_2 in a double-diode circuit. The difference-frequency voltage is applied to a double-pentode circuit, thus deriving short pulses separated by the time

interval $1/(f_1-f_2)$. These pulses trigger a control circuit energizing a commercial electronic counter so arranged that push-button operation gives an immediate indication of the time interval. One of the frequencies is a standard frequency of 100 kc, and the relative frequency difference is of the order of 10^{-8}. Measurement accuracy is within 1 part in 10^9.

621.317.772.087.4 1980
High-Frequency Phase Measurement with Direct Indication: Part 3—Curve Tracers—A. Ruhrmann. (*Arch. tech. Messen*, no. 190, pp. T121–T122; November, 1951.) Description with block diagrams of different instruments for phase recording in polar and Cartesian co-ordinates. Part 2: 3101 of 1950.

621.317.772.089.6 1981
Precision Calibrator for Low-Frequency Phase-Meters—M. P. Wintle. (*Elec. Commun.*, vol. 29, pp. 51–64; March, 1952.) Reprint. See 2770 of 1951.

621.317.784 1982
A General-Purpose Electronic Wattmeter—D. E. Garrett and F. G. Cole. (PROC. I.R.E., vol. 40, pp. 165–171; February, 1952.) Detailed description of a meter for frequencies from zero to 71kc and powers up to 50w, with accuracy to within 3 per cent.

621.317.79:621.318.4 1983
An Instrument for the Measurement of the Number of Turns of Cylindrical Coils—B. Ehlermann. (*Frequenz*, vol. 5, pp. 303–307; November/December, 1951.) Description of an accurate comparator-type test set.

621.396.62.001.4:621.396.619.13 1984
Proposed Test Procedure for F.M. Broadcast Receivers—D. Maurice, G. F. Newell and J. G. Spencer. (*Electronic Eng.*, vol. 24, pp. 106–111; March, 1952.) An outline of a procedure found convenient for testing FM receivers, including tests of sensitivity, selectivity, frequency stability, on-channel suppression ratio, AM suppression, impulsive interference performance, and distortion. Results of tests on three different receivers are analyzed.

621.396.621:621.396.619.13 1985
A New Method for Predicting the Adjacent Channel Performance of Mobile Radio Equipments by Graphical Analysis—(*FM-TV*, vol. 11, p. 6; October, 1951.) In 745 of April please cancel T. S. Eader as author and substitute H. H. Davids.

621.396.822:621.327.43 1986
A Portable, Direct-Reading Microwave Noise Generator—E. L. Chinnock. (PROC. I.R.E., vol. 40, pp. 160–164; February, 1952.). The discharge in an ordinary fluorescent lamp is used as source. The variation, with operating temperature, of the noise-power output and of the impedance match to the associated waveguide are considered. A resistance thermometer is incorporated and calibrated to give a direct-reading db scale for excess noise-power output. Circuit details are given.

621.396.822:621.385.16 1987
A Generator of Electrical Noise—A. P. G. Peterson. (*Gen. Rad. Exper.*, vol. 26, pp. 1–9; December, 1951.) Description of Type 1390-A random-noise generator, which uses a gas-discharge tube with transverse magnetic field as noise source. Three ranges are available, with upper frequency limits of 20 kc, 500 kc and 5 mc respectively and maximum open-circuit output voltage of 1v rms. Various applications are mentioned.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9:621.9.02 1988
New Process for Producing Holes in Hard Materials—A. Kuris. (*Machinery*, (London), vol. 79, pp. 991–992; December 6, 1951.) Description of the "cavitron" process. A boron-carbide abrasive mixed with water is directed on to a blunt-ended tool of the required cross-section. This is made to vibrate 27,000 times per second by means of a magnetostriction oscillator. About 30 minutes is required to make a ½-inch square hole in carbide ¼-inch thick, using a solid tool. With a hollow tool the time would be shorter.

539.16.08 1989
Radiation Measuring Instruments for X Rays to Cosmic Rays—D. Taylor and W. Abson. (*Proc. IEE*, Part II, vol. 98, pp. 760–770; December, 1951. Summary, *ibid*, Part III, vol. 99, pp. 28–30; January, 1952.) A review of techniques. 22 references.

621.317.083.7 1990
Guided-Missile Test Center Telemetering System—J. B. Wynn, Jr., and S. L. Ackerman. (*Electronics*, vol. 25, pp. 106–111; May, 1952.) General description of the equipment and methods used at the U. S. Air Force test center in the Caribbean. The information is transmitted from the missle as FM of a carrier wave, the frequency range being 215–235 mc. Sixteen channels are provided, six being used on a time-sharing basis to give a total of 172 information channels. The effective range is 200 miles at 30,000 feet height. Nine receiving sites are spaced at about 175-mile intervals in the 1,500-mile chain. Receiving, recording and data-presentation equipment are described.

621.317.794:535.61-15/-31 1991
Radiation Measuring Instruments for the Infrared to Ultraviolet Waveband—A. C. Menzies. (*Proc. IEE*, Part II, vol. 98, pp. 771–780; December, 1951. Discussion, pp. 789–792. Summary, *ibid*, Part III, vol. 99, pp. 30–31; January, 1952.) Infrared techniques are considered in some detail and typical electrical methods of emission and absorption spectroscopy in the visible and ultraviolet regions are described. A new type of CdSe photoconductive cell with maximum sensitivity at 7,200 Å and ⅓ that sensitivity at 4,000 Å, decreasing slowly through the ultraviolet, is mentioned. Direct recording of Raman spectra and spectroscopy in the vacuum ultraviolet region are also considered. 33 references.

621.383.001.8 1992
Photoelectric Device for Scanning Curves—H. J. Dreyer. (*Z. angew. Phys.*, vol. 3, pp. 453–456; December, 1951.)

621.384.611.1† 1993
The 31-MeV Betatron (Ray Transformer)—R. Wideröe. (*Brown Boveri Rev.*, vol. 38, pp. 260–272; September/October, 1951.) Description of dual-beam equipment and of the properties and applications of the X-rays produced.

621.384.611.1† 1994
The 31-MeV Betatron Installation at the University Radiological Institute Attached to the Zurich Cantonal Hospital—A. von Arx. (*Brown Boveri Rev.*, vol. 38, pp. 273–280; September/October, 1951.) Description of the apparatus and the control arrangements.

621.384.611.1† 1995
The Betatron and its Applications—M. Bohn. (*Rev. gén. élect.*, vol. 60, pp. 489–494; December, 1951.) General principles of operation and applications in radio therapy and industry are described, with some details of the equipment to be installed at the Cancer-Research Institute Gustave-Roussy, Villejuif, France.

621.384.611.1†:61 1996
Adaptation of 31-MeV Betatron to Medical Applications, and Shielding Problems—G. Joyet and W. Mauderli. (*Brown Boveri Rev.*, vol. 38, pp. 281–291; September/October, 1951.) Problems of dosage and protection of patients and operators are considered.

