Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

September, 1952
Volume 40 Number 9

Sylvania Elec. Prod. Inc.

INSPECTING PICTURE TUBE SCREENS

Initial inspections of the newly formed phosphor screens on picture tubes takes place on a conveyor as they pass a bank of 100-watt fluorescent lamps, immediately after they have been dried by a stream of warm clean air.

PROCEEDINGS OF THE I.R.E.

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

The Institute of Radio Engineers

AMPEREX

AIR-COOLED TUBE

AX9904-R/5924

"...Offers a maximum in kilowatts per dollar..." HARRY R. SMITH, Manager, Television Engineering, Standard Electronics Corporation

STANDARD ELECTRONICS CORPORATION

A SUBSIDIARY OF CLAUDE NEON INC.

TELEVISION AND BROADCAST TRANSMITTERS AND AUXILIARY EQUIPMENT
AMERTRAN ELECTRONIC TRANSFORMERS AND TRANSTATS

TELEPHONE — BIGELOW 3-5540
BIGELOW 3-4444

285 EMMET STREET
NEWARK 5 N J

April 23, 1952

Mr. Sam Norris, Pres.
Amperex Electronics Corp.
25 Washington St.
Brooklyn 1, New York

Dear Mr. Norris:

As you know we have been working with the Amperex Type AX9904-R vacuum tube in the development of the various Standard Electronics television broadcast transmitters. The tube is being used in the currently manufactured Model TH653 and TL 653 Transmitters in both the aural and visual sections.

I believe you will be interested in knowing that we are very well satisfied with the performance of this tube as a broad band linear amplifier on all V.H.F. television channels. The low interelectrode capacitance and low internal impedance of the AX9904-R permit power output levels of 5KW and more, with band widths in excess of 5 megacycles. These conditions are readily obtainable from a single tube operating well within its published tube characteristics. The moderate cost of the tube leads us to believe that it offers a maximum in "kilowatts per dollar".

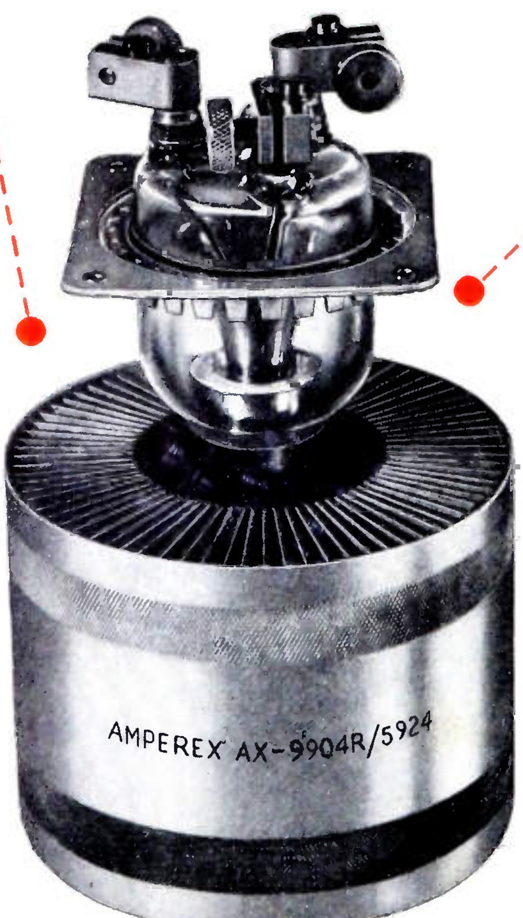
Yours very truly,

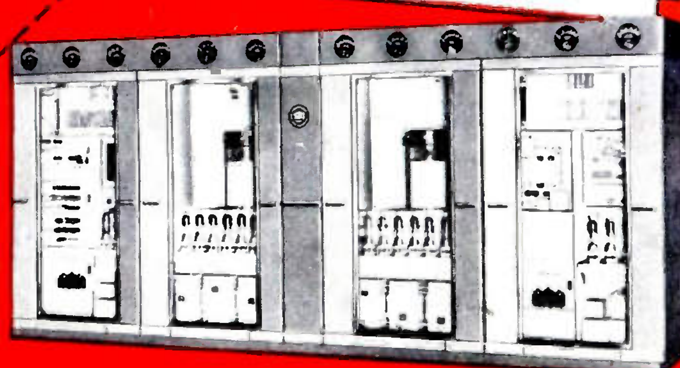
Harry R Smith

Harry R. Smith
Mgr. Television Engineering

HRS:hg

STANDARD ELECTRONICS CORPORATION uses this tube in Models TH653 High Band and TL653 Low Band Transmitters and also in their new 20 Kilowatt Transmitter, built on the exclusive S-E ADD-A-UNIT PRINCIPLE, and with special S-E features that insure dependable operation, maximum convenience, and minimum maintenance.





FEATURES INCLUDE . . . 14 MC band width at 220 MC . . . outputs of 5.7 KW . . . thoriated tungsten filament . . . non-emitting grid . . . disc type grid seal for mimimum inductance . . . minimum capacitance . . . and PROVEN *long life*.

Write for complete data sheets.

This tube is also available in a Water-Cooled Version, Type AX9904-5923.

Re-tube with AMPEREX

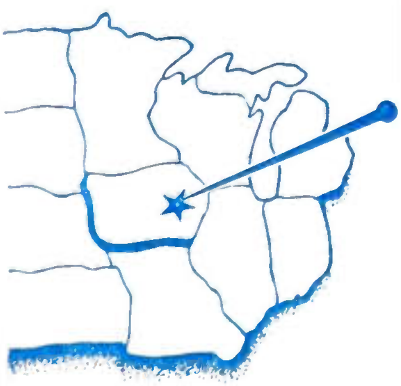
AMPEREX ELECTRONIC CORP.
230 DUFFY AVENUE, HICKSVILLE, LONG ISLAND, N. Y.
In Canada and Newfoundland: Rogers Majestic Limited
11-19 Brentcliffe Road, Leaside, Toronto, Ontario, Canada
Cable: "AMPRONICS"

ISA Instrument SHOW

IRE members are invited to the Seventh National Instrument Exhibit and ISA Conference by the Instrument Society of America.

September 8-12
Cleveland Municipal Auditorium

Sept. 19-20



Cedar Rapids
Iowa
Roosevelt Hotel

Conference on Communications

Sponsored by the Cedar Rapids Section IRE

Exhibits • Papers

National Electronics Conference and Exhibition

September 29, 30, October 1, 1952

Marking the eighth annual National Electronics Conference, in Chicago, sponsored by the local sections of IRE and AIEE and three universities, are these advances: 75 exhibits, and 99 papers in these 21 sessions.

1. Servomechanism Theory, Monday morning, September 29
2. High Frequency Electron Tubes
3. Audio
4. Industrial Measurements
5. Magnetic Amplifiers and Servo Applications, afternoon
6. Television
7. Equipment and Components Reliability
8. Waveguides
9. Transistors, Tuesday morning, September 30
10. Radar and Radio Navigation
11. Circuits I
12. Components, Assembly and Measurements
13. Semiconductors, afternoon
14. Memory Tubes and Tube Reliability
15. Circuits II
16. Computors, Wednesday morning, October 1
17. Antennas
18. Electronic Instrumentation
19. Engineering Management, afternoon
20. Coding and Recording Techniques
21. Delay Lines and H. F. Equipment

We have moved to larger quarters!

The Hotel Sherman, at Randolph, Clark and LaSalle Streets, Chicago, Illinois will be the new site of the 1952 National Electronics Conference.

PROCEEDINGS OF THE I.R.E. September, 1952, Vol. 40, No. 9. 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.

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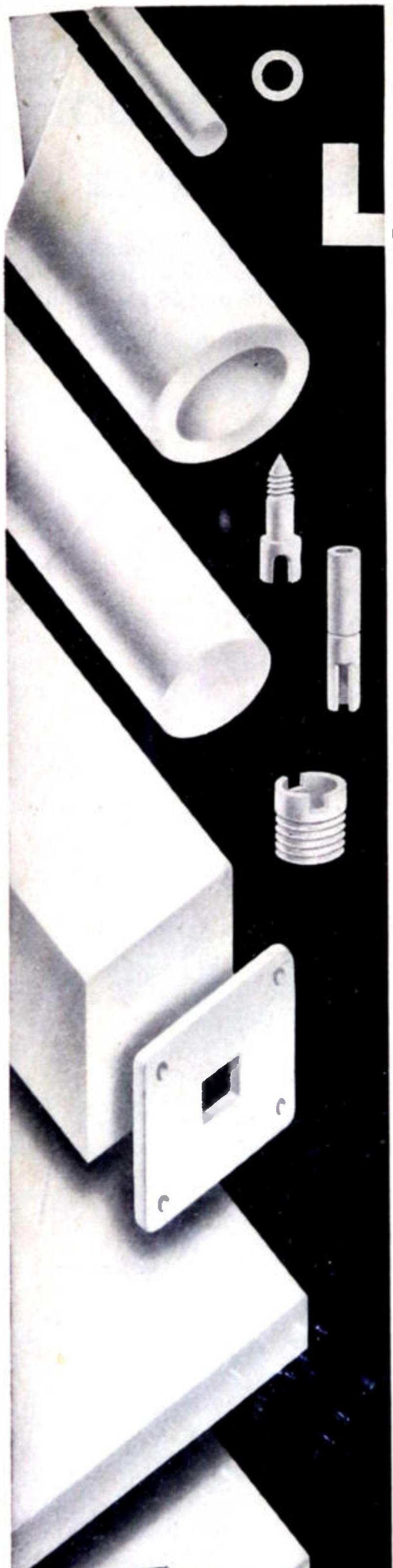


Meetings with Exhibits

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

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.

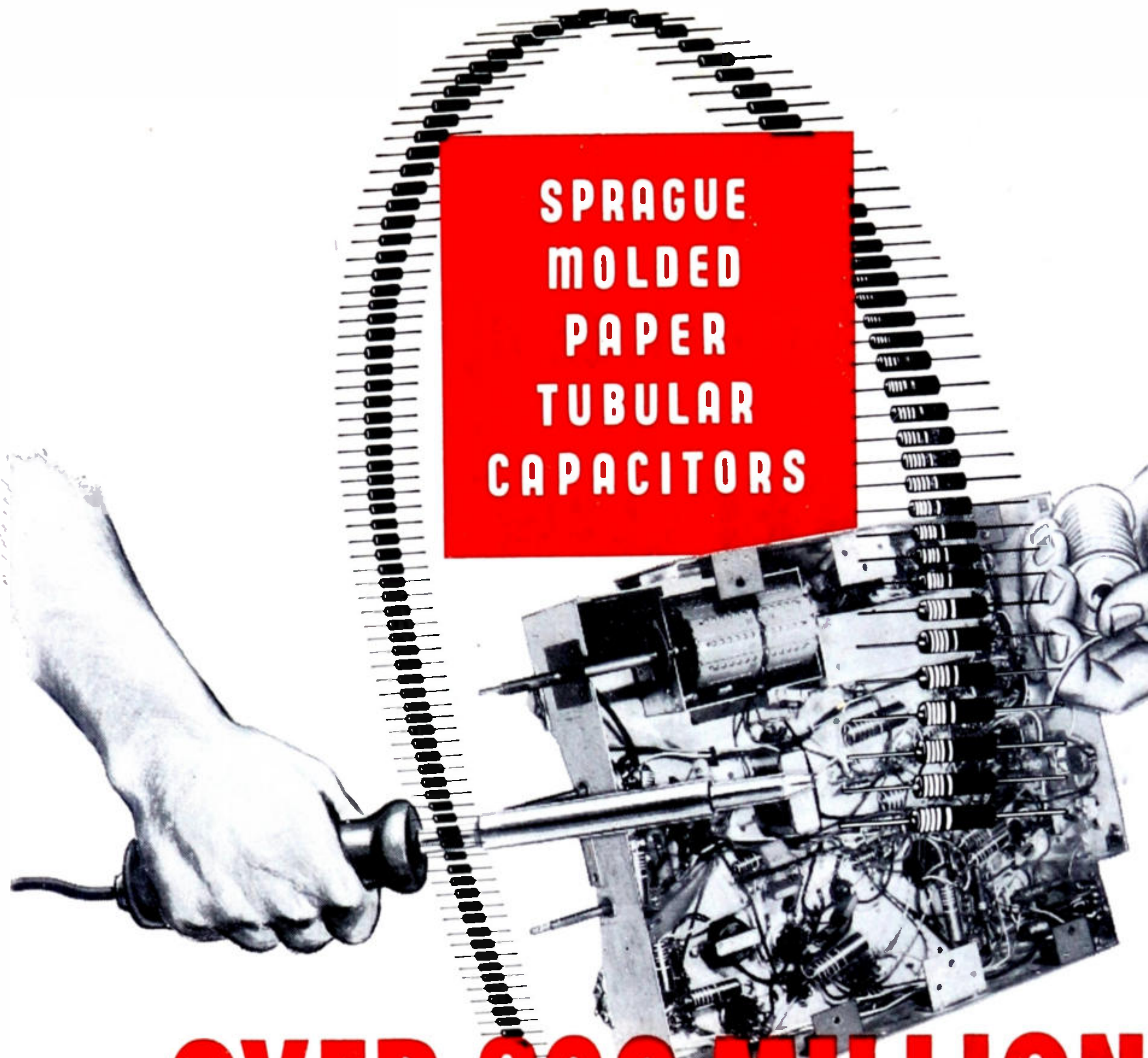
October 29-November 1
Audio Fair Hotel New Yorker, New York, N.Y.

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.

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



OVER 200 MILLION *in use today!*

An unprecedented failure-free service record is the proof of the pudding on the quality of Sprague's Black Beauty phenolic-molded paper tubular capacitors!

And that's why service-conscious TV and radio manufacturers are showing an increasing preference for these dependable capacitors which not only prevent expensive in-warranty service calls but which are insurance for years of set owner satisfaction.

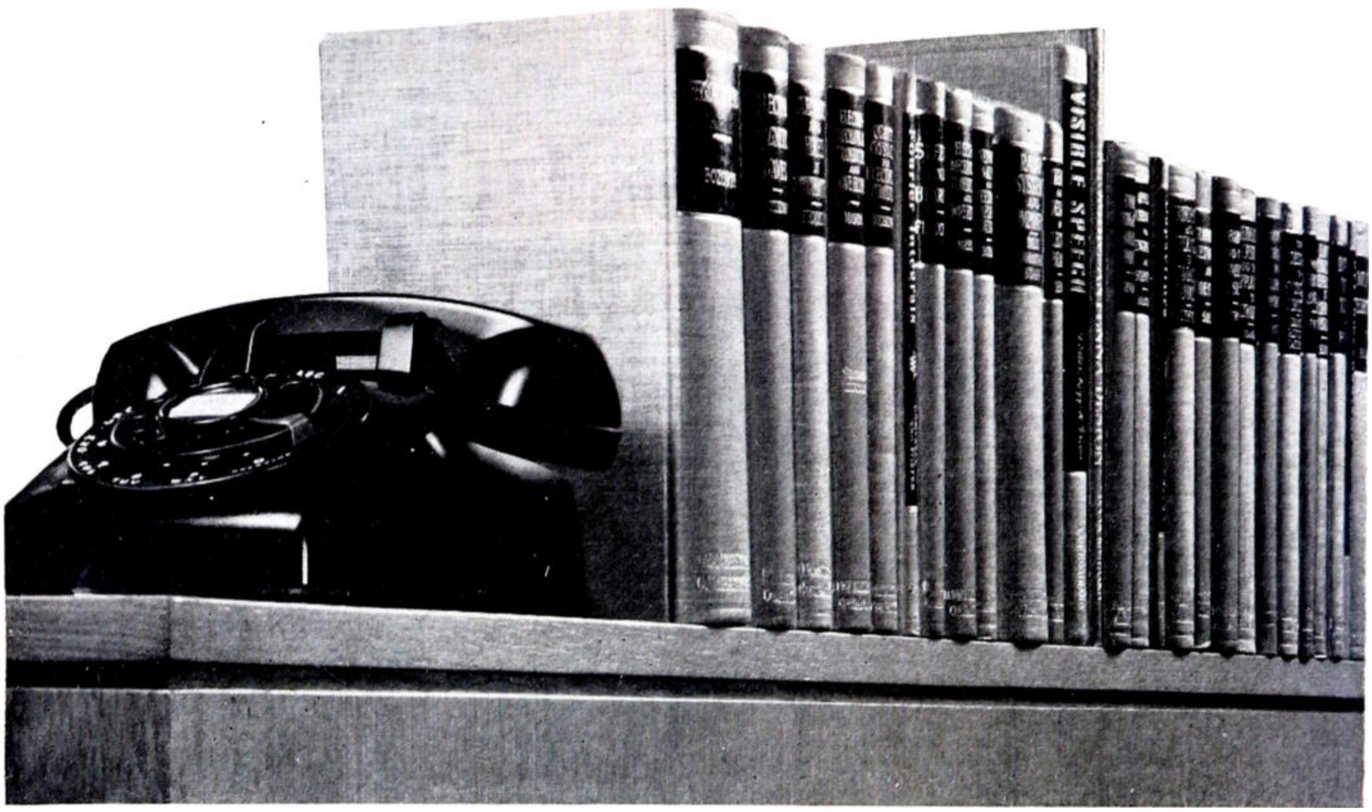
The superiority of Sprague molded capacitors is based on the exclusive Sprague *dry* assembly process, which prevents contamination of capacitor sections during manufacture. Not only is the insulation resistance of these capacitors extremely high, but their capacitance stability and retrace characteristics are unique. The molded housings are non-flammable and offer excellent moisture protection.

Write on your company letterhead for Engineering Bulletins 210-B and 214-A.

PIONEERS IN ELECTRIC AND **SPRAGUE** ELECTRONIC DEVELOPMENT

SPRAGUE ELECTRIC COMPANY • NORTH ADAMS, MASSACHUSETTS

Telephone Science Shares Its Knowledge



The Bell Telephone Laboratories Series of books is published by D. Van Nostrand Company. Other technical books by Laboratories authors have been published by John Wiley & Sons. Complete list of titles, authors and publishers may be obtained from Publication Dept., Bell Telephone Laboratories, New York 14.

List of Subjects: Speech and hearing, mathematics, transmission and switching circuits, networks and wave filters, quality control, transducers, servomechanisms, quartz crystals, capacitors, visible speech, earth conduction, radar, electron beams, microwaves, waveguides, traveling wave tubes, semiconductors, ferromagnetism.

In their work to improve your telephone service, Bell Laboratories make discoveries in many sciences. Much of this new knowledge is so basic that it contributes naturally to other fields. So Bell scientists and engineers publish their findings in professional magazines, and frequently they write books.

Most of these books are in the *Bell Telephone Laboratories Series.* Since the first volume was brought out in 1926, many of the books have become standards . . . classics in their fields. Twenty-eight have been published and several more are in the making. They embody the discoveries and experience of one of the world's great research institutions.

Bell scientists and engineers benefit greatly from the published findings of workers elsewhere; in return they make their own knowledge available to scientists and engineers all over the world.

BELL TELEPHONE LABORATORIES

Improving telephone service for America provides challenging opportunities for individual achievement and recognition in scientific and technical fields.

Built for the toughest service. . . .

Mallory Q Series Wire Wound Controls

If you need a wire wound control that will stand up under the most severe conditions, here's the answer to your problem—Mallory Series Q controls. These new features make the Q series your best choice for military and other exacting applications:

IMPERVIOUS TO MOISTURE AND FUNGUS: all insulation used in this control is made of high resistance material which has exceptionally low moisture absorption . . . treated to prevent fungus growth.

WEATHERPROOF FINISH: nickel plated case, stainless steel shaft, and all other metal parts will pass a 100-hour salt spray test.

LONGER LIFE: hard nickel-silver contacts withstand the wear of thousands of rotations.

SELECTION OF TAPERS: all standard JAN tapers are available.

In addition to these standard features, Q series controls can be supplied in a number of special variations invaluable in applications requiring complete waterproofing or extreme resistance to vibration:

WATERPROOF SHAFT BUSHING: a waterproof gasket between shaft and bushing, sealed with silicone grease, prevents leakage along the shaft.

WATERPROOF PANEL SEAL: gasketed seal prevents leaks at the point of panel mounting.

BUSHING LOCK: a split bushing, when tightened, prevents shaft rotation even under severe shock and vibration.

Mallory carbon controls—with all the construction features of the wire wound units—are also available in the Q series design.

For full information on Q series controls, call or write Mallory today.

Series	Watts	Diameter	Similar JAN Type
QC	2	1$\frac{1}{16}$″	RA15
QR	2	1$\frac{1}{4}$″	RA20
QM	4	1$\frac{3}{8}$″	RA25 & RA30

P.R. MALLORY & CO. Inc.
MALLORY

SERVING INDUSTRY WITH THESE PRODUCTS:
Electromechanical—Resistors • Switches • Television Tuners • Vibrators
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...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

Operations Records Show . . .

. . . Eimac tubes are of the highest electronic standards

TELEPHONE MURRAY HILL 6-7100

CABLE ADDRESS PANAGRA

PAN AMERICAN-GRACE AIRWAYS, INC.
EXECUTIVE OFFICES
CHRYSLER BUILDING - NEW YORK 17, N.Y.

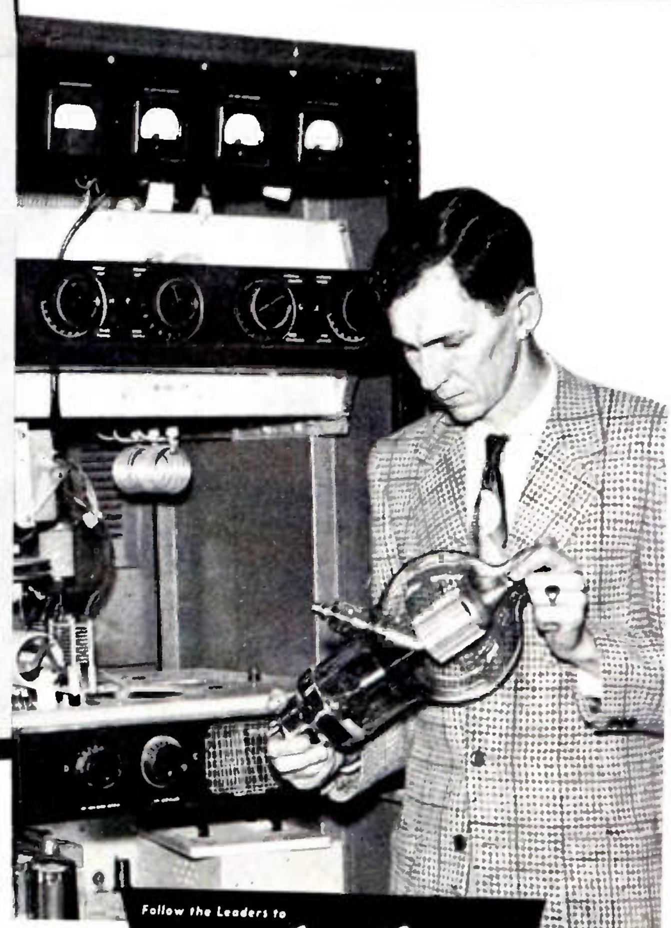
Eitel-McCullough, Inc.
798 San Mateo Avenue
San Bruno, California

Gentlemen:

Panagra operations records show that Eimac tubes have averaged over 10,000 hours of service for us in the ground stations serving our daily, deluxe DC-6 El InterAmericano flights down the west coast of South America. Results in terms of efficiency, economy, and all-around dependability have proven to us that Eimac tubes are of the highest electronic standards.

Very truly yours,

J. H. Dittemore
John H. Dittemore
Assistant Supt. of Communications



Pan American-Grace Airways, Inc., pioneer South American airline has served the west coast countries for 24 years. Find as Panagra has found — that to employ Eimac tubes is to employ the best.

Follow the Leaders to Eimac TUBES

EITEL-McCULLOUGH, INC.
SAN BRUNO, CALIFORNIA
Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

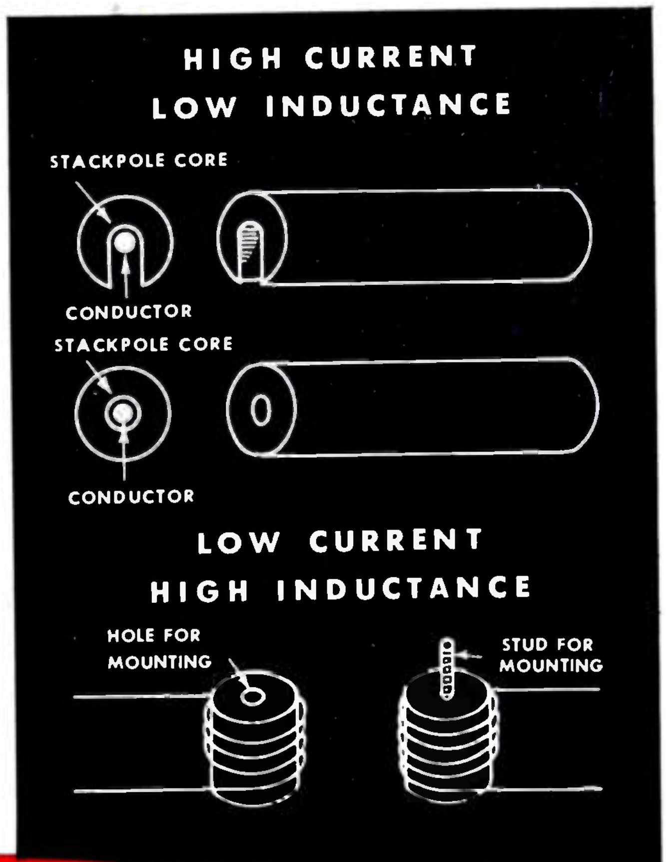
322

Maybe you're missing a bet on

HASH and INTERFERENCE SUPPRESSION!

Recent experience indicates that Stackpole molded iron cores and Ceramag® cores (ferrites) help materially in minimizing "hash" and r-f interference when used in the filter systems of electrical tools and equipment. Their advantages include

- ✔ ***Less IR Drop*** **because of lower d-c resistance**
- ✔ ***Greater R-F Attenuation*** **because of less distributed capacitance**
- ✔ ***Concentrated Field*** **with consequent reduction in coupling to other coils**
- ✔ ***Reduced Space Requirements***
- ✔ ***Increased Inductance*** **for a given amount of wire . . . . or, conversely . . . .**
- ✔ ***Less Wire*** **for the same inductance**



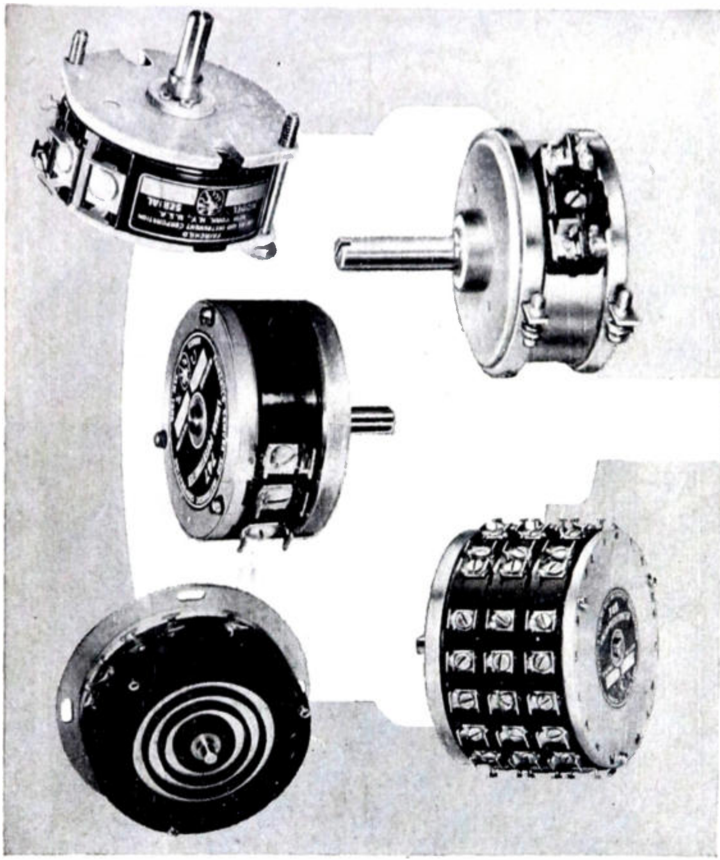
As one of the earliest pioneers in core production, and backed with a complete line of materials, sizes, shapes and production facilities, Stackpole welcomes the opportunity to cooperate in adapting suitable cores to particular noise suppression problems.

Your inquiry entails no obligation and will receive prompt, skilled attention. Write to Electronic Components Division, Stackpole Carbon Company, St. Marys, Pa.

STACKPOLE

MOLDED IRON *Cores* **CERAMAG® (FERRITES)**

Sample Precision Potentiometers now available in 4 to 6 weeks



Better delivery than ever before of Fairchild Precision Potentiometers is the result of recently improved facilities and additions to personnel. Now you can expect delivery of sample standard units with windings to meet your requirements in 4 to 6 weeks after your final approved specifications are received. The same reasonable prices prevail, too.

Enlargement and realignment of facilities and personnel also enable us to start delivery of production orders in 3 to 4 months after receipt of your order.

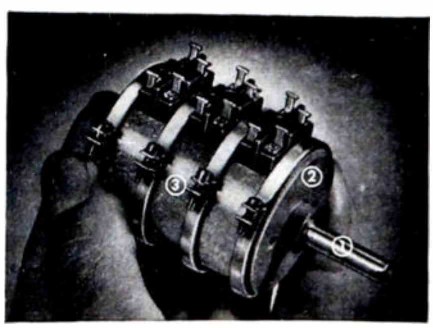
Thus, when you look to Fairchild for your precision - potentiometer requirements you get products built to the highest standards of quality coupled with sound engineering help that starts with your idea and carries through to final delivery.

HOW PRECISION IS DESIGNED AND BUILT INTO FAIRCHILD POTENTIOMETERS

1. Shaft is centerless-ground from stainless steel to a tolerance of +0.0000, –0.0002 in. which, together with precision-bored bearings, results in radial shaft play of less than 0.0009 in.

2. Mounting plate has all critical surfaces accurately machined at one setting to insure shaft-to-mounting squareness of 0.001 in./in. and concentricity of shaft to pilot bushing within 0.001 in. FIR.

3. Housing is precision-machined from



aluminum bar stock. Close tolerance of this construction permits ganging up to 20 units on a single shaft with no eccentricity of the center cups, even though only two bearings are used.

4. Windings are custom-made by an exclusive technique. This, together with precious metal alloy contacts results in guaranteed accuracies of ±0.5% linear and ±1.0% non-linear in standard type potentiometers. Higher accuracies (to 0.05%) are available in other types.

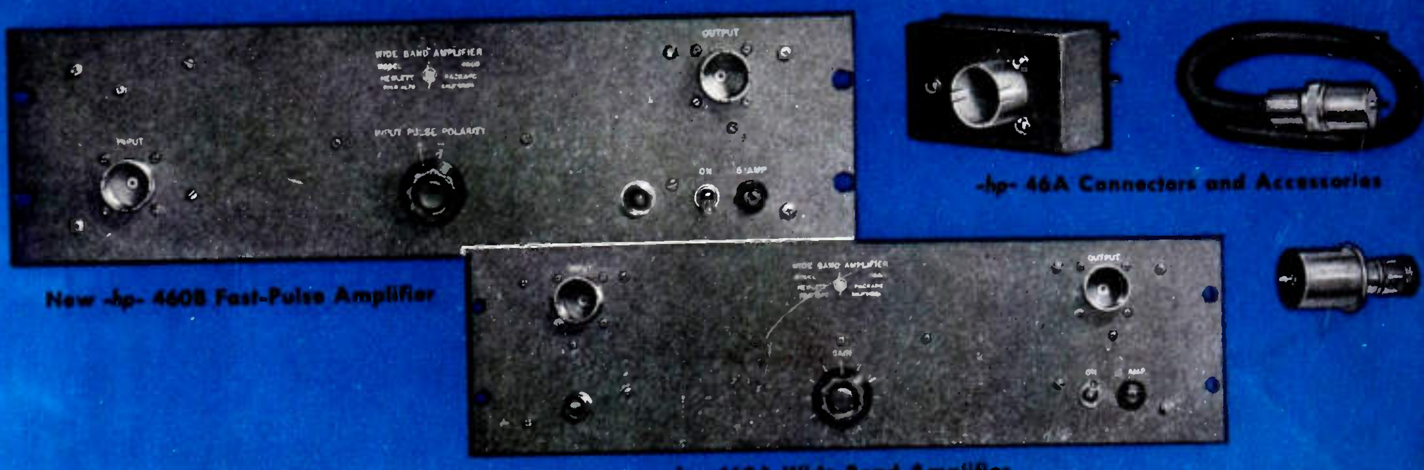
DO YOU HAVE CONTROL PROBLEMS?

Fairchild Sample Laboratory engineers are available to help you with potentiometer problems. To get the benefit of their knowledge and experience write today, giving complete details, to Potentiometer Division, Fairchild Camera and Instrument Corporation, Park Avenue, Hicksville, L. I., New York, Department 140-29H.

FAIRCHILD

PRECISION POTENTIOMETERS

COMPLETE, WIDE-BAND INSTRUMENTATION FOR DISTORTION-FREE, FAST-PULSE MEASUREMENT!



New -hp- 460B Fast-Pulse Amplifier

-hp- 46A Connectors and Accessories

-hp- 460A Wide-Band Amplifier

SPECIFICATIONS

-hp- 460B FAST PULSE AMPLIFIER

FREQUENCY RESPONSE: Closely matches Gaussian curve. Hf 3 db point is approx. 140 mc. Lf 3 db point is approx. 50 kc into 200-ohm load.

MAXIMUM OUTPUT VOLTAGE: High bias, approx. 125 v. negative open circuit. Normal bias (linear amplification) approx. 8 v. peak into 200-ohm load or 16 v. peak open circuit, pos. or neg. pulses.

GAIN: Approx. 15 db into 200-ohm load.

INPUT IMPEDANCE: Approx. 200 ohms.

RISE TIME: Approx. 0.0026 μsec.

DELAY: Approx. 0.016 μsec.

DUTY CYCLE: 0.10 max. for 125 v. output pulse.

LINEARITY PULSE OPERATION: See Figure 1.

MOUNTING: Relay rack. 5¼" x 19", 6" deep.

POWER SUPPLY: 115 v. 50/60 cps. 35 watts.

PRICE: $225.00 f.o.b. factory.

-hp- 460A WIDE-BAND AMPLIFIER

(Specifications same as Model 460B except:)

MAXIMUM OUTPUT VOLTAGE: Approx. 8 v. peak open circuit; 4.75 v. peak into 200-ohm load.

GAIN: Approx. 20 db with 200-ohm load.

DELAY: Approx. 0.012 μsec.

PRICE: $185.00 f.o.b. factory.

-hp- 46A ACCESSORIES

- -hp- **46A-16A PATCH CORD** — 200-ohms, 2' long. $18.50.
- -hp- **46A-16B PATCH CORD** — 200-ohms, 6' long. $25.50.
- -hp- **46A-95A PANEL JACK**—For 200-ohm cables, low capacitance. 1⅛" dia. $7.50.
- -hp- **46A-95B CABLE PLUG**—For 200-ohm cables, low capacitance. $7.50.
- -hp- **812-52 CABLE**—200-ohm cable in lengths to specification. Per foot $1.75.
- -hp- **46A-95C 50-OHM ADAPTOR** — Type N connector for coupling 50-ohm line into -hp- amplifiers. $15.00.
- -hp- **46A-95D ADAPTOR**—Bayonet sleeve for connecting -hp- 410A VTVM to output of 460A/B amplifiers. $15.00.
- -hp- **46A-95E CONNECTOR SLEEVE**—Joins two 46A-95B CABLE PLUGS. $7.50.
- -hp- **46A-95F ADAPTOR**—For connecting to 5XP CRT. $10.00.
- -hp- **46A-95G ADAPTOR**—For connecting to Tektronix type 511 oscilloscope. $12.50.

Data subject to change without notice.

UP TO 90 DB GAIN IN CASCADE! AMPLIFIES MILLI-MICROSECOND PULSES! RISE TIME .0026 μSEC! 125-VOLT OPEN-CIRCUIT OUTPUT! GIVES OVER 100 MC BANDWIDTH TO YOUR STANDARD OSCILLOSCOPE!

Here at last is complete instrumentation for true amplification of fast pulses at high power levels sufficient to operate scalers or counting meters, cathode ray tubes, or to give more than 100 mc band-width to your present oscilloscope. New *-hp-* 460B Fast-Pulse Amplifiers, in cascade with *-hp-* 460A Wide-Band Amplifiers, amplify up to 125 volts, open circuit (limited duty cycle). This permits full deflection of 5XP cathode ray tubes, or 2-inch deflection of 5CP tubes. Ultra-short rise time of 0.0026 μsec, combined with zero overshoot, insures distortion-free amplification of pulses faster than 0.01 μsec.

New *-hp-* 460B Amplifier, cascaded with *-hp-* 460A provides linear amplification of 16 volts peak output and pulse amplification of 125 volts output (slight non-linearity). This combination provides maximum usefulness in fast-pulse study for nuclear radiation work, television or VHF research; for increasing frequency range of your oscilloscope, or general wide-band laboratory amplification. In addition to the above instrumentation, *-hp-* also offers series 46A accessories—a complete set of 200 ohm cables, adapters and fittings for inter-connecting amplifiers or patching to oscilloscopes.

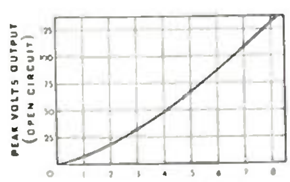


Fig. 1: Linearity of -hp- 460B Fast-Pulse Amplifier

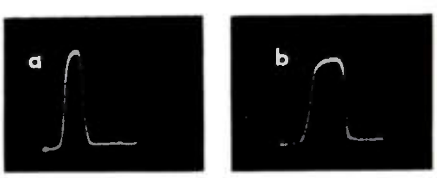


Fig. 2: (a) 0.01 μsec pulse through -hp- 460B Amplifier (b) 0.02 μsec pulse through 3 amplifiers in cascade

Get complete details. Write direct or see your -hp- sales representative.

HEWLETT-PACKARD COMPANY

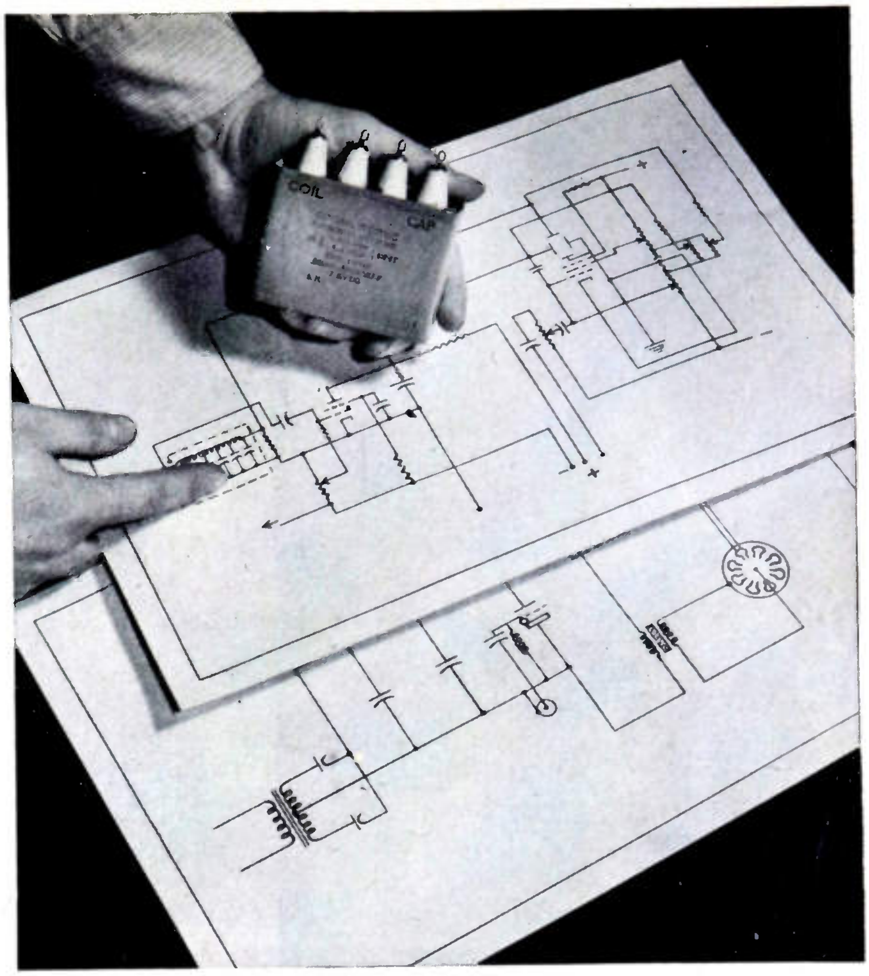
2177D Page Mill Road • Palo Alto, California, U.S.A.

Sales Representatives in all principal cities.

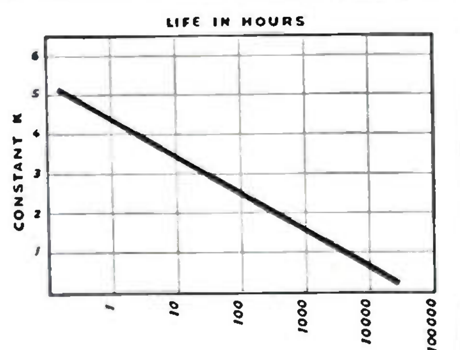
Export: Frazar & Hansen, Ltd., San Francisco, New York, Los Angeles

HEWLETT-PACKARD hp INSTRUMENTS

Got a really tough capacitor network problem for us?



...let our network designers help you solve it!



This curve results from actual life tests of capacitor networks under pulsing conditions in a radar modulator. Service life, for varying values of "K", starts at 2 tenths of one hour and stretches out to something over 40,000 hours (more than 4½ years of continuous operation at full-rated voltages.)

It's just one example of actual performance data G-E engineers use in designing capacitor networks for the job they must do.

Whether your problem deals with guided missiles—aircraft—land or sea radar equipments, General Electric application and design engineers can help you solve it. We've designed and built capacitor networks for every type of pulse radar equipment since the inception of radar.

Take service life for example. You can specify a service life of 10,000 hours—or just 60 seconds. And we'll deliver pulse networks to match your requirements. Here's why:

> Since 1944 General Electric has been running continuous life tests on many types of networks. We've established life limitations, under varying conditions of temperature and voltage, for all types of dielectrics, bushings, materials for coil forms and treating processes.

Let us use this store of information and experience to solve your capacitor network problems. Your inquiry addressed to your nearest Apparatus Sales Office, or to Capacitor Sales Division, General Electric Company, Hudson Falls, N. Y. will receive prompt attention.

General Electric Company, Schenectady 5, New York.

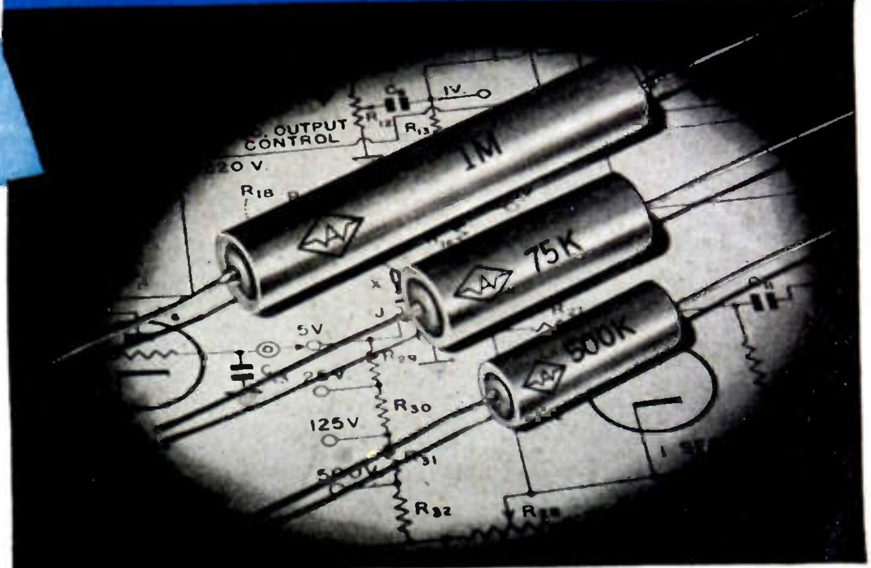
GENERAL ELECTRIC

407-308

Carbofilm® RESISTORS

HERMETICALLY SEALED

TOUGH FOR ROUGH GOING!