621.385.83:521.401.3 1997
Perturbation Characteristic Functions and their Application to Electron Optics—P. A. Sturrock. (*Proc. Roy. Soc. A.*, vol. 210, pp. 269–289; December 20, 1951.)

621.385.833 1998
The Refractive Index in Electron Optics—W. Ehrenberg and R. E. Siday: W. Glaser (*Proc. Phys. Soc.*, vol. 64, pp. 1088–1089; December, 1951.) Further comment. See 1741 of 1951 (Glaser).

621.385.833 1999
New Results on the Electron-Optical Properties of Magnetic Deflection Fields—R. F. K. Herzog. (*Acta Phys. austriaca*, vol. 4, pp. 431–444; May, 1951.) Apart from its deflecting action, a homogeneous magnetic deflection field acts as a system of two crossed cylindrical lenses; this system is analyzed, and some special cases are examined.

621.385.833 2000
A Potential Model for the Study of the Three-Electrode [electron] Lens—M. Bernard. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1438–1440; December 3, 1951.) The equation for the electron trajectories is integrated by successive approximation and simple expressions for the lens parameters are then obtained and evaluated. Comparison with Regenstreif's results (2500 of 1951) shows good agreement.

621.385.833 2001
Theory of Third-Order Rays in the Independent Electrostatic [electron] Lens—E. Regenstreif. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1588–1590; December 19, 1951.)

621.385.833:537.133 2002
The First Images Obtained with a Proton Microscope—P. Chanson and C. Magnan. (*Compt. Rend. Acad. Sci.*, (Paris), vol 233, pp. 1436–1438; December 3, 1951.) The apparatus has three es lenses permitting magnification up to 25,000, the ellipticity error of the central diaphragms being $<0.1\,\mu$. The proton source used is of the hf excitation type, working at 80 mc with a power of 50 w. From photographs obtained at a magnification of 3,000, the resolution limit is under 300 Å.

621.387.424 2003
Xenon-Filled Geiger Counters—G. Barrère. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 233, pp. 1442–1444; December 3, 1951.) Description of the construction and method of filling. The resulting counters were stable, with a plateau range of 400 v and slope of 3 per cent.

621.387.462:549.211 2004
An Explanation of Differences in Counting Properties among Diamond Specimens—G. P. Freeman and H. A. van der Velden. (*Phys. Rev.*, vol. 84, pp. 1050–1051; December 1, 1951.)

621.39.001.11:6 2005
Cybernetics—J. Loeb. (*Onde élect.*, vol. 31, pp. 457–468; December, 1951.) See 233 of February.

PROPAGATION OF WAVES

538.566.2 2006
The Reflection of Plane Electromagnetic Waves in Slightly Inhomogeneous Layers—G. Eckart. (*Arch. elekt. Übertragung*, vol. 5, pp. 555–560; December, 1951.) See 1882 of 1951.

621.396.11 2007

Waves in Nonuniform Propagation Conditions—H. Poeverlein. (*Z. Naturf.*, vol. 5a, pp. 492–498; September, 1950.) In crystal optics and in the case of radio wave propagation in an anisotropic medium such as the ionosphere, MacCullagh's index surface is of the fourth order, so that double refraction occurs, with four solutions for the waves in an anisotropic medium, two increasing and two decreasing waves. An index surface of the fourth order can lead to all sorts of remarkable phenomena in inhomogeneous (layered) media, as experience with radio waves shows.

621.396.11:523.3 2008

An U.H.F. Moon Relay—P. G. Sulzer, G. F. Montgomery and I. H. Gerks. (Proc. I.R.E., vol. 40, p. 361; March, 1952.) Report of the successful transmission of a short message from Cedar Rapids, Iowa, to Sterling, Va., using the moon as a reflector of the 418-mc cw signals. With a transmitter output of 20 kw and high-gain antennas, the estimated received power was about $7+10^{-17}$ w, a value corresponding, for the receiver used, to a signal/noise ratio of 8.6 db, which is in good agreement with the observed ratio.

621.396.11:551.510.535 2009

Group Velocities and Group Heights from the Magneto-ionic Theory—D. H. Shinn and H. A. Whale. (*Jour. Atmos. Terr. Phys.*, vol. 2, p. 200; 1952.) Corrections to paper abstracted in 1405 of June.

621.396.11:621.317.353.3 2010

Experimental Determination of the Resonance Curve in the Gyro-interaction Phenomenon—M. Cutolo, R. Ferrero and M. Motzo. (*Alta Frequenza*, vol. 19, pp. 3–8; February, 1950.) Experiments were made in June–July 1949 on the lines of those previously reported (2328 of 1950). The resonance curves experimentally obtained are regarded as confirming Bailey's prediction of a double-humped curve. The discrepancy between the experimentally obtained value of gyromagnetic frequency and the theoretical value is attributed to the variation of resonance frequency with hour of night.

621.396.11:621.317.353.3 2011

Ionospheric Interaction in Disturbed Conditions—D. A. Bell. (*Proc. Phys. Soc.*, vol. 64, pp. 1053–1062; December 1, 1951.) Describes the use of averaging methods to extract information from observations taken under disturbed conditions. The existence of a "low-frequency anomaly" (Huxley, *Proc. Roy. Soc. A.*, vol. 200, p. 507; 1950) is confirmed, and a possible mechanism suggested. No exceptional interaction was found for a variation in frequency of the disturbing wave around the gyromagnetic resonance frequency.

621.396.11.029.62 2012

The Anomalous Propagation of Radio Waves in the 10-Metre Band—F. H. Northover. (*Jour. Atmos. Terr. Phys.*, vol. 2, p. 200; 1952.) Corrections to paper noted in 1409 of June.

621.396.81 2013

Field-Strength Measurements in the Neighbourhood of a Discontinuity. Application of Millington's Method—P. Pernet. (*HF* (Brussels), no. 12, pp. 317–325; 1951.) In these 3.86-mc measurements, made over a land/sea path near Ostende, special difficulties were encountered due to coastal topography, the presence of local sources of field disturbance and the necessity of taking readings on board a tugboat. Hence the results are not considered sufficient either in quantity or accuracy for a good determination of the characteristics of propagation paths in a polder region, but only to indicate the general trend of the field variation over the paths considered. Agreement with Millington's theory (1758 of 1949) is good, the phenomena of abrupt fall and recovery of signal strength on crossing the land/sea boundary being clearly shown.

RECEPTION

621.396.62.001.4:621.396.619.13 2014

Proposed Test Procedure for F.M. Broadcast Receivers—Maurice, Newell and Spencer (See 1984.)

621.396.621 2015

Flexible Selectivity for Communications Receivers—O. G. Villard, Jr., and W. L. Rorden. (*Electronics*, vol. 25, pp. 138–140; April, 1952.) Details of a single-tube circuit using Q multiplication to achieve IF selectivity equivalent to that obtainable with a quartz-crystal filter.

621.396.621:621.396.619.13 2016

The Theory of Amplitude-Modulation Rejection in the Ratio Detector—B. D. Loughlin (Proc. I.R.E., vol. 40, pp. 289–296; March, 1952.) The procedure for a complete mathematical analysis of the AM-rejection properties of the ratio detector is presented. From the theory, the effect of variations of the parameters of ratio-detector transformers on the AM-rejection properties is predicted. Unbalance effects and their mutual cancellation are briefly considered. The degree of apparent limiting action within the ratio-detector circuit is only incidental and is not related to its AM rejection properties; it thus represents an inadequate design basis for the ratio detector. Summary noted in 1501 of 1950.

621.396.621:621.396.65 2017

Design Considerations for a Radiotelegraph Receiving System—J. D. Holland. (*Elec. Commun.*, vol. 29, pp. 34–50; March, 1952.) Reprint. See 2822 of 1951.

621.396.621.54 2018

Improvement in Gain Stability of the Superheterodyne Mixer through the Application of Negative Feedback—G. E. Boggs. (Proc. I.R.E., vol. 40, pp. 202–207; February, 1952.) Expressions are derived which show the improvement in stability when feedback is applied at IF between the mixer anode and input grid. An increased gain-bandwidth product is obtained by the use of a two-stage mixer. A generalized design procedure and performance details of experimental mixers are given.

621.396.621.54 2019

The Design of the I.F. Stage, with Particular Reference to the MHG [multipath h.f. feedback] Circuit—G. Lander. (*Funk u. Ton*, vol. 5, pp. 638–642; December, 1951.) Description of a variable selectivity arrangement used in Saba receivers, in which the IF stage includes six filter circuits, and multipath feedback is applied to widen the flat top of the response curve.

621.396.828 2020

Suppression of [Negative-] Impulse Interference by means of Gradient Limiting—M. R. Mantz. (*Commun. News*, vol. 12, pp. 43–48; December, 1951.) The gradient referred to is the slope of the input-voltage curve. For pulses the slope is generally greater than for wanted signals; circuits are described which discriminate between signals on this basis. These circuits are useful in m-λ and dm-λ communication reception, and also in broadcast reception if loss of the higher audio frequencies is tolerable.

621.396.828:621.397.62 2021

Suppression of Interference in Radio Receivers caused by Television Receivers—Fighiera. (*Télév. Franç.*, no. 77, pp. 12–13; December, 1951.) Discussion of causes of interference, particularly in the timebase circuit. Remedies suggested are the use of screened cable of minimum length for connecting between certain circuits of the television receiver, screening its cabinet, use of a mains filter, and of a high-pass filter in the antenna lead.

STATIONS AND COMMUNICATION SYSTEMS

621.395.43 +621.396.41:[621.396.619.13 :621.396.813 2022

Echo Distortion in the F.M. Transmission of Frequency-Division Multiplex—W. J. Albersheim and J. P. Schafer. (Proc. I.R.E., vol. 40, pp. 316–328; March, 1952.) Signal intermodulation by echoes is investigated analytically and experimentally. Two types of echo are considered: (a) weak echoes with delays $<0.1\ \mu s$, caused mainly by mismatched long lines; (b) strong echoes with delays $<0.01\ \mu s$, caused by multipath transmission and leading to selective fading. Random-noise signals were used to evaluate the echo distortion as a function of various parameters of the echo, the bandwidth and the rf modulation.

621.396.226:061.75 2023

Anniversary of Transatlantic Radio—R. L. Smith-Rose. (*Nature* (London), vol. 168, p. 980; December 8, 1951.) A short account of Marconi's early experiments and of the successful transmission of radio signals from Poldhu, Cornwall, to Signal Hill, Newfoundland, on December 12, 1901. See also *Electrician*, vol. 147, pp. 1965–1966; December 21, 1951, and *Elec. Times*, vol. 120, pp. 1070–1071; December 13, 1951.)

621.396.44:621.315.052.63 2024

Type-N Power-Line Carrier Equipment—J. McCulloch. (*G.E.C. Telecommun.*, vol. 6, pp. 85–99; 1951.) Full details of the Type-NA equipment designed for duplex operation in the range 136–264 kc and providing up to eight communication circuits. Other assemblies are available with different frequency bands within the range 80–600 kc.

621.396.5 2025

Techniques for Close Channel Spacing at V.H.F. and Higher Frequencies—C. F. Hobbs. (Proc. I.R.E., vol. 40, pp. 329–334; March, 1952.) A vhf communication system is proposed in which the transmitter and local-oscillator frequencies of the individual stations are derived by frequency division and ssb technique from a common reference carrier. This carrier would be radiated by a central station. Use of this technique would permit 5 kc spacing of R/T channels operating at about 1 kmc. Laboratory tests on experimental equipment confirmed this.

621.396.619.11.029.62:621.396.97 2026

Very High Frequency Sound Broadcasting—The Case for Amplitude Modulation—J. R. Brinkley. (*Jour. Brit. IRE*, vol. 11, pp. 585–592; December, 1951. Discussion, p. 593.) The high cost of a vhf broadcasting system in Britain is considered to be unjustified if such a service is used merely to duplicate existing services instead of being used to make available large numbers of new programs and other services not possible in the present medium-wavelength band. The relative merits of FM and AM for such a service are considered; AM is considered preferable because of the smaller frequency band required and because of lower cost of receivers and frequency converters.

621.396.619.13:621.3.018.78 2027

Distortion of a Frequency-Modulated Signal by Small Loss and Phase Variations—F. Assadourian. (Proc. I.R.E., vol. 40, pp. 172–176; February, 1952.) 1950 IRE National Convention paper. General formulas are developed for the harmonic and total distortion in the outputs of linear transmission systems with pure FM inputs, and with amplitude and phase characteristics involving ripples that can be represented approximately by single sine functions of small amplitude.

621.396.65 2028
Radio Links. General Technical Considerations—P. Marzin. (*Ann. Télécommun.*, vol. 6, pp. 363–380; December, 1951.) A survey paper reviewing propagation, types of antenna and transmission line, modulation systems, and tubes. Four actual installations exemplifying various link types are briefly described, (a) a 6- or 12-channel meter-wave ppm portable multiplex equipment; (b) a 24-channel 20-cm-wavelength ppm multiplex unit operating between Deauville and Le Havre; (c) the 60-channel Dijon-Strasbourg FM link with two relay stations, operating on wavelengths of about 1 m; (d) the Paris-Lille FM link with four relay stations, operating at frequencies centered on 3.78 kmc and providing two channels with 240 subdivisions each and one television channel.

621.396.65 2029
A Short-Haul Radio Communication Link Channelized by Time Division—E. M. Mortenson and C. B. Young. (*Elec. Eng.*, *N. Y.*, vol. 70, pp. 1094–1099; December, 1951.) Description of a pam system and equipment providing eight 3-kc channels. Refinements required for long-distance operation, such as repeater relays, diversity receivers, and fault-locating circuits, are eliminated.