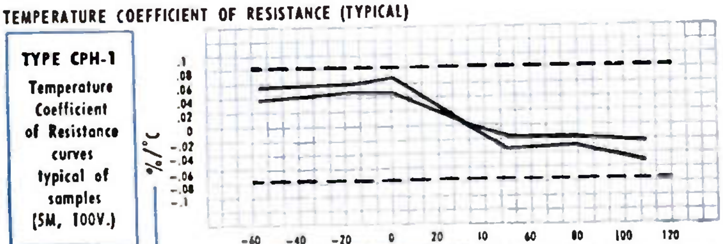
SPECIFICATIONS

1. **Hermetically sealed.** Metal case. Vitrified ceramic end seals, pigtail leads. Thoroughly protected — mechanically, electrically, climatically.
2. **Temperature Coefficient** not exceeding .0003 ohm per ohm per °C. over temperature range of −40°C. to +60°C. up to 15 megohms. Not exceeding .0005 ohm per ohm per °C. up to 100 megohms.
3. **Voltage Coefficient** does not exceed .002% per volt.
4. **Overloads** up to 200% of rated voltage, without showing permanent change in resistance.
5. **Accuracy:** guaranteed tolerance of plus/minus 1% at 25°C. (77°F.).
6. **Aging Changes** negligible. Average change in resistance for self-aging, approximately 0.2% in a year.
7. **Noise:** Silver-to-silver contacts insure very high stability and correspondingly low noise levels.
8. **In Four Sizes:** Two ½ watt, 1 watt and 2 watt. Cased or uncased.

Wilkor, the first licensee under Western Electric patents to produce carbon deposited precision resistors, takes another step forward. Wilkor now offers hermetically-sealed Carbofilm Resistors, the first fully-protected precision resistors available on a production basis.

Primarily intended for circuits calling for the accuracy and stability of wire-wound resistors, yet with the compactness of carbon or composition-element resistors. Excellent for measuring-instrument applications; in test and lab equipment; in oscillography and other critical electronic circuits; in electronic computers and allied techniques; and now, in the encased, hermetically-sealed construction, particularly in applications where resistance values must be critically maintained over long service life, regardless of climatic conditions.

TEMPERATURE COEFFICIENT OF RESISTANCE (TYPICAL)



Literature on request. Let us collaborate in your precision-resistor requirements.

WILKOR DIVISION

AEROVOX CORPORATION
CLEVELAND, OHIO

*Trade Mark

AEROVOX CORPORATION NEW BEDFORD MASS. **HI-Q®** DIVISION OLEAN, N. Y.

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Our Best References Are Our Customers

RMC DISCAPS®

... Specified and Used by Leading TV Makers

The list of satisfied RMC DISCAP customers reads like the "Blue Book" of the TV industry. Few are missing from this top level roster.

RMC offers a full line of by-pass as well as temperature compensating ceramic disc capacitors.

Engineers specify them for their uniform high quality, low inherent inductance and small size.

Purchasing agents specify them because they can depend on RMC to make delivery when needed.

RMC temperature compensating disc capacitors (which meet the RTMA spec for class one ceramic capacitors) are designed to replace tubular ceramic and mica capacitors at a lower cost.

Send for Samples and Technical Data

DISCAP CERAMIC CONDENSERS

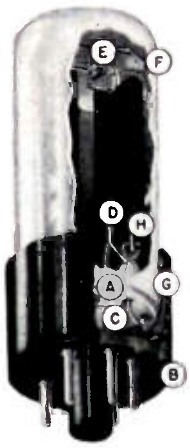
RMC

RADIO MATERIALS CORPORATION
GENERAL OFFICE: 3325 N. California Ave., Chicago 18, Ill.
FACTORIES AT CHICAGO, ILL. AND ATTICA, IND.
Two RMC Plants Devoted Exclusively to Ceramic Condensers

CONSTRUCTIONAL HIGHLIGHTS 5Y3WGTA

For high altitudes: A. Cavity stem (patent pending). B. Barrier base. C. Optimized lead spacing. All three offer maximum isolation and insulation of high-voltage leads for stratosphere operation.

For ruggedization: D. Four-point mount support. E. "Mouse-trap" filament tensioner springs. F. Resilient superstructure cross springs. G. Low-pass mechanical filter between base and mount structure to absorb high-frequency components of shock. H. Cataphoretic-coated filament.



Is your aircraft equipment climbing up . . . up . . . up? Need an all-purpose rectifier — preferably ruggedized — to meet the challenge? High-altitude 5Y3WGTA . . . also the original ruggedized filamentary-type tube . . . is your answer.

At 50,000 feet* CBS-Hytron 5Y3WGTA offers you: Same maximum current and voltage ratings (with safe bulb temperatures) as the standard 5Y3GT at sea level. Plus JAN-1A ruggedization to withstand destructive shock, vibration, acceleration, and impact. And single-ended construction . . . convenient for both new and older equipment. (The 5Y3WGTA is interchangeable with the 5Y3GT or 5Y3WGT.) Check the 5Y3WGTA's ratings . . . its rock-solid construction.

90,000 FEET UP! New CBS-HYTRON 6004

Climbing higher still? Plate connections to top caps of 6004 push ceiling far into stratosphere. CBS-Hytron 6004 operates at 90,000 feet — higher at adjusted ratings — free from arc-over and at safe bulb temperatures. See comparative data for ratings.

COMPARATIVE DATA

Max. Ratings	5Y3WGTA	6004
Operating altitude	50,000 ft.*	90,000 ft.*
Peak inverse plate voltage	1,400 v†	1,000 v††
Peak plate current per plate	400 ma.	400 ma.
Bulb temperature	185° C	185° C
JAN-1A ruggedized	Yes	No
Basing	Single-ended	Double-ended

*Adjusted rating chart available for higher altitudes. † At 50,000 feet. †† At 90,000 feet

CBS HYTRON RADIO TUBES

MAIL TODAY FOR COMPLETE DATA

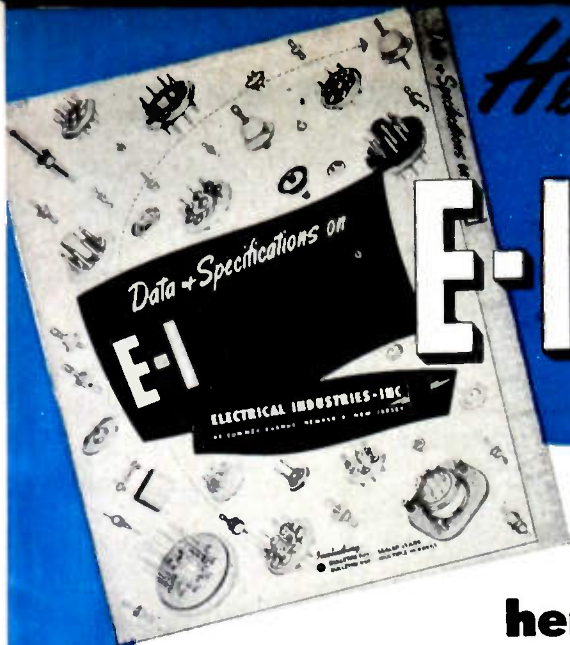
MANUFACTURERS OF RECEIVING TUBES SINCE 1921
HYTRON
RADIO AND ELECTRONICS CO.
A DIVISION OF COLUMBIA BROADCASTING SYSTEM, INC.

SALEM, MASSACHUSETTS

HYTRON RADIO & ELECTRONICS CO.
Salem, Massachusetts

Please send me full data (including adjusted rating chart for higher altitudes) on CBS-Hytron high-altitude rectifiers: 5Y3WGTA and 6004.

YOUR NAME ....
(please print)
COMPANY ....
STREET ADDRESS ....
CITY AND STATE .... P



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E-I STANDARD TYPES — Offering a quick solution to problems relating to sealed terminals. Complete information on standard sealed leads, standard multiple headers, octal plug-in headers and end seals is contained in these helpful catalogs.

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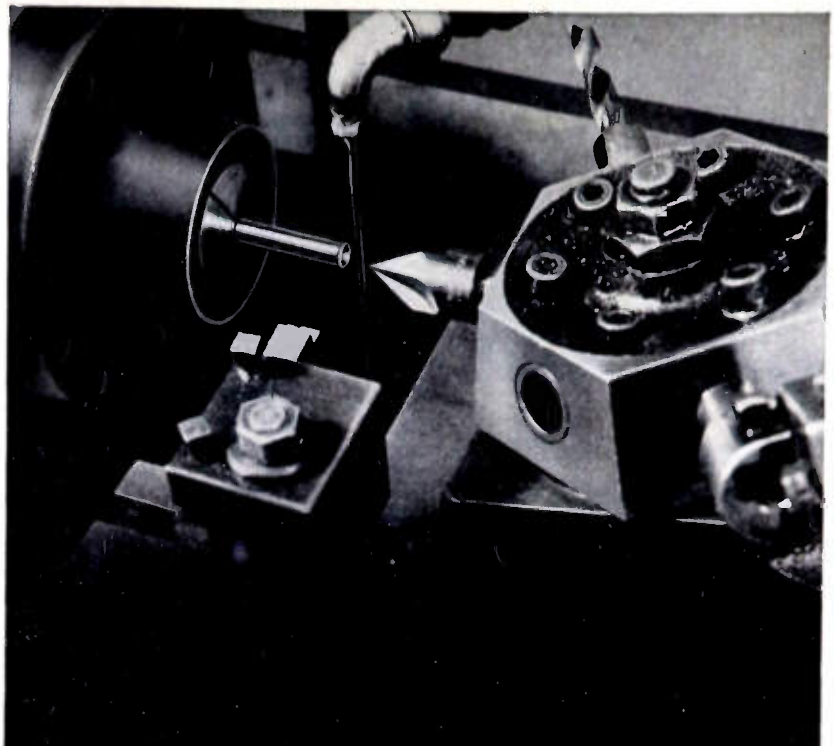


How Superior Fabricates Tubing

to give you the parts you need

Need a tubular part machined, inside or out, at one or both ends?

Like to have it drilled transversely at one or several points?

Want it to meet rigid dimensional and metallurgical specifications?

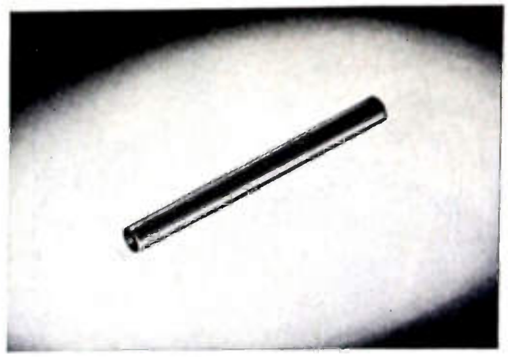
You're reading the right advertisement for all of these are Superior Specialties.

Superior has the experienced men, the specialized, highly developed equipment, the floor space, and the research facilities to produce quantities of drilled and machined tubular parts rapidly and economically.

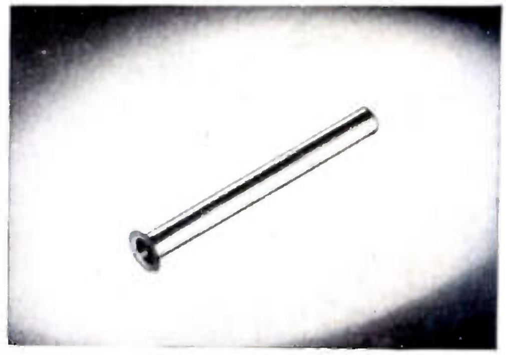
It's a job we like to do and know how to do. But there's more to the story than simple production of fabricated or semi-finished parts, or even top-quality tubing in any analysis and many sizes.

The rest of the story is our willingness, desire and ability to work closely with customers' development engineers and product designers. Frequently we are able to materially assist in design of parts, selection of analysis, and development of processes. Many times we have been able to suggest minor changes in shape or method to effect major economies in assembly time and product cost.

If you are a manufacturer or an experimenter in electronics and have a need for a tubular part of any kind, check with us. We can probably help by giving you quantity production of the parts you need. Write Superior Tube Company, 2506 Germantown Ave., Norristown, Pennsylvania.



Cut and Annealed. Extensive cutting equipment, hand cutting jigs, electronically controlled annealers and other equipment, much of it developed within our own organization results in high speed, precision production of parts.



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NICKEL ALLOYS FOR OXIDE-COATED CATHODES: This reprint describes the manufacturing of the cathode sleeve from the refining of the base metal. Includes the action of the small percentage impurities upon the vapor pressure, sublimation rate of the nickel base; also future trends of cathode materials are evaluated.

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All analyses .010" to ⅝" O.D.
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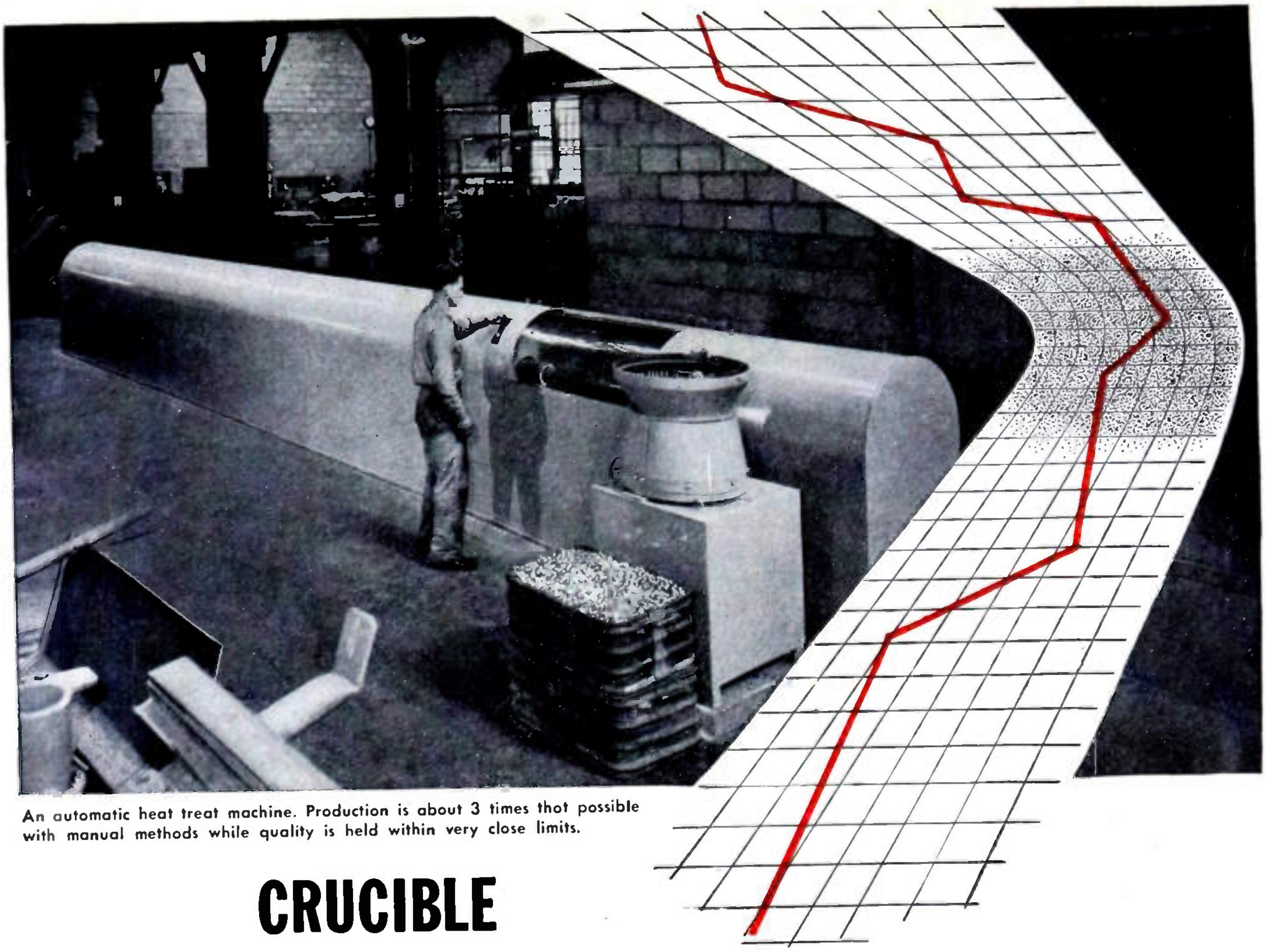
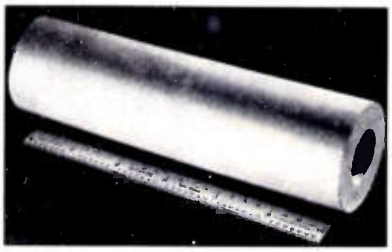


An automatic heat treat machine. Production is about 3 times that possible with manual methods while quality is held within very close limits.

Discharge Capacitor

Centralab, Div. Globe-Union Inc., 900 E. Keefe Ave., Milwaukee 1, Wis., announces the availability of a new type ultra-high speed discharge capacitor which has the characteristics of 30 feet of solid coaxial transmission cable. The size of the tube is 2 inches in diameter, 6½ inches long. When used in the same manner as the coaxial cable, charged to 10,000 volts and discharged across a spark gap, the capacitor tube improves light intensity 900 times.



The capacitor tube is made of hi-K ceramic (K-2000), silvered inside and out. It has a capacitance rating of at least 0.024 μf (24,000 $\mu\mu$f), and immersed in transformer oil is rated at a working voltage of 20,000 volts dc. The unit has a decay time, peak to peak, of 2×10^{-7} second, a rise time zero to peak, of 2×10^{-7} second, and 50 per cent of peak limits occur in a period of 1.8×10^{-7} second. Leakage resistance is in excess of 10,000 megohms, and dielectric strength is approximately 35 volts per mil. Other specifications can be obtained by writing Centralab.

Centralab's capacitor number is DA778-001. The unit may be ordered directly from the factory or through any authorized Centralab distributor, net price of $51.00 each, F.O.B. Milwaukee.

Germanium Diodes

A new range of tapered germanium diodes featuring "polarity at a glance," and designed to replace many of the present type electronic tubes such as detectors and rectifiers, is now being manufactured by **Radio Receptor Co., Inc.,** 251 W. 19th St., New York 11, N. Y.

The unit consists of a germanium wafer soldered to a nickel alloy cathode pin, and an electro-etched tungsten whisker welded to a nickel alloy anode pin, assembled into a glass-phenolic body. The entire assembly is positively impregnated with a polyethylene compound using the vacuum-pressure method.

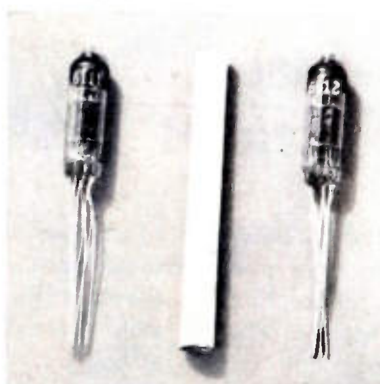
The diode may be clip mounted by the terminal pins, or soldered in by the copper-

tin clad iron "pigtail" leads which are welded into the pins.

Types available include the JAN preferred types, 1N69, 1N70 and 1N81. No. 1N69 is a general purpose and vhf rectifier unit. No. 1N70 is a high voltage diode, and No. 1N81 is a medium voltage diode with low back leakage near 10 volts. Low cost commercial types are the 1N48 and 1N51 general purpose diodes, the 1N64 TV video second detector and 1N65 dc restorer. High performance premium commercial units include the 1N52, 1N63 and 1N75 which are distinguished by low back leakage.

Subminiature Double Triodes

The new subminiature double triodes, types 6111 and 6112, have been announced by the **Radio Tube Division, Sylvania Electric Products Inc.,** 1740 Broadway, New York 19, N. Y.



Both of these tubes are suitable for use at frequencies ranging up into the uhf region.

Type 6111 is a medium-mu double triode in a T-3 envelope, with characteristics similar to those of type 6SN7GT and may be used for similar applications, within the 6111's ratings. Characteristics of the new subminiature 6111 include: Filament, volts—6.3; Filament, current, ma—300.0; Plate, volts (Maximum)—150.0; Plate current, ma. (maximum)—22.0; Plate dissipation, watts (maximum)—1.1; Transconductance, micromhos—5000.0; Amplification factor—20.0.

Type 6112 is a high-mu double triode in a T-3 envelope with characteristics similar to those of type 6SL7GT and may be used for similar applications, within the 6112's ratings. Characteristics of the new subminiature 6112 include: Filament, volts—6.3; Filament, current, ma—300.0; Plate volts (maximum)—150.0; Plate current, ma. (maximum) 1.25; Transconductance, micromhos—2500.0; Amplification factor—70.0.

Static Detector

Keithley Instruments, Dept. 206, 3868 Carnegie Ave., Cleveland 15, Ohio, has developed a highly sensitive static detector, designated as Model 2005. The device clips onto a Keithley vacuum tube electrometer, providing a convenient combination for detecting and locating static charges.

The new electrometer accessory consists primarily of two concentric tubes and an aluminum rod. When clipped over the HI terminal electrode of the electrometer, the tubes provide shielding which gives greater effectiveness to charges along the cylinder axis.

Results are qualitative and observed by noting the deflection of the meter pointer. A wide range of sensitivity can be attained by raising or lowering the inner tube. With the tube lowered, a charged pocket comb throws the pointer off scale from a distance of 10 feet.

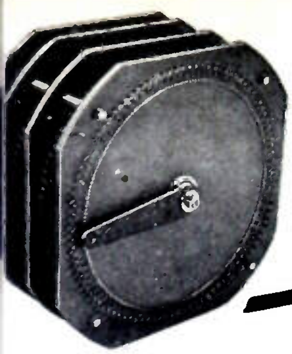
Ion Trap

A new low-priced clip-on ion trap of simplified construction has just been announced by **Heppner Manufacturing Co.,** Round Lake, Ill. The new simplified steel construction lowers manufacturing costs by fully utilizing for the first time the maximum efficiency of the Alnico permanent magnet. This makes Model T-312 the lowest priced ion trap available, according to the manufacturer.

Each trap is stabilized and tested on special equipment designed by Heppner for this specific purpose. Installation time is 2 or 3 seconds. The smooth metal-to-glass contact permits easy adjustment. Model T-312 stays put without wobble or shift during shipment of the completed TV set. It is also light in weight, ⅜ ounce, so the tube's neck cannot be harmed. Gauss readings range from 25 to 60.

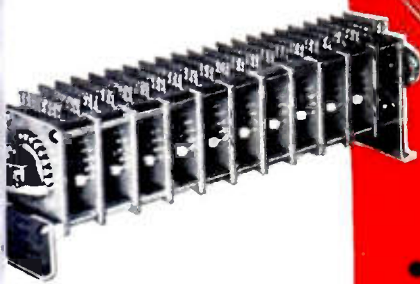
(Continued on page 36A)

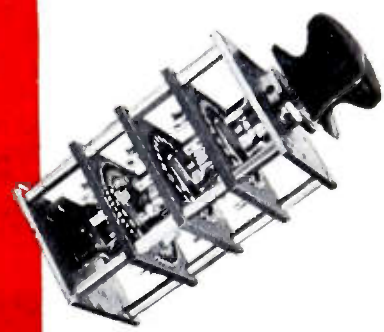
Quick Switches

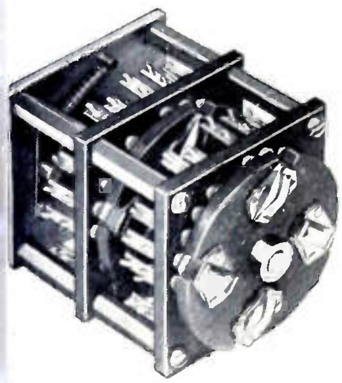


are a

DAVEN SPECIALTY







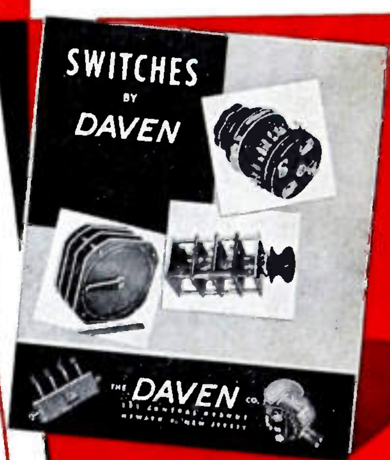
And the "specialty of the house" is double-barreled . . . first, choose from hundreds of standard units to satisfy your needs—for quick switch delivery . . . second, Daven can effect quick 'switches' or changes from standard units to special switches, by using components at hand. Standard parts can be adapted for *your* switch. That too makes for speed, dependability, economy. Write for more detailed data.

Here's Why Daven Switches Excel

- Low and uniform contact resistance.
- Minimum thermal noise.
- High resistance to leakage.
- Trouble-free operation and long life.
- Roller-type positive detent action.
- Depth of unit not increased by addition of detent.

Standard Daven Switches may be the answer to many of your problems. Therefore, check this list below for many of the popular types that are readily available.

Type	Operation	Maximum No. of Positions (per pole)	Maximum Poles per Deck	Deck
G1A	Make before break	24	1	1 3/8"
C1A	Make before break	31	1	1 3/4"
C2B	Break before make	15	1	1 3/4"
D1A	Make before break	47	1	2 1/4"
D7A	Make before break	14	4	2 1/4"
D8B	Break before make	7	4	2 1/4"
D9A	Make before break	9	5	2 1/4"
E3A	Make before break	47	2	2 3/4"
E8B	Make before break	12	4	2 3/4"
E11A	Make before break	15	6	2 3/4"
F1A	Make before break	60	1	3"



It's Free!

Your copy of Daven's complete, new bulletin on switches. Write for it today.

THE DAVEN CO.

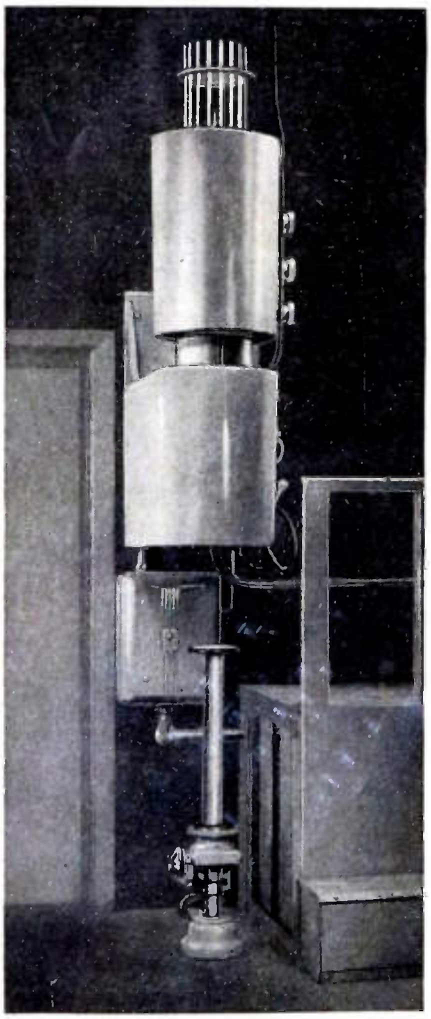
195 CENTRAL AVENUE • NEWARK 4, NEW JERSEY

VISIT DAVEN'S BOOTH 329 AT THE CLEVELAND INSTRUMENTS SHOW—SEPTEMBER 8 TO 12.

LITTON INDUSTRIES NEWS

BETTER CONTROL OF COPPER OR ALLOY BRAZING WITH LITTON HYDROGEN FURNACE

Litton Model 4400 Vertical Hydrogen Furnace is designed for easily observed, accurately controlled production-line brazing of assemblies up to 6½" in diameter and 12" in length. Brazing is performed in a hydrogen atmosphere and work can be inserted into the open bottom either mechanically or hydraulically. Operating temperature range permits copper brazing as well as all types of gold-copper and silver alloy brazing.



Model 4400 Furnace is divided into two chambers. The upper or brazing chamber is equipped with radiant heating for maximum flexibility. The lower or cooling chamber permits rapid cooling to the freezing point of the metal or alloy. The heating chamber has an inconel inner wall surrounded by 3" of thermal insulation. Two replaceable pyrex windows permit a clear view of the work during the heating cycle. Tungsten heating rods are spring-loaded to preserve tautness, and may be easily replaced. The cooling chamber is a double-walled cylinder of stainless steel within which water is circulated.

In operation, work is raised into the upper chamber, heated at the desired rate or rates, and immediately lowered into the cooling chamber. Since power is applied only during the heating cycle (normally less than one-third of loading, heating and cooling time), power consumption is minimized.

SPECIFICATIONS—MODEL 4400 VERTICAL HYDROGEN FURNACE

Work diameter, max.	6½"
Work length, max.	12"
Temperature, max.	1250°C
Voltage to maintain 1250°C	Approx. 22v
Kva to maintain 1250°C	Approx. 23 kva
Overall height	75"
Overall diameter, heater	17"
Overall diameter, cooler	12"

Heater elements: 15 Tungsten rods, .050" dia. x 40" long, connected in parallel.

Time to raise furnace and work to 1000°C: Approx. 17 minutes.

GLASS BAKING OVENS

Litton Glass Baking Ovens are circular and easily mount in any exhaust position. Heating is by Calrod units and ovens are designed for continuous operation at 500°C. Oven models 2, 3 and 4 can be operated in either series or parallel. Ovens range from 5" to 12¾" in diameter, and 12" to 18" in length. Complete details and prices for all models will be supplied on request.

MODEL 5301 BELL JAR

For smaller brazing problems, Litton table-top Bell Jars offer maximum convenience and speed. Visibility through the all-glass jar simplifies alignment and positioning of the work. Vertical movement of the bell is lightened by a counterweight inside the supporting column. Work stand height is variable, and the heater rod can be adjusted and locked in position.

SPECIFICATIONS—MODEL 5301 BELL JAR

Base	11½" x 16½"
Column height	56¾"
Heater stand, height	23½"
Heater stand arm (extended length)	10⅝"
Heater stand, vertical travel	12"
Work stand extensions	2", 4", 6", 8" and 12"
Jar diameter	12"
Height	24"
Travel of jar	28½"

Prices, delivery information on request.

LITTON INDUSTRIES

2538 SAN CARLOS, CALIFORNIA, U. S. A.

DESIGNERS AND MANUFACTURERS of: Glassworking Lathes and Accessories, Vertical Sealing Machines, Burner Equipment, Precision Spotwelders, Oil Vapor Vacuum Pumps, Glass Baking Ovens, Vacuum Tubes and Tube Components, Magnetrons, High Vacuum Molube Oil, Microwave Equipment.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f this Catalog which includes
7 NEW TUBE TYPES
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here are more RAYTHEON SUBMINIATURES
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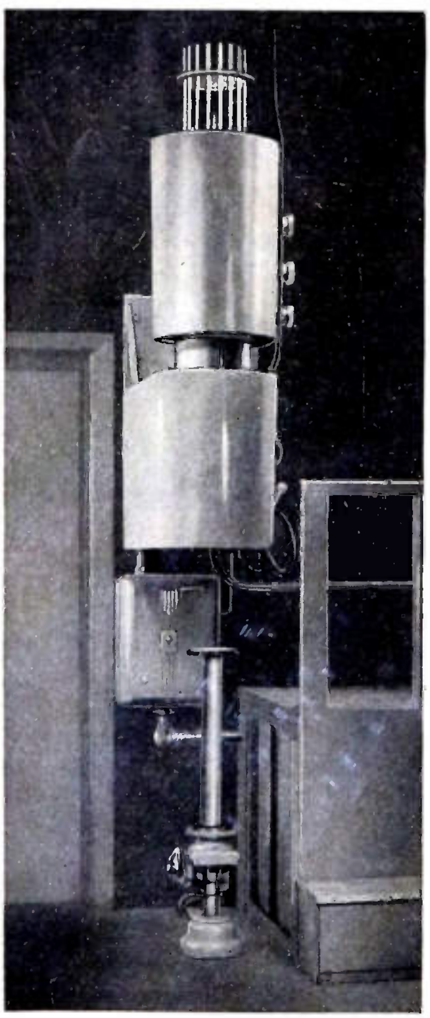

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in world-wide use than all other makes combined

RAYTHEON

RELIABLE

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Thirteen Years of

SUBMINIATURE TUBE DESIGN AND PRODUCTION EXPERIENCE

All meeting military requirements for RELIABILITY based on field and production tests for

SHOCK • VIBRATION
FATIGUE • 5000 HOUR LIFE
CENTRIFUGAL ACCELERATION
HEATER CYCLE LIFE
HIGH TEMPERATURE LIFE
LEAD FATIGUE

Usable in the UHF region

Type	Description	Heater Volts	Heater Ma	Plate Volts	Plate Ma	Grid Volts	Screen Volts	Screen Ma	Amp. Factor	Mut. Cond.
CK5702WA	RF Amplifier Pentode	6.3	200	120	7.5	$R_k = 200$ ohms	120	2.5	—	5000
CK5703WA	High Frequency Triode	6.3	200	120	9.0	$R_k = 200$ ohms	—	—	25	5000
CK5744WA	High Mu Triode	6.3	200	250	4.0	$R_k = 500$ ohms	—	—	70	4000
NEW CK5783WA	Voltage Reference		Operating voltage approximately 86 volts between 1.5 and 3.5 ma.							
CK5784WA	RF Mixer Pentode	6.3	200	120	5.2	−2	120	3.5	—	3200
NEW CK5787WA	Voltage Regulator		Operating voltage approximately 100 volts between 5 and 25 ma.							
NEW CK5829WA	Dual Diode	6.3	150		Max. Peak Inverse 360 volts. $I_o = 5.5$ ma. per plate					
NEW CK6021	Medium Mu Dual Triode	6.3	300	100	6.5	$R_k = 150$ ohms	—	—	35	5400
NEW CK6110	Dual Diode	6.3	150		Max. Peak Inverse 460 volts. $I_o = 4.4$ ma. per plate					
NEW CK6111	Medium Mu Dual Triode	6.3	300	100	8.5	$R_k = 220$ ohms	—	—	20	5000
NEW CK6112	High Mu Dual Triode	6.3	300	100	0.8	$R_k = 1500$ ohms	—	—	70	1800
CK6152	Low Mu Triode	6.3	200	200	12.5	$R_k = 680$ ohms	—	—	15.8	4000

Note: All dual section tube ratings (except heater) are for each section.

Write for Raytheon RELIABLE Subminiature Tubes Catalog R containing complete mechanical and electrical data on these tubes.

RAYTHEON MANUFACTURING COMPANY

Receiving Tube Division — for application information call

Newton, Mass. Bigelow 4-7300 • Chicago, Ill. NAtional 2-2770 • New York, N. Y. WHitehall 3-4980 • Los Angeles, Calif. Richmond 7-5524

RAYTHEON®
Excellence in Electronics

RAYTHEON MAKES ALL THESE

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

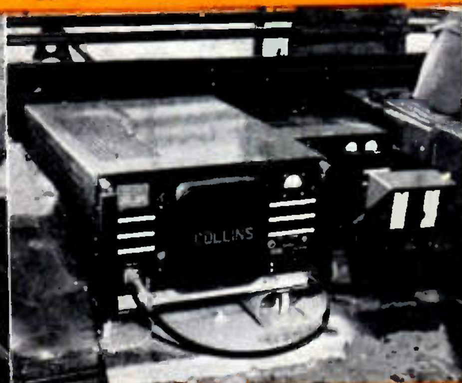
Special Delivery by Air...

Parachuting Signal Corps jeep containing a Collins 18S transceiver upsets on landing.

The jeep's topside cuts a neat furrow into the soft earth and packs the interior with dirt.

Truck rights the jeep. Under other field conditions this would be accomplished by several soldiers.

Protective board covering is removed and the eight-foot whip type antenna released.



Special shock mounts, a heavier dust cover and bracket reinforcements strengthened the 18S.

Radio operator (back to the camera) plugs the microphone and headsets into the transceiver.

Antenna erected and power on, radio operator tunes transmitter and makes test call.

Answer of "receiving loud and clear" proves contact successful.

With contact established, jeep leaves platform.

Tossing radio equipment from airplanes is not generally recommended as standard operating procedure . . . but in recent demonstrations the U. S. Army found the Collins 18S transceiver capable of taking such punishment.

The Collins Radio Company designs and manufactures Communications, Broadcasting, Amateur and Aviation Radio equipment. Write for complete descriptive literature.

For quality in radio equipment, it's . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 W. 42nd St., NEW YORK 18 | 2700 W. Olive Ave., BURBANK | 1930 Carpenter Blvd., DALLAS 2

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 30A)

Radar Tube

Development of a new electronic tube, which makes possible the operation of a beacon radar from a single antenna, is announced by **Sylvania Electric Products Inc.**, 1740 Broadway, New York, N. Y.

Previously, reliable beacon operation required the use of two separate antennae, one for receiving and one for sending.

This model, type 6214, developed to meet the special requirements of beacon radar, is capable of instantaneous operation of the first pulse of a coded system of pulses. Conventional ATR tubes often fail to operate immediately when the transmitter starts, thus preventing proper transmission for this period of time, possibly as long as several seconds.

The instant starting feature of the new tube has been achieved by adding an ignitor electrode to the end plate of the tube. The power supply for this ignitor is taken from the supply for the TR tube ignitor.

Single Phone Headset

Recently approved by the CAA after laboratory and flight tests, a new single-phone type headset has been placed on the market by **Airphone Co.**, Suite 309, Calumet Bldg., Miami, Fla.

The unit is called "AIRPHONE," and according to the CAA Type Certificate

(Continued on page 44A)

New Model 802 Stable Microwave Oscillator

provides a highly stable source of microwave signals

The LFE Model 802 Stable Microwave Oscillator provides a source of highly stabilized microwave frequencies suitable for use as a local oscillator for microwave measurements, or in any other applications where a high degree of stability is required. A dial accurately calibrated directly in frequency is an important feature. The main elements of the unit are a klystron oscillator, a stabilizing monitor loop which consists of a calibrated dual-mode reference cavity, a feedback amplifier and a self-contained power supply.

SPECIFICATIONS

Frequency Coverage
Model 802-X1: 8950 – 9325 Mc
Model 802-X2: 9300 – 9650 Mc
A range of frequencies can also be supplied in the S band or above 9600 Mc in the X band.

Frequency Stability
During short time intervals: One part in 10^9
Long term drift: Less than 100 Kc from original frequency setting.

Dial Calibration
Calibrated directly in frequency – 5 Mc per division.

Power Output
5 milliwatts
Output connection – ½" x 1" waveguide.

Power Consumption
150 watts

Size
12⅜" high x 21¾" wide x 15¼" deep.
The front panel is 10½" x 19" and is designed for rack mounting.

Weight
75 lbs.

For complete information, see your LFE engineering representative or write direct.

L/E

LABORATORY for ELECTRONICS, INC. 75 PITTS STREET BOSTON 14, MASS.

PRECISION ELECTRONIC EQUIPMENT • OSCILLOSCOPES • MAGNETOMETERS • COMPUTERS • MICROWAVE OSCILLATORS • MERCURY DELAY LINES

a NEW instrument and

The NEW Type 304-A, succeeding the world-famous Type 304-H, is more than simply a new instrument — more than a new combination of established circuits. It represents a significant development in the science of instrumentation. The Type 304-A, *a true electronic voltmeter*, reflects *a new concept* of oscillography—

THE DU MONT

TYPE 304-A

The new Type 304-A is in *every* respect a true electronic voltmeter. Every feature of the well-known Type 304-H has been re-evaluated with this concept in mind. All the features that made the Type 304-H so valuable as a *qualitative* instrument have been preserved and augmented to enable not only *qualitative* analysis, but rapid, accurate *quantitative* measurement of amplitude as well.

AMPLITUDE CALIBRATION The novel amplitude calibrating system of the Type 304-A permits signal measurements directly in volts from the screen. Unlike electro-mechanical devices, the new Type 304-A is not restricted to measurement of sinusoidal signals — or to peak-to-peak readings of voltage. The Type 304-A may be used to measure any amplitude portion of the input signal, and has a sensitivity of 0.1 p-p volt full scale, or 0.025 p-p volt per inch.

NEW CATHODE-RAY TUBE A wholly new cathode-ray tube is employed in the Type 304-A. This tube, designated Type 5ADP-, was specifically designed to permit accuracy of measurement. This new flat-faced tube is precision-built to tolerances far more stringent than is the practice in conventional tubes. The angular alignment between x and y deflection systems is held to 90° ±1°, as contrasted to ±3° in conventional cathode-ray tubes. The various distortions and aberrations inherent in all cathode-ray tubes are held to a minimum. The new design of the electron gun and deflection-plate structure assures a deflection sensitivity as much as twice that of equivalent tube types, as well as a smaller spot size, with no sacrifice in brilliance. Also incorporated is an auxiliary focus control which reduces the effects of astigmatism to a minimum. Thus by the inclusion of this new tube and its auxiliary circuitry, an unusually fine, bright trace is achieved, enabling a degree of resolution—and hence a degree of accuracy—heretofore impossible in instruments employing medium accelerating potentials.

HEATER REGULATION Regulation of the heaters of the Y-input stages has been incorporated to promote stability of the amplifier.

SYNC LIMITING Sync limiting, on both recurrent and driven sweeps, assures stable operation, even for varying synchronizing levels, and freedom from horizontal jitter that might tend to interfere with precise analysis.

ILLUMINATED CALIBRATED SCALE A new edge-illuminated scale, calibrated in fifth inches, with every fifth line accentuated, is incorporated in the Type 304-A. Accentuated lines are numbered so amplitude may be read directly.

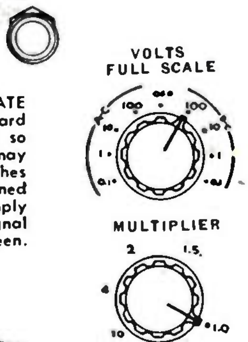
• • • • •

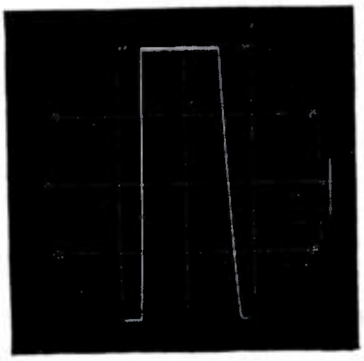
The Type 304-A represents one more step in the development — by Du Mont — of the cathode-ray oscillograph from a purely qualitative instrument to its rightful position as the most versatile, most complete analytical device available.

Domestic Price $333

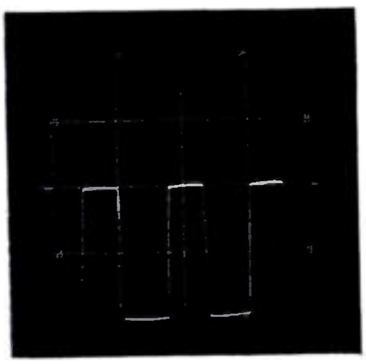
a NEW concept in oscillography!

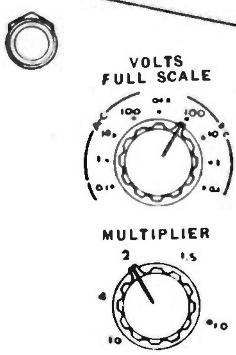
Calibrating the Type 304-A is as simple and easy as "zeroing-in" a vacuum-tube voltmeter.





TO CALIBRATE, depress the front-panel CALIBRATE button to apply the squarewave voltage standard to the screen. Adjust the MULTIPLIER control near 1 so squarewave peaks are at 0 and 100. Amplitude may now be read directly from the scale where 4 inches vertically represents 0.1, 1, 10 or 100 volts, as determined by the VOLTS FULL SCALE selector. Simply depressing the CALIBRATE button returns the signal applied to the Y-input terminals to the screen.





MULTIPLIER CONTROL permits calibration of scales to other values. For example, to calibrate for 200 volts full scale, the multiplier control is adjusted near 2 so peaks of squarewave occupy space from zero to 50. Amplitude may now be measured directly in volts simply by multiplying the settings of the MULTIPLIER control (2) by the product of the scale reading times the VOLTS FULL SCALE setting (100). Use of the MULTIPLIER control extends the range of the Type 304-A to 1000 volts of full scale. Use of precision attenuator, having 1% resistors, permits the accurate calibrating standard to be inserted in back of the attenuator without effect from the attenuator setting.

SPECIFICATIONS:

CATHODE-RAY TUBE — New Flat-Face Type 5ADP.

ACCELERATING POTENTIAL — 3000 volts.

Y-AXIS: ***Deflection Factor*** — 0.1 p-p volt full scale (equivalent to 0.025 p-p volt per inch). Direct to deflection plates. 32-39 p-p volts per inch.
Frequency Response — (at all gain and attenuator control settings) Direct Coupling: Flat at 0 to down not more than 10% at 100,000 cps. Capacitive coupling, down not more than 10% from 10 to 100,000 cps. Down not more than 50% at 300,000 cps. Provision for balanced input on 0.1 volt full-scale range.
Undistorted Deflection — More than 4 inches. Expansion equivalent to 20 inches.
Input Impedance — to amplifier (single ended) 2 megohms, 50 μμf. (Balanced) 2 megohms, 35 μμf. Direct (balanced) 3 megohms, 20 μμf. (single ended) 1.5 megohms, 20 μμf.

X-AXIS: ***Deflection Factor*** — through amplifier, 0.3 p-p volt/in. Direct 40-50 p-p volt/in.
Frequency Response — (at all settings of gain and attenuator controls) Direct coupling: Flat at 0 to down not more than 10% at 100,000 cps.; down not more than 50% at 300,000 cps. Capacitive coupling, down not more than 10% from 10 to 100,000 cps. Down not more than 50% at 300,000 cps.
Undistorted Deflection — More than 4 inches. Expansion equivalent of 30 inches.
Input Impedance — To amplifier, 2.2 megohms, 50 μμf. Direct (single ended 1.5 megohms, 20 μμf. Balanced, 3 megohms, 20 μμf.

LINEAR SWEEPS: ***Sweep Frequency*** — Recurrent and driven sweeps continuously variable in frequency from 2 to 30,000 cps. Maximum sweep writing rate, 1"/μsec. Provision for sweeps of extra-long duration, ½ sec. of sweep secured for each μf of external capacitance.
Synchronization — from signal of either polarity.
Sync Limiting — on both driven and recurrent sweeps.

VOLTAGE MEASUREMENT — Squarewave standard applied for calibration by front panel push button.
Voltage Range: VOLTS FULL SCALE, 0 to 0.1, 1, 10, 100 volts, Multiplier: x1 to x10. Overall accuracy, 5%.

INTENSITY MODULATION — 15 volts blanks beam at normal intensity settings.

CALIBRATED SCALE — Variable illumination. Numbered calibrations for Direct Amplitude measurement.

PRIMARY POWER — 115 or 230 volts, 50-400 cps. 110 w.

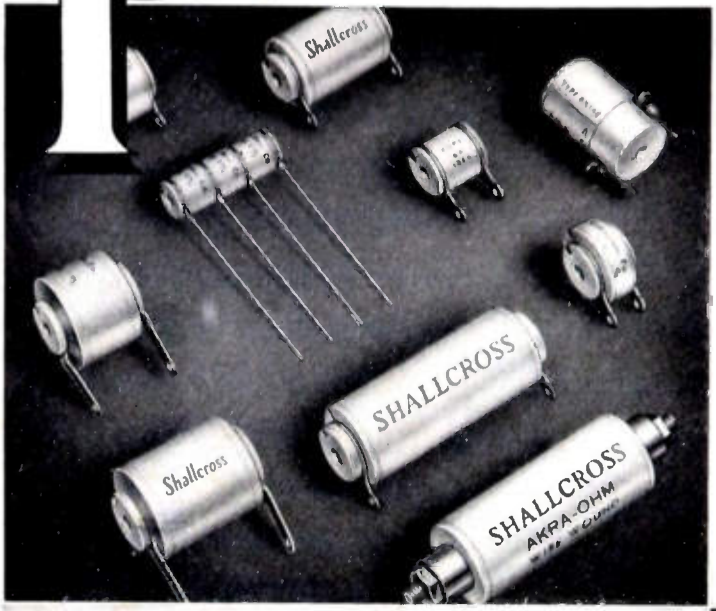
PHYSICAL CHARACTERISTICS — Metal cabinet with grey wrinkle finish. Dimensions: height 13½", width 8¾", depth 19½". Weight 50 lbs.

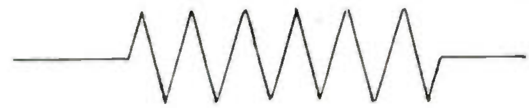
DU MONT *for Oscillography*

Write for technical bulletin A-04-A for complete details.

INSTRUMENT DIVISION ALLEN B. Du MONT LABORATORIES, INC., 1500 MAIN AVE., CLIFTON, N. J.

Precision products





◀ CLOSE TOLERANCE RESISTORS

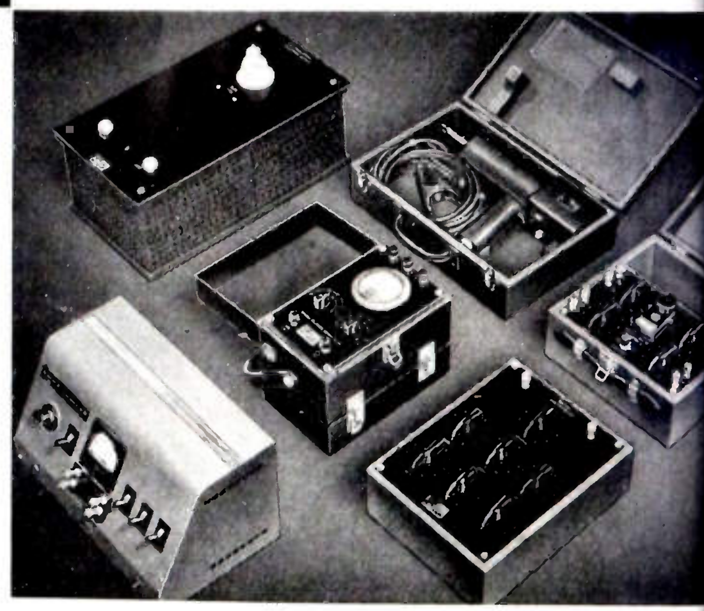
(JAN, Mil, and standard types)

Wire-wound precision resistors have unique characteristics suitable for exacting modern circuits. Shallcross Akra-Ohm resistors meet these requirements and are available in many types, sizes, shapes, and mounting styles. They are noted for high stability, low temperature coefficients, low noise levels, uniformity, long life, and extreme accuracy in matched pairs and sets. Ask for Bulletins R3-C, L-27.

PRECISE ELECTRICAL ▶ MEASURING INSTRUMENTS

Resistance Standards	Megohm Bridges
Decade Potentiometers	Tone Generators
Decade Resistance Boxes	Telephone Test Equipment
Wheatstone Bridges	Low-Resistance Test Sets
Kelvin-Wheatstone Bridges	Galvanometers
Limit Bridges	Ayrton Shunts

Write for Catalog No. 10

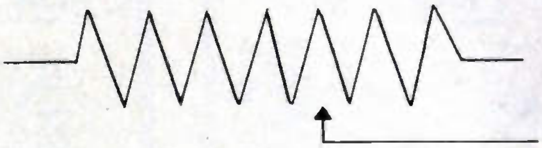


INDUSTRIAL RESEARCH AND DEVELOPMENT SERVICE

Today's complex circuits frequently require the design development, and production of highly specialized components, sub-assemblies, or instruments which fall outside the realm of standard engineering or production facilities. The Shallcross Research Department has been specifically formed to handle such assignments. Composed of electronic, electrical, instrument, mechanical, and chemical engineers of broad experience and backed with adequate modern facilities, this unique service group combines a highly technical as well as an intensely practical engineering-production viewpoint. We invite you to submit your requirements for review and recommendation.

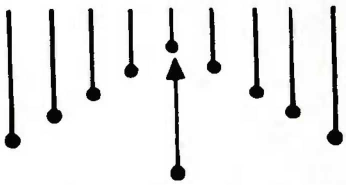
SHALLCROSS MANUFACTURING

by SHALLCROSS



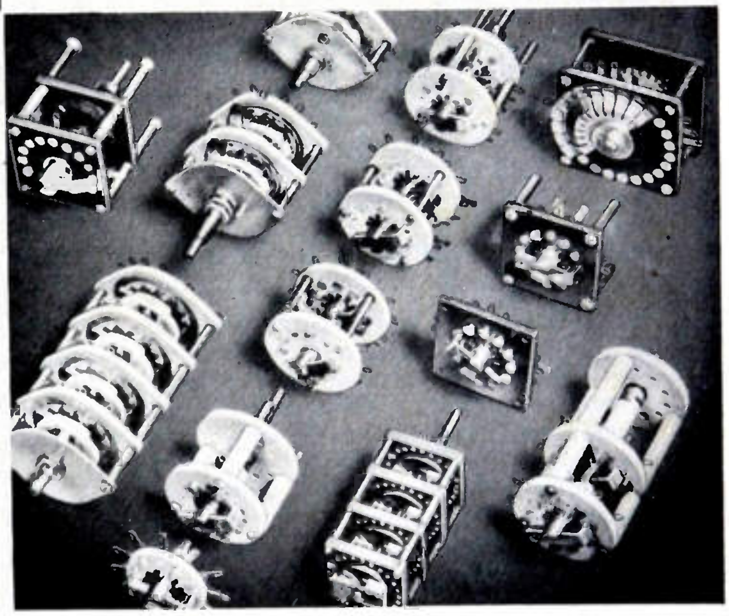
HIGH QUALITY ATTENUATORS

Improved materials and production techniques for Shallcross Attenuators have resulted in a line that sets new higher standards of attenuation performance for practically every audio and communications use. Shallcross Audio Engineering Bulletin No. 4 will be sent on request.



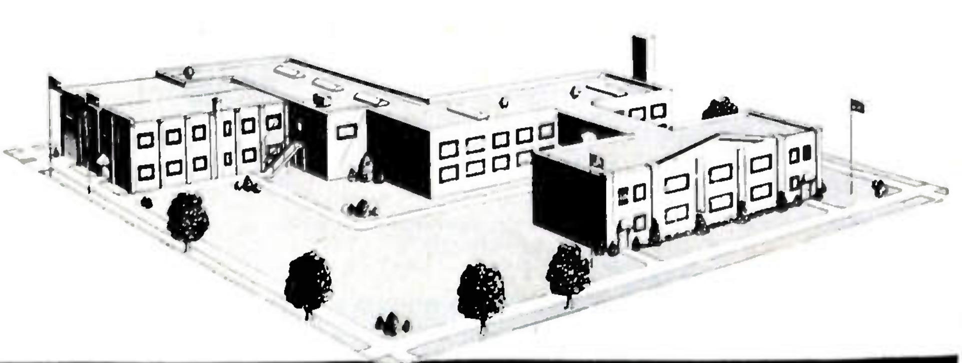
CUSTOM-BUILT SELECTOR SWITCHES

Shallcross builds single or multiple deck selector switches having up to 180 positions. Test units have given satisfactory performance at 250 volts 10 amperes and at 2500 volts 1 ampere A.C. Contact resistance ranges from a low of 0.0005 ohms to a maximum of 0.005 ohms depending upon the size and material of the contact surfaces. You are invited to outline your requirements on Shallcross Specification Sheet No. 6.



HIGH-VOLTAGE
Test and Measuring Equipment

Shallcross kilovoltmeters, kilovoltmeter multipliers, and corona protected resistors provide maximum accuracy, safety and dependability for high-voltage measurement. They are widely used in standards laboratories, and with X-ray equipment, precipitrons, electrostatic generators, and other high-voltage sources. Write for High-Voltage Bulletin L-7.



COMPANY • Collingdale, Pennsylvania

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 36A)

(TC-1-R-12-1), it may be used by scheduled airline, commercial and private pilots and radio operators, as a replacement for the old type headphones.

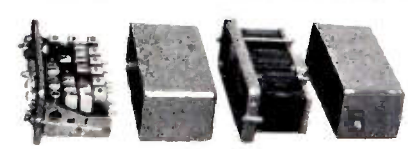
The "AIRPHONE" consists of a specially designed miniature earphone similar to those used in hearing aids, but more rugged in design. The earphone is attached to a supporting, individually fitted earpiece by means of a snap fastener. Component parts include a personal volume control with lapel clip, and a PL-55 plug.

The complete assembly weighs only 2½ oz., but is 12 times more sensitive than conventional headphones. It is used with only one ear, leaving the other clear for cockpit conversation. Price for complete unit, including postage and carry case: $20.98.

Portable-Pack and Semi-Fixed Main Station FM Communications Unit

Mogens Bang & Co., 125 Gates Ave., Montclair, N. J., announces the availability of new FM equipment manufactured by Storno, Copenhagen, Denmark.

The Pack-Set (Type BQP 45) comprises a broadband FM-transmitter, intended for modulation by a high-quality dynamic microphone, and a narrowband FM-receiver for reception of instructions. The set may be carried by hand, over the shoulder, or on the back. It is divided in two parts, clamped together. The upper case contains the transmitter receiver, and



the lower the power supply. The complete unit is water-proof to ensure operation under any climatic condition. Transmitter and receiver are working in full duplex on a common antenna. The front panel is furnished with sockets for microphone, headphone and antenna, a modulation control, a bass cut-off control and a squelch control. Transmitter and receiver are automatically switched-on by inserting the associated plugs. Operation Range: 1 to 25 miles.

(Continued on page 50A)

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.

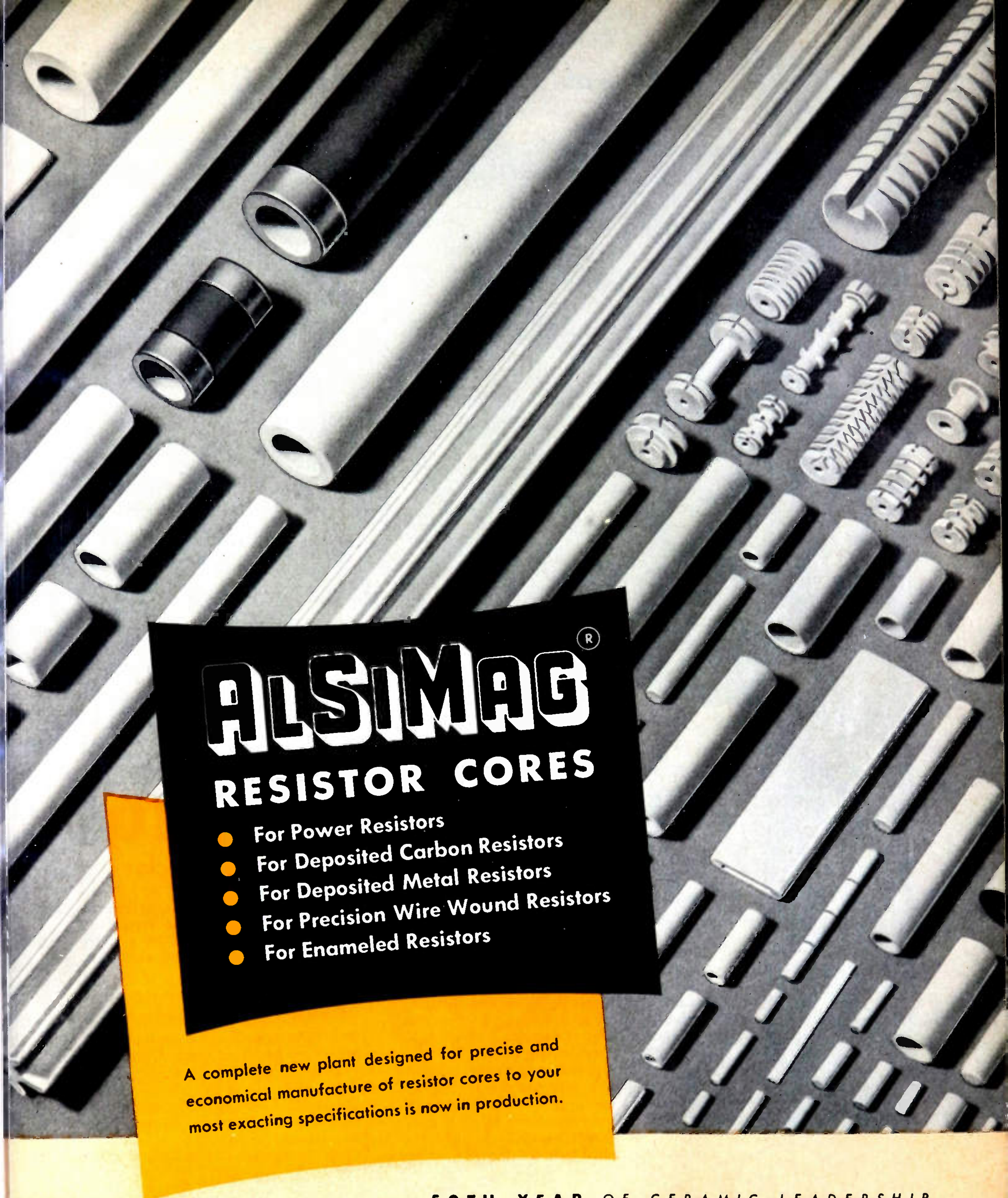
*** can be used with higher source impedances, with corresponding reduction in frequency range. With 200 ohm source, secondary impedance becomes 250,000 ohms . . . loaded response is –4 db. at 300 cycles.

****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"



ALSIMAG®
RESISTOR CORES
For Power Resistors
For Deposited Carbon Resistors
For Deposited Metal Resistors
For Precision Wire Wound Resistors
For Enameled Resistors
A complete new plant designed for precise and economical manufacture of resistor cores to your most exacting specifications is now in production.

INTERNATIONAL RECTIFIER CORPORATION

EL SEGUNDO
CALIFORNIA

HERMETICALLY SEALED

Diameter .... 3/16" to 1-1/4"
Length .... 9/16" to 10"
Current: half-wave .. 1.5 ma to 60 ma
Voltage: DC output ...... 20 volts to 4,000 volts

PHENOLIC CARTRIDGE

Diameter .... 1/8" to 1"
Length .... 1/2" to 12"
Current: half-wave .. 1.5 ma to 60 ma
Voltage: DC output .... 20 volts to 10,000 volts

SELENIUM DIODES

Diameter .... 0.100" to 0.300"
Length .... 0.210" to 0.250"
Output Voltage .... 20V to 80V
Output Current ... 250 ua to 1.5 ma
Temperature Range -50°C to 100°C

Selenium Rectifiers

A recent month's production included Rectifiers to supply 40 microamperes, 1,000 volts, and Rectifiers with a capacity of 140,000 amperes, 14 volts.

POWER STACK
30 Kw DC Power
Considered to be the largest single selenium rectifier stack produced.

Owned and managed by Engineers who are specialists in the design and manufacture of Selenium Rectifiers. Submit your problems for analysis and we will be glad to offer our recommendations.

INTERNATIONAL RECTIFIER CORPORATION

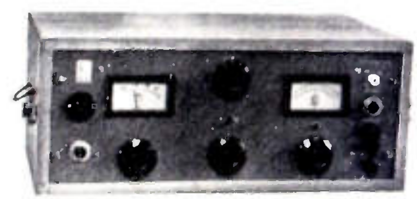
General Offices: 1521 E. Grand Ave., El Segundo, Calif. • Phone: El Segundo 1890
Chicago Branch Office: 205 West Wacker Drive • Phone: Franklin 2-3889

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)

The FM-Receiver (Type RX46) is intended for reception of the signals from the pack-sets or the relay transmitter. It is continuously tunable over the entire band 88–108 Mc. It has built-in power supplies for both ac mains and 12 Volt dc. Either can be selected from a front panel equipped with an AF-level control, a meter with associated switch, terminals for connection to a telephone line, and sockets for power,



antenna and headphone. The level control allows audio output from +12 db to −12 db in steps of 3 db and from −12 to −44 db with increasing intervals. Zero db corresponds to 2.6 volts across 600 Ohms equal to 100 per cent modulation (0.775 volts equals 30 per cent modulation). The meter arrangement allows control of all important tensions and currents.

An automatic frequency-correcting network is employed, eliminating the normal temperature drift and thus keeping the receiver tuned-in on the transmitter frequency. The receiver is intended for 19 inch rack-mounting. For portable use it can, however, be delivered in a mahogany cabinet.

The Talk-Back Transmitter Type TX46-1 is used for giving orders and instructions to the persons carrying the pack-sets. It is sufficiently powerful to maintain the contact outside the range of the pack-transmitter in order to secure that the instructions are carried through under any circumstances. Four-spot frequencies are available allowing individual contact with four groups of pack-sets without disturbing the others. The transmitter is a 10 watt phase-modulated FM-transmitter, intended for operation on four crystal controlled frequencies within 300 kc in the band 88–102 Mc. Built-in power supplies allow operation on either 220-volt ac mains or 12 volt dc. The transmitter may be modulated either from a carbon microphone or from a telephone line with usual line level. The front panel is equipped with sockets for power, microphone and antenna, as well as terminals for telephone line and ground. Selector switches are provided for power and the four-spot frequencies. A meter with associated switch allows control of all stages as well as the antenna current. A special tuning knob for the power amplifier is provided, since tuning is dependent on the antenna conditions. The transmitter is intended for 19-inch rack mounting, but may also be delivered in a mahogany cabinet.

(Continued on page 60A)

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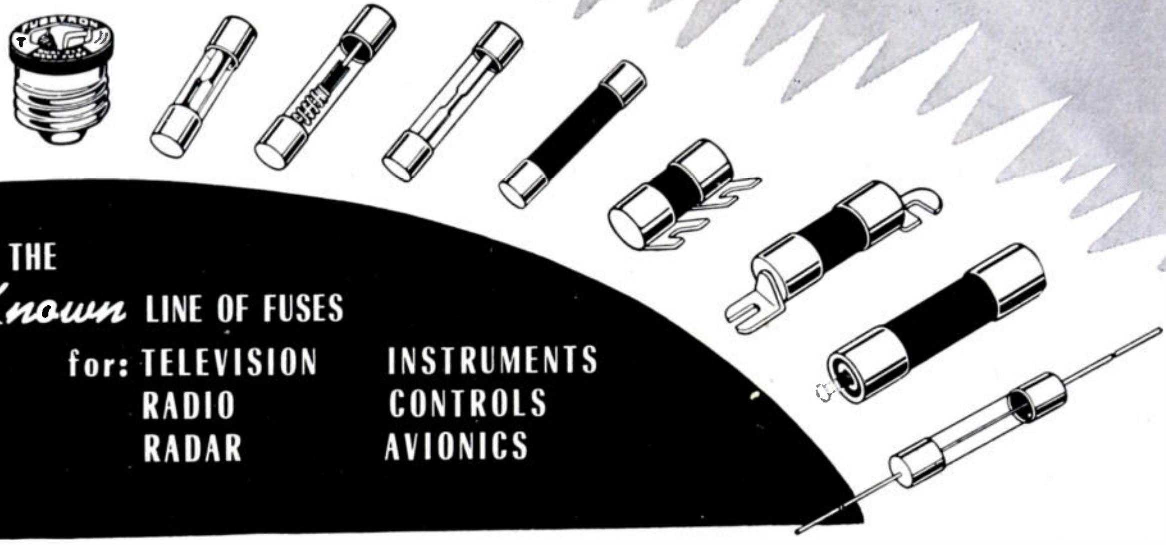

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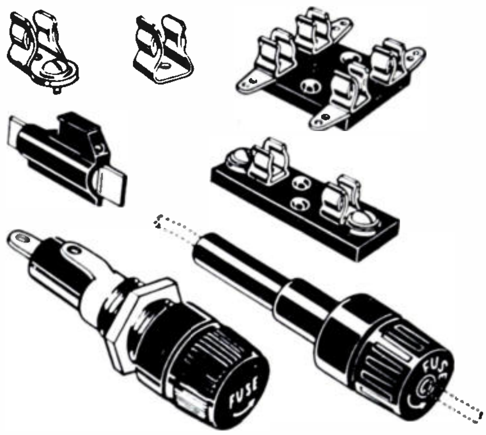

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University at Jefferson, St. Louis 7, Mo.
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City & Zone.......................State....................... 952

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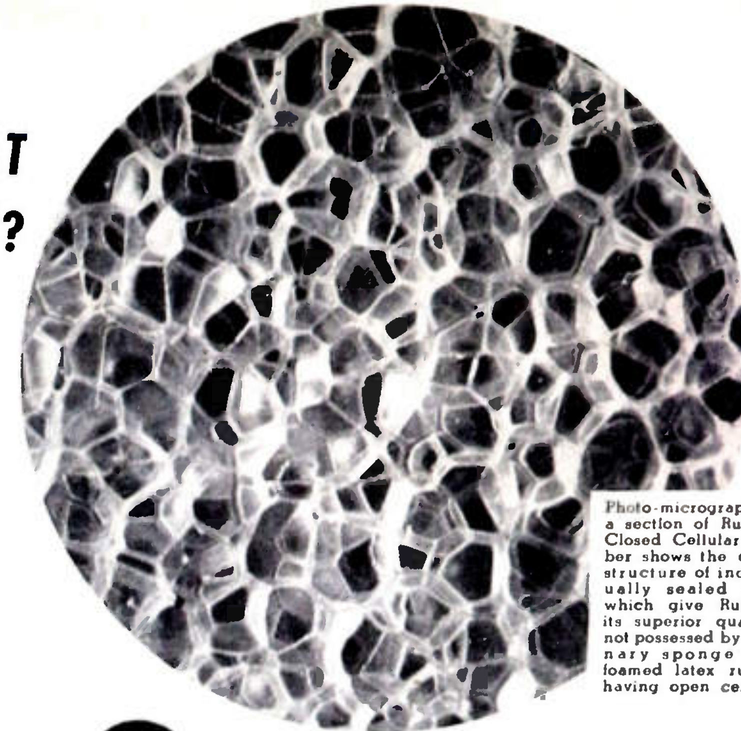

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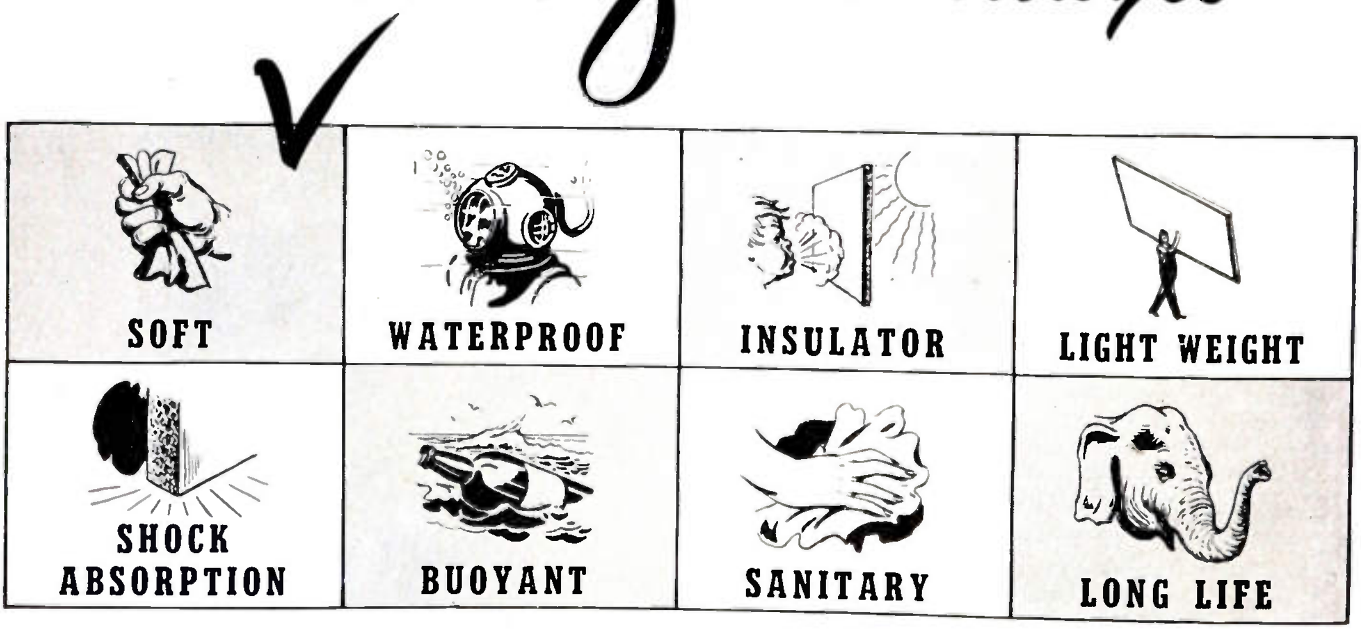

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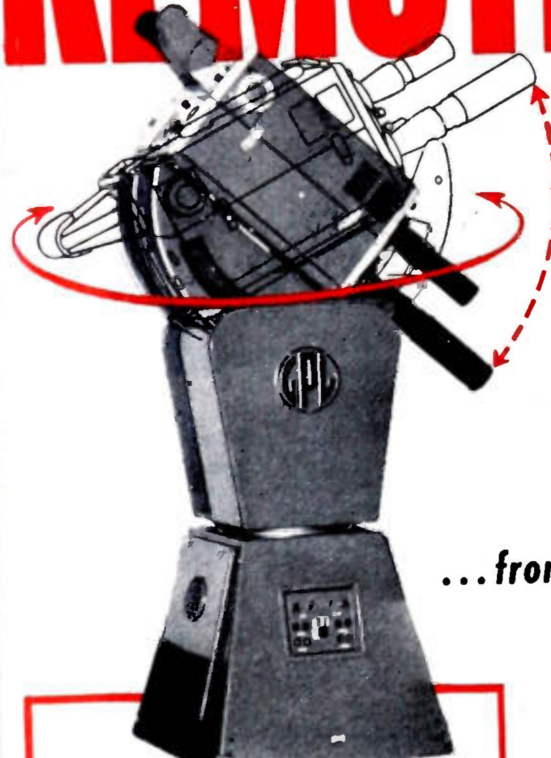

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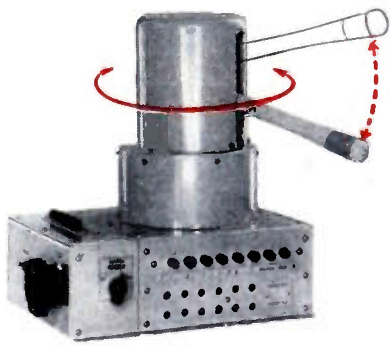

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You are now assured immediate delivery of the Microtorque* Potentiometer. As a new service to customers, a complete stock of resistance values as listed, is maintained to assure immediate delivery for prototypes, experimental work or emergency production. The Microtorque* is the solution where remote indicating, low torque (.003 oz. in.), jewel bearings and instrument quality are required.

SPECIFICATIONS

LINEARITY: ± 0.5% of total resistance.

MAXIMUM OPERATING SPEED: 100 rpm.

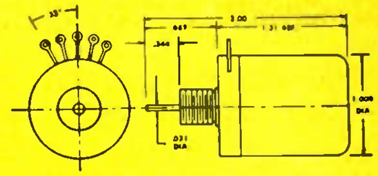
ACCELERATION: Will withstand 50G steady state acceleration in best axis.

VIBRATION: Will withstand 0.06" double amplitude sinusoidal vibration from 10 to 55 cps in best axis.

AMBIENT TEMPERATURE: Will function mechanically from −54° C. to + 71° C.

MOMENT OF INERTIA: 2×10^{-4} oz-in.^{2} (approx.)

TEMPERATURE COEFFICIENT OF RESISTANCE: .0006/° C. Max.



Following Microtorques* are available from stock in quantities of six or less:

RES. OHMS	STARTING TORQUE IN-OZ.	TURNS OF WIRE TYPE 2	TURNS OF WIRE TYPE 9	CURRENT** MA.	PRICE***
250	.006	350	450	57	$45.00
1,000	.004	500	650	28	$40.00
2,000	.004	700	750	20	$40.00
5,000	.003	900	1200	14	$40.00
10,000	.003	1,000	1300	10	$40.00
25,000	.003	1,000	1300	7	$45.00

**Must be de-rated for ambient temperature over 60° C

***Prices apply to quantities of six or less. For quotation on larger quantities or specialtypes, please write

Above Microtorques* are available in the following two types
Type 2: 270° ± 10° Electrical Rotation, Mechanical Rotation Limited by internal stops

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Brush does not short ends of coil

Giannini also produces potentiometers of various types, including non-linear functions, and tapped windings

Other Giannini Potentiometers that are available on special order; soon to be stocked.

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Linear and functional outputs.
Ball bearings; 1/4" shaft,
0.1 oz. in. torque.
500 to 100,000 ohms.
1.125" diameter x 1.16" long.

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Straight-line motion along axis.
Linear or functional outputs.
200 to 60,000 ohms.
5 sizes, 1" diameter from
2.33 to 6.54" long.

Syncromount JUNIOR

Linear and functional outputs.
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Foremost manufacturer of toroidally-wound potentiometers. Where linearity, stability, rigidity, power dissipation and precision are required, toroid windings are outstanding performers.

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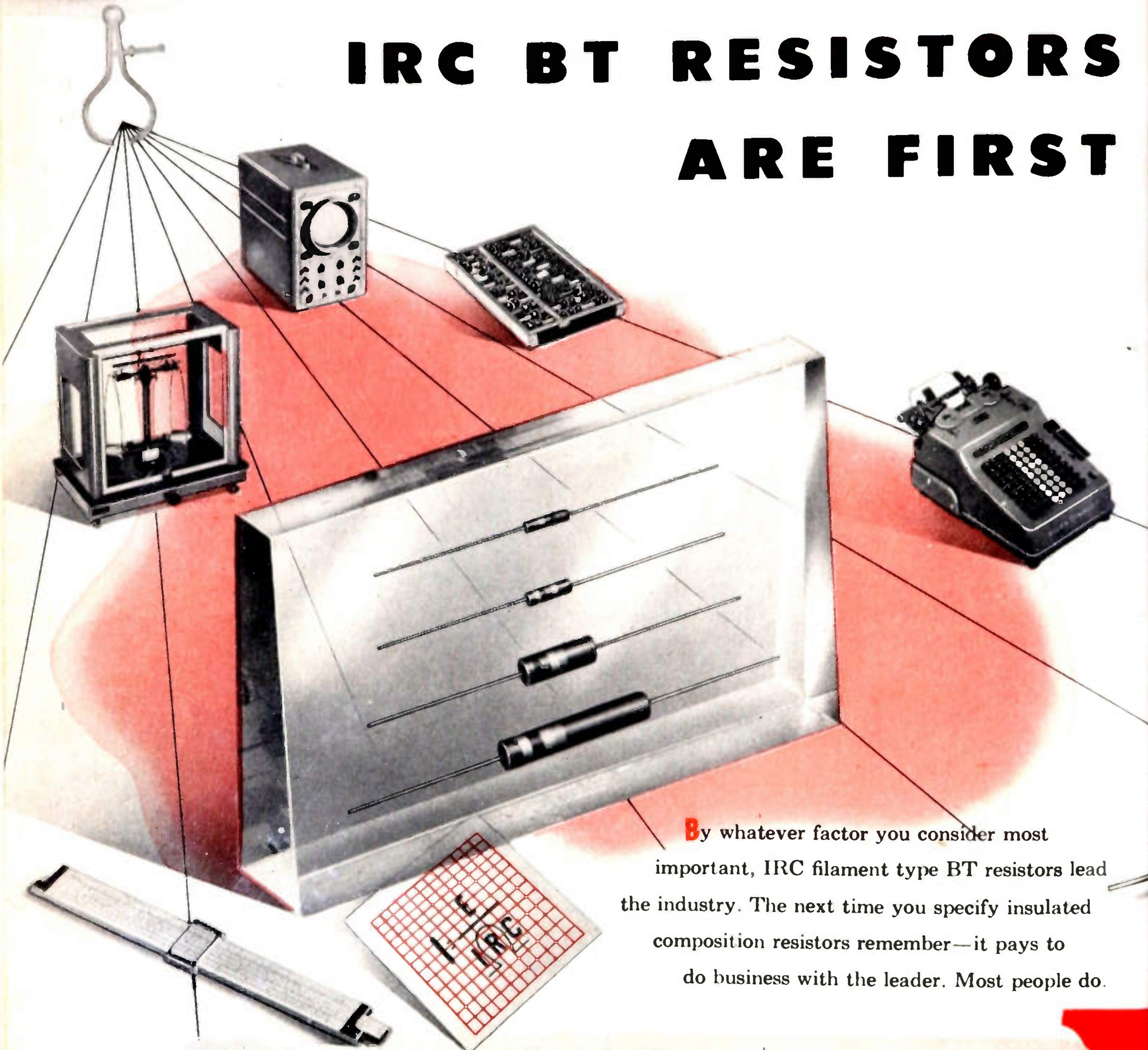
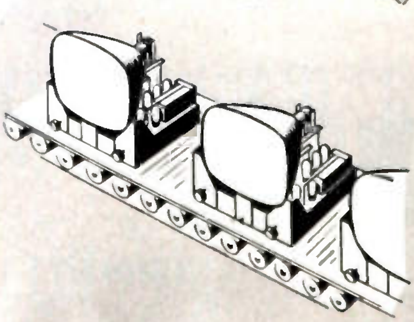
However you compute

IRC BT RESISTORS ARE FIRST



By whatever factor you consider most important, IRC filament type BT resistors lead the industry. The next time you specify insulated composition resistors remember—it pays to do business with the leader. Most people do.



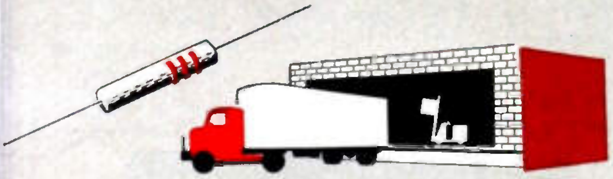
IF QUANTITY PRODUCTION INDICATES LEADERSHIP— ***remember*** more IRC BT resistors are used in radio and TV sets than any other brand. During the last five years IRC supplied 40% of the resistors used in radio and TV set production.

IF QUALITY STANDARDS DENOTE LEADERSHIP— ***remember*** IRC Advanced Type BT resistors meet and beat rigid JAN-R-11 specifications. Nearly all producers of government equipment have tested and approved IRC's advanced BT resistor.

leadership

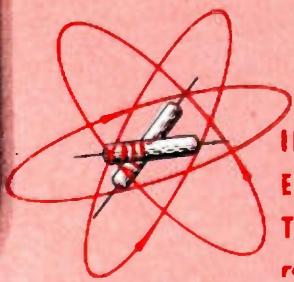
IF GLOBAL ACCEPTANCE REFLECTS LEADERSHIP—

remember IRC filament type BT resistors are favored in every major market of the world. Licensee plants in Canada, England, Denmark, Belgium, Italy and Australia produce IRC BT's for international electronics.



IF DEPENDABLE DELIVERY REPRESENTS LEADERSHIP—

remember IRC's long record of strike-free labor relations protects your assembly lines. Dependable delivery of Advanced Type BT resistors is further assured by IRC's financial ability to maintain large stocks of wanted ranges and to draw from foreign licensees when demand warrants it.



IF RESEARCH AND ENGINEERING TESTIFY TO LEADERSHIP—

remember IRC BT resistors are produced by the largest resistor manufacturer in the world. The finest accumulation of resistance know-how has been pooled in the perfection of these filament type resistors.

Smallest insulated resistor available anywhere

This coupon brings you full data on IRC BT Resistors

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Please send me full data on IRC Advanced Type BT Resistors:—

☐ Also name and address of nearest IRC Distributor who can furnish speedy delivery of BT resistors in small quantities.

NAME ____________________

TITLE ____________________

COMPANY ____________________

ADDRESS ____________________

CITY ____________________ ZONE ______ STATE ______

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 50A)

Range Calibrator

Tel-Instrument Co., 50 Paterson Ave., East Rutherford, N. J., announces the availability of a new range calibrator (Type 2010), specifically designed for production testing and calibrating of radar receivers. This instrument embodies several very basic circuit features which obsolete the TS-102/AP range calibrator for applications requiring accurate pulse repetition rates and absolute freedom from marker jitter.

Type 2010 offers: crystal-controlled pulse rates of 200 to 2,000 pps, derived from a 327.80 kc marker crystal by the use of a binary divider system, "Master"-"Slave" operation permitting the use of a number of calibrators for multi-position test systems, narrow and stable 500 yard marker pulses, and a completely regulated power supply for operation from 105–125 v, 100 w 50–60 cps.

Motor Starting Capacitors

A new line of motor starting capacitors designed to operate under adverse conditions has been announced by the **Illinois Condenser Co.**, 1616 N. Throop St., Chicago 22, Ill.

Built-in features of these capacitors include low power factor and wide temperature range. Shock proof construction assures long life under severe vibration.

Connecting terminals and foil are of heavy construction and corrosion proofed.

Every unit is fully guaranteed for a period of one year.

These electrolytic motor starting capacitors for 50–60 cps operation are supplied in round aluminum containers with insulating sleeve.

The capacitors are available in a wide range of capacities in voltage rating from 24 to 220 vac.

Complete details may be secured by writing the manufacturer.

(Continued on page 76A)

electronic wire and cables for standard and special applications

Whether your particular requirements are for standard or special application, choose *LENZ* for the *finest* in precision-manufactured electronic wire and cable.

GOVERNMENT PURPOSE RADIO AND INSTRUMENT HOOK-UP WIRE,
plastic or braided type, conforming to Government Specification JAN-C-76, etc., for radio and instruments. Solid or flexible conductors, in a variety of sizes and colors.

SPECIAL HARNESSES,
cords and cables, conforming to Government and civilian requirements.

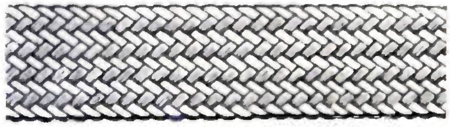
RADIO AND INSTRUMENT HOOK-UP WIRE,
Underwriters Approved, for 80° C., 90° C. and 105° C. temperature requirements. Plastic insulated, with or without braids.

SHIELDED JACKETED MICROPHONE CABLE
Conductors: Multiple—2 to 7 or more conductors of stranded tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. Tough, durable jacket overall.

RF CIRCUIT HOOK-UP AND LEAD WIRE
for VHF and UHF, AM, FM and TV high frequency circuits. LENZ Low-Loss RF wire, solid or stranded tinned copper conductors, braided, with color-coded insulation, waxed impregnation.

JACKETED MICROPHONE CABLE
Conductors: Extra-flexible tinned copper. Polythene insulation. Shield: #36 tinned copper, closely braided, with tough durable jacket overall. Capacity per foot: 29MMF.

SHIELDED MULTIPLE CONDUCTOR CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Closely braided tinned copper shield. For: Auto radio, indoor PA systems and sound recording equipment.



TINNED COPPER SHIELDING AND BONDING BRAIDS
Construction: #34 tinned copper braid, flattened to various widths. Bonding Braids conforming to Federal Spec. QQ-B-S75 or Air Force Spec. 94-40229.

SHIELDED COTTON BRAIDED CABLES
Conductors: Multiple—2 to 7 or more of flexible tinned copper. Insulation: extruded color-coded plastic. Cable concentrically formed. Closely braided tinned copper shield plus brown overall cotton braid.

PA AND INTERCOMMUNICATION CABLE
Conductors: #22 stranded tinned copper. Insulation: textile or plastic insulated conductors. Cable formed of Twisted Pairs, color-coded. Cotton braid or plastic jacket overall. Furnished in 2, 5, 7, 13 and 25 paired, or to specific requirements.

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A NEW WAY TO GET GREATER CIRCUIT RELIABILITY!

Now, when designing equipment, you can freely specify 5-Star Tubes knowing they will be available in quantities when you need them. Greatly expanded G-E output offers you . . . for the first time . . . an assured supply of these famous types that are *designed and built for highest reliability.*

Take advantage of 5-Star availability, to develop new electronic circuits that excel in their dependable performance . . . in freedom from tube replacements . . . in lower maintenance needs.

Gain the benefits of

- Buyer preference because your equipment is more dependable.
- Lower designing costs! 5-Star Tubes come to you uniformly predictable in performance.
- Lower manufacturing costs in your plant! Fewer rejects from tube causes mean fewer units to be reworked.
- Lower warranty-servicing costs on your equipment in users' hands.