621.396.65 2030
Radio Relay Design Data, 60 to 600 Mc/s—R. Guenther. (Proc. I.R.E., vol. 40, p. 337; March, 1952.) Corrections to paper abstracted in 495 of March.

621.396.65:621.311 2031
Radio Links for Power Stations—L. Persson. (*Ericsson Rev.*, no. 2, pp. 42–47; 1951.) Description of a particular link developed in Sweden, combining one telephony channel with a number of af tone channels for telemetry and telecontrol. The equipment comprises a 20-w 3-m transmitter operated in the 160-mc band and providing a total af channel from 300 to 7,500 cps, and a double-frequency-conversion receiver.

621.396.65.029.62 2032
Transportable U.S.W. Directional-Link Equipment—R. Siegert. (*Telefunken Ztg*, vol. 24, pp. 204–212; December, 1951.) This 250-w FM equipment is considered to afford the best compromise between output power and equipment weight considerations. Designed for 12-channel operation, it can be modified for 24-channel operation, and is continuously tunable over the 41–75-mc band. Effective range is normally 120–200 km, depending on terrain; for greater distances relay stations must be used, the over-all distance being limited to about 300 km if CCIF requirements are to be met.

621.396.65.029.63/.64:[621.396.5+621.397.5 2033
Planning Radio Links (Section Planning)—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 43, pp. 531–536; December, 1951.) A general discussion of communication-channel aspects, route layout, design of antenna towers etc., with particular reference to plans for relaying television on decimeter wavelengths between Hamburg and Cologne and also between Cologne and Frankfort, in combination with a multichannel telephone system. A zig-zag route is used, and the length of the path between adjacent relay stations does not exceed about 40 km. See also 1431 of June.

621.396.712:621.396.66 2034
The Design of Automatic Equipment for Programme Routing and Sequential Monitoring—H. D. M. Ellis and J. C. Taylor. (*BBC Quart.*, vol. 6, pp. 241–256; Winter, 1951/1952.) A description of the automatic system used at the BBC sw transmitting station at Skelton, Cumberland, for routing six incoming (different) programs to the eighteen separate transmitters and for monitoring the radiated programs. The programs are directed to the transmitters via a preset switching mechanism controlled from the station master clock in accordance with the desired 24-hour schedule; connections to the sequential monitor are similarly controlled. The monitor enables a single operator to listen alternately to a program line and then to the output of each transmitter connected to that line, each program line (with associated transmitters) being examined in turn. At worst a fault is discovered in slightly less than 90 seconds. The automatic apparatus is designed to provide reliability of a high order.

621.396.931:621.396.619.13 2035
Mobile Radio Equipment, Type SRR 178—D. J. Braak. (*Commun. News*, vol. 12, pp. 57–67; December, 1951.) The equipment, designed for duplex operation, has a choice of 10 crystal-controlled frequencies in the 156–174-mc or the 70–87.5-mc band. Phase modulation is used. Special coaxial-line filters enable a single antenna to be used for both transmission and reception.

621.396.931.029.6 2036
Switching Methods for V.H.F. [radio-communication] Networks—E. P. Fairbairn. (*Jour. Brit. IRE*, vol. 11, pp. 576–584; December, 1951.) A description is given of vhf and uhf types of communication network. The simplex, duplex, and two-frequency simplex methods of operation are described and consideration is given to methods for covering large areas of country and to the principles used in selective calling systems.

621.396.933 2037
Some Future Developments in Aeradio—G. R. Scott-Farnie and M. I. Forsyth-Grant. (*Jour. Brit. IRE*, vol. 11, pp. 595–606; December, 1951.) Developments up to 1939 are briefly reviewed and postwar developments are considered in greater detail. Future developments in air/ground communications are discussed, with particular reference to the effect of extended use of radio teletype equipment. Radio aids to navigation, approach, and landing are considered briefly and an outline is given of an ideal "aeradio" plan which could possibly materialize within the next 10 years.

SUBSIDIARY APPARATUS

621-526 2038
Servomechanisms—H. Chestnut and R. W. Mayer. (*Gen. Elec. Rev.*, vol. 54, pp. 39–46; December, 1951.) Text of the first chapter of a book (1816 of April) dealing with stability and accuracy in automatic control systems.

621-526 2039
A Logarithmic Plotting Technique for the Design of Closed-Loop [control] Systems—J. A. Tanner. (*Trans. Soc. Instr. Tech.*, vol. 3, pp. 170–181; December, 1951. Discussion, pp. 181–182.) Description of a design technique based on steady-state response. An explanation is given of the resolution of the gain/frequency characteristic (plotted on logarithmic scales) into asymptotes whose intersection points are determined by the time constants of the control system considered. Reference is made to the gain-phase interrelations developed by Bode. Correlation between the asymptotic gain characteristic and the Nyquist diagram is established and the stability margin is considered. The procedure for determining the characteristics of stabilizing networks is outlined and examples are given of stabilization by means of series elements and by minor-loop elements. Determination of system time constants from experimental frequency response curves is considered briefly.

621-526 2040
A Generalized Method for Analyzing Servomechanisms—A. A. Hauser, Jr. (Proc. I.R.E., vol. 40, pp. 197–202; February, 1952.)

621-526:621.3.016.352 2041
"Hereditary" Phenomena in Servomechanisms; a General Criterion of Stability—J. Loeb. (*Ann. Télécommun.*, vol. 6, pp. 346–352; December, 1951.) If the point $(-1, 10)$ lies outside the area occupied by the family of curves $A(x, \omega)$ representing a single serve element, the system is stable. If it lies inside this area, the system may oscillate. Nyquist's and Kochenburger's stability criteria are shown to be particular cases of this general criterion. The case of stable oscillations is considered. See also 502 of March.

621.314.63:537.311.33 2042
The "Unforming" of Electrolytic Rectifiers and of some Barrier-Layer Rectifiers—W. C. van Geel and B. C. Bouma. (*Philips Res. Rep.*, vol. 6, pp. 401–424; December, 1951. In French.) Unforming or removal of rectifying properties, is performed by passing current in the opposite direction to the forming current; reforming can be performed by passing current in the same direction as the original forming current, or by passing ac. These processes are studied experimentally for various electrolytic rectifiers and for $Zr\text{-}ZrO_2-(CuI+I)$, Se, and resin-layer types. It is suggested that in all these cases the rectification is due to a two-layer barrier, one layer being an n-type and the other a p-type semiconductor; the unforming current removes the conductivity from one of these layers, thus destroying the rectifying properties.

621.314.632.1 2043
On the Deviation from Ohm's Law at the Anode Surface of Cuprous-Oxide Rectifiers—M. Ono. (*Jour. Phys. Soc. Japan*, vol. 6, p. 397; September/October, 1951.) A discussion of the effects of nonlinear contact resistance between the graphite anode and the cuprous oxide.

621.314.634 2044
Creep Phenomena of Selenium Rectifiers—M. Tomura. (*Jour. Phys Soc. Japan*, vol. 6, pp. 357–361; September/October, 1951.) A study of the initial decay of current in the hard-flow direction after application of voltages ranging up to 30 v.