Prompt study of G-E 5-Star advantages will strengthen your competitive position and point the way to important savings. Ask for the facts . . . by return mail, or visit from a G-E tube engineer! *General Electric Company, Tube Department, Schenectady 5, N. Y.*

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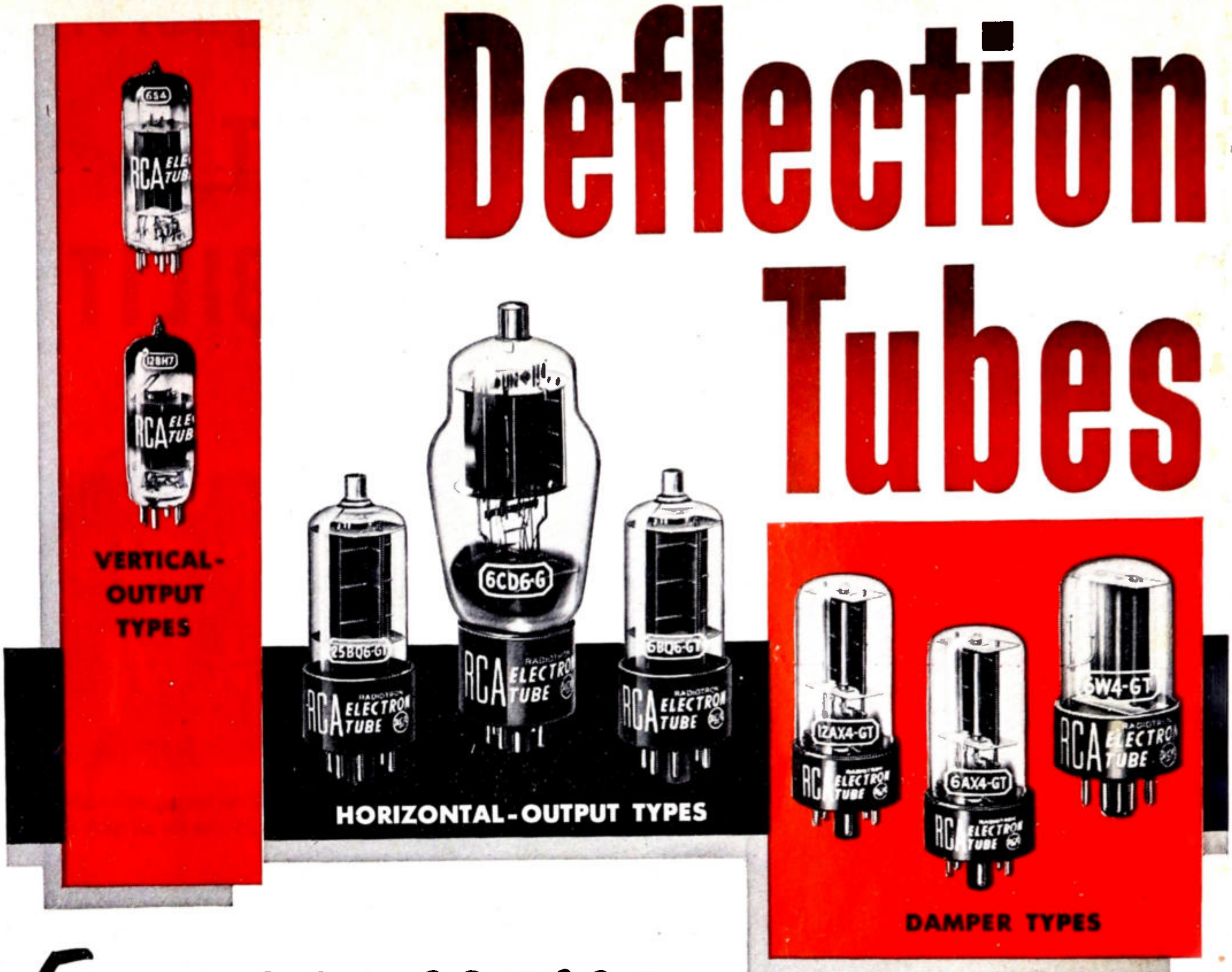
2C51	GL-5670— h-f medium-mu twin triode.
2D21	GL-5727— thyratron.
5Y3-GT	GL-6087— full-wave rectifier.
6AK5	GL-5654— sharp-cutoff r-f pentode.
6AL5	GL-5726—twin diode.
6AQ5	GL-6005—beam power amplifier.
6AS6	GL-5725—dual-control sharp-cutoff r-f pentode.
6AU6	GL-6136—sharp-cutoff pentode.
6BA6	GL-5749—remote-cutoff r-f pentode.
6BE6	GL-5750—pentagrid converter.
6C4	GL-6135—medium-mu triode.
6SK7	GL-6137—remote-cutoff r-f pentode.
12AT7	GL-6201—high-Gm medium-mu twin triode.
12AU7	GL-5814—medium-mu twin triode.
12AX7	GL-5751—high-mu twin triode.
12AY7	GL-6072—low-noise medium-mu twin triode.
......	GL-5686—beam power amplifier.

Booklet ETD-548 contains a cross-reference table of ratings and characteristics for application use when substituting 5-Star Tubes for standard types. Wire or write for it!

GENERAL ELECTRIC

163-1A3

Deflection Tubes



From one source...

8 types to fill your deflection-system needs

RCA now offers these eight volume-type, deflection-system tubes having performance range to take care of most circuit requirements efficiently and economically.

Vertical-Output Tubes: The RCA-6S4 is a miniature medium-mu triode for circuits operating from the boosted B-voltage supply and is capable of supplying ample deflection for kinescopes having diagonal-deflection angles up to 70 degrees and operating at voltages up to 18 kilovolts. The RCA-12BH7 is a miniature twin triode used in similar circuits having more modest deflection-voltage requirements and permits the economy of using one triode unit as the vertical oscillator.

Horizontal-Output Tubes: The RCA-6BQ6-GT beam power tube is capable of fulfilling most requirements for horizontal deflection and high voltage in receivers having B voltages in the order of 230-300 volts. RCA-25BQ6-GT, the same as the 6BQ6-GT except for its 25-volt, 0.3-ampere heater, is for use in receivers having series heater strings. The huskier RCA-6CD6-G has the reserve power desired in deluxe receivers operating at high voltages up to 18 kilovolts.

Damper Tubes: The RCA-6W4-GT has the high perveance needed for efficient operation and good linearity in horizontal-deflection circuits. The RCA-6AX4-GT features higher heater-cathode voltage ratings so that a separate heater supply is not required. RCA-12AX4-GT, identical with the 6AX4-GT except for its 12.6-volt, 0.6-ampere heater, is for use in receivers having series heater strings.

Application Notes: Application Notes covering "Design Considerations for Minimizing Ripple and Interference Effects in Horizontal-Deflection Circuits," and "Horizontal-Deflection-Output and High-Voltage Transformer RCA-230T1 for 18-Kilovolt Kinescope Operation" are yours for the asking. For your copies —and data on these RCA tubes for deflection systems—write RCA, Commercial Engineering, Section IR47, Harrison, N.J.

For additional information on using these RCA tubes in your circuits contact the nearest RCA Field Office.

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

RADIO CORPORATION of AMERICA TMK. ®

ELECTRON TUBES — HARRISON, N. J.

PROCEEDINGS OF THE I.R.E.®

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 40 — *September, 1952* — NUMBER 9

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

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INSTITUTE NEWS AND RADIO NOTES SECTION

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

Chairmen of New IRE Professional Groups

G. T. ROYDEN

L. H. MONTGOMERY

Communications Group

G. T. Royden was born on June 20, 1895, in Fort Clark, Tex. He received the B.A. and engineering degrees in 1917 and 1924, at Stanford University, Calif.

From 1919 to 1925, Mr. Royden was associated with the Navy at the Mare Island Navy Yard, Calif., where his assignments included the development of frequency measurement techniques. In 1925, he joined the Federal Telegraph Company to develop a radio receiver for operation on alternating current, suitable for reception of broadcast stations.

When the Mackay Radio and Telegraph Company was organized to operate Federal's radiotelegraph station in 1927, Mr. Royden became division engineer at the San Francisco office. In 1936 he returned to Federal in Newark, N. J., and in 1946, he transferred to the engineering department of the Mackay Company in New York, where he has remained.

A number of patents have been issued in Mr. Royden's name in the fields of radio receivers, direction finders, antennas, modulators, and quartz-crystal-controlled oscillators.

Mr. Royden joined the Institute in 1919 as an Associate, transferred to Member in 1927, and was elected Fellow of the IRE in 1933. He was Chairman of the IRE San Francisco Section in 1933 to 1934, and has served on numerous IRE Committees and the IRE Board of Directors.

Mr. Royden is a member of the American Institute of Electrical Engineers and Sigma Xi.

Medical Electronics Group

L. H. Montgomery was born on January 18, 1907, in Nashville, Tenn. He entered the radio field by joining the technical staff of Station WSM, Nashville, in 1925, remaining with that station up to the present.

In 1929, Mr. Montgomery was instrumental in the design and building of sound equipment for the Vanderbilt University stadium. In 1937 to 1938, he assisted in the research and manufacture of degenerative feedback amplifiers for RCA, and in 1939 and 1941, Mr. Montgomery participated in the design and building of several broadcast stations. As a radio class instructor at Vanderbilt in 1942 to 1943, he later became an electronics consultant at the medical school there. In this work he designed and constructed special electronic equipment for medical research at the school and Vanderbilt Hospital. At present, he is a research assistant working in the field of neurology in the anatomy department at the medical school.

In addition to his other duties, Mr. Montgomery is vice president in charge of the electronics manufacturing department of the Metal Products Company in Nashville, has been consulting engineer for a number of stations in the Southeastern states, and is chief facilities design engineer for WSM and WSM-TV.

Mr. Montgomery was associated with the Institute in 1927 and became a Senior Member in 1950. He is a member of the American Institute of Electrical Engineers committee on electrical techniques in medicine and biology, and Sigma Xi.

The Engineer in World Affairs

ELLERY W. STONE

In this world, the attaining of satisfactory results depends upon at least two basic factors. One of these factors involves the implements and supplies needed to produce the results. The other factor includes the purpose and skill of the user of the available tools and materials.

Engineers have skilfully and commendably developed the first of these essential elements. But in many and important instances, mankind has so far deplorably failed to provide either the moral purpose or the intelligent application of available instrumentalities.

The resulting problems are clearly set forth in the following guest editorial from a Fellow of the IRE, who is President of the American Cable and Radio Corporation, and a Rear Admiral of the United States Naval Reserve.—*The Editor.*

For the last three centuries, engineers have probably had greater influence than any other group of men, not excluding politicians; yet, the average engineer seldom considers the broad aspects of his work—or if he does, he tends to misunderstand them.

The field of communications will illustrate my point. After Samuel F. B. Morse perfected his telegraph system in 1844, the westward expansion of the United States went hand in hand with telegraphy. The world's most extensive international network was an important factor in building the world's greatest empire—the British. The economic development of our Latin American neighbors was immensely stimulated by the southward extension of the United States cable service, which gave to many countries their first telegraphic contact with the outside world. The growth of telegraphic communication helped to reduce the price of service for all. For example, between the United States and Argentina, the price per word went from seven dollars and fifty cents, in 1891, to twenty-seven cents in 1952.

But communications have done more than help build nations, empires, and trade; they have deeply influenced the social, political, and military organization of the world around us, and have shaped our lives as individuals, as well. For if there is any one trend in the world more important than another today, it is the rise of centralized authority; a thing wholly impossible in its present scope without modern communication networks.

Our dwindling importance as individuals might be accepted more gracefully if there were any reason to believe that the world is becoming a better and more peaceful place in which to live. Unfortunately this is not the case. We know from painful experience that cable and radio circuits and broadcasting stations, while they physically bind the world together, do not necessarily contribute to understanding even among free nations, and still less between the free and Soviet portions of the world.

It used to be fashionable to say that if the heads of governments could just sit down and talk with one another, there would never be another war. But in our time we have seen international relations plumb new depths while the efficiency of international communications achieved new heights.

This points up the fact that communication facilities, like any other engineering product, are merely tools of men; the benefits to be derived from their use depend on how we use them. In the last analysis, the fate of man rests on men—not on material considerations or facilities.

And so it is that we as engineers can never be content with our achievements merely as engineers, or rely on them to save the world in spite of human frailty. We are fortunate that, living in a democracy, we count politically as well as scientifically, and can help determine how our creations shall be used. It is up to us to make the most of this priceless opportunity, lest ruthless men enslave us with the help of our own inventions.

The IRE Professional Group System*

W. R. G. BAKER,† FELLOW, IRE

IN PROFESSIONAL ORGANIZATIONS, trade associations, and even in industry, we hear the term "splinter group" or "splinter organization." Generally, what is meant is that a small group of an old and established organization splits off and sets up its own operation.

Such action is, to a large extent, generated by the attitude of the older organization toward change. The old or parent organization may fail to recognize the birth of a new idea, a new profession, a new business, and in some instances, a new horizon.

Business, and especially big business, has recognized this phenomenon of birth and growth. It has applied several curatives, or perhaps they should be termed "organizational opportunities," one of which is called decentralization.

There is nothing really new in decentralization. It is simply the wrapping up in one package, like-things, products, functions, and perhaps even men and philosophies.

Nature herself practices decentralization in perhaps the most extreme form. The idea is that when something has been nourished or developed to a point where it should stand on its own feet, then it should do so.

There are many different aspects of decentralization. With human beings, the parental care may continue through the life of the parents although the child may, to all intents and purposes, make his own way. In the animal kingdom we have the example of the survival of the fittest.

Decentralization as applied by industry merits a further examination. In this instance, a decentralized unit deals with items that are related in engineering, manufacturing, and distribution. That is, industry would not decentralize a unit including both diesel engines and receiving tubes.

A decentralized unit in industry must conform with the basic over-all policies established by the parent organization. Generally such policies are few in number and are broad in nature, i.e., fiscal, labor, community relations, legal, and the like. Most frequently, such policies are to be used as a guide, and are an indication of what has been found desirable in the past.

It is probably safe to say that American industry would not have reached its present heights if the principal of decentralization had not been employed quite generally. Decentralization in industry not only permits grouping like-products, but permits men mutually interested in such products to work out the destination of the business in which they are interested. More important, it allows the parent organization to assign the authority and fix the responsibility, and permits the men concerned to accept the accountability.

By now I suppose you are wondering what all this, in general, and decentralization, in particular, has to do with the Professional Group System.

Let us first look at our Institute. Originally, when radio was wireless and the membership was very small, there was a mutuality of interest, which I suppose one could say was point-to-point communication based on telegraphy.

As our industry developed, and wireless became radio and radio turned into electronics, then what happened to the Institute? The Institute grew to about 30,000 members, scattered all over the world. The mutuality of interest based on one particular application was replaced by thousands of products and many, many different types of services.

This, then, is the almost ideal type of climate to develop splinters: a large parent body, great diversity of interests within its membership, and the desire of men interested in the same product, system, or application to work together in the particular section of the electronics field with which they are mutually concerned.

At this point we should examine the Institute's structure to determine if any part of the existing organization would be useful in solving the problem. For example, could the present technical committee system be used?

Actually, these technical committees are primarily concerned with the functions of standardization and definitions. These functions themselves are of great importance to the industry, and represent a direct responsibility of the Institute. We could ill afford to divert any effort from these primary obligations.

Other professional organizations have technical committees whose major task is to obtain papers and operate symposia. Such committees perform only minor standardizing functions. The Institute has no comparable committee system.

Just a few days ago I read an article by Dr. Karl T. Compton on the founding of the American Institute of Physics. I would like to quote from this article. "In the late 1920's another problem presented itself to the American Physical Society. This was the emergence of groups of physicists who felt that the main current of interest in the American Physical Society was not meeting their particular professional requirements. These groups undertook to establish new societies and new publications devoted to their important special interests. Consequently, the American Physical Society was concerned over the centrifugal tendency to separate the basic science of physics into a number of independent groups. Very naturally, each of these groups had its own financial problems of publication."

The article goes on to say that, as a result of a study of the problem "the American Physical Society, the Optical Society of America, and the Acoustical Society of America co-operate in establishing the American Institute of Physics."

The fact is that, through the operational plan indicated by the quotations above, a solution was found to the problem of splintering. That is not to say that in certain instances splintering is not a beneficial action nor that this particular solution is applicable to the resolution of all such problems. Perhaps in this particular instance we might say that the splinters were collected under one roof.

Now let us consider the I.R.E. Professional Group System.

First let me say that I had nothing to do with originating the idea or, in fact, with the initial planning which was necessary to implement it. So far as I know, the idea originated with Dr. R. A. Heising, aided and abetted by Dr. W. L. Everitt.

If we are to apply the principles of decentralization, as developed by industry, to the Institute and the Professional Group System, two important fundamentals must be established. First, the Professional Groups must operate under, and conform with the basic policies of the Institute. The control of the Professional Groups must not be such as to stifle the initiative of the groups, yet it must be such as to prevent any adverse effect on the reputation, prestige, or standing of the Institute as one of the outstanding professional bodies in the world.

Second, the authority delegated by the Institute to the Professional Groups must be balanced by the acceptance on the part of each group of the responsibility and accountability.

That these conditions have been met is due in large part to the wisdom, patience, and judgment of the Executive Committee.

The structure of the Professional Group System is very simple. Each Professional Group comprises a group of I.R.E. members with a mutuality of interest in some particular aspect of electronics. The Chairman of each group is a member of the Professional Group Committee. The Chairman of the Professional Group Committee is a member of the Executive Committee of the Institute.

The form of organization shown in Table I results in a close interlocking of all the elements vital to the development of the Professional Group idea.

The general direction of the Professional Groups and the supervision of the policies established by the Executive Committee is at present executed by the Professional

* Decimal classification: R060. Original manuscript received by the Institute, April 25, 1952. Speech given at the President's Luncheon of the I.R.E. National Convention, New York, N. Y., March 4, 1952.
† General Electric Co., Electronics Park, Syracuse, N. Y.

Group Committee. Since the Chairman of the Professional Group Committee is a member of the Executive Committee, he functions as a direct liaison between these two committees. As the Professional Group System grows, it may be desirable to establish a Co-ordinator of Professional Groups who, as a member of the Executive Committee, will be concerned primarily with policy matters relating to the Professional Groups. Such an arrangement would permit the Professional Group Committee and its chairman to devote their entire attention to operating matters.

Perhaps the most effective way to show the present status of the Professional Groups is by a tabulation of some of the more perterinent data in Table II.

I have not attempted to give information on the operating details of the Professional Group System since information on each group is available. I do want to point out that the Professional Group System is still very much in the developmental state. The development has been slow, as perhaps it should be, in order that we may make the minimum of mistakes and grow soundly. There remain to be resolved many problems in both the operations and policy areas.

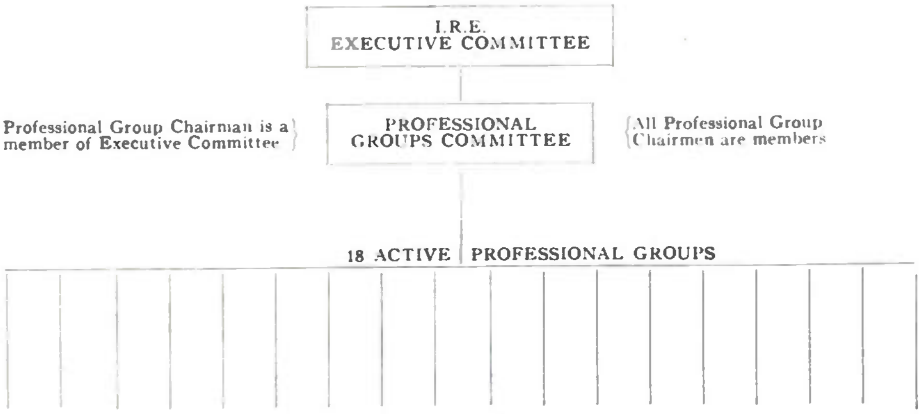
Some of the more important of these problems are:

1. The policy and operations problem concerning the relations of the Institute, the sections, the Professional Groups, and Professional Group Chapters.
2. The fiscal policies applying to the Professional Groups and their Chapters. This involves assessments and other means of obtaining group funds.
3. The publication policies relating to the PROCEEDINGS, group transactions, group newsletters, and symposia transactions.
4. The policy relating to advertising in group publications.
5. The operational problems related to symposia, especially where a symposium is a joint activity with some other organization.

There are many minor operational problems, most of which are under development by the Institute Headquarters. To implement this work the Technical Department has set up a separate organization headed by an assistant to the Technical Secretary. The sole purpose of this organization is to serve the Professional Groups and to concentrate on the solution of the many problems that arise almost daily.

In conclusion, let me repeat that the Professional Group system is distinctly in the formative state. It is important that its development proceed as aggressively as is consistent with sound and conservative Institute policy. The development of the Professional Group System is, in my opinion, of sufficient importance to the Institute to justify more than a casual interest on the part of every member of the Institute of Radio Engineers.

TABLE I



G-1 AUDIO
G-2 BROADCAST TRANSMISSION SYSTEMS
G-3 ANTENNAS & PROPAGATION
G-4 CIRCUIT THEORY
G-5 NUCLEAR SCIENCE
G-6 VEHICULAR COMMUNICATIONS
G-7 QUALITY CONTROL
G-8 BROADCAST & TELEVISION RECEIVERS
G-9 INSTRUMENTATION
G-10 RADIO TELEMETRY & REMOTE CONTROL
G-11 AIRBORNE ELECTRONICS
G-12 INFORMATION THEORY
G-13 INDUSTRIAL ELECTRONICS
G-14 ENGINEERING MANAGEMENT
G-15 ELECTRON DEVICES
G-16 ELECTRONIC COMPUTERS
G-17 MICROWAVE ELECTRONICS
G-18 MEDICAL ELECTRONICS

TABLE II
PROFESSIONAL GROUPS REPORT FOR THE YEAR 1951

Group no	*Name of group*	*No of members*	*Trans, newsletters papers, etc. published*	*No of symposia or conference meetings*	*Attendance at each meeting*	*Financial outcome of meeting*	*Papers presented and published as a result of meetings*	*Date group started*
1	Audio	1,330	4 Transac 4 Newslet 2 Papers	2 Symp			16	6–2–48
2	Broadcast and trans systems	1,059		2 Symp 1 Conf	135 at Phila symp	$93.00 pro Phila	20	7–7–48
3	Antennas and propagation	1,367		4 Conf			64	2–1–49
4	Circuit theory	1,625		3 Symp 1 Conf			33	4–5–49
5	Nuclear science	658	2 Newslet	2 Symp 1 Conf		$166.50 prof Brookhaven symp	18	4–5–49
6	Vehicular communications	524	3 Newslet	2 Conf	185 Chicago	$218.17 profit at Chicago	12	4–5–49
7	Quality control	425	1 Newslet	1 Symp 1 Conf			7	7–12–49
8	Broadcast and telev rec	1,042						8–9–49
9	Instrumentation	1,703	4 Bulletins	1 Symp 1 Conf	549 at Wash	$240.19 prof Wash	35	1–30–50
10	Radio telemetry and rem con	490		1 Symp			6	6–12–50
11	Airborne electronics	436	2 Transac	1 Conf			72	1–9–51
12	Information theory	404		1 Symp 1 Conf			11	5–8–51
13	Industrial electronics	1,307		1 Conf			1	5–8–51
14	Engineering management	382						6–12–51
15	Electron devices	48		1 Conf				10–9–51
16	Electronic computers	201		2 Conf	872 at Phila		16	10–9–51
16	TOTALS	13,001	22	28			301	

High-Frequency Crystal Units for Primary Frequency Standards*

A. W. WARNER†, MEMBER, IRE

This paper is published with the approval of the IRE Professional Group on Instrumentation.
—*The Editor*

Summary—A new approach to the design of crystal units for primary frequency standard use has resulted in crystal units in the 3- to 20-mc frequency range characterized by high Q and low capacitance in the series arm of the equivalent electrical circuit.

By utilizing the overtone frequency of specially shaped AT-cut quartz plates, both Q and the rate of impedance change with frequency are enhanced together, and in addition the stability with time of the crystal unit is increased because of a larger frequency-determining dimension. Additional characteristics of the crystal units include small size, stability under conditions of vibration and shock, and low-temperature coefficient.

Crystal-oscillator stabilities of one part in 10^9 per month have been achieved without recourse to stabilized circuits.

REFINEMENTS in crystal-unit processing techniques during the last decade have made it desirable to re-examine the quartz-crystal frequency spectrum in search of designs which show promise of maximum frequency stability. Desirable features would include small size, ability to withstand physical shock (including portability), and ability to be mass produced,[1] all with no sacrifice in precision as a frequency standard.

It is common practice to obtain good frequency stability by employing a large piece of quartz in the shape of a bar, plate, or ring oscillating at a relatively low frequency, usually 0.1 mc. These quartz plates, bars, or rings must be supported by ingenious devices which minimize energy loss through the mounting, in order to maintain the high Q necessary for a frequency standard. These mountings include resonant wires, cords, pins, and rods, fastened to or supporting the quartz at nodal points. The result is often that the vibrating system on which the frequency stability depends includes such materials as solder, steel, silver, and the like, and variations in frequency occur because of migration or shift of these materials with time and circumstances.

High-frequency, shear-type crystal units, such as the *AT* and *BT* cuts, are well suited to overcome these limitations, for the edges of the crystal plate can be made dormant by proper shaping of the quartz plate and have support wires ridgedly fastened thereto. In addition they are small in size and are a type that is normally manufactured in large numbers.

A development project was established to determine the ultimate practical primary frequency standard crystal unit of the high-frequency shear type, with the following objectives in mind, in the order of importance:

1. Raise Q to 10^6 or more, for good circuit isolation,
2. Achieve an ultimate frequency stability of 5 parts in 10^{10} per day, comparable to the finest frequency standards now in use.
3. Preserve the same production techniques established for the manufacture of the plated crystal unit.[2]
4. Achieve a temperature coefficient of frequency less than 1 part in 10^7 per degree centigrade, to simplify oven design and construction.
5. Shorten as far as possible the initial aging period (2 months to 1 year now required for most primary frequency standards).
6. Adjust the frequency to within 1 cycle per megacycle of the intended operating frequency, to remove the burden of adjustment from the associated circuit.
7. Use an impedance level to match available transmission lines and transformers, i.e., 75 to 300 ohms.

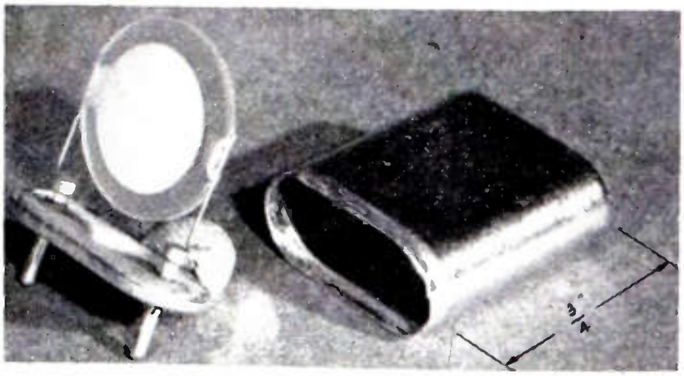


Fig. 1—A typical overtone crystal unit for frequency standard use, with the cover removed to show the simplicity of design.

The results to date of this development can be seen in Fig. 1, showing the small size and simplicity of a typical crystal unit; Fig. 2, showing a typical frequency versus

* Decimal classification: R214. Original manuscript received by the Institute, January 17, 1952; revised manuscript received, May 14, 1952. This work is the result of fundamental studies undertaken during 1949 and early 1950 to determine if the high-frequency plated crystal unit could be used as a primary standard.

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

[1] There is no known method of communicating a frequency accurate to 1 part in 10^8 or better from one place to another hundreds of miles away; therefore, the need for many primary frequency standards, which use a time standard as a reference, usually the mean solar day or a derivative thereof.

[2] R. A. Sykes, "High-frequency plated quartz crystal units," PROC. I.R.E., vol. 36, pp. 4–7; January, 1948.

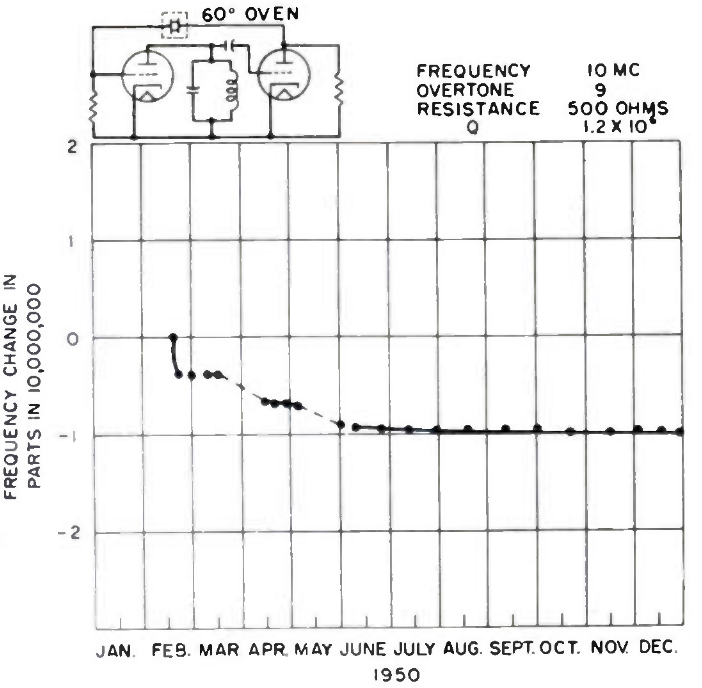


Fig. 2—Aging data taken on a crystal unit, oven, and simple oscillator with reference to the Bell Telephone Laboratories primary frequency standard. An accuracy of 1 part in 10^8 was held for 7 months.

time curve of one of the crystal units in a one-stage oven and simple circuit; and Fig. 3, showing a detailed graph of the aging of one crystal unit for the first 3 weeks after it was made. Q's between 1 and 3 million are readily produced and resistance values are between 100 and 150 ohms at 5 mc.

It is the purpose of this paper to describe briefly the design principles and some of the problems associated with the development of the high-frequency crystal unit for primary frequency standard use and to show the method of solution in some cases.

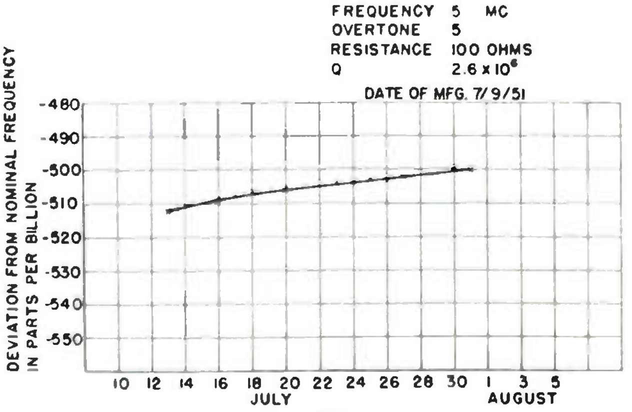
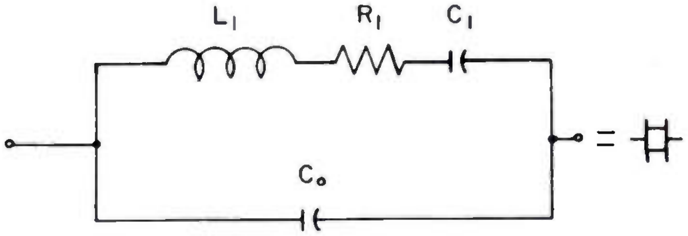


Fig. 3—Aging data taken on a crystal unit alone, showing a change in frequency of about 5 parts in 10^{10} per day, 2 weeks after manufacture.

The design principles are briefly as follows: (1) A spherical contour, to be explained later, is generated on one side of an otherwise conventional plane-parallel quartz plate. The effect of this contour on the finished crystal unit is to restrict the characteristic mechanical vibration to the center of the crystal plate, effectively isolated from the mounting wires. This results in greatly increased Q; increased inductance, L_1; and lower series resonant resistance, R_1, where Q, L_1, and R_1 are as shown in the equivalent electrical circuit of Fig. 4.

$$r \equiv \frac{C_0}{C_1} \doteq \frac{f_r}{2(f_a - f_r)} \quad \text{where}$$



f_r = series resonant frequency, f_a = antiresonant frequency

$$L_1 = \frac{r}{C_0\omega_r^2} \quad \text{since} \quad \omega_r^2 = (2\pi f_r)^2 = \frac{1}{L_1C_1}, \qquad \omega_r = 2\pi f_r$$

$$Q = \frac{r}{C_0\omega_r R_1} \quad \text{since} \quad Q = \frac{\omega_r L_1}{R_1}$$

At series resonance,

$$\frac{dX_c}{d\omega} = 2L_1, \qquad X_c = \text{crystal reactance.}$$

OVERTONE OPERATION

$$r \propto n^2 \quad \text{where} \quad n = \text{overtone order}$$

$$L_1 \propto n^3 \quad \text{and} \quad C_0 \propto \frac{1}{n} \quad \text{for} \quad \omega_r \text{ constant}$$

$$R_1 \propto n^3 \quad \text{for} \quad Q \quad \text{and} \quad \omega_r \text{ constant}$$

At cut quartz,

$$f_r = \frac{n1670}{t} k_c, \qquad t = \text{thickness in mm.}$$

Fig. 4—Equivalent electrical circuit and characteristics, applicable to the design of high-frequency overtone crystal units.

(2) The high Q thus achieved allows the use of overtone operation[3] without excessively high values of R_1, resulting in a thicker crystal plate which is more easily calibrated to frequency and affected less by contamination.

The problems were mainly: (1) to remove the unwanted effects of grinding and lapping the quartz plate, (2) to reduce the ever-present contamination of the crystal plate effectively, (3) to produce compact, pure, precise gold films, and (4) to achieve a better temperature coefficient of frequency through closer control of the crystal-plate orientation with respect to its crystalographic axis.

To eliminate the unwanted effects of grinding and lapping (i.e., loose, strained, or powdered material and cracks, fissures, scratches, and the like), it was found necessary to polish the quartz plate optically. In this way it is possible to observe that the faulty material

[3] Overtone operation as used in this article refers to the operation of the crystal plate on one of its odd approximately harmonically related overtones. The crystal plate behaves very much as though it were a stack of n fundamental plates, n being the overtone order. At a given frequency, the thickness of the crystal plate is almost directly proportional to the overtone employed.

has been removed. By optical means even the smallest scratches, called "strakes," may be readily seen on a polished plate, and a continual check kept on the finishing process.

The polished surface is also more easily kept clean since the contaminants are not imbedded and can be removed by relatively short exposures to the commonly used solvents. The polished plates also have less surface area than lapped plates so that the mass of foreign material per unit volume of quartz is less, by a ratio of as much as 5 to 1.

The method of depositing the gold electrodes onto the quartz plates may be divided into two phases: first, the production of a base electrode common to all units and deposited on a whole group of crystal plates at one time; and second, the addition of a precise amount of gold to each crystal unit individually, so that the series resonant frequency of the crystal will be correct. The first phase is readily accomplished since conventional vacuum systems and relatively high temperatures may be employed. A predetermined weight of gold is evaporated in a vacuum onto the clean, masked, surfaces of the quartz plates. For the final addition of gold to bring the crystal to the desired frequency, the frequency of the crystal unit must be continuously measured to determine the correct weight of gold, to an accuracy equivalent to about 1/100 of a microgram. The use of high temperatures is ruled out and the time of pumping must not be excessive since only one unit may be adjusted at a time. The apparatus for this second phase was designed for extremely rapid pumping, with large ducts and small, easily degassed chambers for the evaporation process. An elaborate system of traps was added to reduce the possibility of foreign material contaminating the gold film, and multiple chambers were used so that the pumps would never be idle. All controls are automatically operated electrically for speed and safe operation. The result is the production of a compact, pure, gold electrode with a high order of precision.

The problem of a better temperature coefficient was solved by better X-ray measuring equipment and quartz processing methods capable of accuracies to better than 2 minutes of arc.

As pointed out above, a spherically contoured crystal plate is used in the design of these crystal units. Specifically, an *AT*-cut, contoured, overtone crystal plate is used, and Fig. 4 gives some useful equations to relate physical crystal unit characteristics and equivalent electrical characteristics. The shaping consists of a convex spherical contour generated on an otherwise conventional plane-parallel, crystal blank. The inductance L_1 and Q as a function of the radius of curvature of this type of shaping are shown on Fig. 5 for a 9-mc, third overtone crystal unit. As the contour first departs from flatness, the Q is improved (because the energy loss through the mounting is less) and the inductance L_1 is little affected. But as greater contours are employed, the effect on the inductance becomes much greater and the rate of improvement in Q tapers off. Since $Q=\omega L_1/R_1$, the result is that the resistance R_1 is a minimum at some particular value of contour where the increase in inductance balances the increase in Q, as shown in Fig. 6. Fig. 7 is a design chart giving the radius of curvature to obtain this minimum R_1 for a frequency range of 2 to 30 mc and operation on overtones 1 to 9. It should be borne in mind that these values of contour for minimum R_1 are for a particular design and that crystal-plate size, mounting, electrode size, and the like all affect, to some extent, the point at which the balance between L_1 and Q previously referred to, is obtained.

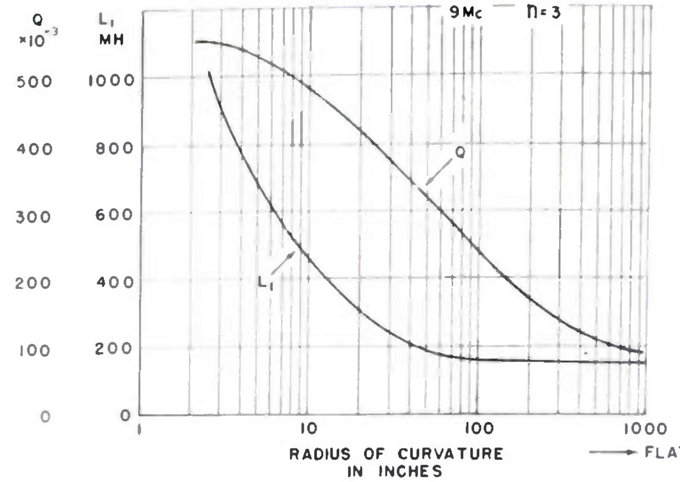


Fig. 5—Variation in Q and inductance with the degree of contour on a 9-mc, third overtone crystal unit.

The principal reason for choosing that point of minimum impedance is to allow the use of a high-order overtone without exceeding the desired impedance dictated by circuit conditions since the crystal unit impedance increases with the cube of the overtone. For a given frequency, an overtone mode crystal plate is thicker, higher in Q, more stable, and more easily adjusted to

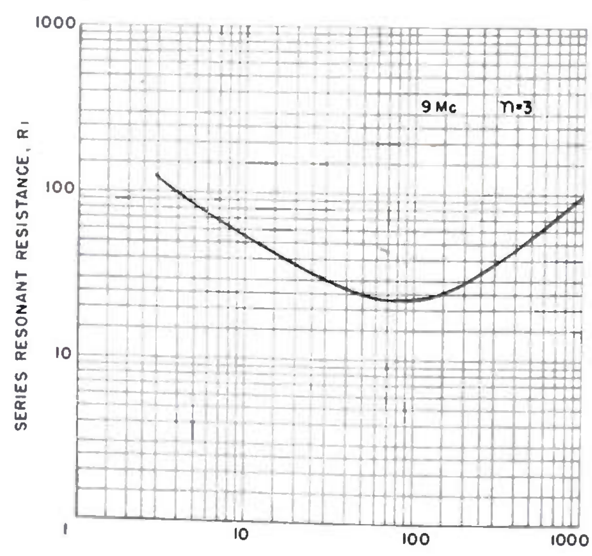


Fig. 6—Variation in series resonant resistance with the degree of contour on a 9-mc, third overtone crystal unit.

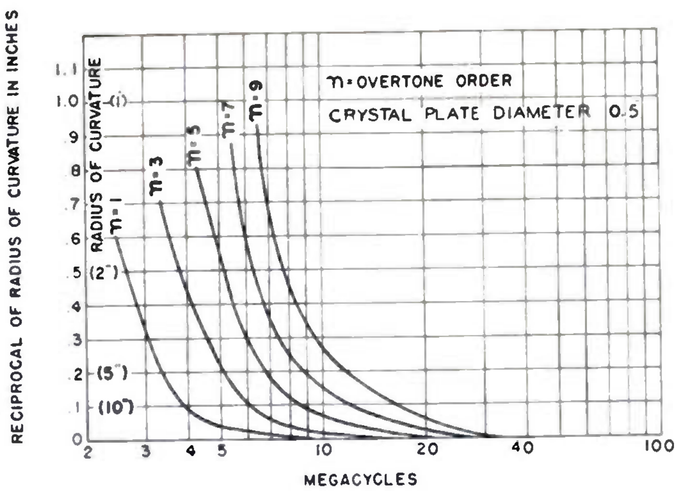


Fig. 7—Typical design chart of the degree contour to obtain minimum series resonant resistance for crystal units 3 to 20 mc and overtones 1 to 9.

frequency as the overtone order is increased. Therefore, a higher and more beneficial overtone may be used if the crystal plate is ground with the contour of lowest impedance.

In summary, by designing the crystal unit so as to take advantage of optimum overtone, contour and surface finish, as well as advanced manufacturing techniques, performance has been made 10 to 1,000 times better in all respects, i.e., Q, aging, frequency tolerance, and temperature coefficient. The crystal unit still retains much of the original plated crystal-unit design, basically the same type of tools and techniques set up for lapping, plating, mounting, and adjusting conventional high-frequency plated crystal units being used. The result is a new and valuable primary frequency standard crystal unit that is small, rugged, and more adaptable to mass-production techniques.

A Coaxial Power Triode for 50-KW Output up to 110 MC*

R. H. RHÉAUME†, MEMBER, IRE

Summary—**The introduction of all coaxial ring-seal terminals, a thoriated cathode, a re-entrant anode with integral coolant jacket, and a novel assembling technique has facilitated the development of a new power triode for 50-kw rf output up to 110-mc frequency. Increased power output ratings are available at lower frequencies. The bandwidth is suitable for television broadcasting.**

Design requirements are reviewed for optimum electrode geometry, heat-dissipating ability, minimum lead inductance, high rf conductivity of the vacuum seals, and other desiderata for high-power, high-frequency service. Mechanical features are described, and the circuit performance of the tube is discussed for lower frequencies as well as the vhf region.

Introduction

A RIGOROUS ANALYTICAL METHOD is not available for designing high-frequency, high-power triodes completely from basic physical and mathematical considerations. Several different avenues of approach are explored, and the derived information is judiciously correlated to arrive at the desired design objectives. Some fundamental requirements are:

1. Electrode dimensions must be small in the direction of wave propagation compared with a wavelength for uniformity of applied potential. Other dimensions must not be large enough to induce spurious oscillations.
2. Electrode dimensions must also be made small to minimize electrode capacitances, especially since electrode spacings must be reduced for minimum transit-time losses and maximum perveance.
3. Grids and anodes must have high heat-dissipating capability, and cathodes must be able to supply adequate emission for space-charge operation with a minimum of heating power.
4. Electrode leads and terminals must be designed for minimum inductance and for optimum coupling into grid-separation concentric lines.
5. Rf conducting surfaces, including vacuum seals, must pass very large charging currents without overheating. Excessive dielectric heating of insulation must be avoided.

High-Frequency Design of the ML-5681

A. Significant Dimensions

Fig. 1 is a section view taken along the principal axis of the ML-5681. The multistrand, free-hung, thoriated-tungsten cathode has a vertical length of about 6 inches, or $\lambda/16$ at 110-mc frequency. This is an empirically established dimension, one quarter of a quarter wavelength, short enough to ensure uniform instantaneous potential along each of the electrodes even at the shortest wavelength. The effective cathode-grid spacing is approximately one eighth of an inch, and the grid-anode distance is about three times as great. The grid is a molybdenum helix spot-welded to a set of vertical molybdenum stays which are spaced equally around the grid-bolt circle. The largest diameter of the tube is

* Decimal classification: R334. Original manuscript received by the Institute, April 12, 1951; revised manuscript received, April 23, 1952.

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8 inches, or one-quarter the diameter to excite the 110-mc TE_{11} mode of oscillation. This would be the lowest frequency spurious mode encountered in concentric circuitry. The grid and anode terminal flanges are separated by $3\frac{1}{4}$ inches of glass, and the associated kovarglass seals are $6\frac{1}{2}$ inches in diameter.

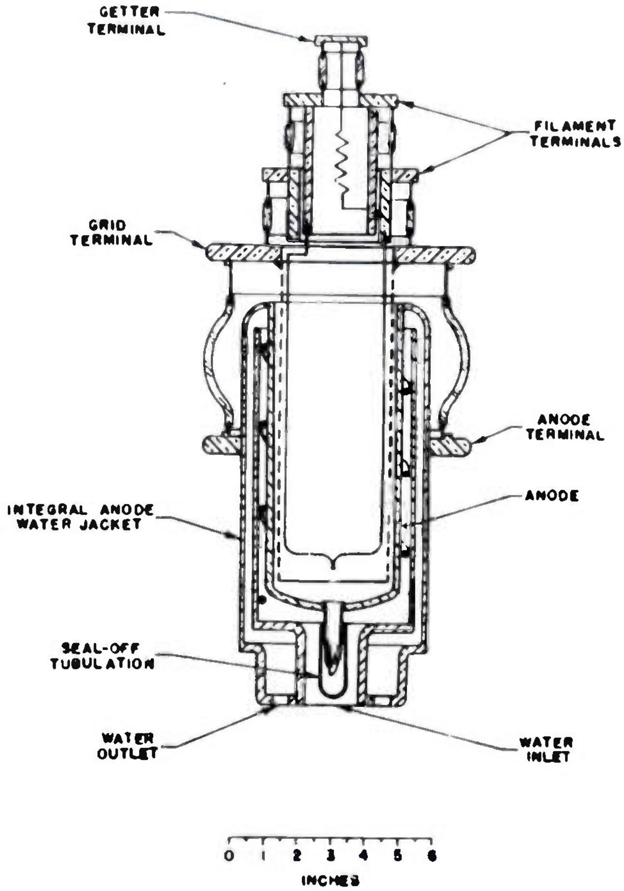

Fig. 1—Section view along the principal axis of the ML-5681 power triode.

B. Anode Cooling

With the full-rated plate dissipation of 75 kw, the anode heat-flow density of the ML-5681 is roughly twice that of earlier tubes. It has been accomplished by combining several design principles:

1. using a large number of equally spaced cathode strands to distribute the heating more uniformly over the total anode surface,
2. employing a heavy-wall copper anode to prevent the formation of hot spots,
3. taking advantage of the integral coolant jacket to obtain optimum coolant flow characteristics over the anode surface.

Uniform heat distribution and cooling of the anode must be achieved without impairing its structural strength. Referring again to Fig. 1, a family of parallel spiral partitions is inserted in the cooling jacket, in contact with both the baffle and the anode surface, with each spiral spaced equally from its neighbors around the anode circumference, and with a pitch of approximately two-thirds of one revolution along the entire length. The anode itself remains smooth and is not grooved or modified in any way. The advantages of this construction are:

1. Alternate hot and cold bands of moving liquid in the coolant jacket are eliminated by properly directing the flow.
2. A uniform depth of liquid passage and velocity of flow are obtained.
3. The mechanical strength of the anode is entirely preserved.
4. Uneven cooling at the ends of the anode is eliminated.
5. Larger increments of temperature of the cooling liquid may be used with correspondingly greater anode heat dissipations.

With a flow of 25 gallons per minute of cooling water at an entering temperature of 20°C, audible steam hissing does not occur until 70 kw are being dissipated on the anode, compared with 55 kw for a plain jacket.

C. Grid Design

Overheating of grids from rf charging currents, as well as electron bombardment heating, may occur in high-frequency, high-power triodes. Charging current is directly proportional to frequency.

$$I_c = 2\pi f C_{gp} E. \quad (1)$$

At 110 mc with $E = 8$-kv swing between grid and plate, the charging current is 260 rms amperes. Most of this current flows up the grid stays, so they are made as large and as numerous as possible. Since the charging current will be densest at the foot of the grid, the junction with the grid terminal flange is of massive design for good electrical and thermal conduction, as may be seen in Fig. 2 (see page 1035).

The distance between adjacent grid helix turns has to be made at least as small as the cathode-grid spacing to maintain satisfactory grid control. Therefore, a relatively fine molybdenum wire and short pitch are used for the helix.

D. Cathode Structure and Perveance

The ML-5681 cathode is a free-hung multistrand thoriated-tungsten cage suspended from the lower ends of the two large, concentric copper cylinders which are the cathode terminal leads, as shown in Fig. 1. This construction eliminates the necessity for a spring-loaded central supporting mast and allows an unusually close cathode-grid spacing. Single-phase heating power is used, with alternate filament strands connected in phase opposition to minimize magnetic force distortions. The rated heating power, 2.7 kw, provides a usable peak emission current of 65 amperes.

The following relation between instantaneous plate current and electrode potentials is approximately true for class C operation of a space-charge limited triode.

$$i_p = \frac{kA}{d_{c_g}^2}\left(e_c + \frac{e_b}{\mu}\right)^{3/2}, \qquad (2)$$

where A is cathode area,
d_{c_g} is cathode-grid spacing,
k is a constant,
e_b is instantaneous plate potential,
e_c is instantaneous grid potential,
μ is amplification factor, measured with slightly negative grid.

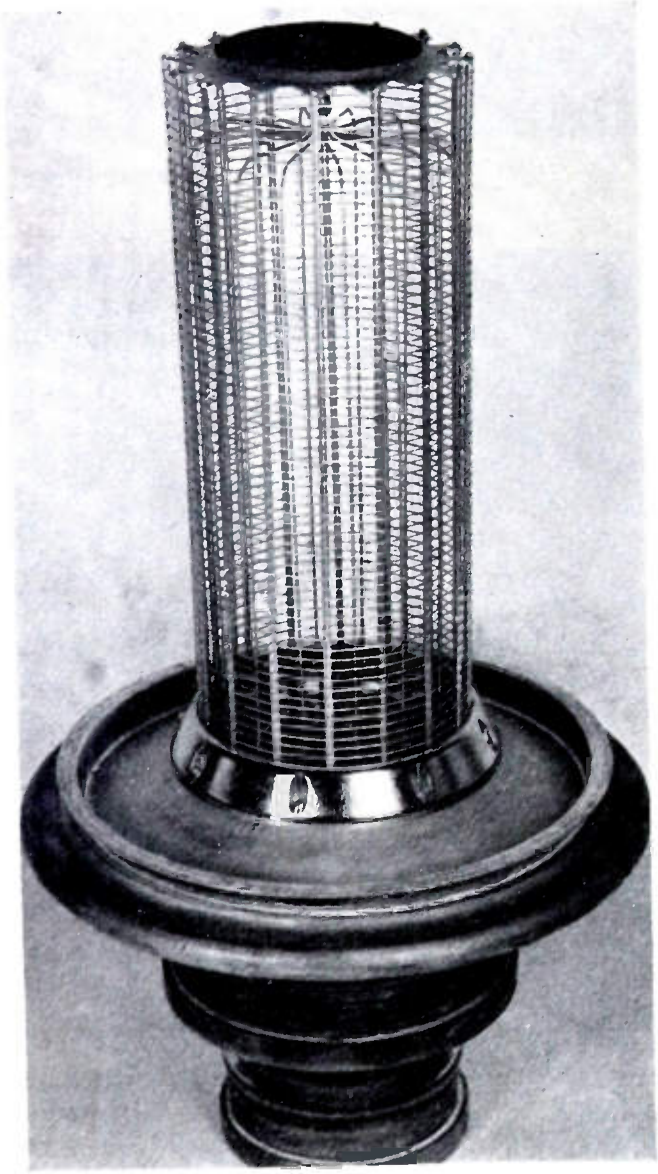

Fig. 2—ML-5681 grid-cathode assembly.

To achieve large values of i_p with small e_b and e_c the "perveance" factor, $kA/d_{c_g}^2$ must be made as large as possible. This objective is attained in spite of high-frequency size limitations upon A by minimizing d_{c_g}. In this way a higher frequency ceiling, larger power output, improved bandwidth, better plate efficiency, smaller dielectric loss, and lower driving power are obtained than with former tubes.

To take full advantage of high perveance it is essential for the cathode to emit enough electrons to maintain space-charge limited operation at all times. In this regard the thoriated-tungsten cathode of the ML-5681 has certain advantages over equivalent pure tungsten cathodes:

1. A higher cathode electron emissivity may be used, 1 amp/cm² compared with 0.5 amp/cm².
2. Less heater power is needed to supply the same total electron emission.
3. Less grid heating occurs from cathode radiation since the thoriated filament runs about 600°C cooler than pure tungsten and requires less total heating power.

The magnitude of the amplification factor is not very critical for optimum class C operating characteristics, although too low a μ will result in excessive loss of driving power in the grid bias while too high a μ will cause undesirable driving power loss through grid bombardment at high output levels. The magnitude of μ which has been selected for the ML-5681, twenty-three, is believed to be a reasonable compromise between these two power-gain limiting conditions.

E. Terminal Lead Inductances

By wrapping copper sheets around adjacent pairs of tube terminals and inserting the probe of a megacycle meter through a small hole in the copper, it was found that the grid-anode self-resonant frequency was 167 mc and the grid-cathode self-resonant frequency was 147 mc with both cathode terminals strapped together. The effective grid-anode and grid-cathode lead inductances are therefore approximately 0.014 μh. Accordingly, the $\lambda/4$ mode may be utilized at, and even above, 110 mc in both input and output concentric lines; this is advantageous for physical compactness of the circuit, for low energy storage, and for ample frequency bandwidth.

When ML-5681's are used with ordinary lumped constant circuitry at low and medium frequencies, these small lead inductances will render the neutralization of grid-anode capacitance less frequency-sensitive, and the tendency of input admittances to increase with rising frequency will be less rapid than with earlier types of power tubes.

F. Vacuum Seals

Ordinary $\frac{3}{4}$-inch diameter, kovar-glass vacuum seals will pass up to 30 rms amperes at 40 mc without seriously overheating the metal-glass junctions; $6\frac{1}{2}$-inch seals such as those of the ML-5681 at 110 mc would therefore be expected to handle 200 amperes, but the actual grid-anode charging current at 110 mc will be 260 amperes. Therefore, the entire area of contact between metal and glass of the grid-anode seals is gold-plated to reduce rf resistance and heating at the higher frequencies.

The larger of the grid sealing rings consists of two deep-drawn kovar parts shaped to fit into each other for the purpose of forming a silver-soldered, metal-to-metal final seal instead of the conventional glass-to-glass type. With ordinary glass-to-glass final seals, the necessary heat for softening is applied by a gas torch over internal parts which have been previously cleaned or outgassed. It is difficult to prevent oxidation within the tube even though an internal protective atmosphere is employed. The glass must be annealed afterward to remove working stresses, at the risk of de-aligning glass-supported electrodes. With metal seals, on the other hand, no annealing is necessary because all glassed parts have been previously heat treated. The final seal is made by induction heating in a protective atmosphere within a bell jar; owing to the poor heat conductivity of the kovar the adjacent tube parts remain cool and stress free.

G. Electron Transit-Time Effect

Flight time of electrons between parallel planes under space-charge limited conditions may be expressed:[1]

$$T = 6.6 \times 10^{-4}\left(\frac{d}{J}\right)^{1/3} \mu\text{sec}, \tag{3}$$

where d is the distance between the planes in centimeters and J is the electron emissivity of the cathode plane in amperes/cm². For the ML-5681, T becomes 4.5×10^{-10} seconds or 0.3 radians in the cathode-grid region at 110 mc. Equation (3) emphasizes the importance of high-emissivity cathodes and small cathode-grid spacings in high-frequency, high-power tubes. Fig. 3 shows the computed effect of electron transit time upon the output power of this tube with frequency.

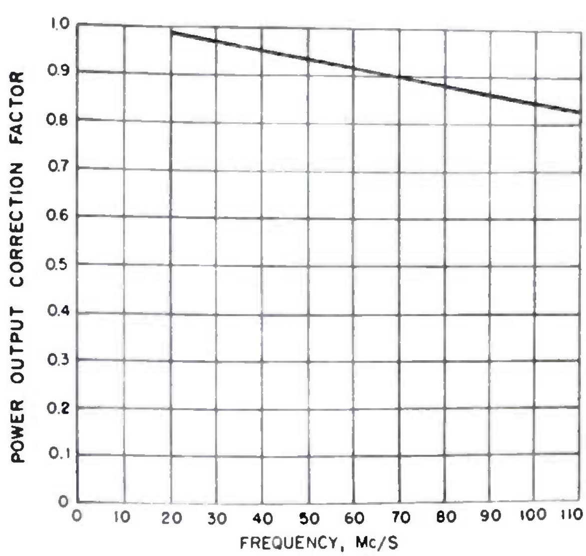

Fig. 3—ML-5681 correction of power output with frequency (transit-time effect).

H. ML-5681 Mechanical Features

Tubes fulfilling the previously outlined high-frequency, high-power design principles are smaller and lighter in weight than medium frequency tubes of equivalent power, as may be seen from Fig. 4. Both tubes illustrated are capable of 100-kw rf output at 20 mc. The ML-5681 is only 18 inches long, 8 inches in diameter, and weighs only 43 pounds. It may be disengaged from its socket merely by twisting it through 60 degrees and lifting it straight upward about an inch. A new one is inserted by reversing the process, automatically making the only water connection, as shown in Fig. 5.

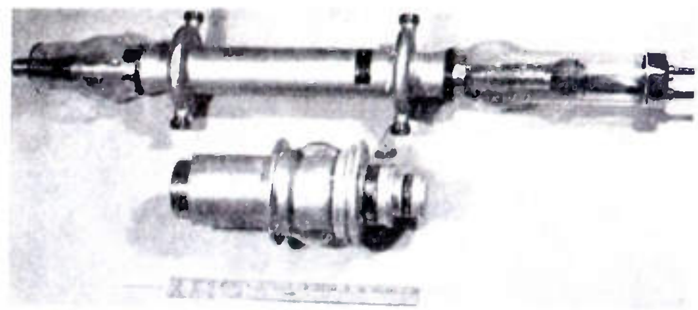

Fig. 4—Size comparison of the ML-5681 with an equivalent medium-frequency tube.

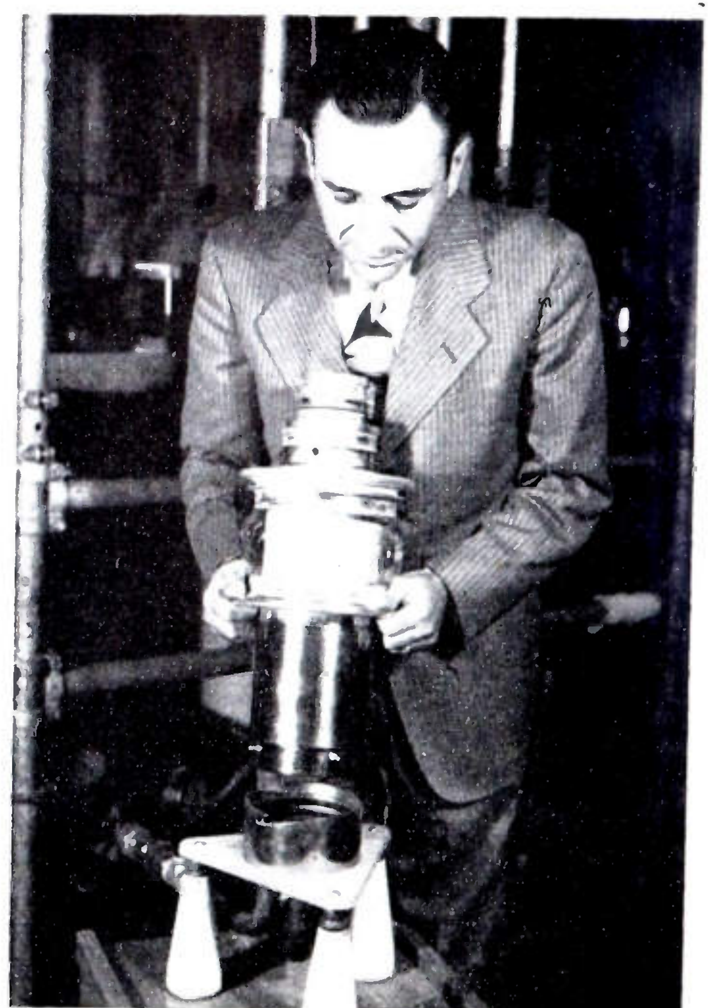

Fig. 5—Method of inserting an ML-5681 in its mounting socket.

[1] H. D. Doolittle, "Design Problems in Triode and Tetrode Tubes for U.H.F. Operation," Machlett Cathode Press, Machlett Laboratories Inc., Springdale, Conn., vol. 6, p. 11; 1949.

Circuit Performance

Constant current characteristics derived from pulsed measurements are given in Fig. 6. As a result of this tube's high perveance, it will be seen that relatively small increments of grid voltage are required for large increases of plate current at constant plate potential.

Prototype models of this tube were oscillated at Bell Telephone Laboratories up to 45-kw power output at 100 mc in a grid-separation coaxial circuit. Difficulties

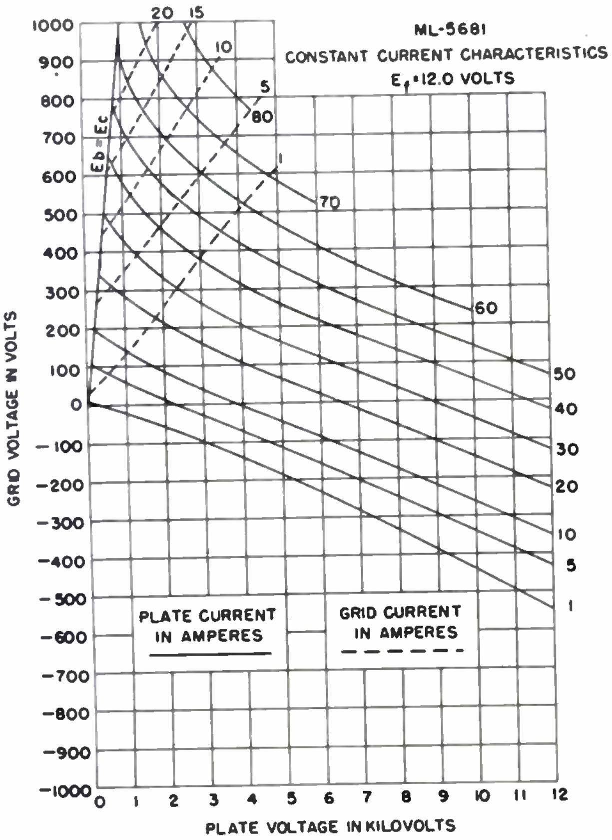

Fig. 6—ML-5681 constant current characteristics.

with high-dissipation water loads at very high frequencies prevented further testing up to 50 kw and 110 mc. Other engineering groups who are currently studying performance of the ML-5681 may wish to publish additional results at a later date.

In the 15–18-mc region, a pair of these tubes have been tested as push-pull plate-modulated power amplifiers in an international short-wave transmitter operating at a carrier level of 100 kw. Good stability was achieved, with a plate efficiency of 70 per cent, even though less than optimum excitation was available for the specified tube-plate voltage. Parasitics appearing as harmonics of the carrier frequency up to 60 mc or higher were suppressed by means of single-turn inductors at the grid terminal flanges.

Factory testing is performed in a general-purpose Hartley power oscillator at a frequency of 1 mc. While no concerted effort has been made to establish circuit parameters for optimum plate efficiencies, values up to 78 per cent have been recorded.

An approximate class C analysis indicates that the ML-5681 should be a good broad-band linear amplifier for television broadcast bands two through six, giving 50-kw synchronizing peak-power output with 10 kw of driving power.

Conclusion

It is not convenient to design high-frequency, high-power triodes rigorously from basic mathematical and physical considerations. However, it has become possible to fabricate commercially a triode capable of 50-kw rf output up to 100-mc frequency with a bandwidth suitable for television broadcasting. The introduction of all coaxial ringseal terminals, a thoriated cathode, a reentrant anode with integral coolant jacket, and a novel assembling technique have facilitated the achievement of such desiderata as a minimum physical size and weight, low interelectrode capacitances and small lead inductances, superior grid and plate heat-dissipating capability, with improved cathode electron emissivity, perveance, and rf seal conductivity. It has also been possible to achieve very good mechanical features.

Tubes designed in this way for optimum high-frequency, high-power service are also likely to give better performance in the medium- and low-frequency regions than earlier tubes which were designed with only these latter purposes in view.

Acknowledgment

It is a pleasure to acknowledge the contributions and suggestions of Dr. H. D. Doolittle to this paper, and to point out that Mr. G. J. Agule executed the mechanical design and developed novel fabricating procedures for the ML-5681 power triode. Early prototype models were built and tested by Messrs. C. E. Fay and D. A. S. Hale at Bell Telephone Laboratories.

Inverted Magnetron*

JOSEPH F. HULL†, MEMBER, IRE

Summary—A radically different type of magnetron is described in which the positions of the cathode and the anode segments are inverted from those in ordinary magnetrons. The direction of curvature of the interaction space of this magnetron is therefore opposite that of conventional magnetrons. Electronic efficiencies of 50 to 80 per cent have been measured on these structures and static input impedances as low as 60 ohms have been observed on high-power pulse tubes. Sound scientific basis is provided for the use of the parallel-plane magnetron interaction space in new microwave devices.

INTRODUCTION

THIS PAPER deals with a radical variation of the magnetron oscillator. The conventional magnetron, as used today, consists of a rugged cylindrical cathode surrounded by a set of equally spaced anode segments. These segments are connected to some radio-frequency resonant tank circuit in such a way that the phase difference between adjacent segments is ordinarily π electrical radians. Only a few years after the magnetron had become a high-power microwave radar tube did it occur to workers in this field that it might be possible to build a magnetron consisting of a large cylindrical cathode with the emission coating on its inner surface, inside of which, suitably spaced, would be the anode segment structure and tank circuit resonant cavity or cavities. An obvious advantage of this construction is the same as that of inversion (or turning "inside out") of any type of electron tube: namely, increasing the cathode-emission surface area. Also, in this case there would be an increase of anode area presented to the cathode since the anode segments are located outside, instead of inside, the resonant cavity structure. However, the most important reason for investigating the inverted magnetron is to study the behavior of a high-density space-charge cloud under the influence of magnetron-type RF fields when the configuration of the interaction space is radically different from that of a conventional magnetron oscillator.

Inversion of the magnetron, however, is not as simple in theory as inversion of a tube with simpler electron trajectories, such as the triode or tetrode, since a much more complicated interaction between the rotating space charge and the electromagnetic field presented by the anode segments must take place. In fact, earlier attempts to operate inverted magnetrons have been essentially unsuccessful. Several suggested reasons for these failures have been given. One is that the interaction of the electrons with the fields was unfavorable in the case of the inverted magnetron, thus forbidding oscillation. Another suggestion is that when the multicavity vane-type magnetron was turned inside out the mode separation decreased and the tank circuit efficiency decreased so much that the tube would not oscillate. The experiments of the author indicate that the latter explanation of previous failures is the more plausible since inverted magnetrons with new and different types of resonant cavities, which do not suffer from inversion, have been found to operate stably and with reasonably high efficiency.

Two previous papers[1,2] have analyzed and demonstrated the practicability of using the conventional single-cavity interdigital magnetron in the simplest mode. Due to the fact that the interdigital magnetron is a single-cavity resonator and the vane-type magnetron is a multicavity resonator, it is easy to show why the interdigital magnetron structure, but not the vane-type magnetron, can be inverted without detrimental effects on mode separation and circuit efficiency. Adequate mode separation in vane-type magnetrons is ordinarily achieved by means of straps which connect alternate vane tips together. The effectiveness of strapping is greatest for the shortest strap length. For the inverted vane-type magnetrons the vane tips are relatively far apart so that the strapping is ineffective and sufficient mode separation is therefore difficult to achieve. In the interdigital magnetron, a single-cavity resonator, the number of modes is essentially independent of the number of anode segments. Therefore, a large-diameter cavity can be built with a large number of segments and a large anode area having sufficient mode separation.

DESCRIPTION OF THE EXPERIMENTAL TUBES

The prototype of the tube used in these experiments is shown in Fig. 1. Two scalloped copper discs, *A* and *B*, are mounted on and brazed to a hollow center post, *C*. To the outer surfaces of the discs are attached an interleaving set of fingers, *D*, which form the anode segments. In the simplest and lowest frequency mode of oscillation of this cavity, the radio-frequency magnetic field lines are concentric with the center post. The magnetic flux density is maximum at the center post, decreasing to its smallest value at the tooth structure. Except in the immediate vicinity of the tooth structure the electric field exists in the axial direction only, and is zero at the center post, increasing to its maximum value at the tooth structure. Since the fingers are attached alternately to the top and bottom plate, the fingers present a plus-minus, or π mode potential distribution, to the cathode. Other higher-order, higher-frequency modes which may exist in this cavity have sinusoidal angular variations of the fields. In all the higher-order modes

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† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

[1] J. F. Hull and A. W. Randals, "High power interdigital magnetrons," PROC. I.R.E., vol. 36; November, 1948.

[2] J. F. Hull and L. W. Greenwald, "Modes in interdigital magnetrons," PROC. I.R.E., vol. 27; November, 1949.

the magnetic field threads through the tooth structure so that a considerable portion of the fields exist outside the cavity.

To prevent electromagnetic energy leakage out of the cavity in the cavity mode, the folded choke, *G*, is pro-

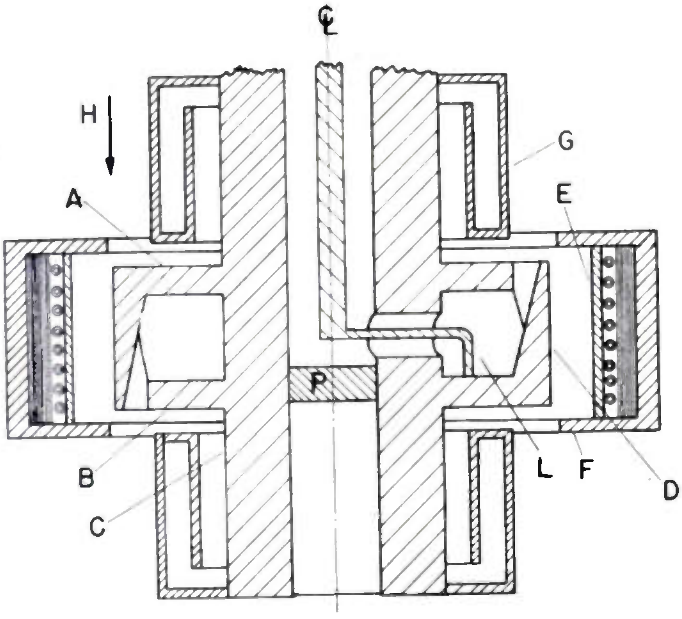

Fig. 1—Cross-sectional view of essential elements of a single-cavity inverted interdigital magnetron. The main vacuum envelope and coaxial output seal are external to all the parts shown.

vided on both ends of the cavity. To prevent leakage of the electrons axially along the dc magnetic field lines, *H*, end hats, *F*, at cathode potential, screen the interaction space from the outer portion of the tube.

Many different methods could be used to couple power out of this magnetron, but the method chosen for these experiments was a loop, *L*, connected to the inner conductor of the coaxial line whose outer conductor is the hollow center post. Water cooling is provided through the other end of the hollow center post which is separated from the vacuum by the plug, *P*.

A longer anode with a large capacity for heat dissipation can be made by properly stacking a number of cavities as shown in Fig. 2. The anode bars, *D*, now replace the fingers, and each bar joins alternate discs together. Adjacent bars connect opposite sets of discs so that, at resonant frequency, the bars present the desired π mode potential distribution to the cathode. In order to prevent radio-frequency unbalance between opposite sets of anode bars, there must be an odd number of cavities. Water cooling and power-output coupling are provided in the same manner as in Fig. 1. Fig. 3 is a photograph of one of the tubes used in the experiments.

Design Considerations

In designing this magnetron it was necessary to determine the cavity dimensions for the desired resonant wavelength and external *Q*, as well as the correct interaction space dimensions for the desired operating voltage, anode current, magnetic field, and electronic efficiency. The calculations for the resonant wavelength and external *Q* were very similar to the analysis of the conventional interdigital magnetron which is well

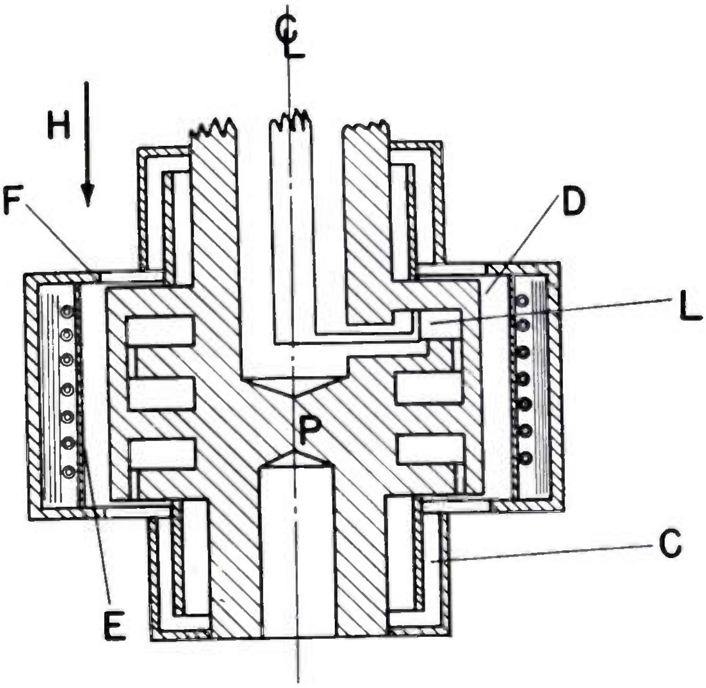

Fig. 2—Cross-sectional view of a stacked inverted interdigital magnetron.

known.[3,4] The equation for resonant wavelength in the mode of zero angular variation in the cavity is given by

$$\frac{2\pi r_t^2\epsilon}{\alpha c h}\,\frac{\lambda}{2\pi r_t} = \frac{J_0(K_1r_t) - \dfrac{J_0(K_1r_i)}{N_0(K_1r_i)}\,N_0(K_1r_t)}{J_1(K_1r_t) - \dfrac{J_0(K_1r_i)}{N_0(K_1r_i)}\,N_1(K_1r_t)}. \qquad (1)$$

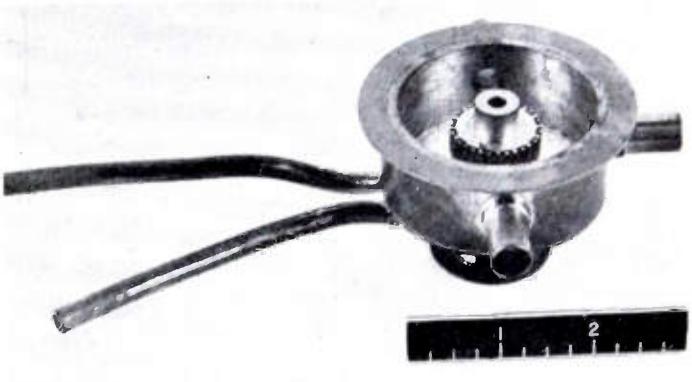

Fig. 3—A partially assembled single-cavity inverted magnetron used for basic study of magnetron space-charge behavior.

(A list of symbol definitions is given at the end of this paper.) A graphical solution of this equation for a wide range of parameters is given in Fig. 4.

[3] F. Lüdi, "Eigenfrequenzen des E-typus eines Kapazitats-belasteten Zylindrischen Hohlraumes," *Helv. Phys. Acta*, vol. XVII, Fasciculus Sextus, pp. 429–436; 1944.
[4] J. F. Hull and L. W. Greenwald, *op. cit.*

The external Q for the mode with no angular variation in the fields may be calculated from (2)

$$Q_x = \frac{\omega h\left[Z_0 + \frac{\omega^2 L^2}{Z_0}\right]\left[\pi\epsilon r_t(r_t - r_i) + h\alpha c\right]\left[J_0(K_1 r_t) - \frac{J_0(K_1 r_i)}{N_0(K_1 r_i)} N_0(K_1 r_t)\right]^2}{A^2 K_1^2\left[J_1(k_1 r_i) - \frac{J_0(K_1 r_i)}{N_0(K_1 r_i)} N_1(K_1 r_i)\right]^2} \tag{2}$$

The interaction of the electron stream with the radio-frequency fields is essentially the same as in the parallel-plane magnetron since the curvature of the tooth structure is not great. Therefore, to obtain design equations the cathode and anode radii and the number of anode segments were allowed to approach infinity in the well-

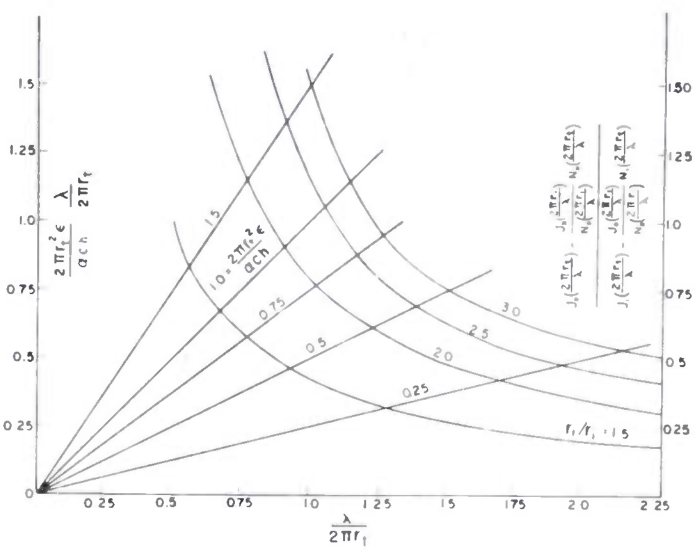

Fig. 4—Resonant wavelength equation plot for inverted interdigital magnetrons in the mode of no angular variation.

known Hartree magnetron starting-voltage equation for conventional magnetrons. The following relationship between operating voltage, V, and magnetic field, B, was obtained:

$$\frac{V}{V_0} = \frac{2B}{B_0} - 1, \tag{3}$$

where $V_0 = 2s^2f^2m/e$ is the lowest possible anode operating voltage and $B_0 = 2sfm/de$ is the lowest possible magnetic field for oscillation. The maximum electronic efficiency is given by

$$\eta_{e(\max)} = 1 - \frac{V_0}{V}.$$

It may be seen that the lowest value of magnetic field for oscillation, B_0, is dependent only on the frequency and d/s, the ratio of cathode-anode spacing to the distance between tooth centers. The optimum value of the parameter, d/s, was determined experimentally to be between 0.7 and 1.0 for the inverted magnetrons described in this paper. In order to realize a reasonable electronic efficiency, the tube was designed so that the ratio V/V_0 was about 10.

A total of 10 inverted magnetrons have been built, 8 of which operated with over-all efficiency over 30 per cent. The first of these were the single-cavity type shown in Fig. 1, with typical dimensions as follows:

anode radius	1.27 cm
α (number of segments)	32
d (cathode-anode spacing)	0.198 cm
s (distance between segment centers)	0.250 cm
center-post diameter	0.64 cm

The modes of this cavity were found at the following frequencies:

cavity mode	2,730 mc
first-order mode	3,100 mc

When three cavities were stacked together to make a cavity of the type shown in Fig. 2, the first three modes were found at essentially the same frequencies as the modes in a single isolated cavity. This showed that longitudinal modes were sufficiently removed in frequency so that they were unimportant. Stacking together as many as six cavities did not bring the frequency of longitudinal modes within the frequency range of the first three cavity modes. (Radio-frequency unbalance due to use of an even number of cavities was permitted while checking higher-order modes.) Typical dimensions of a multicavity magnetron used in these experiments are as follows:

number of cavities	3
anode radius	1.9 cm
α (number of segments)	32
d (cathode-anode spacing)	0.292 cm
s (distance between segment centers	0.374 cm
center-post diameter	1.02 cm

The modes of this magnetron were at the following frequencies:

cavity mode	2,800 mc
first-order mode	3,100 mc
second-order mode	3,400 mc

The cathodes which were used were all of the oxide-coated type with a nickel base, and were heated by an insulated tungsten heater wire wound on the outside surface of the emitter sleeve. The end hats were made of oxidized 18-8 stainless steel.

Experimental Results

The highest over-all efficiency attained with a single-cavity inverted magnetron was 51 per cent. The measured circuit efficiency of this tube was 68 per cent and the electronic efficiency of this tube 75 per cent. A performance plot of a typical single-cavity tube is shown in Fig. 5. This performance plot was taken with 5-per cent duty-cycle pulse so that the operation was essentially

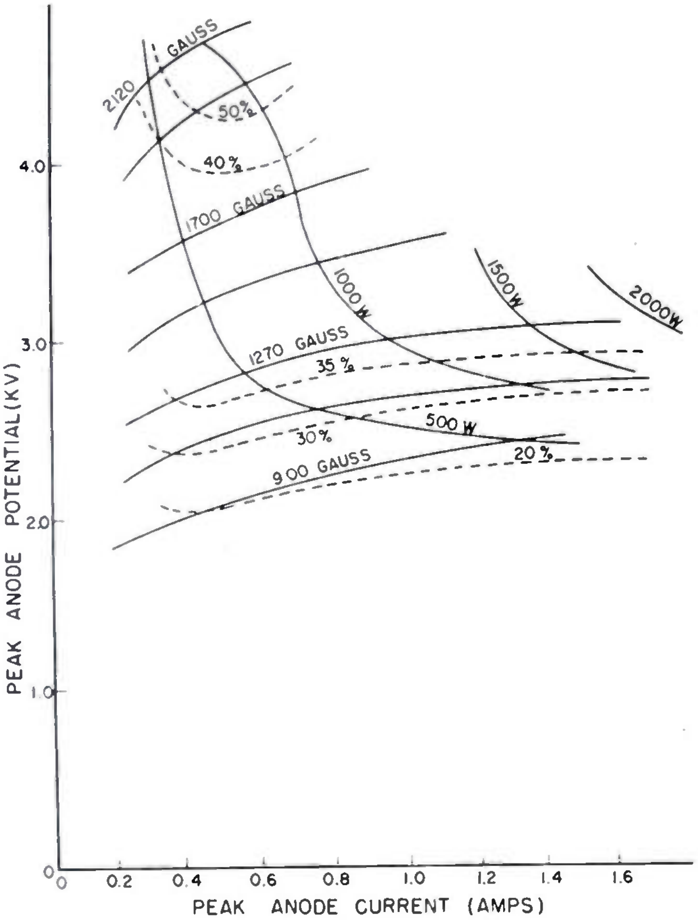

Fig. 5—Performance plot of a single-cavity inverted interdigital magnetron. $\lambda = 12$ cm. Duty cycle 5 per cent.

continuous wave except for anode dissipation and cathode back-bombardment. A power output of 1,500 watts was obtained at 2,500 mc and 51-per cent efficiency. It may be seen that the low-current operation of this tube is approximately the same as that of conventional magnetrons with regard to the cutoff characteristic. This tube also was operated continuous wave with a power output of 300 watts at an over-all efficiency of 35 per cent, and it also was operated on 0.001 duty-cycle pulse with 50-kw power output and 30-per cent over-all efficiency. A Rieke diagram was also taken with 0.05 duty-cycle pulse operation. The frequency instability region and pulling figure were the same as for a conventional magnetron with the same Q_x and line length.

A tube of the type shown in Fig. 2 was also built and operated with 0.0005 duty-cycle pulse. An over-all efficiency of 35 per cent was observed at 0.25-mw output. This tube was operated demountably with no bakeout, and the efficiency and power output were limited by gas generated during the pulse. A power output of 0.42 mw was obtained with an anode voltage of 10,000 volts, 160-amperes anode current, and 2,000 gauss.

Conclusions

It has been demonstrated on more than a half-dozen inverted magnetrons that over-all efficiencies between 25 and 55 per cent and electronic efficiencies between 50 and 85 per cent may be achieved. Circuit efficiency and adequate mode separation may be achieved with inverted magnetron cavities if the interdigital-type resonator is used. Either a single-cavity resonator or a plurality of stacked resonators may be used.

In comparison with the conventional magnetrons, the inverted magnetron structure at a given frequency allows larger cathode emission surfaces to be used. Inverting the cavity structure also provides for greater anode surface since the outer surface of a toroidal cavity structure is greater than its inner surface. This reduces the surface dissipation density on the anode.

Due to the fact that a relatively low anode voltage and high anode current are required, it may be possible to operate a high-power pulse inverted magnetron directly from a gas modulator tube without any pulse transformer.

Most important of all, it has been demonstrated that high electronic efficiency can be achieved in a magnetron interaction space whose curvature is opposite that of the conventional magnetron. These experiments indicate that the electronic efficiency of a planar magnetron, neglecting end effects, is comparable to that of conventional magnetron oscillators. This provides sound scientific basis for designing devices utilizing the planar magnetron interaction space with high-space charge, such as pulse magnetron amplifiers.

Symbol Definitions

All symbols used in this paper and not listed here are standard in the mks system of units.

r_c = radius of center post
λ = cavity-resonant wavelength
r_t = mean radius of tooth structure
α = number of teeth, or segments
h = axial length of the cavity from one end surface to the other
c = total equivalent capacitance between adjacent teeth in the tooth structure per tooth
A = loop area
L = loop inductance
z_0 = characteristic impedance of output line
s = circumferential distance between segment centers
f = frequency of oscillation
m = mass of the electron
e = charge of the electron
d = cathode-anode spacing
V_0 = lowest possible operating anode voltage
B_0 = lowest possible magnetic field for oscillation
$K_1 = 2\pi/\lambda$

Accuracy of Bolometric Power Measurements*

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Summary—When the RF power distribution along a bolometer wire differs from the low-frequency power distribution, the substitution method of measurement may give an error unless certain special conditions are satisfied. These conditions are most closely fulfilled, in practice, by a convectively cooled wire whose length to diameter ratio is very large. The possible error for the case of a Wollaston wire mounted in air at atmospheric pressure is analyzed and compared with that obtained with wires mounted in vacuo. It is shown that the air-mounted Wollaston wire is subject to a smaller error than are the evacuated units and that this advantage increases as the wire length becomes an appreciable fraction of a wavelength. It is concluded that Wollaston wire bolometers, when properly designed and mounted, can be used to measure cw power over a frequency range extending into the millimeter wavelength region with an accuracy approaching that of low-frequency measurements.

I. Substitution Method of Power Measurement

In power measurements at uhf and microwave frequencies, a bolometer is often used to absorb the RF power. The RF power heats the bolometer and produces a change in its resistance. This resistance change then serves as an indication of the RF power absorbed. In the substitution procedure, the RF power W is measured by replacing it with a measurable amount of low-frequency power $\overline{W}$, which produces the same resistance change. The two types of power are then assumed equal to each other.

The equivalence of bolometer resistance when low-frequency power is substituted for RF power may be expressed mathematically as follows:

$$\int_0^L \frac{dx}{\int_0^a 2\pi\rho\sigma(\rho, x)d\rho} = \int_0^L \frac{dx}{\int_0^a 2\pi\rho\bar{\sigma}(\rho, x)d\rho}, \qquad (1)$$

where the bolometer is a wire of radius a and length L, having a dc conductivity function σ with RF power in the bolometer and a dc conductivity function $\bar{\sigma}$ with low-frequency power only in the bolometer. The functional dependence of the conductivities σ and $\bar{\sigma}$ on the radial co-ordinate ρ and axial co-ordinate x is governed by the steady-state distribution of temperature in the bolometer under RF and low-frequency heating, respectively.

The question of the radial temperature distribution in the wire may be quickly disposed of. Gainsborough[1] has shown that the maximum temperature difference in the cross section of the wire, whether under low-frequency or RF heating conditions, must be less than $w/2\pi k$, where w is the power per unit length (watts per cm) and k is the heat conductivity of the metal in watts per cm per degree C. A typical Wollaston wire bolometer having a power input of 60 mw per cm, corresponding to a temperature rise of about 200°C above ambient, will, according to this formula, have a radial temperature variation of less than 0.02°C or less than one part in 10^4. Feenberg[2] has shown that in a tungsten wire dissipating 50 watts at 3,000°C the dc resistance under RF heating differs by only 1/30 per cent from the value under dc heating. Thus, for all practical bolometers, the radial temperature distribution is extremely uniform. Accordingly, the resistance of the wire is a function only of the lengthwise distribution of temperature and (1) may be simplified to

$$\Delta R = \int_0^L \Delta r(x)dx = \int_0^L \Delta\bar{r}(x)dx = \Delta\overline{R}, \qquad (2)$$

where $\Delta r(x)$ is the change in dc resistance per unit length due to RF power W superimposed on some fixed low-frequency bias power in the bolometer and $\Delta\bar{r}(x)$ is the change in dc resistance per unit length due to the replacement of the RF power by an additional amount of bias power $\overline{W}$. In practice, the procedure frequently followed is to retract a portion of the low-frequency bias power after the RF power has been applied so as to maintain a constant bolometer resistance as indicated by a balanced Wheatstone bridge. Equation (2) does not necessarily imply that the total RF power in the bolometer equals the total low-frequency power used to replace it when the resistance change resulting from each of these is the same. That is,

$$\Delta R = \Delta\overline{R}$$

does not necessarily imply that

$$W = \int_0^L w(x)dx = \int_0^L \bar{w}(x)dx = \overline{W}, \qquad (3)$$

where $w(x)$ is the RF power per unit length and $\bar{w}(x)$ is the low-frequency power per unit length which is substituted for it. Clearly, (3) will be satisfied identically when $\Delta R = \Delta\overline{R}$ if the RF and low-frequency power distributions w and $\bar{w}$ are the same. However, if the two distributions are different, it becomes necessary to evaluate separately the two integrals of (3) to determine the error. This error is given by

$$E = \frac{\overline{W} - W}{W}, \qquad (4)$$

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[1] C. F. Gainsborough, "Some sources of error in microwave milliwattmeters," *Jour. IEE* (London), pt. III, pp. 229–238; July, 1948.

[2] E. Feenberg, "The Frequency Dependence of the Power-Resistance Relation in Hot Wire Wattmeters," Sperry Gyroscope Co., Great Neck, L. I., N. Y., Report No. 5220-108; March 10, 1943.

a positive algebraic sign meaning that the power reading is high.

II. Physical Basis of Substitution Errors

In order that a metallic bolometer wire be free of substitution errors, it is sufficient that it satisfy the following criteria:[3]

1. The temperature rise at any point along the wire is proportional to the power dissipation at that point.
2. The proportionality factor is independent of the length co-ordinate.