621.316.722.078.3 2045
The Design of Series-Parallel Valve Voltage Stabilizers—F. A. Benson. (*Electronic Eng.*, vol. 24, pp. 118–119; March, 1952.) Discussion of various factors affecting the achievement of high stability with this type of stabilizer.

621.316.722.078.3 2046
An A.C. Stabilizer—B. Collinge and T. N. Marsham. (*Jour. Sci. Instr.*, vol. 28, p. 374; December, 1951.) "A voltage stabilizer is described which provides a continuously variable alternating voltage which is independent of mains frequency variations and free from waveform distortion. Electronic control of a 2-kva Variac is used to provide an output voltage constant to within $\pm\frac{1}{2}$ v."

621.319.35 2047
The Production of High Direct Voltages by Charging Mercury Drops—A. Dobrowsky. (*Elektrotech. u. Maschinenb.*, vol. 68, pp. 577–580; December 15, 1951.) Discussion of a proposed method in which Hg is sprayed under pressure into an evacuated vessel, the drops

being charged during passage through a grid before falling into an iron container, where they coalesce, the voltage consequently rising. Theory is developed, and calculations of performance made.

621.352.39 2048

Miniature High-Capacity Battery Cells—R. R. Clune. (*Electronics*, vol. 25, pp. 216, 222; April, 1952.) Description of the construction of the RM (Ruben-Mallory) type of cell. The RM-1 is 0.625 inch in diameter, 0.65 inch high, weighs 0.43 ounce, and is rated at 1,000 mah. It will operate efficiently at currents up to 100 ma. A longer cell (1.95 inches) is rated at 3,600 mah and can furnish currents up to 200 ma. Special construction avoids leakage and renders the cells self-sealing after any escape of gas.

778.37 2049

The Photographic Study of Rapid Events. [Book Review]—W. D. Chesterman. Publishers: Oxford University Press, London, Eng., 167 pp., 21s. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 368–369; December, 1951.) A valuable reference book for research workers in all spheres of scientific investigation. Methods and equipment for high-speed photography are classified according to the character, speed and duration of the phenomena investigated.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5:06.053 2050

Television at the Sixth Full Meeting of the C.C.I.R. in Geneva, 1951—Kirschstein. (*Fernmeldetech. Z.*, vol. 4, pp. 542–544; December, 1951.) A brief report of the proceedings. Tables are given showing the stage of development reached and the different operating standards used in the various participating countries.

621.397.5:535.62 2051

Color Television and Colorimetry—W. T. Wintringham. (PROC. I.R.E., vol. 40, p. 357; March, 1952.) Corrections to paper noted in 825 of April.

621.397.5+621.396.5]:621.396.65.029.63/.64 2052

Planning Radio Links (Section Planning)—Schmidt. (See 2033.)

621.397.5(204.1) 2053

Underwater Television—(*Engineer* (London), vol. 192, p. 764; December 14, 1951. *Engineering* (London), vol. 172, p. 765; December 14, 1951.) Describes the rapid adaptation of the Marconi image-orthicon camera to work under water in the search for the submarine "Affray." Instrumental modifications included remote operation of camera heating and cooling, remote level indication, and water-leak detection. No electronic modifications were necessary. The camera worked with a 2-inch f-1.29 lens (set at f4) at a depth of 280 feet. Lighting was provided by a 1.5-kw diver's lamp. Effective range under these conditions was 15 feet and the equipment could be used continuously for two hours or more, whereas a diver could only work for a few minutes.

621.397.61 2054

Development Problems of Television Transmitters—W. Burkhardtsmaier. (*Telefunken Ztg*, vol. 24, pp. 193–203; December, 1951.) A survey of the principal problems arising in the design of video transmitters, and discussion of the merits of various solutions proposed. The 1-kw transmitter built for Hamburg N.W.D.R. is described briefly. It operates in the frequency range 174–216 mc, has a bandwidth of 5.75 mc, and the frequency is constant to within ±1 kc.

621.397.611.2 2055

Improvements in Image Iconoscopes by Pulsed Biasing the Storage Surface—R. Theile and F. H. Townsend. (PROC. I.R.E., vol. 40, pp. 146–154; February, 1952.) The disadvantages associated with high-velocity electron scanning in storage-type television camera tubes are minimized by periodically irradiating the storage surface with high-velocity electrons, while simultaneously reducing the collector potential. The method is most successful in the transmission of intermittently projected pictures such as "memory-scanned" cinema films.

621.397.611.1 2056

Portable Pickup Equipment for 625-Line Television—H. Hewel. (*Funk u. Ton*, vol. 5, pp. 643–657; December, 1951.) The equipment described consists of four units, (*a*) pulse generator, (*b*) mixer, (*c*) camera and control unit and (*d*) 30-m cable on drum, with camera stand; it is operated from a 50-cps supply at 220 v, 125 v or 110 v, and takes 450 w. The output from the 70-Ω wide-band cable is a 21-mc carrier modulated with the 0–6-mc vision signal.

621.397.62 2057

Television Receiver with Interchangeable Units—(*Télévision*, no. 19, pp. 285–289, 295; December, 1951.) Description of a receiver for 819- or 441-line television in which servicing is facilitated by constructing the various functional units on separate chassis.

621.397.62 2058

Choice of Intermediate Frequency in Television Receivers—J. Harmans. (*Frequenz*, vol. 5, pp. 307–311; November/December, 1951.) Harmonic and second-channel interference in the vision signal may be avoided if the IF is sufficiently high. The highest IF possible in the case of stagger-tuned circuits is determined as a function of bandwidth, number of stages, and tube type. Limiting IF values for six different tubes are tabulated.

621.397.62:621.396.622.72 2059

Constant-Input-Impedance TV Second Detector—W. K. Squires and R. A. Goundry. (*Electronics*, vol. 25, pp. 109–111; April, 1952.) A triode with low grid-cathode capacitance is operated with a high-value cathode resistor so that it is self-biased nearly to cut-off. Detection is effected by means of the nonlinearity of the grid-anode characteristic in the cut-off region. Under these conditions the input impedance is practically constant at any given frequency, and for all frequencies is the same as that of a typical IF amplifier with the same input capacitance. When this type of detector is used in a receiver with suitably designed video and IF circuits, performance is superior to that of an equivalent receiver with a diode detector, particularly as regards transient response.

621.397.621 2060

The Theory and Design of Television Frame Output Stages—E. T. Emms. (*Electronic Eng.*, vol. 24, pp. 96–101; March, 1952.) Two modes of operation are discussed which may be more efficient than the commonly used transformer method with sawtooth-current input. These modes are termed (*a*) the "minimum mean anode current" condition and (*b*) the "zero initial slope" condition. Analysis based on an equivalent circuit is given for the transformer circuit, and the special features of the two modes are described, with particular reference to anode-current wave form. Design examples are outlined for each mode and comparison is made of designs for a peak current of 35.5 ma. This shows that mode (*b*) leads to larger transformer inductances, and greater mean anode current and voltage swing, than mode (*a*). The use of a negative-feedback network to obtain the desired parabolic anode-current wave form is considered and a practical frame timebase circuit is given, with component details.