Since the resistance is a linear function of the temperature, the two criteria stated above define an error-free bolometer whose resistance per unit length is *the same linear function of power density everywhere*, regardless of how the power is distributed. Such an ideal bolometer will have an absolutely uniform distribution of temperature for a uniform distribution of power, the temperature falling abruptly to the fixed ambient temperature prevailing at the ends. Moreover, the point-by-point rise in temperature above ambient will, in general, faithfully reflect the distribution of power along the wire. In addition, the total resistance change of such a wire will be linearly related to the total power input. Conversely, the degree to which a bolometer approaches the above ideal is qualitatively indicated by the uniformity of its temperature distribution under low-frequency bias power heating and by the linearity of its steady-state resistance-power relation.

These two characteristics—uniformity of temperature and resistance-power linearity—depend on the relative importance of the following processes in the cooling of the bolometer: metallic conduction of heat along the wire, flow of heat from the surface of the wire by conduction through the surrounding medium (referred to as *convection*[4] in this paper), and radiation.

If radiation is the predominant bolometer cooling process, then considerable nonlinearity errors are possible when a standing wave of RF power exists on the wire.[1] However, most bolometers used in practice, because of their small diameters (1 to 10μ) and low operating temperatures, have negligible radiation loss and are therefore almost entirely free of error from this source. For example, in the case of a 1 micron diameter Wollaston wire in air, the radiation loss is less than 1 per cent of the convection loss up to the temperature of melting platinum (1,774°C).

A thin bolometer wire mounted in an evacuated envelope is almost entirely cooled by metallic conduction and exhibits a temperature distribution under dc heating conditions which is essentially parabolic rather than constant in character. Such a bolometer is subject to a significant substitution error if the RF power distribution differs appreciably from the uniform dc power distribution which produces the same resistance change. Bleaney,[5] Gainsborough,[1] and Broc[6] have treated this type of bolometer and show that substitution errors as high as 35 per cent are possible if the bolometer is approximately a half wavelength.

A thin wire bolometer mounted in air is predominantly cooled by convection. Except for end effects, the convective cooling along the wire is substantially constant. Moreover, the convective process is characterized by a temperature-power relationship which is nearly linear. Thus, the two criteria stated are approximately satisfied and small errors are therefore to be expected for this case. The remainder of this paper presents a quantitative evaluation of the substitution errors which may occur when a convectively cooled bolometer of the Wollaston wire type is used for power measurements.

III. Substitution Errors in Convectively Cooled Thin Wire Bolometers

The following solution for the substitution errors consists of two distinct parts. In the first part, the error is computed as though the convective cooling were a strictly linear process, that is, the heat loss by convection is assumed as linear in temperature rise above ambient. An error arises from the fact that some metallic conduction of heat takes place along the wire and that the convective cooling through the surrounding air is not strictly radial, particularly near the ends of the wire. In the second part, the heat loss by metallic conduction is neglected and the convective cooling is assumed to be strictly radial but at the same time appropriately nonlinear in the temperature rise. This nonlinearity will also produce an error.

From these two separate solutions, the bounds for the total error can be stated. By this technique, a very difficult nonlinear boundary value problem is simplified so that a useful numerical solution is obtained.

A. Solution of Heat Equation with Linear Convective Cooling

According to the well-known theory of Langmuir,[7] the convective cooling of a thin wire takes place by conduction of heat through a stagnant air sheath of finite diameter surrounding the wire. The diameter of the Langmuir sheath is a function of the wire diameter, being about 1 mm at atmospheric pressure for a 1 micron diameter wire. This is comparable to the length of the wire itself for many Wollaston wire bolometers.

The steady-state heat balance equation takes the following form:

$$- ks \frac{d^2\theta}{dx^2} + \gamma\theta^{\mu} = w(x), \tag{5}$$

[3] H. J. Carlin, "Broadband bolometric measurement of microwave power," *Radio and Telev. News*, Electronic Engineering Edition; July, 1949.

[4] This conduction process constitutes "convection" as defined by Langmuir, the expression "forced convection" being reserved for the cooling of the wire by currents in the surrounding fluid medium.

[5] B. Bleaney, "Radio-Frequency Power Measurements by Bolometer Lamps at Centimeter Wavelengths," *Jour. IEE* (London), pt. III-A, vol. 93, pp. 1378–1382.

[6] J. Broc, "Measurement of low power at centimetric wavelengths," *Onde Elec.*, vol. 30, pp. 108–120; March, 1950 (in French).

[7] I. Langmuir, *Phys. Rev.*, vol. 34, p. 401; 1912.

where the first term on the left-hand side represents the loss of heat per unit length by metallic conduction along the wire and the second term the heat loss per unit length by radial convection (neglecting for the moment any axial flow of heat in the surrounding Langmuir sheath). The right-hand side represents the excitation power input per unit length. The symbols in (5) are defined as follows:

$\theta(x)$ = temperature rise above ambient at point x measured from the midpoint of the wire in degrees C

k = heat conductivity in watts per cm per °C = 0.70 for platinum

s = cross-sectional area of wire in cm^2

γ = convection constant in watts per cm per °C $\cong 3\times 10^{-4}$ for wires 1 to 3 microns in diameter

μ = empirical exponent for convection loss term.

To first approximation, μ may be taken as unity to give a convection loss linear in temperature rise. (The radiation term has been omitted since the discussion is confined to thin wires.) If $w_0(x)$ represents low-frequency (bias) heating, it is given by

$$w_0(x) = i_0^2 r_0(1 + \alpha\theta(x)), \tag{6}$$

where

i_0 = low-frequency bias current in amperes

r_0 = resistance in ohms per cm at ambient temperature

α = temperature coefficient of resistivity in ohms per ohm per °C = 0.0037 for platinum.

For an impressed power as defined by (6), the solution of (5) is

$$\theta(x) = \frac{i_0^2 r_0}{\gamma - \alpha i_0^2 r_0}\left[1 - \frac{\cosh cx}{\cosh cl}\right], \tag{7}$$

where

$$c = \sqrt{\frac{\gamma - \alpha i_0^2 r_0}{ks}} \simeq \sqrt{\frac{\gamma}{ks}}$$

and $l = L/2$, the half length of the wire, and where the ends of the wire as well as the periphery of the cylindrical Langmuir sheath are assumed at ambient temperature. Even for relatively large bias powers, the term $\alpha i_0^2 r_0$ has a small effect on the value of c, and its omission does not significantly affect the calculated bolometer error. Therefore, in all computations, the change in power distribution resulting from the change in bolometer resistance with temperature is disregarded and the approximation shown above for c is used.

Following the discussion of Section I, the fractional substitution error may be defined as the ratio of the difference between the true RF power and retracted bias power to the true RF power. It can be shown that in a linear system this ratio is also given by

$$E = \frac{\Delta R - \Delta \overline{R}'}{\Delta \overline{R}'}, \tag{8}$$

where ΔR and $\Delta \overline{R}$ are the respective changes of the bolometer resistance corresponding to an amount of RF power W and replaced by an equal amount of bias power $\overline{W}'$.

$\Delta \overline{R}'$ can be computed from the temperature distribution which results when the low-frequency replacement power $\overline{W}'$ is applied, with constant power density P_0 per unit length, so that $\overline{W}' = P_0 L$. This temperature distribution is given by (7), as modified by omitting the second-order term, $\alpha i_0^2 r_0$, from the denominator, as already discussed. Similarly, ΔR is obtained from the temperature distribution which exists along the bolometer when the RF power W is introduced. In this case, the distribution of temperature is obtained from the solution of (5) where the forcing function, $w(x)$, is an appropriate RF power distribution. It will be assumed that this power distribution is of the form

$$w(x) = \frac{2P_0}{1 + \sin\phi/\phi}\cos^2(2\pi x/\lambda), \tag{9}$$

where x is the length co-ordinate with the mid-point of the wire as origin and λ is the wavelength of the applied frequency. Equation (9) corresponds to the power resulting from a standing wave of RF current with current loop at the mid-point of the wire. The amplitude factor of (9) is such that

$$W \equiv \int_0^L \frac{2P_0}{1 + \sin\phi/\phi}\cos^2(2\pi x/\lambda) = P_0 L \equiv \overline{W}'. \tag{10}$$

It can be shown[3] that the occurrence of a current maximum at the wire mid-point corresponds to the case of maximum error. Proceeding in the manner outlined above, the substitution error of a bolometer wire, which is cooled both by metallic conduction and purely radial convection, is given as a function of the electrical length of the bolometer, $\phi = 2\pi L/\lambda$, by

$$E = \frac{1 + \dfrac{1}{1 + (\phi/cl)^2}\,\dfrac{\sin\phi/\phi - \dfrac{\cos\phi\tanh cl}{cl}}{1 - \tanh cl/cl}}{1 + \sin\phi/\phi} - 1. \tag{11}$$

In the preceding analysis, curvature of the heat-flow lines in the surrounding Langmuir sheath was neglected, but the metallic conduction loss was included. Actually, this latter term, because of the extreme thinness of the wire, is small in relation to the convection loss. For example, ks/γ, for a 1 micron diameter wire is approximately 0.5×10^{-4}. Accordingly, in the ensuing treatment of the error due to the curvature of the heat-flow lines, the wire will be represented as a filament of zero thickness along which heat is generated with a distribution corresponding to a dc or RF forcing function. The wire is surrounded by a sheath of stagnant air whose diameter is approximately a thousand times the wire diameter. The wire is therefore represented as the axis of a closed cylindrical box, with ends infinitely close to the circular end plates of the box and thermally insulated from them. The box is filled with a homogeneous conducting medium and all walls are at ambient temperature. In representing the wire and its surroundings in this fashion, loss of heat by conduction along the wire, by radiation, and by convective currents in the surrounding medium is neglected; loss by conduction through the air alone is taken into account. Since the thermal problem is described by Laplace's equation in the air medium, an electrostatic analogue to this problem is a line charge situated in a homogeneous dielectric with boundary conditions as shown in Fig. 1. The strength of the heat source per unit length is equivalent to the electric charge per unit length, the temperature rise above ambient to the electrostatic potential, the heat-flow lines to the electric field lines, and the thermally conducting medium to the homogeneous dielectric.

Referring to Fig. 1, it should be noted that the boundary conditions are severe, inasmuch as the entire plane normal to the wire at its ends is fixed at ambient temperature. This results in a greater axial component of heat flow than would be the case if the end plates were not constrained to be at ambient temperature; therefore, the errors computed from this analogue are somewhat pessimistic.

The validity of representing the bolometer wire as a filament rather than as a thin cylinder on the surface of which the charge distribution is placed is easily justified in view of the extreme thinness of the wire. Thus, if a uniform distribution of charge is assumed on the filament, the radial component of the electric displacement at a distance from the axis of 0.001 L (corresponding to the surface of a wire 2 microns in diameter and 1 mm long) is constant within about 1 over 99 per cent of the wire length. This means that so far as the potential (i.e., temperature) distribution on and outside the bolometer wire is concerned, the line charge may be used in place of the charged thin cylinder.

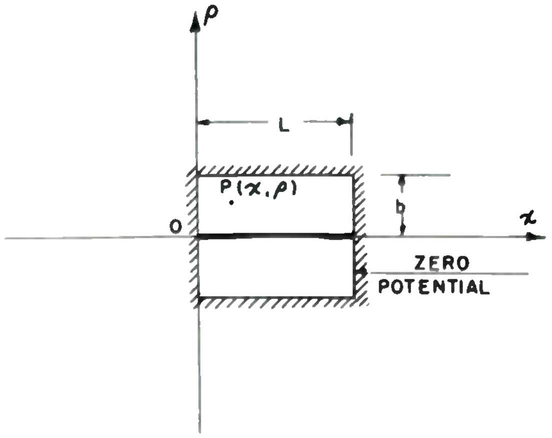

Fig. 1—Electrostatic analogue for an infinitesimally thin heated wire surrounded by a stagnant air sheath. Finite line charge on axis of closed cylinder.

The procedure for evaluating the error by means of the electrostatic analogue parallels exactly the one described in connection with (11). In the present case, the temperature change θ at any point x at a distance "a" from the axis due to an impressed power $w(x')$ along the wire is given by the expression[8]

$$\theta(a, x) = \frac{\pi}{L}\sum_{n=1}^{\infty} \sin\left(\frac{n\pi x}{L}\right) A_n \int_0^L w(x') \sin\left(\frac{n\pi x'}{L}\right) dx', \tag{12}$$

where

x is the axial co-ordinate of the point at which the temperature rise θ is evaluated

x' is the axial co-ordinate of the point at which power $w(x')$ is generated

n is an integer

$$A_n = \frac{j\left[J_0\left(j\frac{n\pi b}{L}\right)H_0^{(1)}\left(j\frac{n\pi a}{L}\right) - H_0^{(1)}\left(j\frac{n\pi b}{L}\right)J_0\left(j\frac{n\pi a}{L}\right)\right]}{J_0\left(j\frac{n\pi b}{L}\right)} \tag{13}$$

b is the radius of the cylindrical box

J_0 and $H_0^{(1)}$ are respectively zero-order Bessel and Hankel functions of the first kind.

The origin of co-ordinates is here taken as one end of the wire instead of the mid-point.

The impressed power, $w(x')$, is either a uniformly distributed low-frequency power $\overline{W}'$ of amount P_0 per unit length, so that $\overline{W}' = P_0 L$, or an equivalent amount of RF power in the form of a standing wave of power, described by the expression

$$w(x') = \frac{2P_0}{1 + \sin\phi/\phi}\cos^2\left(\frac{2\pi x'}{\lambda} - \frac{\phi}{2}\right), \tag{14}$$

where the amplitude factor has been chosen to give equality of RF power W and substitution power $\overline{W}'$ as in (10).

The above distribution of RF power corresponds to a current maximum at the mid-point of the wire and gives the maximum error.

Upon performing the indicated integrations for both the low-frequency and RF cases, the following expression is obtained for the fractional error as a function of the electrical length of the bolometer wire, $\phi = 2\pi L/\lambda$, using (8):

[8] For this solution, thanks are due L. Felsen of the Microwave Research Institute, who obtained it by application of the characteristic Green's function technique. Thanks are also due to L. Sweet of the Microwave Research Institute for his computational assistance. An alternate, but more slowly converging, expression may be obtained by integrating the solution for a point charge in a cylinder, given in by W. R. Smythe, "Static and Dynamic Electricity," p. 174.

$$E = \frac{\displaystyle\sum_1^{\infty} \frac{[1-(-1)^n]^2}{n^2} A_n - \frac{1}{1+\sin\phi/\phi} \sum_1^{\infty} \frac{[1-(-1)^n]^2}{n} A_n B_n}{\displaystyle\sum_1^{\infty} \frac{[1-(-1)^n]^2}{n^2} A_n}, \tag{15}$$

where

$$B_n = \frac{1}{n} + \frac{\cos\phi}{2}\left[\frac{1}{n-2\phi/\pi} + \frac{1}{n+2\phi/\pi}\right]. \tag{16}$$

In Fig. 2 are plotted curves of bolometer error as a function of L/λ. Curve 1 gives the calculated error when metallic conduction of heat along the wire is taken into account, ignoring the axial component of heat-flow lines in the surrounding air (11). The calculated error when both the axial and radial components of the heat-flow lines in the surrounding air are considered (15), but neglecting any metallic conduction of heat almost coincides with Curve 1. The sum of these two errors is plotted as Curve 2. This curve, therefore, constitutes an upper bound on the total error which may result from the simultaneous operation of both mechanisms of longitudinal heat conduction, and it is reasonable to suppose that the actual error will be somewhat less than the arithmetic sum of the two. Curve 3 represents the error (dependent only on electrical length) in a bolometer wire cooled entirely by metallic conduction (e.g., a bolometer mounted in an evacuated envelope) as given by Bleaney.[5] Broc's[9] curve agrees with Bleaney's for values of $L \leqq 0.4\lambda$, but gives somewhat smaller errors for $L > 0.4\lambda$.

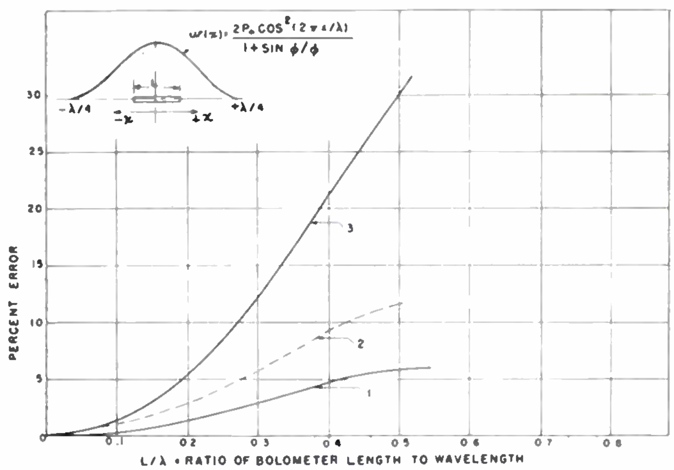

Fig. 2—Curves of error for a cosine-square type RF power distribution. Curve 1 represents the error due to metallic conduction of heat along the wire, but with the axial flow of heat in surrounding air neglected. Curve 1 is practically indistinguishable from the curve of error due to axial flow of heat in surrounding air, but with metallic conduction neglected. Curve 2 gives the arithmetic sum of the above two types of error, representing an upper bound for the error due to total lengthwise conduction of heat. Curve 3 represents the error for a bolometer cooled entirely by metallic conduction. Curves 1 and 2 are for a Wollaston wire, 1 micron in diameter and 1.5 mm long, in air.

[9] J. Broc, *loc. cit.* This is because Broc used Wollaston wires in incompletely evacuated envelopes (10^{-2} to 10^{-4} mm Hg), where a finite heat loss by convection could still occur. Thus, the value of cl for a 1-micron diameter wire, 3 mm in length (as used by Broc), is about 2.5. This value, when used in (11), results in a 25-per cent error for $L = 0.5\lambda$ as compared with a 30-per cent error if convection loss is entirely neglected. For values of $L \leqq 0.4\lambda$ the small amount of convective cooling does not materially affect the error.

Curves 1 and 2 of Fig. 2 were obtained for a Wollaston wire of representative dimensions, 1 micron in diameter and 1.5 mm in length. The characteristic constant, cl, of such a bolometer is approximately 17 and determines the error shown in Curve 1. The temperature distribution of this bolometer under low-frequency bias conditions is shown in Fig. 3. This curve was computed from (7) for the case of metallic conduction and radial convection. The temperature distribution computed from (12), which takes into account axial heat flow, is similar in character. The temperature distribution in a bolometer cooled only by metallic conduction is shown for purposes of comparison.

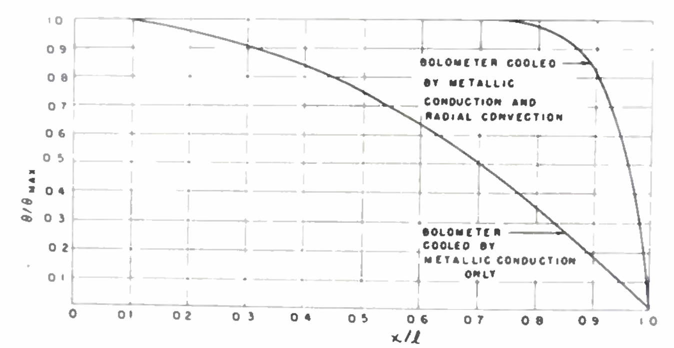

Fig. 3—Calculated temperature distributions for uniformly heated bolometer wires. Fraction of maximum temperature rise above ambient plotted against length co-ordinate expressed as a fraction of half-length l. Upper curve computed for a Wollaston wire, 1 micron in diameter and 1.5 mm long, in air, $cl = 17$.

B. Solution of Heat Equation with Nonlinear Convective Cooling, Neglecting Conduction

It is well known from empirical data that Wollaston wire bolometers exhibit a resistance-power characteristic which follows an exponential law of the form

$$\Delta\overline{R} = A\overline{W}^{\nu} \tag{17}$$

for ranges of power which approach the burn-out value. In this equation $\Delta\overline{R}$ is the total dc resistance change produced in the bolometer by the heating effect of an amount of bias power $\overline{W}$, while A and ν are characteristic constants of the bolometer.

The above equation may be very simply derived from the bolometer heat-balance equation by recognizing that the predominant heat loss takes place by convection through the surrounding air. Thus, dropping the metallic conduction term in (5), but retaining the exponent μ in the convection term, the heat-balance equation may be written as

$$\gamma\theta^{\mu} = w, \tag{18}$$

assuming also that the convective loss is constant along the bolometer, an assumption which is justified by the nearly constant temperature along the wire under bias conditions, as shown in Fig. 3.

In general, w is a function of the length co-ordinate, so that the total resistance change will depend on how the power w is distributed. Solving for θ in (18) and remembering that

$$r = r_0(1 + \alpha\theta), \tag{19}$$

the total resistance change becomes

$$\Delta R = \int_0^L \alpha r_0 \left(\frac{w}{\gamma}\right)^{1/\mu} dx. \tag{20}$$

For a low-frequency bias power of $\bar{w}$ per unit length, (20) becomes

$$\Delta \bar{R} = \alpha R_0 \left(\frac{\bar{W}}{\gamma L}\right)^{1/\mu}, \tag{21}$$

where $\bar{W} = \bar{w}L$, the total bias power. On comparing (21) with (17), the following identifications can be made:

$$\nu = 1/\mu \tag{22}$$
$$A = \alpha R_0/(\gamma L)^{1/\mu}.$$

Equation (22) affords a convenient means for evaluating μ and γ. The bolometer static resistance-power characteristic is plotted on log-log paper, as has been done for two different bolometers in Fig. 4, and the exponent ν and constant A are directly obtained from the plot. The excellent straight-line fit of the data in Fig. 4 is in itself

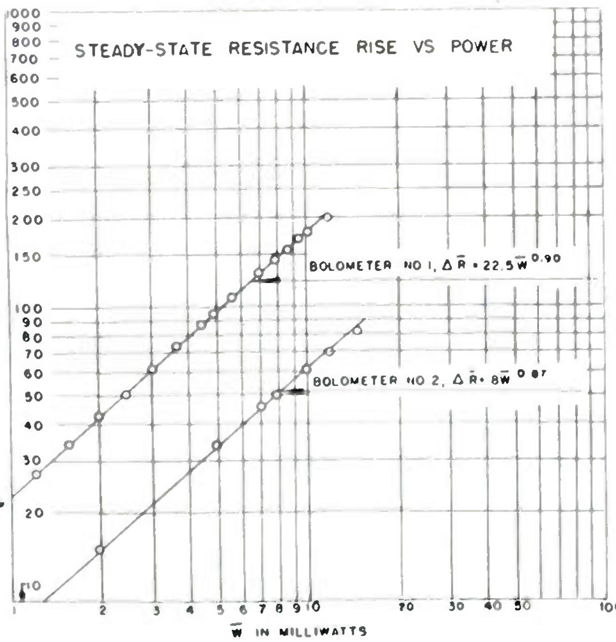

Fig. 4—Steady-state resistance-power characteristics for two different Wollaston wire bolometers.

good experimental evidence that Wollaston wires mounted in air at atmospheric pressure are mainly cooled by convection. Table I compares the values of μ and γ as obtained from Fig. 4 with the values to be expected from the data of Langmuir.[7]

TABLE I

	μ		γ	
	From Langmuir's data	From static characteristic	From Langmuir's data	From static characteristic
Bolometer 1	1.21	1.11	0.9×10^{-4}	2.4×10^{-4}
Bolometer 2	1.21	1.15	1.0×10^{-4}	1.4×10^{-4}

The experimental values of γ agree as to order of magnitude, with the calculated values being 1.4 to 2.6 times larger than the latter. This is in part due to the deviation of the observed value of μ from Langmuir's μ, (column 1) and to the fact that Langmuir's data was obtained for wires much thicker than Wollaston wires and much longer in relation to the diameter of the stagnant air sheath surrounding the wire which is postulated by Langmuir.

As has already been pointed out, the fact that μ deviates from unity is responsible for a type of error which has been referred to as arising from bolometer "resistance-power nonlinearity." If the RF power distribution is nonuniform, an error results which depends on the ratio of RF power to total bias power as well as on the degree of nonuniformity of the RF power distribution. This follows from the fact that a point on the bolometer where the power density is greater than the average loses proportionately more heat by convection than it would if the power density were less.

Applying the definition of (4), the error has been calculated for two nonuniform power distributions, one triangular, the other rectangular, in order to obtain a quantitative notion of the relative magnitude of the nonlinearity error as compared with the errors discussed in the preceding section. These two distributions are shown in Fig. 5, superimposed on the residual bias power

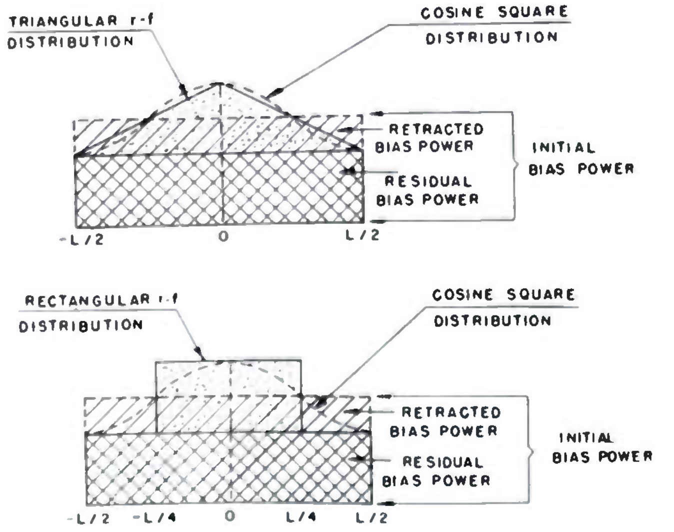

Fig. 5—Bias and assumed RF power distributions in bolometer used to compute the error due to resistance-power nonlinearity.

after withdrawal of the bias power required to keep the total dc bolometer resistance constant. It is seen that a cosine-square type of power distribution for a one-half wavelength bolometer lies intermediate between the triangular and rectangular distributions, being very close to the triangular. Two different values of ν were used, 0.9 and 0.8, although the former is more typical for Wollaston wires than the latter. The error is negative for these distributions and the power reading will be low. Upon performing the necessary operations in (20) the error (4) may be given by the following expressions:

$$E\,(\text{triangular}) = \frac{2}{q}\left\{\left[\frac{(1+q)^{\nu+1}-1}{q(\nu+1)}\right]^{1/\nu}-1\right\}-1 \qquad (23)$$

$$E\,(\text{rectangular}) = \frac{2}{q}\left\{\left[\frac{(1+q)^{\nu+1}+1}{2}\right]^{1/\nu}-1\right\}-1, \qquad (24)$$

where q is twice the ratio of the RF power W to the residual bias power.

These errors are shown in Figs. 6 and 7 plotted against M, the ratio of RF power to the initial bias power. In practice, this ratio rarely exceeds 1/3. For a bolometer with $\nu=0.9$ (which is a typical value) and $M=1/3$, the indicated error is less than 1 per cent for the triangular distribution and less than 2 per cent for the rectangular distribution. These two distributions correspond to a half wavelength bolometer with a full variation of RF power along the wire, and therefore the indicated errors are extreme. For smaller electrical lengths, the nonlinearity errors are negligible compared to those resulting from the lengthwise flow of heat which were discussed in the previous section and shown in Fig. 2.

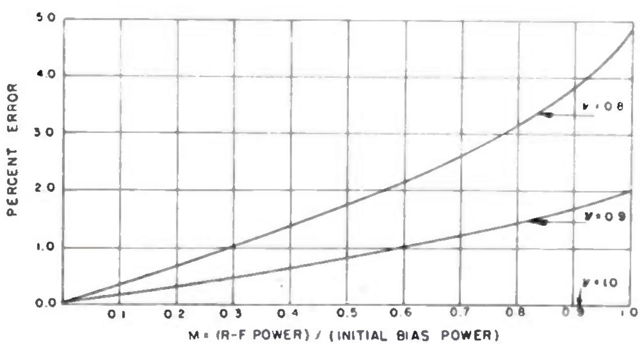

Fig. 6—Calculated resistance-power nonlinearity error as a function of ratio of RF power to initial bias power for triangular distribution of Fig. 5.

Conclusions

It has been shown that the portion of the bolometer substitution error caused by resistance-power nonlinearity is small compared with the error caused by the lengthwise flow of heat. For thin wires (1 to 3 microns in diameter), the portion of the error arising from axial flow of heat in the air surrounding the wire is essentially a function only of bolometer length. Curve 1 of Fig. 2 represents this contribution to the error for a thin wire 1 mm long. Longer wires will give even smaller errors.

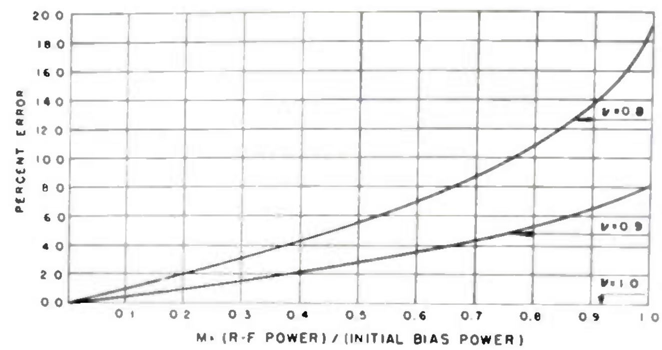

Fig. 7—Calculated resistance-power nonlinearity error as a function of ratio or RF power to initial bias power for rectangular distribution of Fig. 5.

Fig. 8 summarizes the error which is due to metallic conduction and radial heat flow for a bolometer mounted in air at atmospheric pressure. The curve of Fig. 8 relates the bolometer length, expressed as a fraction of a wavelength, to L/d, the ratio of bolometer length to diameter, when the error due to metallic conduction and radial heat flow is equal to or less than 2 per cent. If the bolometer length is an appreciable fraction of a wavelength, a bolometer mounted in an evacuated envelope has a substantially greater error than a convectively cooled element of similar dimensions.

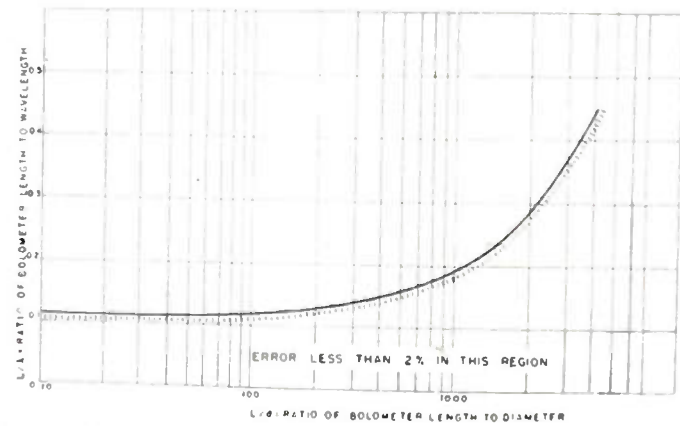

Fig. 8—Maximum permissible L/λ for varying L/d to limit error arising from metallic conduction and radial convection to no more than 2 per cent. Computed for Wollaston wires, 1 to 3 microns in diameter.

Based on the analysis presented in this paper, it is concluded that Wollaston wire bolometers, when properly designed and mounted, afford a means of measuring cw power, over a frequency range extending to the millimeter wavelength region, with an accuracy approaching that of low-frequency measurements.

Identification of Tornadoes by Observation of Waveform Atmospherics*

HERBERT L. JONES†, SENIOR MEMBER, IRE AND PHILIP N. HESS‡, ASSOCIATE, IRE

Summary—**Research on the characteristics of atmospherics peculiar to tornadoes has been in progress at the Oklahoma Agricultural and Mechanical College since 1947. It has been discovered that high-energy thunderstorms which develop into tornadoes generate discharges with a preponderance of frequencies in the 200- to-400-kc band. The number of these discharges increases as tornado time is approached.**

I. Introduction

THE RESEARCH on tornado identification and tracking was undertaken at Oklahoma Agricultural and Mechanical College in the hope of developing some method by which cognizance of such storms might be established in sufficient time to permit a warning to be effective. From the time that the project was started in 1947, the author made an effort to obtain accurate information concerning the characteristics of the tornado type of storm.

As such evidence accumulated, it became extremely probable that the atmospherics resulting from tornadoes would possess characteristics which would be different from atmospherics resulting in other types of storms. This conclusion was supported by other known conditions relevant to this type of storm. The available energy in an incipient tornado type of cumulus cloud must be considerably greater than that for an ordinary type of cumulus cloud from which a thunderstorm develops. The energy producing the initial whirl, and permiting the development of the characteristic funnel, must necessarily be enormous. The updraft in the inner chimney can be conceived as made up of air currents of considerably higher velocities, both before and during the formation and progress of the funnel, with a resulting increased rate of separation of electrical charges, which in turn suggests higher electrical potentials, more energy to dissipate in each lightning discharge, and a definite increase in the number of strokes.

For an effective study of atmospheric wave shapes, it was necessary to have linear amplification over a broad range of frequencies. In order to conserve film, it was considered essential to have a camera that would be operated automatically by the incoming atmospheric.

It has been noted that the great majority of lightning strokes in the vicinity of tornado cloud formations occur in a small area which includes the tornado funnel. Since charge flows along an ionized air column between the cloud and earth, these lightning strokes may be thought of as vertical, grounded radiators of electromagnetic energy.

Observation has shown that the radiation from lightning strokes contains frequencies ranging from visible light to the very low radio frequencies. Therefore, any analysis of the electromagnetic wave at a distance from the radiating element must be considered on a single-frequency basis.

When Sommerfeld's equation is used to determine the frequency range of the atmospheric detection equipment, it is found that the various frequencies will be attenuated at different rates. The low frequencies diminish at a rate proportional to the reciprocal of the distance, while the higher frequencies diminish at a rate proportional to the reciprocal of the distance squared. Fig. 1

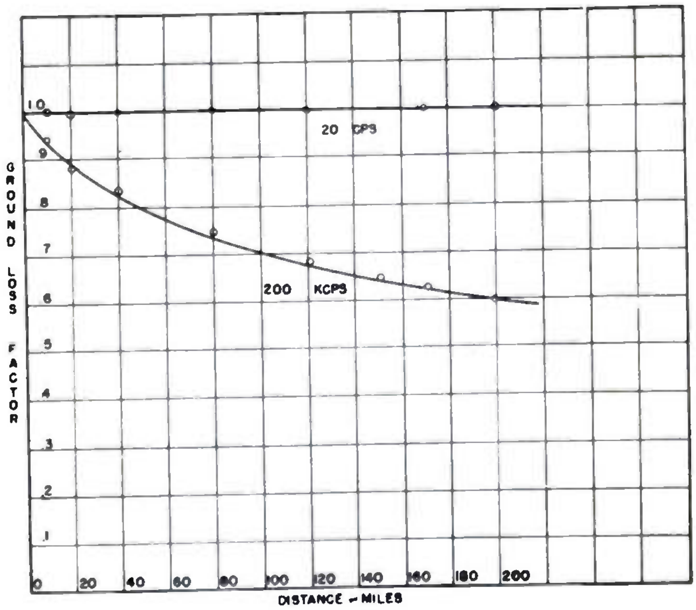

Fig. 1—Ground loss factor versus distance.

is a plot of the ground-loss factor versus distance for a low-frequency wave and a medium-frequency wave. This fact has a definite bearing on the type of detection equipment necessary, as will be shown later.

II. The Atmospheric Detection Apparatus

The atmospheric detection apparatus, as shown in Fig. 2, consists of a detecting element, an amplifying

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element, a recording device, and an instantaneous direction finder. Since these elements have different purposes, they will be discussed separately.

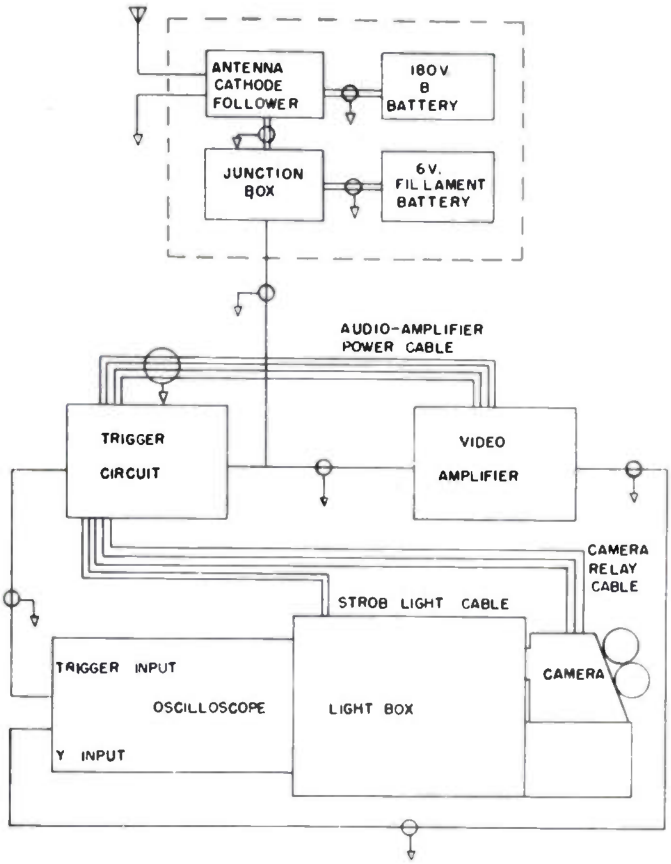

Fig. 2—Cabling diagram for atmospheric detector.

A. The Detecting Element

The detecting element consists of a vertical antenna coupled to a cathode follower. The cathode follower is designed to have a very large input impedance to match the high impedance of the short antenna. This impedance match is necessary to prevent attenuation and phase shift at the lower frequencies. It was originally thought that compensation amplifiers would be needed to aid the cathode follower in obtaining a flat response to the input signal at the lower frequency limit. However, the analysis of the arriving wave showed that the various frequency components under consideration would be attenuated by varying amounts. Thus a flat response would still fail to present an accurate waveform of the radiation from the lightning stroke. This is of slight importance since a difference in waveforms is being sought and all lightning is subject to this varying attenuation rate.

B. The Amplifying Element

Since compensation amplifiers were found to be unnecessary, the amplifying element consists merely of a one-stage video amplifier of conventional design, and the vertical deflecting amplifiers of the recording oscilloscope. Both of the amplifier sections have a flat response between the limits of 40 cps and 2 mc. The output of the video amplifier is fed to the oscilloscope and to the trigger circuit through a 50-ohm coaxial cable.

C. The Recording Device

The recording system consists of an oscilloscope, an automatic camera, and the necessary control circuits. Associated with the recording system is a trigger circuit whose function is to initiate the sweep on the recording oscilloscope, unblank the direction-finder cathode-ray tube, flash a strob light to illuminate a clock and a date device, and operate the automatic camera.

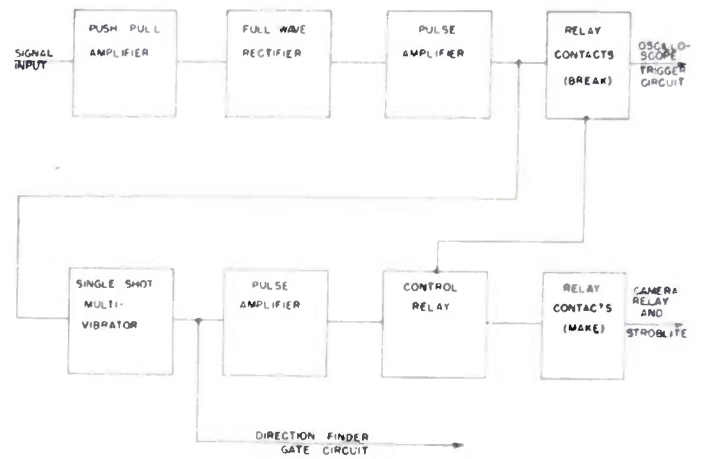

Fig. 3—Trigger circuit.

The trigger circuit is shown in Fig. 3. It consists of a push-pull input amplifier coupled to a duodiode so that a negative pulse will appear at the diode output regardless of the polarity of the incoming atmospheric. This circuit was included because the polarity of the incoming tornado atmospheric may be either positive or negative. This pulse is amplified and used to trigger a single-shot multivibrator. Square waves from the multivibrator are used to unblank the direction-finder cathode-ray tube and to cause a relay tube to conduct, closing a double pole, double-throw relay. These square waves have a period of approximately 500 msec.

The relay contacts are used to open the circuit to the recording oscilloscope, to apply voltage to the strob light, to ground the input to the multivibrator, and to close the camera relay. The first operation is necessary to prevent other atmospherics from interrupting the multivibrator cycle and thus interfering with the correct operation of the automatic recording camera.

The automatic recording camera is a modified 35-mm motion-picture camera with an f2 lens fixed-focused at 2 feet. The film advance mechanism is modified to complete one cycle of operation, that is, advance the film one frame when actuated by a rotary solenoid. In turn, the rotary solenoid is controlled by the aforementioned camera relay. Thus, an arriving atmospheric will trigger the multivibrator which will cause the camera relay to close and remain closed for 500 msec; a period of time long enough to allow the camera rotary solenoid to complete one cycle and advance the film one frame. The inertia effect of the camera mechanism is sufficient to allow the waveform to be photographed and the oscilloscope to go dark before the film begins to advance. As a result there will be no blurring of the image.

The camera lens is enclosed in a light-tight box with the oscilloscope, the clock, the date device, and the strob light. Since the scope sweep and the strob light are triggered only by the arriving atmospheric, no shutter on the camera is necessary.

The schematic diagram of the camera and strob light control circuits is shown in Fig. 4. In order to show all of the atmospheric waveform, a small delay line was included between the video amplifier and the vertical deflection amplifiers in the oscilloscope. This delay compensates for the delay in the sweep triggering caused by the trigger circuits.

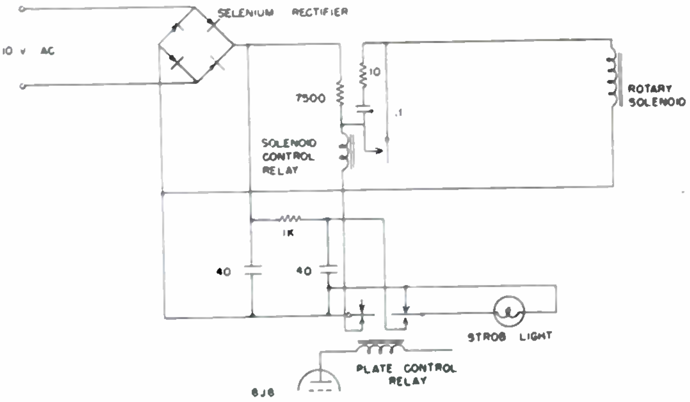

Fig. 4—Relay circuit.

III. Results for Season of 1950

At approximately 6:20 P.M., on June 9, 1950, a tornado funnel was reported about 60 miles west of Stillwater, Okla. The detection equipment had been in operation several hours before and after the critical time. It was noted that the triggering of the equipment was very rapid at this time, in fact, almost continuous. The trigger volume control was set at a very low value. Representative waveforms of the storm are shown in sequence in the photographs of Figs. 5 and 6. It was noticed in the middle of the afternoon, by visual observation, that an occasional high-frequency atmospheric of high amplitude would show up. The number of these high-frequency atmospherics increased as the critical time of 6:20 P.M. was approached. By 7:00 P.M., the number of these high-frequency atmospherics had decreased markedly and only showed up occasionally from 10:00 P.M. until midnight. It was estimated that for a 25-minute period, while the funnel was active, the ratio of high-frequency atmospherics to ordinary atmospherics was about 1-to-1, while before and after the time of

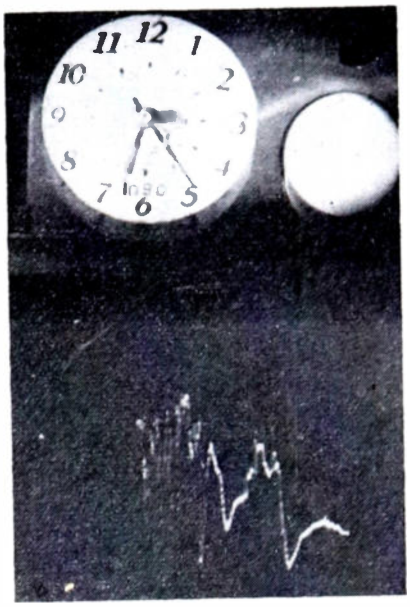

Fig. 5—Tornado, June 9, 1950, cro sweep 20k.

activity of the funnel, it was 1-to-15 or 1-to-20. This condition also held for the next afternoon, June 10, when a whirl started to form over the tracking station. Observers from the town said that the funnel actually came

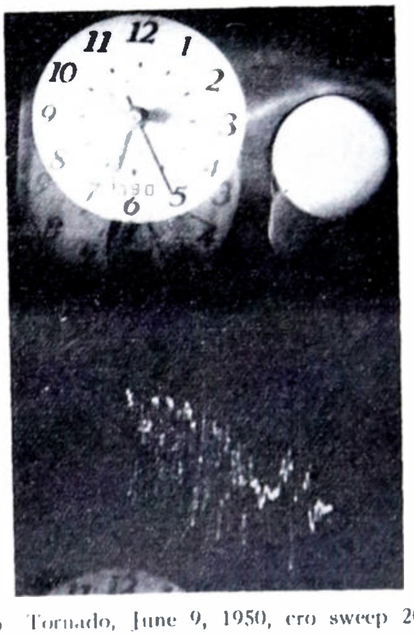

Fig. 6—Tornado, June 9, 1950, cro sweep 20k.

down about a fourth of the way. Again the relay went wild, and the preponderance of the high-frequency components was noticeable. Fig. 7 was taken during the progress of this storm. Figs. 8 and 9 are pictures of atmospherics taken during a thunderstorm which occurred a number of hours prior to the tornado of June 9. These waveforms are typical of records taken during a large number of ordinary thunderstorms. It is significant that no high-frequency components appear during thunderstorm conditions. The authors have spent

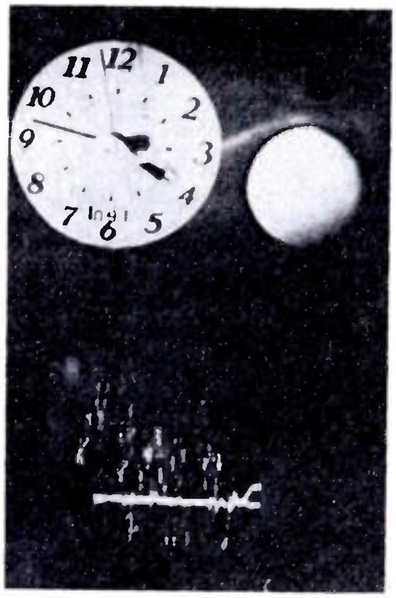

Fig. 7—Tornado, June 10, 1951, cro sweep 20k.

many hours at the scope during thunderstorm activity, and have yet to observe a high-frequency component during this type of storm. It is significant that the high-frequency components appeared only during periods of tornado activity. Observations during the relatively inactive tornado season of 1951 confirmed the results obtained in 1950.