621.397.621.2:621.396.615 2061

Blocking-Tube Oscillator Design for Television Receivers—A. F. Giordano. (*Elec. Eng.* (N. Y.), vol. 70, pp. 1050–1055; December, 1951.) Essentially full text of 1951 A.I.E.E. Fall General Meeting Paper. When used in a receiver vertical-deflection circuit, the oscillator may be adjusted with its free-running frequency below the synchronizing frequency, so that timing is controlled by the synchronizing pulse. Loss of synchronization may occur due to excessive drift of the free-running frequency; this problem and others relating to interlacing and noise are discussed.

621.397.645 2062

The Contribution of the I.F. Amplifier to the Signal Build-Up in Television Receivers—H. Zimmermann. (*Fernmeldetech. Z.*, pp. 537–542; December, 1951.) A theoretical investigation is made of the closeness with which the optimum transmission curve required by the 625-line television standard can be approached using a stagger-tuned three-stage IF amplifier. The corresponding build-up transients are shown for ideal conditions and for actual operating conditions. The significance of the phase build-up in the intercarrier process is indicated.

621.397.828 2063

Suppression of Harmonics in Radio Transmitters—G. T. Royden. (*Elec. Commun.*, vol. 28, p. 321; December, 1951.) Correction to paper abstracted in 3144 of 1951.

TRANSMISSION

621.396.61 2064

A Range of 100 and 150-Watt Transmitters for the M.F., H.F. and V.H.F. Bands—J. Campbell. (*GEC Telecommun.*, vol. 6, pp. 72–84; 1951.) Technical details of five communication transmitters for which 15 types of standardized unit were developed.

621.396.61:621.396.65 2065

The LD-T2 Radio Transmitter—N. F. Schlaack. (*Bell Lab. Rec.*, vol. 29, pp. 561–564, 570; December, 1951.) Description of a multi-channel ssb transmitter used in the Bell System long-distance radiotelephony network. Operating frequency is in the range 4–23 mc. The transmitter accepts two independent af bands from 100 cps to 6 kc and uses a low-amplitude triple-modulation system followed by a six-stage linear amplifier.

621.396.619.13 2066

Realizability of the Point of Inflexion of a Modulation Characteristic for F.M. by means of Reactance Valves—W. Mansfeld. (*Frequenz*, vol. 5, pp. 317–323; November/December, 1951.) Operating conditions for a parallel-connected reactance-tube circuit are discussed. The greatest linearity is obtained in operation about the point of inflexion in the frequency curve. A considerable improvement in linearity can be obtained with a push-pull circuit. The ratio of the voltages across the two parts of the voltage divider for the reactance tube should be smaller in the capacitive than in the inductive type of circuit.

621.396.619.13 2067

Frequency Modulation by means of a Capacitor with Controlled Charging Cycle—R. Otto. (*Frequenz*, vol. 5, pp. 323–327; November/December, 1951.) The modulation voltage is applied across two rectifiers connected in series, one of which is directly in series with the fixed capacitor (C) in the hf circuit. Control of the charging time can thus be achieved and the effective capacitance of C varied between 10 per cent and 90 per cent of its nominal value, with little distortion. The system works well provided the resistance of the rectifier when conducting is $<\frac{1}{5}$ of the hf impedance of the capacitor C. Circuits have been operated satisfactorily at frequencies up to over 100 mc with a frequency swing of 100 kc and distortion <1 percent.

621.397.828 2068
Suppression of Harmonics in Radio Transmitters—Royden. (*Elec. Commun.*, vol. 28, pp. 321; December, 1951.) Correction to paper abstracted in 3144 of 1951.

TUBES AND THERMIONICS

621.383 2069
The Photoelectric Effect in Cs-Ga, Cs-In and Cs-Tl Photocathodes—N. Schaetti, W. Baumgartner and C. Flury. (*Helv. Phys. Acta*, vol. 24, pp. 609–613; December 31, 1951. In German.) 1951 Société Suisse de Physique Lucerne Meeting paper. Spectral sensitivity and conductivity data are presented and the effects on Cs-Sb cells of additions of In and Tl are reported.

621.383 2070
The Properties of Cs-Sb Photocathodes at Different Temperatures—N. Schaetti and W. Baumgartner. (*Helv. Phys. Acta*, vol. 24, pp. 614–619; December 31, 1951.) 1951 Société Suisse de Physique Lucerne Meeting paper. Measurements at temperatures down to $-196°C$ show in all cases a backward shift of the long-wave limit, with conductivity curves of the semiconductor type.

621.384.52 2071
The Cold-Cathode Glow Discharge Tube—D. S. Peck. (*Bell Lab. Rec.*, vol. 29, pp. 550–553; December, 1951.) A simple account of the construction and operation of diodes and triodes, with particular reference to Western Electric types, which contain a small quantity of radioactive material as ionization source. Measurements of ionization times under different conditions are reported.

621.385 2072
Control of the Current Distribution in Electron Beams—E. Gundert. (*Telefunken Ztg*, vol. 24, pp. 223–236; December, 1951.) Detailed analysis of control by an auxiliary negative grid between two positive electrodes such as the screen grid and anode. The electron trajectories, particularly for the limiting case of oscillation near the grid plane, are calculated to a first approximation. Assuming a uniform electron stream at some distance from the control grid, the control-grid characteristics are determined from the path separation in the limiting case. Three distributions of current about the angle of incidence of electrons on the grid are considered: (*a*) uniform distribution, (*b*) distribution following a trapezoidal law, (*c*) distribution following a bell-shaped curve. Calculations to a second order of approximation give the dependence of the slope of the characteristic on the ratio of the field strengths on the two sides of the grid. Mathematical calculations are mainly confined to appendices.

621.385:621.396.822:621.317.7 2073
Direct-Reading Instrument measures Tube Noise—van der Ziel. (See 1967.)

621.385-713 2074
Evaporation-Cooled Power Tubes—C. Beurtheret. (*Electronics*, vol. 25, pp. 106–107; March, 1952.) Description of the cooling system used in some high-power French broadcasting transmitters. Cooling is so effective that tubes can be operated at about three times their normal power rating for conventional water cooling.

621.385.029.6 2075
A Travelling-Wave Valve without Retarding Line—H. Kleinwachter. (*Elektrotech Z.*, vol. 72, pp. 714–717; December 15, 1951.) In usual types of traveling-wave tube the signal wave is greatly retarded to match its velocity to that of the beam electrons. In the tube described here, a density-modulated beam is caused to traverse a path along which the direction of the electric field is alternately positive and negative (e.g. by using a cylindrical waveguide divided into short cylinders, all the odd and all the even members being connected together). An idealized chart showing electron positions at different instants illustrates how signal amplification is achieved. Calculation indicates that there will be three progressive wave components in the beam current, one of which can have an infinitely high phase velocity and can hence supply energy to an unretarded E wave. The tube is superficially similar to that described by Field *et al.* (2068 of 1951), but operates on a different principle.

621.385.029.63/.64+621.392.2 2076
Slow Electromagnetic Waves—Akhiezer and Faynberg. (See 1829.)

621.385.029.64 2077
On the Theory of Electron-Wave Tubes—O. E. H. Rydbeck and S. K. H. Forsgren. (*Acta polyt.* (Stockholm), no. 84, 31 pp.; 1951.) Reprint. See 2866 of 1951.

621.385.032.216 2078
Space-Current Changes in Thermionic Valves following Small Pulses of Current—J. R. Tillman. (*Proc. Phys. Soc.*, vol. 64, pp. 1046–1052; December 1, 1951.) Amplitude-modulated pulses of voltage were applied to the control grid of a commercial tube working conventionally, and produced space-current changes of up to 0.01 a/cm² from the oxide-coated cathode. An unmodulated pulse train sampled the anode current after a delay t, and showed a small component at the modulation frequency which decayed with a time constant of about 1 ms as t increased. The component was sometimes in phase with the component at $t=0$ (emission enhancement) and sometimes in antiphase (fatigue). These effects were often related to the history of the tube, but not, apparently, to the impedance between cathode core and coating. Fatigue may be largely a surface effect.

621.385.032.216 2079
The Influence of the Core Material on the Thermionic Emission of Oxide Cathodes—H. A. Poehler. (Proc. I.R.E., vol. 40, pp. 190–196; February, 1952.)

621.385.032.216 2080
The L-Cathode Structure—G. A. Espersen. (Proc. I.R.E., vol. 40, pp. 284–289; March, 1952.) Basic features of this type of cathode are reviewed and methods of measuring the rate of Ba evaporation are described. Data on cathode life of various types of tube are tabulated and briefly discussed. See also 773 of 1951 (Lemmens *et al.*).

621.385.032.216 2081
On Poisoning of Oxide Cathodes by Atmospheric Sulphur—H. A. Stahl. (*Appl. Sci. Res.*, vol. B1, pp. 397–412; 1950.) Detailed account of work noted in 2051 of 1951.

621.385.032.216:537.525.92 2082
Space-Charge Effect in the Oxide-Cathode Layer—T. Shindo. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 352–356; September/October, 1951.) The existence of space charge in the oxide layer, near the surface, is deduced from a theoretical study of a one-dimensional model. The effect of this space charge on the emission properties is indicated.

621.385.032.216:537.582 2083
On the Equation of Thermionic Emission of the Oxide-Coated Cathode—Y. Watanabe, E. Takagi and S. Katsura. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 26–45; 1949.) A summary of the emission theories, with discussion of the experimental results obtained by Takagi on the initial current characteristic of diodes with cathodes activated to varying degrees.

621.385.2:[537.315.6+621.3.012.6 2084
A Method of Calculating the Space Distribution of Potential and the I/V_a Curve for a Diode—H. Bonifas. (*Bull. Soc. franç. élect.*, vol. 1, pp. 741–757; Correction, *ibid.*, vol. 2, p. 115; March, 1952. December, 1951.) For calculation of the space distribution of potential see 1162 of May. The I/V_a curve can be conveniently divided into three sections for purposes of calculation. The first comprises the lower bend and the part that is practically a straight line, i.e. covers low values of I. It is determined from the relation $V_a/l = K'\{(I - I_0)/I_{th}\}^{1/2}$, where l = anode-cathode distance, K' is a coefficient depending on the nature and operating temperature of the cathode, I_0 = anode current for zero voltage, I_{th} = saturation current. The second section comprises the upper bend in the curve, and the third the upper part which is nearly straight and horizontal. Both are calculated from variants of the equation given above, modified to take account of the effect of the cathode field. The method used reproduces the shape of the experimental curve even for high values of I. Appendices establish a theorem applied in the calculation and give a rigorous derivation of Schottky's equation.

621.385.3 2085
Effective Grid Potential and Cathode Current Density of a Planar Triode, taking account of "Island" Formation—W. Dahlke. (*Telefunken Ztg*, vol. 24, pp. 213–222; December, 1951.) The triode is represented by the sum of elementary three-plate capacitors, the grid being replaced by an electrode consisting of a solid plate whose shape and effective potential determine the discharge properties of the tube. This potential is calculated, (*a*) neglecting, (*b*) taking account of space charge. The plate shape is dependent on the degree of "island" formation, approaching more closely in longitudinal cross-section to a sine wave, of wavelength the same as the pitch of the grid wires, the greater the "island" formation. The mean penetration factor and its relative variation, as well as the cathode current density and the slope, are shown in a series of diagrams.

621.385.3.029.62/.63 2086
Triode Amplifiers in the Frequency Range 100 Mc/s to 420 Mc/s—D. C. Rogers. (*Jour. Brit. IRE*, vol. 11, pp. 569–575; December, 1951. *Elec. Commun.*, vol. 29, pp. 12–19; March, 1952.) "It has been found possible, by careful attention to the geometry of electrodes and leads, to design triode tubes capable of operation at frequencies up to 420 mc in grounded-grid circuits, and yet mounted on conventional pressed-glass bases, and using only the recognized techniques of receiving-tube manufacture. This paper outlines the design features requiring special consideration and gives some of the results achieved with various special tube types."

621.385.832 2087
Elementary Theory of the Generation of Electron Beams by means of Triode Systems: Part 1—Properties of the Static Field of Commonly Used Beam Systems—M. Ploke. (*Z. angew. Phys.*, vol. 3, pp. 441–449; December, 1951.) Electron guns with pillbox and with hairpin cathodes are investigated. The potential field near the cathode exhibits a singularity which is practically independent of the electrode shapes and which gives rise to the observed conical shape of the beam. The axial potential and the dependence of the cathode field on gun dimensions and voltages are determined.

621.385.832 2088
Correction of Deflection Defocusing in Cathode-Ray Tubes—J. E. Rosenthal. (Proc.

I.R.E., vol. 40, pp. 353–357; March, 1952.) Discussion on paper noted in 1806 of 1951.

621.396.615.141.2 2089

The Cut-Off Characteristic and the Potential Distribution of the Magnetron Tube—Y. Watanabe and S. Katsura. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 10–25; 1949.) The curvature of the anode-current/field-coil-current characteristic is discussed in terms of the distribution of initial velocities of the emitted electrons and the change of potential distribution under the magnetic field.

621.396.615.142.2 2090

Recent Developments in Klystrons—R. H. Varian. (*Electronics*, vol. 25, pp. 112–115; April, 1952.) A review of progress in the design of reflex, amplifier, and floating-drift klystrons, with mention of an amplifier developed at Stanford University with an output of 10 mw at 3 kmc.