IV. Program for 1952

The records for tornado activity in 1950 and 1951 cover only a few storms, most of these being of low intensity. Conclusions drawn from these records must be substantiated by future results. Apparently the high-energy conditions that exist as the tornado situation builds up are manifested by the high-frequency components. Just when these high-energy levels are first attained depends on the meteorological conditions. During the season of 1952, it is planned to start observations at the first sign of thunderstorm activity, as predicted by meteorological methods, and to continue the observations until the situation is terminated. In this way there will be obtained a complete record of the atmospheric activity for a number of thunderstorms, and, it is hoped, for a number of tornadoes. These results can then be correlated with meteorological observations and should give a better understanding of tornado phenomena.

It would be interesting to know just how the violent high-frequency activity is generated during the existence of the funnel. It has been suggested by one authority that there may be a special type of generator in the whirl itself. This is a problem for the future which will necessitate special equipment in a portable laboratory, possibly airborne.

A new approach to the problem of tornado tracking will be attempted during the tornado season of 1952. The United States Signal Corps has provided a 3-cm radar and atmospheric direction finder to augment the present equipment. It is expected that these additions will contribute materially to the final solution.

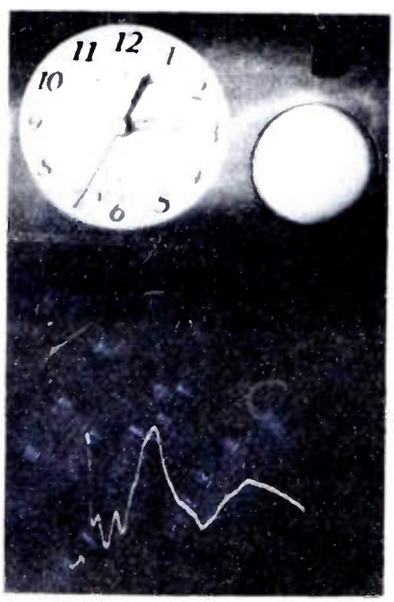

Fig. 8—Thunderstorm, June 9, 1950, cro sweep 20k.

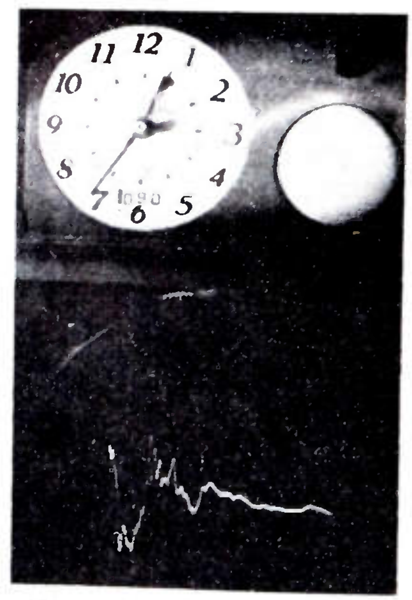

Fig. 9—Thunderstorm, June 9, 1950, cro sweep 20k.

Cosmic Radio Noise Intensities in the VHF Band*

H. V. COTTONY†, SENIOR MEMBER, IRE AND J. R. JOHLER†, ASSOCIATE, IRE

Summary—During 1948 and 1949, the National Bureau of Standards conducted continuous, broad-directivity measurements of the cosmic radio noise intensities at frequencies between 25 and 110 mc. Their purpose was to evaluate the importance of this noise from the standpoint of its interference with radio communication. The results show a regular daily variation in noise corresponding to the movement of the principal sources of cosmic radio noise across the antenna receiving pattern. This normal cosmic noise intensity pattern was found to be constant within the limits of the accuracy of the measurements. It was found convenient to present the results in terms of daily maxima and minima which bracketed the daily variations. No measurable change in these limits was observed in the course of these measurements.

Besides the normal cosmic radio noise, periods of abnormal high noise levels, generally associated with periods of unusual solar activity, were observed and recorded.

I. Introduction

At frequencies below approximately 30 mc, at which long-distance radio communication is normally carried on, terrestrial radio noise is likely to determine the minimum useful signal strengths. The terrestrial radio noise or "static" is generated in the tropical thunderstorm areas and propagated by ionospheric transmission. At the upper end of the hf band where dependable ionospheric propagation ceases, the terrestrial radio noise intensity rapidly decreases and seems to disappear altogether. However, radio noise emanating from extra-terrestrial sources, known as "cosmic radio noise," constitutes one of the limiting factors for radio communication in the upper portion of hf and in the vhf bands. This type of radio noise was identified by Jansky in 1931[1,2] as a characteristic hissing noise apparently originating at a fixed point in space near the center of our galaxy. Subsequently, a number of other observers, notably Reber,[3,4,5,6,7] investigated the distribution of cosmic noise with frequency and direction in space. Fig. 1 (on the following page) is a sky map showing the contours of noise intensities from the different portions of the sky as determined by Reber using directive receiving equipment at 160 mc.

In addition, Southworth[8] and Reber, independently, found that the sun itself is a radiator of noise in the radio-frequency spectrum. The intensity of this noise is considerably in excess of that to be expected on the basis of thermal radiation by a black body at the temperature of the sun's surface (6,000°K). By convention, the term "cosmic radio noise" includes both the solar radio noise originating in the sun and the galactic radio noise which arrives from interstellar space.

Since the date of Reber's early measurements, considerable work has been done on the astronomical aspects of these phenomena, both in the field of physical measurement of noise intensities and in the realm of speculation as to the nature and the character of the noise itself. The references 9 to 13, 15, 16, and 21 give a representative sample of the work accomplished to date. The importance of the cosmic noise to radio communication was discussed in some detail by Norton.[14]

In order to investigate the diurnal, seasonal, and frequency characteristics of cosmic radio noise as it would affect the operation of hf and vhf radio-communication systems, a program of measurements was initiated at the National Bureau of Standards in 1946. During 1947, some preliminary measurements were performed and reported upon by Herbstreit and Johler.[15,16] Continuous measurements were begun in March, 1948. A description of the work, including equipment and some data, was presented by Johler at URSI-IRE meetings on May 3, 1948 and November 1, 1949 in Washington, D. C.

This paper presents a more complete description of the equipment than heretofore presented, the data on cosmic noise collected to January 1, 1950, and a discussion of the results obtained from these data.

* Decimal classification: R113.413. Original manuscript received by the Institute, May 23, 1951; revised manuscript received, April 30, 1952.

† National Bureau of Standards, Washington 25, D. C.

[1] K. G. Jansky, "Directional studies of atmospherics at high frequencies," Proc. I.R.E., vol. 20, pp. 1920–1932; December, 1932.

[2] K. G. Jansky, "Electrical disturbances apparently of extraterrestrial origin," Proc. I.R.E., vol. 21, pp. 1387–1398; October, 1933.

[3] G. Reber, "Cosmic static," Proc. I.R.E., vol. 28, pp. 68–70; February, 1940.

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[5] G. Reber, "Cosmic static," *Astrophys. Jour.*, vol. 100, pp. 279–287; November, 1944.

[6] G. Reber, "Solar radiation at 480 mc/sec," *Nature*, vol. 158, p. 945; December 28, 1946.

[7] G. Reber, "Cosmic static," Proc. I.R.E., vol. 36, pp. 1215–1218; October, 1948.

[8] G. C. Southworth, "Microwave radiation from the sun," *Jour. Frankl. Inst.*, vol. 239, pp. 285–297; April, 1945.

[9] J. G. Bolton, "Discrete sources of galactic radio frequency noise," *Nature*, vol. 162, pp. 141–142; 1948.

[10] J. S. Hey, S. J. Parsons, and J. W. Phillips, "An investigation of galactic radiation in the radio spectrum," *Proc. Royal Soc. A.*, vol. 192, pp. 425–445; 1948.

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[12] M. Ryle and D. D. Vonberg, "Solar radiation on 175 mc," *Nature*, vol. 158, p. 339; September 7, 1946.

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[14] K. A. Norton, "Propagation in the FM Broadcast Band," "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. 1; 1948.

[15] J. W. Herbstreit and J. R. Johler, "Frequency variation of the intensity of cosmic radio noise," *Nature*, vol. 161, p. 515; April 3, 1948.

[16] J. W. Herbstreit, "Cosmic Radio Noise," "Advances in Electronics," Academic Press Inc., New York, N. Y., vol. 1, 1948.

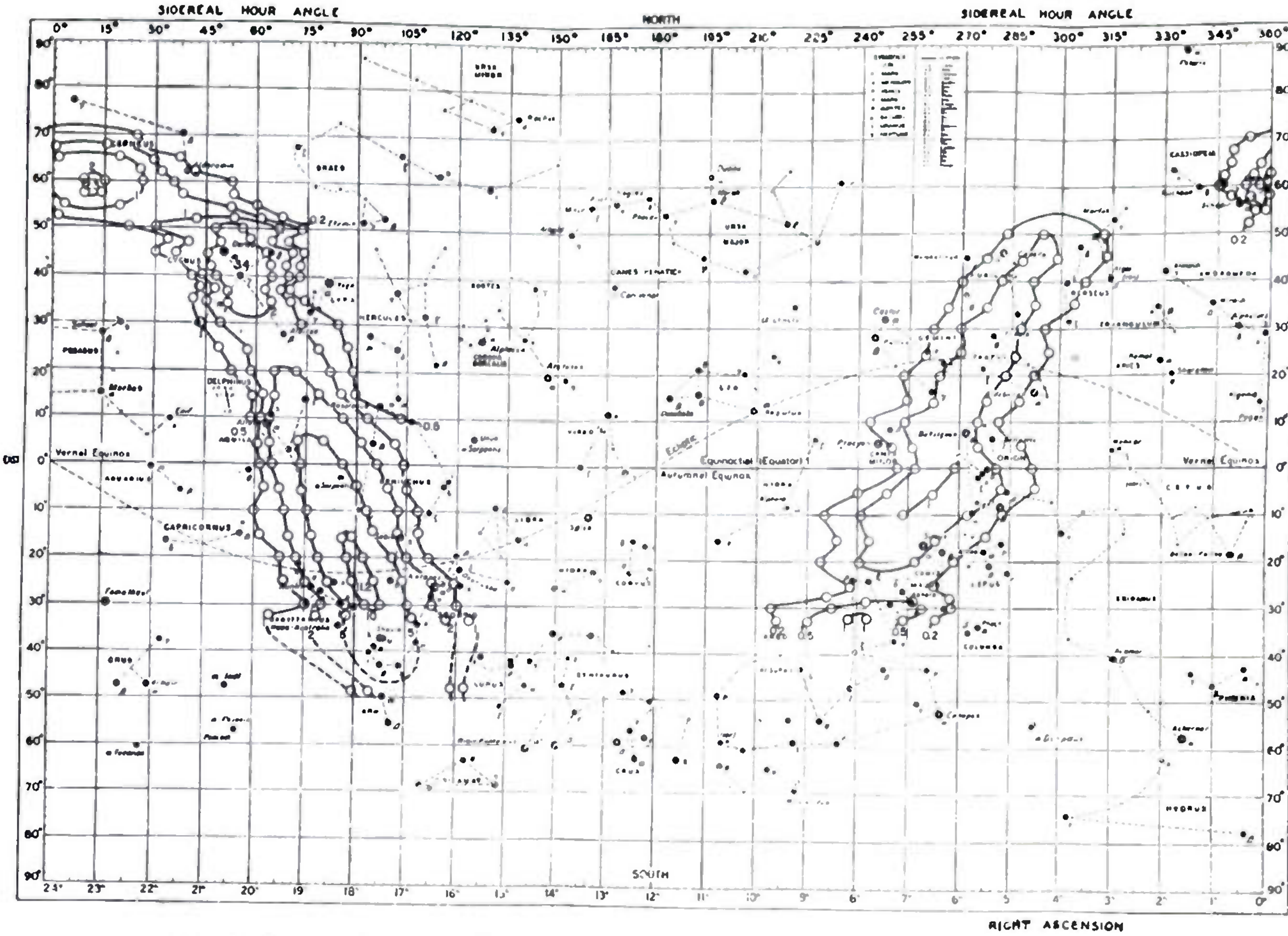

Fig. 1—Sky map showing the contours of cosmic radio noise observed by Grote Reber on 160 mc.

II. Instrumentation

The design of the receiving equipment for cosmic radio noise measurements presents special requirements: (1) high gain, (2) low internal noise, and (3) high degree of gain stability. High gain is required because the cosmic radio noise field strengths are relatively low compared with the normal atmospheric radio noise values, and also because the cosmic noise is measured at higher frequencies where the antenna delivers lower power to the receiver. Low internal noise is a consideration because the voltages to be measured are lower and, with the present knowledge of receiver design, the internal noise rapidly increases with frequency. Both of these requirements were met by the use of special two-stage preamplifiers in conjunction with modified commercial receivers. High degree of gain stability was necessary in order to obtain the desired degree of accuracy in the results, of the order of 1 db. This was met by employing well-regulated power supplies and by housing the equipment in a shelter the temperature of which was maintained constant to within $\pm\frac{1}{2}$°F.

The antennas used for measurements were half-wave, horizontal dipoles one-quarter wavelength above ground oriented in the east-west direction. In order to permit a direct comparison between the results obtained at different frequencies, all antennas were erected and oriented in an identical manner.

Fig. 2 is a block diagram which shows the interconnections between the major components of the equipment. The preamplifier-converter units include two stages of preamplification using the cascode circuit. Using this circuit, it was found possible to obtain a noise figure of 2 at 110 mc and proportionately better values at lower frequencies.

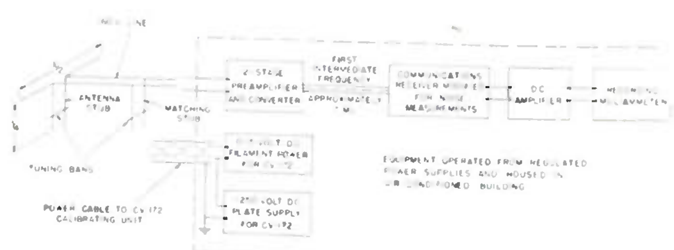

Fig. 2—Block diagram showing the major components of one receiving system and the calibrating equipment.

Fig. 3 illustrates the principle of calibration. A type CV-172 noise diode is mounted across the terminals of each dipole. The length of each dipole is adjusted for half-wave resonance. The length of the metal stub is adjusted to tune out the capacitance of the diode and the terminals. When calibrating, the dipole elements are removed and replaced by a resistor, R_a, equal in magnitude to the radiation resistance of the dipole. The space current I_s through the diode is varied by varying the filament current. The diode is then a source of shot noise current i_n which flows through the resistor R_a. The value of the shot noise current in amperes is given by the relationship

$$\overline{i_n^2} = 2eI_s\Delta f, \tag{1}$$

where e is the charge of the electron, 1.602×10^{-19} coulomb, and Δf is the bandwidth over which measurements are made in cycles per second.

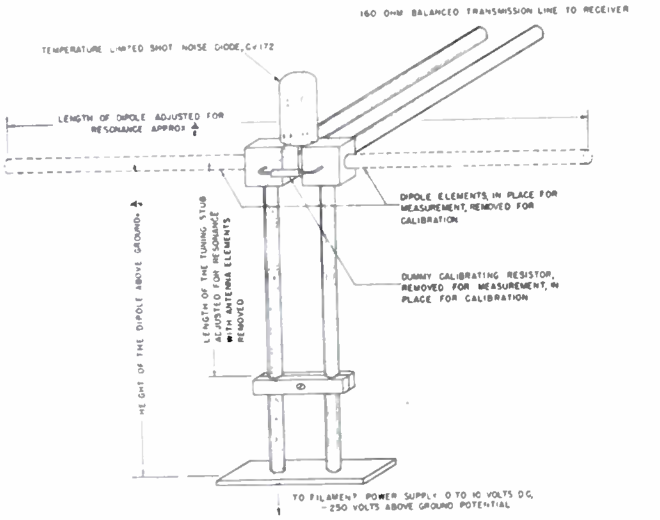

Fig. 3—A simplified view of the dipole antenna used to measure cosmic radio noise intensity, showing the method of absolute calibration employed. The diagram is not to scale. Impedance matching stub is not shown.

The open-circuit voltage developed across the terminals of the dummy antenna resistor is then equal to V given in volts by the relationship

$$\overline{V^2} = 4kTR_a\Delta f + 2eI_sR_a^2\Delta f, \tag{2}$$

where k is the Boltzmann constant, 1.37×10^{-23} joules per degree K and T is the absolute temperature of the calibrating resistor, degrees K. (For purposes of numerical computations, T is assumed to be equal to 300°K or, approximately, 80°F.)

The first term on the right side of the equation is the Nyquist's term, giving the mean-squared voltage due to the thermal agitation of the electrons within the calibrating resistor.[17,18]

It is convenient to express the intensity of the noise to be measured in terms of the fictitious temperature T' at which the resistor R_a must be to develop equal noise voltage thermally.

Then

$$4kT'R_a\Delta f = 4kTR_a\Delta f + 2eI_sR_a^2\Delta f$$

and

$$T' = T + \frac{eR_a}{2k}\cdot I_s. \tag{3}$$

[17] J. B. Johnson, "Thermal agitation of electricity in conductors," *Phys. Rev.*, 2nd ser., vol. 32, pp. 97–109; July, 1928.

[18] H. Nyquist, "Thermal agitation of electric charge in conductors," *Phys. Rev.*, 2nd ser., vol. 32, pp. 110–113; July, 1928.

Using type CV-172 diodes, which can, without burning out, carry a space current of 100 ma, it was possible to calibrate up to effective temperature T' of approximately 60,000°K. For R_a of 100 ohms and for 10-kc bandwidth, this equals approximately 1.8 μv. This was adequate for normal comic radio noise measurements.

III. Units Employed for Presenting the Results

It has been a generally accepted practice among the physicists interested in the cosmic-noise measurements to express the intensity of such noise in terms of temperature, degrees K, at which a resistor would generate thermally an equal available noise power. This, to a considerable extent, is a matter of convenience since, using temperature units, neither source impedance nor bandwidth employed need be specified. Such representation is legitimate only when the noise voltages being considered are random in character. Since this paper is intended for engineers interested in evaluating the interference value of cosmic noise, the results were also converted to terms of power intensity in watts per square meter and those of field strengths in microvolts per meter. The conversion to power intensity is accomplished by the use of the Jeans-Rayleigh black-body radiation law,

$$P = \frac{8\pi kTf^2\Delta f}{c^2}, \tag{4}$$

where P is the power radiated by a black body in the frequency interval Δf, in watts per square meter, f is the frequency at which the measurements are being made, in cycles per second, and c is the velocity of propagation in meters per second, 3×10^8.

This expression gives the power radiated by a black body in both planes of polarization. The black-body radiation is randomly polarized so that the power is equally distributed in each plane. Since the receiving dipole is sensitive to only one plane of polarization, it would receive only half the power radiated by a black body. Hence, the true power received by the dipole is half of that given by the above relationship. For a bandwidth of one cycle per second and when frequency f is in megacycles, this becomes

$$P = 1.91\times10^{-27}\times f^2\times T \tag{5}$$

The electric field strength is obtained directly from power intensity by the relationship

$$\overline{E^2} = P\times Z_0 \tag{6}$$

where E is the electric field intensity, in volts per meter, and Z_0 is the characteristic impedance of space, 376.7 ohms. If E is expressed in microvolts per meter for a bandwidth, Δf, of 1,000 cycles per second, and frequency f, is in megacycles per second, then

$$E_{rms} = 2.68\times10^{-5}\times f\times\sqrt{T}. \tag{7}$$

IV. Results of Measurements—Normal Cosmic Radio Noise

In the course of the two years' measurements, it was found that the normal cosmic radio noise intensities have a very regular diurnal pattern. Fig. 4 presents a typical record of one week's measurements. As all measurements described here, this record was made with the dipoles oriented in the east-west direction. This orientation makes the system more sensitive to radiation from low angles in the southern and northern directions and less sensitive to low-angle radiation from east and west. The most intense source of galactic radio noise is located near the constellation of Sagittarius. At the latitude of Washington, 39°N, this source attains its highest elevation above the horizon at approximately 25° due south. Therefore, such orientation of the antenna may be expected to produce higher maximum noise intensity than the north-south orientation of the antenna. At the same time the reduced sensitivity of the antenna to low-angle radiation coming from east and west leads one to expect that when the major sources of galactic noise are east or west of the meridian the response of the east-west oriented antenna would be lower than that of an antenna oriented north and south, and it may have a lower minimum. This difference in maxima and minima was confirmed by Herbstreit[16] experimentally.

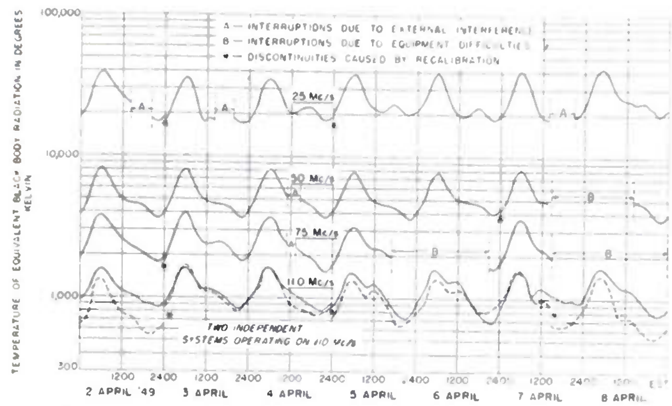

Fig. 4—A typical record of one week's measurements on four frequencies with five receiving systems plotted to a common temperature scale.

Fig. 4 shows that in early April, 1949 the maximum noise intensity was recorded on all frequencies at approximately 0600 EST, while the minimum was observed at approximately 2200 EST. These times correspond to 1900 and 1100, sidereal time, respectively. By reference to Fig. 1, it can be seen that the maximum corresponds to the time when the constellation of Sagittarius is just west of the meridian and the constellation of Cygnus is approximately the same distance east of the meridian; the antenna is thus in a position to receive the maximum of energy from the two sources. At sidereal time of the minimum, 1100, the sky is relatively free of the more intense sources of cosmic noise.

It was found convenient to present the data obtained during the twenty-two months of measurements by plotting the daily normal maxima and minima, in this way presenting the normal upper and lower limits for the noise. The yearly averages of daily maxima and minima are presented in Table I in degrees K, in micro-

TABLE I

AVERAGES OF NORMAL DAILY MAXIMA AND MINIMA FOR 1948 AND 1949

	1948 (March—December)					1949 (January—December)					(1948–1949)				
	Equiv black body temp degrees K	Field strength μv/m for 1,000 cps	Power intensity watts/sq m for 1 cps	Std deviation db	No of observ	Equiv black body temp, degrees K	Field strength μv/m for 1,000 cps	Power intensity watts/sq m for 1 cps	Std deviation db	No of observ.	Equiv black body temp degrees K	Field strength μv/m for 1,000 cps	Power intensity watts/sq m for 1 cps	Std deviation db	No of observ
			$\times 10^{-20}$					$\times 10^{-20}$					$\times 10^{-20}$		
25-mc normal daily maxima	41,800	0.137	5.00	0.7	181	34,700	0.125	4.17	0.8	304	37,300	0.130	4.48	—	485
25-mc normal daily minima	19,800	0.094	2.38	1.0	176	16,600	0.087	1.99	0.8	308	17,700	0.090	2.13	—	484
35-mc normal daily maxima	—	—	—	—	—	16,500	0.121	3.88	0.4	109	16,500	0.121	3.88	0.4	109
35-mc normal daily minima	—	—	—	—	—	8,430	0.086	1.98	0.4	112	8,430	0.086	1.98	0.4	112
50-mc normal daily maxima	7,320	0.115	3.52	0.3	231	7,550	0.117	3.63	0.3	269	7,440	0.116	3.57	—	500
50-mc normal daily minima	3,440	0.079	1.65	0.3	240	3,510	0.080	1.69	0.4	274	3,470	0.079	1.67	—	514
75-mc normal daily maxima	2,790	0.107	3.01	0.5	218	3,490	0.119	3.77	0.4	235	3,160	0.113	3.41	—	453
75-mc normal daily minima	1,230	0.071	1.33	0.6	182	1,580	0.080	1.71	0.6	222	1,430	0.076	1.55	—	404
#1 110-mc normal daily maxima	1,160	0.101	2.70	0.6	185	1,300	0.107	3.02	0.6	207	1,230	0.104	2.86	—	392
#1 110-mc normal daily minima	510	0.067	1.19	0.9	144	530	0.068	1.23	0.9	200	520	0.068	1.21	—	344
#2 110-mc normal daily maxima over ground	1,250	0.105	2.90	—	34	1,380	0.110	3.21	—	126	1,380	0.110	3.21	0.6	160
#2 110-mc normal daily minima over ground	450	0.063	1.05	—	32	600	0.072	1.39	—	113	570	0.071	1.32	1.0	145
#2 110-mc normal daily maxima over mat	—	—	—	—	—	1,310	0.107	3.04	0.6	83	1,310	0.107	3.04	0.6	83
#2 110-mc normal daily minima over mat	—	—	—	—	—	560	0.070	1.30	0.8	71	560	0.070	1.30	0.8	71

volts per meter for 1,000-cps bandwidth, and in watts per square meter per cycle per second bandwidth. Statistical analysis of the data showed that the distribution of the observations is very nearly normal. The standard deviations obtained as a result of this analysis are presented in columns 4, 9, and 14 of Table I. Examination of the tabulated values of standard deviations shows that the standard deviation for any frequency in either year never exceeds 1 db, and generally is considerably smaller.

It is a generally accepted assumption that the intensity of the galactic radio noise, at least when averaged out over the visible area of the sky, is constant. If this were so, the variations in the measured values of cosmic noise intensity must be attributable to errors in measurements, or are due to absorption by the ionosphere. To verify this, an effort was made to evaluate the accuracy of the measurements by estimating or computing the errors from the various sources. The analysis itself is too lengthy to be presented here in full; however, a summary of the results is presented in Table II. The table shows that the root-sum-square error from all the sources considered is of the same order of magnitude as the standard deviation of the measured values. This confirms the assumption that the variations in the measured values of cosmic radio noise are not the result of variations in the phenomenon being measured, but are introduced by the instruments and methods used in measurement.

Examination of the cosmic radio-noise data, averaged month by month, revealed no sign of absorption by the ionosphere at frequencies of 50 mc and higher. The 35-mc equipment was operated for too short a period of time for any conclusion to be made for this frequency. However, 25-mc equipment showed definite signs of variations attributable to the ionospheric absorption. Because of the earth's movement around the sun, the times of daily occurrence of the maximum and minimum cosmic radio-noise intensity change, being approximately four minutes earlier each succeeding day. Thus, the time of the maximum coincides with noon at approximately December 31, while the time of the minimum coincides with noon at approximately September 1. The 25-mc records reveal that around November and December in either year, when the maximum is measured at the time of maximum ionospheric absorption, the daily maximum values of cosmic-noise intensity are lower than at other times of the year. The daily minimum values of cosmic radio noise have a corresponding trough around September and October when the daily minimum is observed in the late morning hours. By using the departures of monthly mean values of daily maxima and minima from the annual mean values, root-mean-square errors were computed for variations due to ionospheric absorption. These errors appear in line 9 of Table II.

In Fig. 5 the cosmic noise intensities in microvolts per meter for 1,000-cps bandwidth are presented together with the atmospheric radio-noise data. The latter were derived from the National Bureau of Standards circular No. 462, "Ionospheric Radio Propagation," June 25, 1948. It should be noted that the early tentative values of cosmic radio noise intensities presented in that circular were in error, being 9 db too low. The cosmic noise values in this figure are normal daily maxima and minima averaged over the 22 months of measurements. Included are also some observations, curve *E* made in the Arctic in 1947, of atmospheric radio noise,

TABLE II

ESTIMATES OF THE MAGNITUDES OF ERRORS, IN DECIBELS, IN THE MEASURED VALUES OF COSMIC RADIO NOISE

Sources of Errors	25 mc		35 Mc		50 mc		75 mc		110 mc	
	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min
1. Inaccuracies in reading recorder chart	0.05	0.08	0.06	0.08	0.05	0.07	0.08	0.17	0.08	0.17
2. Changes in sensitivity of equipment	0.30	0.28	0.24	0.26	0.25	0.25	0.28	0.39	0.43	0.70
3. Inaccuracies in reading calibrating diode current	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.03	0.03	0.02
4. Errors in the measured values of radiation resistance*	(0.21)	(0.21)	(0.21)	(0.20)	(0.20)	(0.19)	(0.19)	(0.17)	(0.16)	(0.09)
5. Variations in radiation resistance with weather	0.21	0.21	0.21	0.20	0.20	0.19	0.19	0.17	0.16	0.09
6. Variations in absorption by the ground	0.20	0.20	0.20	0.20	0.20	0.19	0.18	0.16	0.15	0.08
7. Variations in temperature of calibrating resistor**	±0.00 ∓0.05	±0.00 ∓0.05	±0.00 ∓0.05	±0.01 ∓0.05	±0.01 ∓0.05	±0.02 ∓0.05	±0.02 ∓0.05	±0.04 ∓0.04	±0.04 ∓0.04	±0.10 ∓0.02
8. Interference	nil	nil	nil	nil	nil	nil	nil	nil	nil	nil
9. Absorption by ionosphere	0.62	0.74	nil	nil	nil	nil	nil	nil	nil	nil
10. Natural fluctuation in galatic noise	unknown	unknown	unknown	unknown	unknown	unknown	unknown	unknown	unknown	unknown
Root-sum-square error due to all random effects	0.75	0.85	0.39	0.39	0.34	0.38	0.39	0.49	0.49	0.74
Mean standard deviations (From Table I)	0.75	0.90	0.40	0.40	0.30	0.35	0.45	0.60	0.60	0.90

* Errors in measured values of radiation resistance are systematic and are not included in the summation.
** The two errors due to this cause are in opposite direction and their difference is used in the root-sum-square summation.

in vlf and lf bands. In addition to the cosmic and atmospheric noise there is plotted the noise field intensities produced by black-body radiation at the temperature of the earth's surface (taken to equal 300°K or, approximately, 80°F). However, this noise does not necessarily exist at that level of intensity since most of the surroundings, notably the ground itself, depart considerably from being perfect black bodies. Also, the surrounding objects including the ground occupy only half of the sphere. Fig. 5 also shows, for comparison purposes, the noise field strengths, curve *B*, corresponding to the internal noise of well-designed receivers. This latter curve was obtained from the empirical relationship presenting best available noise figures for a range of frequency which appeared in a paper by Norton and Omberg.[19] Fig 5 displays the fact that for a well-designed, high-gain, low-noise receiver, cosmic radio noise may well present the limit to communications up to approximately 200 mc. For receiving systems using directive antennas the interference value of cosmic noise may be important at a considerably higher frequency for such times as the direction of the maximum sensitivity of the antenna coincides with the direction of the more intense sources of cosmic radio noise.

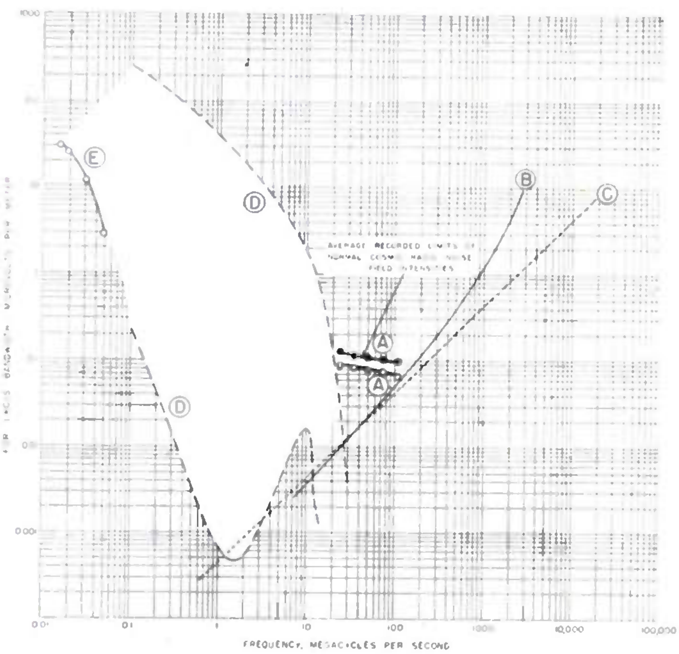

Fig. 5. Comparative intensities of radio noise of natural origin in the spectrum of radio frequencies.
A—Average daily upper and lower limits of the normal cosmic radio noise field intensities.
B—Noise field intensities corresponding to internal noise of well-designed receiver (derived from Norton and Omberg[19]).
C—Noise field intensities (in one plane of polarization) produced by "black-body" radiation at 300°K (80°F).
D—Upper and lower limits of atmosphere radio noise intensities. Derived from the National Bureau of Standards Circular No. 462. Also radio propagation unit report RPU-5.
E—Atmospheric radio-noise intensities measured in Arctic (Cottony[20]).

Herbstreit,[16] in reporting the early phases of this work, attedpted to correct for the absorption by the ground. He found it necessary to add approximately 1 db to the measured results at 25 mc and 1.7 at 110 mc to obtain the incident noise intensities. During this program of measurements, an attempt was made to verify these deductions by operating two 110-mc receiving systems, one over the ground, the other over a metallic screen. No significant difference was observed in the results. A review of Herbstreit's computations and a measurement of the ground constants showed that the relative dielectric constant of the ground at Sterling, Virginia was 23 rather than the assumed 4, and that the actual distribution of noise sources resulted in reflection at more nearly a grazing angle than with a uniform distribution of noise sources assumed by Herbstreit. Both of these factors contributed to a significantly lower recomputed value of absorption by the ground. It is now estimated that at 110 mc the absorption by the ground should lower the observed noise intensity by approximately 0.65 db. At lower frequencies the correction is still smaller. Because the corrections are only estimates, and since the possible error is of the same order of magnitude, these corrections were not applied to the results in this paper.

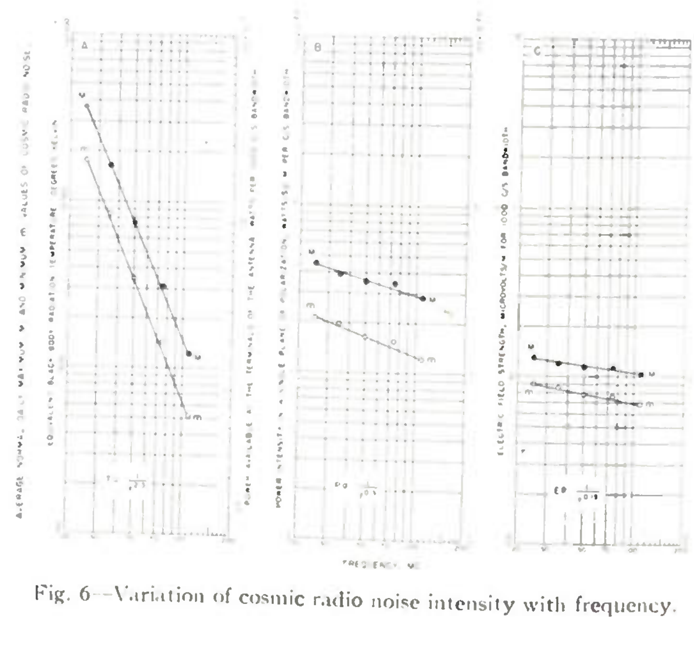

Fig. 6—Variation of cosmic radio noise intensity with frequency.

One of the aims of this investigation has been to determine, with some accuracy, the frequency law of cosmic noise intensity. Fig. 6 presents the comparison of noise intensities with frequency graphically in the three sets of units. The graphs are plotted to a logarithmic scale in each case. The intensities are averages for the duration of the measurements.

[19] K. A. Norton and A. C. Omberg, "The maximum range of a radar set," PROC. I.R.E., vol. 35, pp. 4–24; January, 1947.
[20] H. V. Cottony, "Observations of Atmospheric Radio Noise in Arctic Regions," Memorandum Report; January 15, 1948. (Not published, not available for distribution.)

It can be seen that when the intensities are expressed in terms of temperatures of equivalent black-body radiation, they vary inversely as the 2.3 power of the frequency, when in terms of electric field strength, inversely as the 0.15 power and, when in terms of power intensity, inversely as the 0.3 power. Moxon,[21] who investigated the variations of cosmic radio noise intensity with frequency in the range of 40 to 200 mc, found the intensity (expressed as temperature of equivalent black-body radiation) to vary inversely as the 2.7 and as 2.1 power of the frequency in the plane of galaxy and away from the plane of the galaxy, respectively. Comparison of daily maxima and minima obtained in this investigation does not disclose any indication of difference in the frequency law of radiation.

V. Results of Measurements—Abnormal Phenomena

In addition to the normal cosmic radio noise intensities characterized by their regularity, there have been observed from time to time abnormal phenomena which in all observed instances appear to be associated with solar disturbances. Two such observations are described here as illustrative of such phenomena.

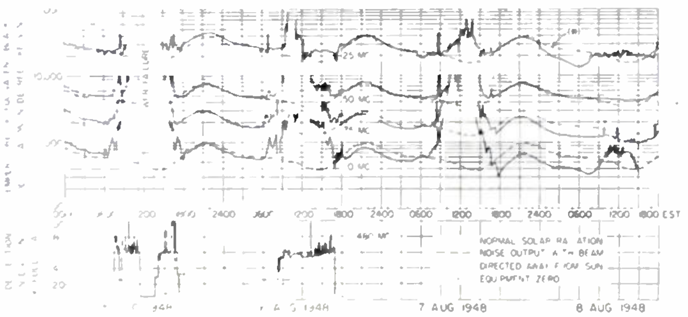

Fig. 7—Transcribed records of abnormal cosmic radio noise obtained during a period of high solar activity. (A) Broad directivity measurements of cosmic radio noise on 25, 50, 75, and 110 mc. (B) Directional measurement of solar radio noise on 480 mc. (*) Normal level of cosmic radio noise.

Fig. 7 presents the cosmic noise measurements made on August 4–8, 1948. On those dates a large group of sunspots travelled across the face of the sun passing close to the center of the solar disc. This phenomenon was accompanied by an enhanced level of cosmic noise. The figure shows two sets of observations. The upper portion of the figure presents the transcribed record made by the equipment described in this paper on the frequencies of 25, 50, 75, and 110 mc. The lower portions of the figure present the transcribed record made by a radiometer located near the site of the cosmic-noise equipment and operated by Reber in connection with solar-noise studies. This radiometer consisted of a 480-mc receiver with a directive antenna consisting of a dipole located in the focus of a 25-foot parabolic reflector. It is automatically directed at the sun in the daytime. The radiometer measures the relative level of solar activity as evidenced by the intensity of radio-frequency radiation from the sun.

The two sets of records show that, while the broad directivity measurements show considerable increase in the noise in the middle of the day, amounting to several times the normal noise power level, the radiometer shows only a moderate increase in the noise level intensity accompanied by relatively short-duration, high-intensity bursts. An examination of the detail of the noise records shows that there is a close qualitative correspondence between the bursts of noise as recorded by the radiometer on 480 mc and those recorded by the broad-directivity equipment. This indicates that in each case the noise is apparently of solar origin. However, the general increase in noise is greater in the case of broad-directivity measurements. Since the broad-directivity receiving systems receive the radiation from a relatively great area in the sky, they are relatively insensitive to radiation from a small area unless the intensity of radiation from that area is very high. Thus a quiet sun, assuming equivalent black-body radiation temperature of 1,000,000°K at radio frequencies, would produce a maximum increase in the measured noise intensity equivalent to approximately 10 to 15°K, which is below the resolving power of the equipment. The fact that the broad-directivity equipment at 25–110 mc is affected to a greater degree by the solar radiation than the relatively directive equipment pointed at the sun at 480 mc may be explained by the fact that the radiation from a disturbed sun is far more intense at the lower frequencies or that, possibly, the radio noise is generated in the vicinity of the earth although induced by solar emission.[22] In either case the phenomenon invites further investigation.

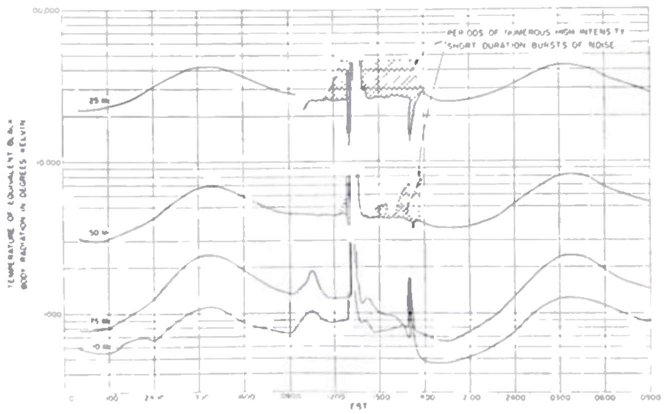

Fig. 8. Transcribed record of sudden ionospheric disturbance of May 7, 1948, with the accompanying solar bursts plotted to a common temperature scale.

Another type of disturbance recorded by the equipment is illustrated by Fig. 8. This presents cosmic noise measurements of May 7, 1948. These show a series

[21] L. A. Moxon, "Variation of cosmic radiation with frequency," *Nature*, vol. 158, no. 4021, pp. 758–759; November 23, 1946.

[22] H. V. Cottony, "Radio noise of ionospheric origin," *Science*, vol. 111, no. 2872, p. 41; January 1950.

of sudden bursts of noise beginning about 1300 EST and lasting for approximately one hour. These bursts of noise are also accompanied by a sharp drop in noise at 25 mc and, to a lesser extent, at 50 mc. It is believed that this drop in cosmic noise intensity at lower frequencies is caused by increased absorption in the ionosphere. The period of increased absorption is, in this case, of a few minutes in duration. On the same day, a few hours later at approximately 1700 EST, there is a record of a lesser noise burst on 75 and 110 mc with a simultaneous sudden decrease in cosmic noise intensity on 25 and 50 mc. In this case there are no noticeable bursts of noise at the lower frequencies. However, it may be a phenomenon similar to the first, the lack of the noise bursts being possibly explainable by the more oblique path of radiation from the sun which lengthened its path through the absorbing medium. This type of phenomenon is invariably associated with sudden ionospheric disturbances, a phenomenon which was first reported in 1935 by Dellinger.[23,24] The sudden ionospheric disturbances (SID) consist of failures in radio communication due to disappearance or fading-out of all signals presumably due to high absorption in the ionosphere. Their connection with eruptions on the sun and the normal presence of bursts of nonatmospheric radio noise were likewise noted by Dellinger. On May 7, 1948, three SID were observed at Washington, D. C. at 1000 to 1025, 1248 to 1440, and 1704 to 1755 EST. These coincide with the periods during which intense bursts of noise on 75 and 110 mc and drop in noise on 25 mc were noted on cosmic radio noise recorders. The phenomenon illustrated by the observations of May 7, 1948 appears to be distinct in character from that observed in August 5–8, 1948; but both are apparently closely connected with solar disturbances.

[23] J. H. Dellinger, "A new cosmic phenomenon," *Science*, vol. 82, no. 3028, p. 351; October 11, 1935.