621.396.615.142.2 2091

The Duplex Traveling-Wave Klystron—T. G. Mihran. (Proc. I.R.E., vol. 40, pp. 308–315; March, 1952.) Theory and description of an experimental model designed to combine the properties of the two-cavity klystron and the "distributed" amplifier.

621.396.615.142.2 2092

Power-Amplifier Klystron for Air Navigation—V. Megoned. (*Electronics*, vol. 25, pp. 156, 166; May, 1952.) Description of the construction of the three-cavity Type-SAL39 klystron, with technical operating data. The frequency range is 0.96–1.215 kmc, with corresponding peak output power from 25 to 10 kw.

621.385 2093

Elektronenröhren und ihre Schaltungen (Valves and Valve Circuits). [Book Review]—M. Kulp. Publishers: Verlag Vandenhoeck & Ruprecht, Göttingen, 1951, 346 pp., 26.50 DM. (*Fernmeldetech. Z.*, vol. 4, p. 523; November, 1951.) "The book will prove extremely useful for scientists of all technical branches."

621.385.032.216 2094

The Oxide-Coated Cathode, Vol. 1: Manufacture. [Book Review]—G. Hermann & S. Wagener. Publishers: Chapman & Hall, London, 1951, 148 pp., 21s. (*Z. angew. Math. Phys.*, vol. 2, p. 497; November 15, 1951.) Translation by Wagener of the second volume (see 2603 of 1951) of a standard German work, with additional data on magnetron cathodes.

621.396.615.142 2095

Les Tubes Electroniques à Commande par Modulation de Vitesse (Velocity-Modulation Valves). [Book Review]—R. Warnecke & P. Guénard. Publishers: Gauthier-Villars, Paris, 773 pp., 7000 francs. (*Wireless Eng.*, vol. 29, pp 112–113; April, 1952.) "In a work with 385 references the authors have set out to present all that is known about velocity-modulation tubes, from the point of view both of theory of performance and design and of practical operation and application." . . . They "are to be congratulated on producing this very comprehensive book, not only for the quality of the material presented, but also on account of the enormous effort which must have been necessary to produce such a work."

MISCELLANEOUS

061.3:621.39 2096

1952 I.R.E. National Convention, New York, 3rd–6th March—(Proc. I.R.E., vol. 40, pp. 212–234; February, 1952.) Schedule of the various sessions and summaries of the 211 papers presented.

061.4:621.396 2097

Telecommunications, Radio Navigation and Electronics at the XIXth Salon international de l'Aéronautique (Paris, 15th June–1st July 1951).—M. Adam. (*Génie civil*, vol. 128, pp. 401–405; November 1, 1951.) Report on equipment and components on show.

621.3(083.71/.74) 2098

Standardisation of Technical Terms and Circuit Presentation—J. Scott-Taggart. (*Electronic Eng.*, vol. 24, pp. 63–65; February, 1952.) Condensed version of paper in "Naval Radio and Electrical Review" discussing the most recent recommendations made by the BSI, the British Services, and other authorities regarding the choice of technical terms and the use of symbols and abbreviations in publications relating to radio and allied subjects.

I.R.E., vol. 40, pp. 353–357; March, 1952.) Discussion on paper noted in 1806 of 1951.

621.396.615.141.2 2089

The Cut-Off Characteristic and the Potential Distribution of the Magnetron Tube—Y. Watanabe and S. Katsura. (*Tech. Rep. Tohoku Univ.*, vol. 14, pp. 10–25; 1949.) The curvature of the anode-current/field-coil-current characteristic is discussed in terms of the distribution of initial velocities of the emitted electrons and the change of potential distribution under the magnetic field.

621.396.615.142.2 2090

Recent Developments in Klystrons—R. H. Varian. (*Electronics*, vol. 25, pp. 112–115; April, 1952.) A review of progress in the design of reflex, amplifier, and floating-drift klystrons, with mention of an amplifier developed at Stanford University with an output of 10 mw at 3 kmc.

621.396.615.142.2 2091

The Duplex Traveling-Wave Klystron—T. G. Mihran. (PROC. I.R.E. vol. 40, pp. 308–315; March, 1952.) Theory and description of an experimental model designed to combine the properties of the two-cavity klystron and the "distributed" amplifier.

621.396.615.142.2 2092

Power-Amplifier Klystron for Air Navigation—V. Megoned. (*Electronics*, vol. 25, pp. 156, 166; May, 1952.) Description of the construction of the three cavity Type-SAL39 klystron, with technical operating data. The frequency range is 0.96–1.215 kmc, with corresponding peak output power from 25 to 10 kw.

621.385 2093

Elektronenröhren und ihre Schaltungen (Valves and Valve Circuits). [Book Review]—M. Kulp. Publishers: Verlag Vandenhoeck & Ruprecht, Göttingen, 1951, 346 pp., 26.50 DM. (*Fernmeldetech. Z.*, vol. 4, p. 523; November, 1951.) "The book will prove extremely useful for scientists of all technical branches."

621.385.032.216 2094

The Oxide-Coated Cathode, Vol. 1: Manufacture. [Book Review]—G. Hermann & S. Wagener. Publishers: Chapman & Hall, London, 1951, 148 pp., 21s. (*Z. angew. Math. Phys.*, vol. 2, p. 497; November 15, 1951.) Translation by Wagener of the second volume (see 2603 of 1951) of a standard German work, with additional data on magnetron cathodes.

621.396.615.142 2095

Les Tubes Electroniques à Commande par Modulation de Vitesse (Velocity-Modulation Valves). [Book Review]—R. Warnecke & P. Guénard. Publishers: Gauthier-Villars, Paris, 773 pp., 7000 francs. (*Wireless Eng.*, vol. 29, pp 112–113; April, 1952.) "In a work with 385 references the authors have set out to present all that is known about velocity-modulation tubes, from the point of view both of theory of performance and design and of practical operation and application." . . . They "are to be congratulated on producing this very comprehensive book, not only for the quality of the material presented, but also on account of the enormous effort which must have been necessary to produce such a work."

MISCELLANEOUS

061.3:621.39 2096

1952 I.R.E. National Convention, New York, 3rd–6th March—(PROC. I.R.E., vol. 40, pp. 212–234; February, 1952.) Schedule of the various sessions and summaries of the 211 papers presented.

061.4:621.396 2097

Telecommunications, Radio Navigation and Electronics at the XIXth Salon international de l'Aéronautique (Paris, 15th June–1st July 1951).—M. Adam. (*Génie civil*, vol. 128, pp. 401–405; November 1, 1951.) Report on equipment and components on show.

621.3(083.71/.74) 2098

Standardisation of Technical Terms and Circuit Presentation—J. Scott-Taggart. (*Electronic Eng.*, vol. 24, pp. 63–65; February, 1952.) Condensed version of paper in "Naval Radio and Electrical Review" discussing the most recent recommendations made by the BSI, the British Services, and other authorities regarding the choice of technical terms and the use of symbols and abbreviations in publications relating to radio and allied subjects.