[24] J. H. Dellinger, "Sudden disturbances of the ionosphere." Proc. I.R.E., vol. 25, pp. 1253–1290; October, 1937.

VI. Conclusions

On the basis of the two years' radio noise measurements in the vhf band the following conclusions are reached:

1. The normal cosmic radio noise in the vhf band, although relatively low in intensity, may, under conditions of good receiver design and proper antenna match, be the limiting factor to communications.
2. With the present knowledge of receiver design and for broad-directivity antenna systems, the cosmic radio noise may be the limiting factor to communication in the vhf band up to approximately 200 mc.
3. For receiving systems employing directive antennas the range of diurnal variation in cosmic radio-noise intensity may be expected to be much greater than that measured with the broadly directive antenna systems described here. Under these circumstances, if the direction of the signal to be received coincides with the direction to the more intense sources of the galactic radio noise, the frequencies at which the cosmic radio noise can be the limiting factor may be considerably higher than 200 mc.
4. Under the condition of abnormal solar activity, which is not an infrequent phenomenon, the level of the radio noise is greatly enhanced, and may, on occasion, be expected to present serious interference to radio communication in the vhf range.

The Effective Bandwidth of Video Amplifiers*

F. J. TISCHER†

In dealing with problems concerning video amplifiers with very wide bandwidth, the bandwidth as defined in the usual way does not seem to be useful for characterizing the utility of the amplifier for pulses and video reproduction. The attempt is made to derive an "effective" bandwidth which takes into account the phase characteristic but is, within certain limits, independent of the direct shape of the amplitude characteristic.

This "effective" bandwith, which is determined directly from the complex transfer function on the steady-state basis, is defined as the bandwidth of an "ideal" amplifier, giving the same steady final state and the same rise time of the transient response if excited by an unit step signal. The effective bandwidth as a new figure of merit can replace the double indication of bandwidth and rise time, as usually applied heretofore.

As most amplifier coupling networks are "minimum phase-shift" networks, the phase characteristic is at the same time defined by the amplitude characteristic, and it is therefore sufficient to know this characteristic in order to determine the effective bandwidth. This is very important in the case of amplifiers with very wide bandwidth, because the amplitude of amplification is the only value that can be measured in a comparatively simple way at all frequencies. The values of the effective bandwidth for some theoretical standard transfer functions give a good idea of the values that can be expected in practice for amplifiers with normal transfer functions and influence of phase distortions.

The improvement of the effective bandwidth by compensation of the phase error and the possible reduction of the number of stages of an amplifier in cascade coupling, constant gain being assumed, and the limits of this improvement are also investigated.

* Decimal classification: R363.4. Original manuscript received by the Institute, February 28, 1951; revised manuscript received April 7, 1952. In print for publication in the *Transactions of the Royal Institute of Technology*, Stockholm, Sweden.

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Interaction Between Surface-Wave Transmission Lines*

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Summary—**An important question connected with surface-wave transmission lines is the interaction between them or with other wires which act like surface waveguides. An analysis is made of two parallel lines with the provision that the coupling is small. When the two lines are identical, there is maximum interaction, and under suitable conditions, complete power transfer from one line to the other occurs. The analysis is supplemented by typical examples.**

I. Introduction

One of the problems connected with surface-wave transmission lines of the type described by Goubau[1,2] is the interaction between two such lines. This interaction may be of consequence when there are two active lines parallel to each other, as in the case of transmitter and receiver antenna feeds for radio relay, or when there is another wire line, perhaps a telephone line, parallel to an active surface-wave line. In either case, we impose the restrictions that the distance between the two lines is large compared with their diameters and that each line is in the form of a highly conductive wire with a dielectric coating. The first restriction allows us to assume that the coupling is brought about only by the longitudinal electric field components and that the field associated with one line is essentially constant in the vicinity of the other line. The second restriction enables us to find easily the parameters of surface waves on the isolated lines by the method given by Goubau. Then the corresponding parameters when interaction is present can be expressed in terms of small deviations from these parameters.

II. Analysis

Consider first two infinitely long lines without losses. If the lines are separated enough, the longitudinal electric field component associated with one line can be considered constant over the cross section of the other. The corresponding radial electric field component is neglected since its only effect is to cause a slight modification of the symmetry of the field. Furthermore, the impressed displacement current in the dielectric is small compared with the current in the conductor as long as the coating is not very thick.

* Decimal classification: R117.1. Original manuscript received by the Institute, June 29, 1951; revised manuscript received May 8, 1952. Presented in abridged form at the 1951 IRE National Convention.

† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

[1] G. Goubau, "Surface waves and their application to transmission lines," *Jour. Appl. Phys.*, vol. 21, pp. 1119–1128; November, 1950.

[2] G. Goubau, "Single conductor surface-wave transmission lines," Proc. I.R.E. vol. 39, pp. 619–624; June, 1951.

With these considerations we proceed to an analysis of the surface waves existing on the system of two lines indicated in Fig. 1. For each line acting alone, the radius of the conductor a, the outer radius of the dielectric

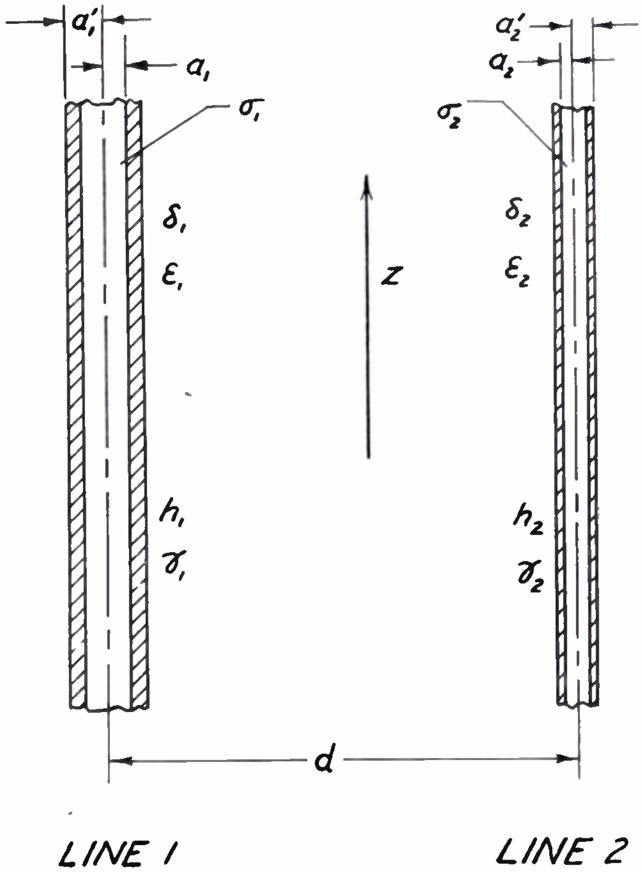

Fig. 1—Two parallel surface-wave transmission lines.

coating a', and the dielectric constant ϵ of the coating are sufficient for the determination of the propagation constant h. This is related to the quantity γ appearing in the expressions for the field components outside the line,

$$\begin{aligned} E_r &= A\frac{h}{\gamma}H_1^{(1)}(j\gamma r)e^{j(\omega t-hz)} \\ E_z &= AH_0^{(1)}(j\gamma r)e^{j(\omega t-hz)} \\ H_\phi &= A\frac{k^2}{\omega\mu\gamma}H_1^{(1)}(j\gamma r)e^{j(\omega t-hz)}, \end{aligned} \tag{1}$$

by the relation, $h^2=k^2+\gamma^2$, where k is the free-space propagation constant.

In order to find the propagation constants of surface waves of the system of two lines and to find the relative amplitudes in the two lines for each such wave, we divide the total or actual field, E_T, H_T, within and in the vicinity of one line, say line 2, into a primary field, E_P, H_P, due to the actual current in line 1, and a secondary field, E_S, H_S, due to the actual current in line 2 itself. The primary field and the total field, E_T, H_T, existing when line 2 is present, individually satisfy the source-free Maxwell's equations

$$\begin{aligned} \nabla \times H_P - j\omega\epsilon_0 E_P &= 0 \\ \nabla \times E_P + j\omega\mu H_P &= 0 \\ \nabla \times H_T - j\omega\epsilon E_T &= 0 \\ \nabla \times E_T + j\omega\mu H_T &= 0. \end{aligned} \tag{2}$$

Since $E_T = E_P + E_S$ and $H_T = H_P + H_S$,

$$\nabla \times H_S - j\omega\epsilon E_S = j\omega(\epsilon - \epsilon_0)E_P; \tag{3}$$

$$\nabla \times E_S + j\omega\mu H_S = 0, \tag{4}$$

where ϵ may be complex to account for conductivity.

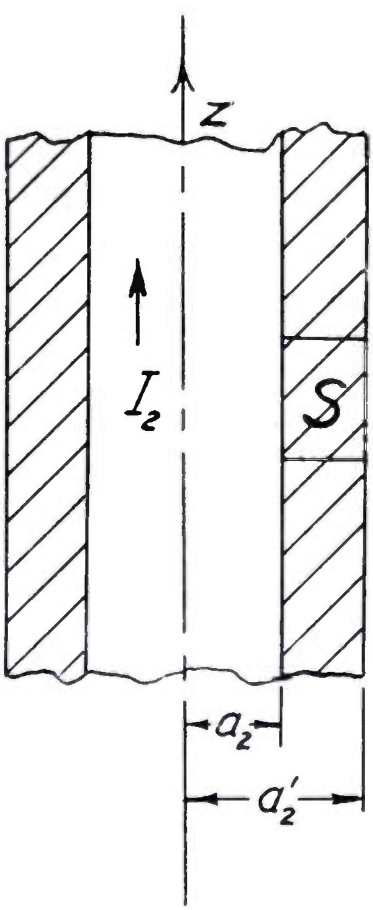

Fig. 2—Enlarged section of line 2.

In order to obtain a relation between the currents in the two lines, consider the enlarged section of line 2 shown in Fig. 2. Equation (4) is integrated over a surface S of the cross section of the dielectric coating extending longitudinally a distance δz

$$i\omega\mu \int_S H_S \cdot d\sigma + \int_S (\nabla \times E_S) \cdot d\sigma = 0. \tag{5}$$

The integral of H_S is proportional to the current in line 2, I_2. The integral of E_S is transformed into a line integral along the boundary of S. Actually, we are expressing Faraday's induction law. The evaluation of the latter integral is carried through by the use of the boundary conditions at the inner and outer surfaces of the dielectric coating so that E_S is expressed everywhere on the path of integration in terms of surface-wave components which, in turn, are related to I_1 and I_2, the currents in the lines. The result of this evaluation, the details of which are given elsewhere,[3] is

$$\left[\ln\left(\frac{a_2'}{a_2}\right)\frac{1}{2\pi\omega\epsilon_2}(k_2^2 - h'^2) + \frac{1}{2\pi}\ln(0.89\gamma' a_2')\frac{\gamma'^2}{k}\sqrt{\frac{\mu}{\epsilon_0}}\right] I_2 - j\frac{\gamma'^2}{4k}\sqrt{\frac{\mu}{\epsilon_0}}\, H_0^{(1)}(j\gamma' d) I_1 = 0, \tag{6}$$

where h' is the propagation constant of the actual field, $\gamma'^2 = h'^2 - k^2$, and $k_2 = \omega\sqrt{\mu\epsilon_2}$. Using (47) of footnote reference 1, neglecting γ'^2 with respect to k^2 and rearranging terms, we obtain the equation,

$$-j\pi H_0^{(1)}(j\gamma' d) I_1 + \left[2\ln(0.89\gamma' a_2') - 2\frac{\gamma_2^2}{\gamma'^2}\ln(0.89\gamma_2 a_2')\right] I_2 = 0.$$

A similar relation is obtained by treating line 2 as the primary. Then,

$$a_{11} I_1 + a_{12} I_2 = 0, \tag{7}$$

$$a_{21} I_1 + a_{22} I_2 = 0, \tag{8}$$

where

$$a_{ii} \equiv 2\ln(0.89\gamma' a_i') - 2\frac{\gamma_i^2}{\gamma'^2}\ln(0.89\gamma_i a_i'), \quad [i = 1, 2]$$

and

$$a_{12} = a_{21} \equiv -j\pi H_0^{(1)}(j\gamma' d).$$

(γ' pertains to the system of two lines while γ_1 and γ_2 are the parameters for the isolated lines).

Equations (7) and (8) are solved in the Appendix for γ'. There are two solutions: γ_1' close to γ_1 and γ_2' close to γ_2. Thus the system may have two surface waves, to each of which corresponds a current in each line. The expressions obtained for the evaluation of γ_1' and γ_2' are $\gamma_i'^2 = \gamma_i^2(1 + \eta_i)$, where

[3] A. A. Meyerhoff, "Interaction between surface-wave transmission lines," Signal Corps Engineering Laboratories Technical Memorandum No. M-1376; May, 1951.

$$\eta_i = \frac{c_i^2}{b_i[\sigma_{ji}(1 - b_j) + \ln s_{ij}]}, \quad \text{when } |\sigma_{ij}| \gg 0, \quad (9)$$

$$\eta_i = -\frac{\sigma_{ij}}{2}\left(1 - \sqrt{1 + \frac{4c_i^2}{\sigma_{ij}^2 b_i b_j}}\right),$$

$$\text{when } |\sigma_{ij}| \ll 1, \quad (10)$$

$$\eta_i = (-1)^i \frac{c}{\sqrt{b_1 b_2}}, \quad \text{when } \sigma_{ij} = 0 \quad (11)$$

and

$$\eta_i = (-1)^{i+1} \frac{c}{b}, \quad \text{when } \sigma_{ij} = 0 \text{ and } a_1' = a_2'. \quad (12)$$

The s's, η's, σ's, b's, and c's are defined in (17) and (19).

In a given case some question may arise as to whether (9) or (10) should be used. In most cases the two expressions reduce to nearly the same value for values of $|\sigma_{ij}|$ near 0.1, but in extreme cases, when the coupling is high, the transition value is somewhat greater for the best approximation.

The current ratios are obtained from (7) and (8)

$$\frac{I_j}{I_i} = -\frac{a_{ii}}{a_{ij}} \quad [i, j = 1, 2 \text{ or } 2, 1] \cdot$$

Each expression should be evaluated for one of the two values of γ'. We take the expression I_j/I_i to mean the ratio for the wave characterized by γ_i'. As in the Appendix, a_{ii} and a_{ij} are replaced by their approximate equivalents, $\eta_i b_i$ and c_i, respectively. Then

$$\frac{I_j}{I_i} \simeq -\frac{\eta_i b_i}{c_i}. \quad (13)$$

Equation (13) holds for any σ_{ij}. If $\sigma_{ij} = 0$, it reduces, with (11), to

$$\frac{I_j}{I_i} = (-1)^i \cdot \sqrt{\frac{b_i}{b_j}}, \quad (14)$$

or if $\sigma_{ij} = 0$ and $a_1' = a_2'$,

$$\frac{I_j}{I_i} = (-1)^i. \quad (15)$$

So far the analysis has been carried through for the dissipationless case. To include dissipation we assume that the field distribution over any cross section is undisturbed. Then the attenuation with respect to this undisturbed field is computed in much the same way as it is done for a single line. Equations (58) to (62) of footnote reference 1 are used with the proper values of γ'. The approximation introduced by this procedure is not serious as long as the lines are not so far apart that a cross section is no longer an equiphase plane over an area including both lines. However, this method does not take into account the contribution to the power of the second and fourth integrals in the expression,

$$V = \int E_1 \times H_1^* \cdot d\sigma + \int E_1 \times H_2^* \cdot d\sigma + \int E_2 \times H_2^* \cdot d\sigma + \int E_2 \times H_1^* \cdot d\sigma.$$

for the propagating power.

III. Applications

Since we have solved only for the case of infinitely long lines, the conditions at the launching site must be taken into account. Furthermore, both the amplitude and phase of the currents of each wave may be chosen at will for some value of z, the co-ordinate in the direction of propagation. Normally, the excitation is done by launching a wave by means of a horn on one line, say line 1. The current in line 1 has a certain value and the current in line 2 at this plane is nearly zero. Knowing the current ratios for the two waves and the total current in each line at the plane of the launching site, we can compute the individual current in each line for each wave. In order to have these currents at this plane we should, technically, provide the correct field distribution over an infinite cross section, perhaps by a suitable dipole distribution. The horns provide a close approximation to the major part of the correct field, and the radiated power associated with this approximation is of no more consequence than in the excitation of a single line.

Let us consider some special cases. In the case where the lines are identical, the current ratios from (15) are $+1$ for one wave and -1 for the other, and the two waves, under the above excitation conditions, have equal amplitudes. That is, for one wave the currents are in phase in the two lines and for the other the currents are 180° out of phase. At the plane of the launching site all the power is associated with line 1. Since, by (12), the two waves have slightly different phase velocities, eventually, at some distance from this plane the currents in line 1 cancel while line 2 has all the power. The power continues to shift between the two lines. We shall see later, however, that in practical cases the distance required for a complete shift is very large. If the propagation constants of the isolated lines are equal but the radii are different, there is still a complete energy shift. By (14) the current, I_2, is greater than I_1 for each wave if a_1' is smaller than a_2'. If, in this case, we excite line 1 with a certain current, when the current reaches a maximum in line 2, the total there will be greater than the original current in line 1. However, by (57) of footnote reference 1, the power ratio is still unity.

As an example, assume that the two lines are identical, $a = 0.1$ cm, $a_1 = 0.105$ cm, $\epsilon = 3$, $\tan \delta = 8 \times 10^{-3}$, the conductor is of copper, and the frequency is 3,300 mc. These lines might be, except for uncertainty about the loss factor, enameled wires. Some of the results of this example are plotted in Fig. 3. For this case of identical lines where one is excited with a certain current, I_1, at $z = 0$, at some value of z, say l, this same current exists in line 2 and there is no current in line 1. The curve of l in Fig. 3 shows the distance required in kilometers for one complete transfer as a function of d, the separation of the lines in centimeters. In the usual situation the separation might be 100 cm. Then l is more than 14 km.

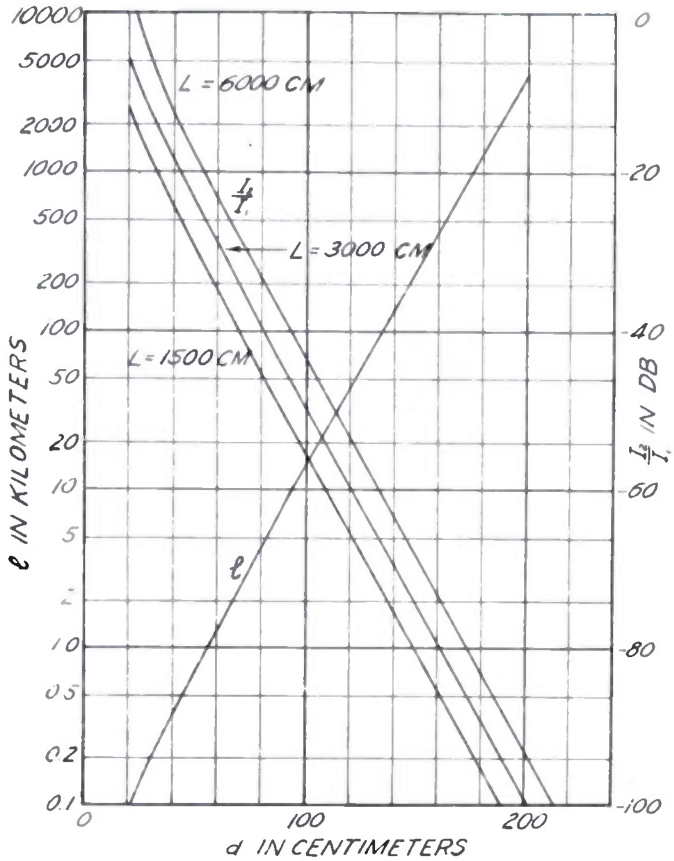

Fig. 3—The propagation distance l required for complete energy transfer versus separation d, and the total current ratio versus d for three line lengths L.

If the lines are used as antenna feeds, say 30 meters long, very little energy will be transferred while propagating along this length. The ratio of the total current, I_2, in line 2 to the total current, I_1, in line 1, 1,500, 3,000, and 6,000 cm from the launching site is plotted in decibels in the other three curves of Fig. 3. The curve for the 3,000-cm length, for instance, shows that for 100-cm separation the secondary current is about 50 db below the primary current. This value has been verified experimentally.

Another way to reduce the coupling to line 2 is to make the two lines different from each other. A surprisingly small change from the identical case decreases the current ratios considerably. For example, a change in the dielectric thickness of one line of two-tenths of a per cent reduces the current ratios, if the separation is 150 cm, from 100 per cent to a little over 1 per cent. However, the maximum value of secondary current is achieved much sooner than in the case of identical lines.

This example, as a whole, demonstrates some interesting facts about two-wire waves. As was mentioned before, of the two waves, one has currents in phase and the other has currents 180° out of phase in the two lines. The latter is a two-wire wave, but the current amplitudes for this wave are equal only when the lines are identical. In this symmetrical case, the initial conditions can easily be adjusted to provide only this wave. But when the lines are at all different, the current ratio for the out-of-phase wave is no longer unity, and it becomes impossible to maintain this wave with equal amplitudes in the two lines. Thus, we see that the ideal two-wire wave is a special case of surface waves along two lines, requiring perfect symmetry.

The question now arises, is it possible to put to practical use the phenomenon of complete power transfer? Theoretically, for long-distance transmission where the field extension is made large to admit of very low attenuation, the wave could be launched on a line, running parallel to the long line for the distance required for complete power transfer, having the same propagation constant as the long line but a smaller field extension, and thus a smaller required horn size. However, for any reasonable reduction in horn size, the auxiliary line must assume impractically small diameters. Nevertheless, the phenomenon may be useful when a line already in service for some other purpose is to be used as a surface-wave transmission line, and it is not possible to place horns directly on this line for tapping some of the energy from an active line.

As another example, consider a long line with a radius of 1 cm and a loss of 3 db per mile at 200 mc, having parallel to it a No. 12 enameled wire, used perhaps as a telephone line. With a separation of 1 meter, the maximum current in the enameled wire, if the transmission line is excited, is about one-tenth the current in the excited line. The current in the primary associated with the normal wave there is reduced by less than 1 per cent, and the additional attenuation caused by the presence of the enameled wire is 0.16 db per mile compared with the 3-db per mile attenuation for the undisturbed lone line. Thus, the enameled wire has very little effect on the transmission. If these lines are on telegraph poles, it would be easy to separate them even more than 1 meter if we desired to reduce the disturbance further.

IV. Acknowledgment

Thanks are due Dr. Georg Goubau for originally suggesting the subject of this paper and for making many valuable suggestions in the course of its preparation.

V. Appendix

For a solution to exist for the ratios of the currents in (7) and (8), the determinant of the coefficients must vanish, i.e.,

$$a_{ii}a_{jj} = \alpha_{ij}^2. \tag{16}$$

Equation (16) is to be solved for γ'. We define

$$s_{ij} \equiv 1 + \sigma_{ij} \equiv \frac{\gamma_i^2}{\gamma_j^2} \quad [i, j = 1, 2 \text{ or } 2, 1], \tag{17}$$

$$1 + \eta_i \equiv \frac{\gamma_i'^2}{\gamma_i^2}, \tag{18}$$

$$b_i \equiv 1 + 2 \ln (0.89\gamma_i a_i'),$$

and (19)

$$c_i \equiv -j\pi H_0^{(1)}(j\gamma_i d).$$

With the assumption, valid in all practical cases, that $\eta_i \ll 1$, a_{ii} reduces approximately to $\eta_i b_i$ and a_{ij} can be evaluated at $\gamma' = \gamma_i$. There are several cases:

Case I, $|\sigma_{ij}| \gg 0$

Here a_{jj} can be evaluated at $\gamma' = \gamma_i$. Then (16) reduces to

$$\eta_i b_i [2(1 - s_{ji}) \ln (0.89\gamma_j a_j') + \ln s_{ij}] = c_i^2,$$

which, with (17) and (19), becomes (9).

Case II, $|\sigma_{ij}| \ll 1$

This time both σ_{ij} and η_i are small, so that a_{jj} is evaluated in the same manner as a_{ii}, and we have

$$\eta_i b_i(\eta_i + \sigma_{ij}) b_j = c_i^2,$$

the solution of which is

$$\eta_i = -\frac{\sigma_{ij}}{2}\left[1 \pm \sqrt{1 + \frac{4c_i^2}{\sigma_{ij}^2 b_i b_j}}\right]. \tag{20}$$

The minus sign is chosen in (10) because η_i must be zero when c_i is zero (no interaction). The other root gives the relation existing between γ_j' and γ_i, equivalent to using the other value of i.

Case III, $\sigma_{ij} = 0$; *and Case IV,* $\sigma_{ij} = 0$ *and* $a_1' = a_2'$.

These follow directly from (10). The difference in the exponents of (11) and (12) is due to the fact that b_i is negative. The minus sign is arbitrarily assigned to the root for $i = 1$.

Part III—Investigations of High-Frequency Echoes*

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Summary—**A high-frequency sky-wave propagation experiment, carried out on November 19, 1944 by the Institute of Physics, Luftkriegsakademie, Gatow, in co-operation with the Deutsche Reichspost, is described. Signals of duration of 10 to 12 msec were transmitted at 0.5-sec intervals from DLO, Rehmate (near Berlin) on 19,947 kc so that echoes which travelled repeatedly around the earth might be studied. The signals were simultaneously observed and recorded with a cathode-ray oscillograph on moving film at Randers, Denmark, 480 km, and at Gatow, 50 km, from the transmitter. The time intervals measured ranged from 0.1376 to 0.1384 sec between first and second circuits, and are shown graphically together with their individual amplitudes. Periodic fades of the echoes are correlated with multiple paths of propagation and the vertical motions of the ionospheric reflector. The high field intensity of repeated signals is evidence of strong focusing of the hf energy since the propagation seems to occur in a narrow great-circle beam**

I. Introduction

IN HIS PREVIOUS PAPERS[1,2] the author has reported hf echoes which occurred in long-distance ionospheric propagation between 10 and 20 mc, during the years 1941 to 1945, at the minimum of the sunspot cycle. These echoes were characterized by a rather long time interval, since indirect or reverse signals reached the receiving position along the opposite great-circle path; moreover, repeated circuits of direct and indirect signals were frequently observed. The accuracy of the measured time intervals was 5.10^{-5} sec, and the individually measured values for complete circuits of the earth were between 0.1376 and 0.1381 sec. In a recent publication of the author[3] an account is given of investigations of those signals which travelled once, twice, and three times around the globe. During the winter 1949 to 1950, approximately the maximum of the sunspot cycle, measurements of this kind were repeated by Kootwijk-Radio, Holland.[4] The radio frequencies used varied between 10 and 22 mc, and the measured time intervals were between 0.137 and 0.139 sec, to an accuracy of 10^{-3} sec. During the past maximum no measurements were made which surpassed an accuracy of 10^{-4} sec; therefore, no conclusions can be reached about the stability of the average value of 0.13778 sec for a complete circuit. This problem is of utmost interest because, after essentially changed ionospheric conditions, displacement of the optimal frequency spectrum for occurrence of round-the-world echoes and different effective F_2-layer heights should be expected.

II. Observations and Measurements

On November 19, 1944 short signals were sent out each half-hour for a 5-minute period from the hf commercial station DLO, Rehmate (near Berlin) on 19,947 kc, starting at 0655 to 0700 and closing at 1025 to 1030 GMT. The transmitting power was 40 kw, and a directional aerial toward the northeast (Japan) was used to suppress reverse circulating signals. The transmitted signals were simultaneously observed at Randers, Den-

* Decimal classification: R112.4. Original manuscript received by the Institute, June 13, 1951; revised manuscript received, April 22, 1952. For Parts I and II, see H. A. Hess, "Part I—investigations of high-frequency echoes," vol. 36, pp. 981–992; August, 1948 and H. A. Hess, "Part II—investigation of high-frequency echoes," vol. 37, pp. 986–989; September, 1949.

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[1] H. A. Hess, "Investigations of high-frequency echoes," Proc. I.R.E., vol. 36, pp. 981–992; August, 1948.

[2] H. A. Hess, "Investigations of high-frequency echoes—part II," Proc. I.R.E., vol. 37, pp. 986–989; September, 1949.

[3] H. A. Hess, "Studien an mehrfachen Kurzwellenumläufen," *Fernmeldetechn. Z.*, no. 7, pp. 243–248; July, 1950.

[4] A. H. de Voogt, "Les échos radioélectriques autour de la terre," *Onde Elect.*, no. 283, pp. 1–5; October, 1950.

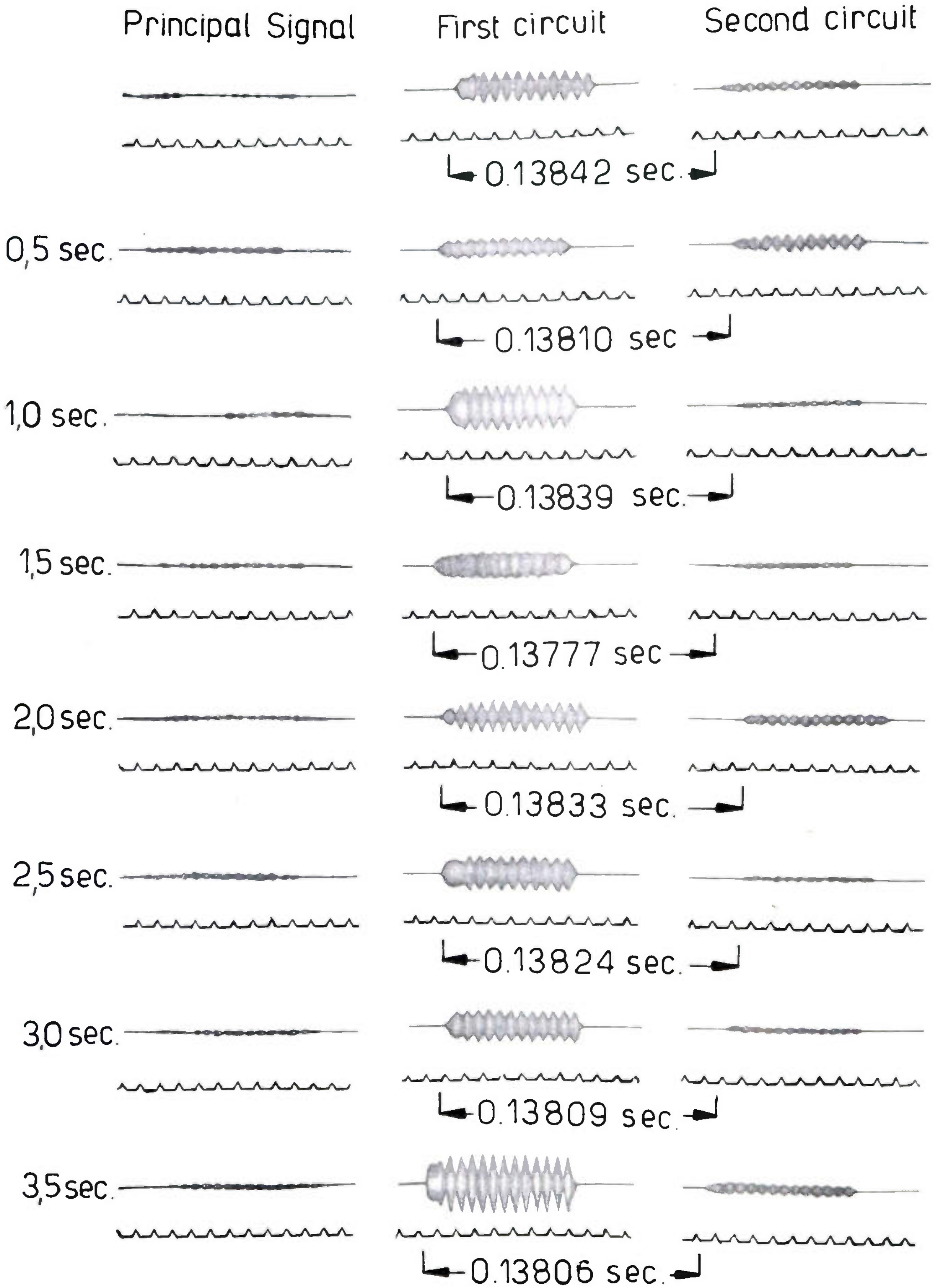

Fig. 1—Recordings made at 0.5-second intervals on DLO, 19,947 kc on November 19, 1944, 0855 GMT, showing detour signals and first and second circuits.

mark 56°31′N, 10°02′E, and at Gatow 52°27′N, 13°11′E. For the experiment at Randers a long wire aerial of 7λ at 20 mc directed from northeast to southwest was used to give maximal amplitudes of repeated circuits.

Fig. 1 is an original record made on November 19, 1944, 0855 GMT, at Randers during the period of maximum occurrence of multiple echo circuits, in which are shown successively the principal or detour signals and the first and second circuits during an interval of 3.5 sec. No direct signal could be received at Randers, 480 km from the transmitter, since the skip zone extended to 2,000 km at this time. The first arriving, indistinct, weak signal was like a scattered reflection which had detoured approximately 4,000 km and had been reflected from the ionosphere outside of the skip zone, as found in the author's earlier investigations.[5] The strongest signal was the first circuit. It was followed by the second circuit after 0.138 sec, and the third circuit appeared on a few film records after the same interval. At Gatow, 50 km from the transmitter, the direct signal was received with a rather strong intensity, immediately accompanied by scattered reflections, and the first circuit was received 0.13778 sec after the direct signal. Second and third circuit signals were not observed at Gatow, evidently because the level of local disturbances was stronger than at Randers, and suitable antennas were not used for the reception. The signals of DLO were modulated about 50 per cent with 900 cps, which made possible further conclusions, since the marked distortion of the signals connected with selective fading is frequently perceived. The lengths of the individual circuit signals seemed to be equal within a fraction of 1 msec, while signals formed by scatterings are frequently characterized by a length much greater than that of the circuits.

Direct signals; detour signals; first, second, and third circuit signals are, if present, recorded on film every 0.5 sec. All time-interval measurements are referred to the start of the individual signals to avoid errors due to the continuous fluctuations in the amplitudes and the deformation of the signals.

Fig. 2 shows a graphical evaluation of four films recorded at intervals of 30 minutes during the optimal occurrence of the echoes, giving the time intervals between the first, second, and third circuits, and their relative amplitudes. The measurements indicate a variation of the time intervals between 0.1376 and 0.1385 sec, and variations up to 0.5 msec characteristically occur in the 0.5-sec period. The average value of the time interval between the first and second circuit was 0.13784 sec at 0756 GMT, and rose uniformly to 0.13805 sec at 0927 GMT. The amplitudes of the circuits show strong fluctuations with different high minima and maxima. It was found occasionally that the intensities of first and second circuits were approximately the same, while the average amplitudes of the first and second circuits were in the ratio 3 to 1.

[5] H. A. Hess, "Untersuchungen an Kurzwellen-Echosignalen, II Teil," *Z. Naturf.*, vol. 2(a), no. 9, p. 528; September, 1947.

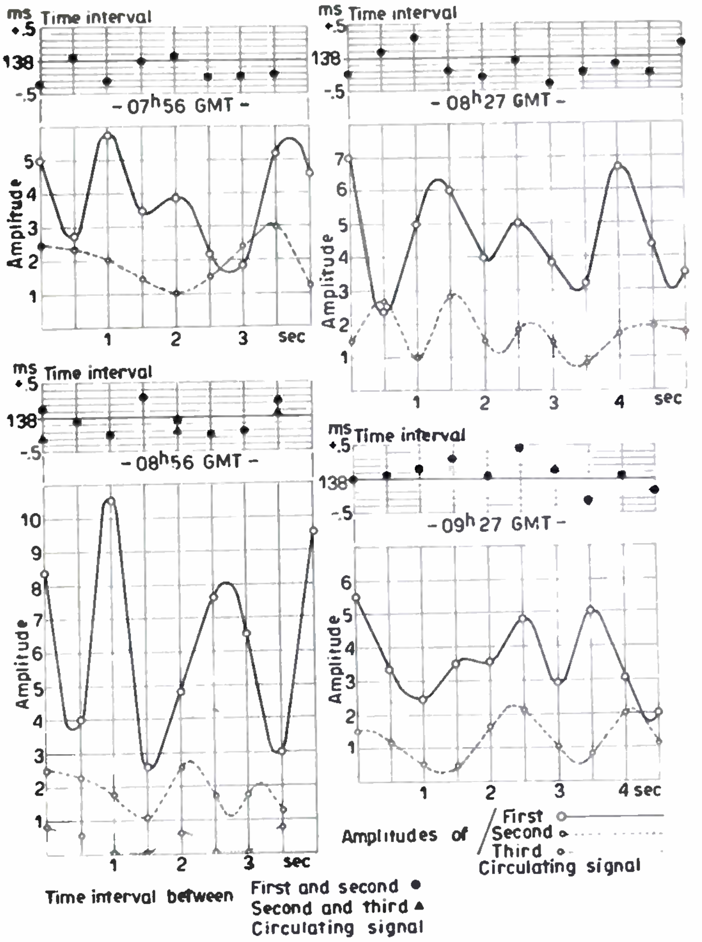

Fig. 2—Graphical illustrations of the measured time intervals between first and second circuits and their amplitude ratios.

In Fig. 3 are shown postulated propagation paths for the circuits for November 19, 0900 GMT, when the Berlin-Randers line was approximately perpendicular

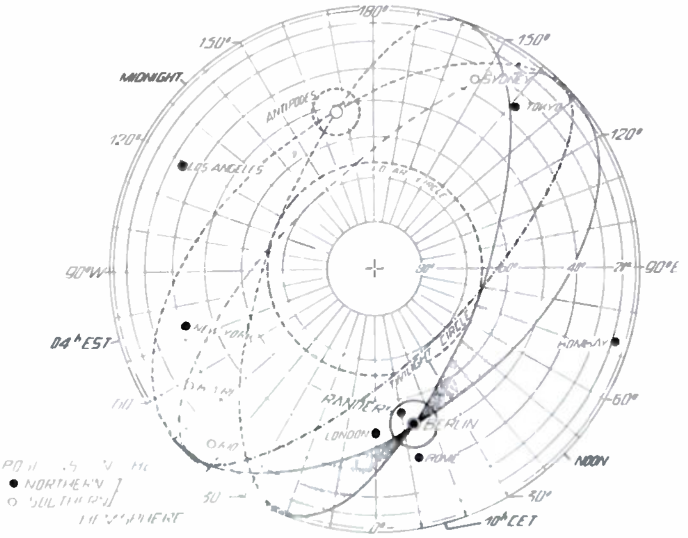

Fig. 3—Propagation on a spherical sector within the echo girdle on November 19, 0900 GMT.

to the great circle line of the twilight zone. The signals sent out from Berlin with a directive antenna to the northeast (Japan), toward the direction of the twilight zone, reached Randers from the southwest (South America). This was confirmed by experiments with individual directive antennas. On the route of about 41,300 km around the globe, the circuit signal had evidently deviated from its original direction since it was received both at Berlin and at Randers. A smaller deviation is to be expected for those multiple circuits which must have travelled within the space between Berlin and Randers on a serpentine path around the globe. Considering the general ionospheric conditions near the twilight zone, it is evident that the propagation of the circuits never occurs along only one path; a beam of great circle lines, starting from Berlin, cutting the antipodes, and returning to the transmitting position, probably exists. Evidently, strong focusing is effected at the antipodes and at the transmitting position. In consequence of its insignificant deviation from the great circle line, Randers, 480 km distant from Berlin, is considered to be situated within the focusing zone.

The fact that the intervals between the individual circuits differ by only a half millisecond is surprising, but can be explained by means of geometrical optics[5] if very low angles of arrival are assumed in sky-wave propagation.

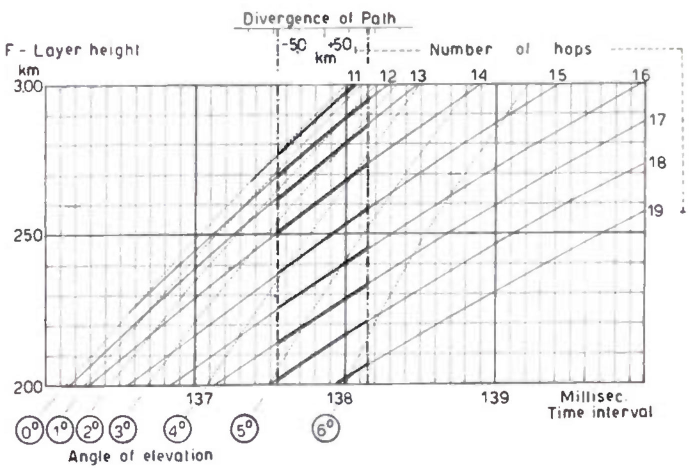

Fig. 4—Time intervals of circuits at 11 to 19 hops between the F_2-layer and earth's surface, layer-heights 200 to 300 km, and angles of arrival between 0° and 6°.

Fig. 4 shows the ratio between the circuit period and the height of the reflecting F_2-layer at 11 to 19 hops around the earth. The angles of arrival (measured between the horizon and the point of the ionospheric reflection) are also indicated, as is the range of the periods of a complete earth's circuit, 0.1376 to 0.1381 sec, which was found by many hundreds of individual measurements. The graphs illustrate the possibility of a complete circuit with 11 to 15 hops with an F_2-layer height between 260 and 280 km; angles of arrival between 2° and 5° may be expected for this case.

The small variations of the time interval between the first and second circuits observed on November 19, 1944, 0800 to 0930 GMT, are probably caused by the movement of the echo girdle, which during the morning moves from southwest to northeast, at noon from east to west, and during the afternoon from southeast to northwest.

All reflected sky waves are characterized by phase shifts because of the continuous up-and-down movement of the ionosphere. The values of the so-called Doppler shifts depend on the velocity of the ionospheric layer, the angle of incidence at the ionosphere, and the number of ionospheric reflections. The phase shifts are the cause of periodic amplitude fluctuations if interference is obtained due to multiple paths of the propagation. The curves of Fig. 2 probably manifest these facts.

An experimental fact of importance may be mentioned relative to the fading period and signal amplitudes. One always has to wait for favorable amplitudes in order to make a photographic record. These conditions were usually obtained within a few minutes. The average recorded amplitudes of the circuits, therefore, exceeded the actual average. A maximum in echo strength means the accidental near-equality of the phases of many of the numerous interfering wave components.

III. Conclusions

The periodic fading and distortion of the modulation of signals which travelled once or twice around the globe indicate propagation by multiple paths between ionosphere and earth's surface. An analysis of the signals in terms of individual wave components corresponding to the various paths of propagation was impossible because of the very great number of such components and the extremely small time difference between their arrival. With regard to the possibility of later experiments using short pulses, separation of the various paths is doubtful. The strikingly high field strength of repeated circuits is mainly caused by the effect of multiple paths at different hops between the ionosphere and earth's surface, and by strong focusing of the radiated hf energy in a narrow beam of great circle lines within the echo girdle. The measurements were carried out during the minimum of the eleven-year sunspot cycle, when 20 mc was approximately the maximum usable frequency for ordinary-ray F_2-reflections.

[5] K. Rawer, "Quelques effets importants de l'optique géométrique de l'ionosphère," *Rev. Sci.* (Paris), vol. 86, no. 3296, p. 481; 1948.

The Electric Polarizability of Apertures of Arbitrary Shape*

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Summary—An electrolytic-tank method of measurement developed in a previous paper has been used to obtain extensive data on the electric polarizability of apertures of various practical shapes. These shapes include rectangular slots, rounded slots, crossed slots, rosettes, and dumbbells. The measured values are presented in this paper in both graphical and tabular form. The accuracy is believed to be of the order of one or two per cent.

Introduction

In a previous paper by this author, a technique for accurately measuring the magnetic and electric polarizabilities of apertures was described and extensive magnetic-polarizability data were given.[1] Similar data are now available on electric polarizabilities, and are presented in this paper.

The significance of the magnetic polarizability M and the electric polarizability P is discussed in the paper referred to above, and a number of other references are listed there that show how these two parameters may be used in the calculation of electromagnetic coupling between any two regions separated by a thin wall containing an aperture small compared to a wavelength. The measurement technique, which utilizes an electrolytic cell, is fully described in the earlier paper, and the necessary formulas are derived for the electric polarizability as well as the magnetic polarizability.

In addition to the applicability of the electric polarizability data to aperture design, the data may also be used to compute the effective permeability of an array of thin conducting obstacles. The method is described in another paper by this author.[2]

Measurement Technique

The electrolytic cell utilized for electric-polarizability measurements is shown in Fig. 1. The inside dimensions are 6 × 6 × 6 inches, and the height of the solution is maintained at approximately 5½ inches. The two shaded walls are internally plated with rhodium, while the remaining walls are nonconducting lucite. A thin nonconducting obstacle having the shape of the aperture for which data is desired is suspended by two fine nylon threads at the center of the solution in a plane parallel

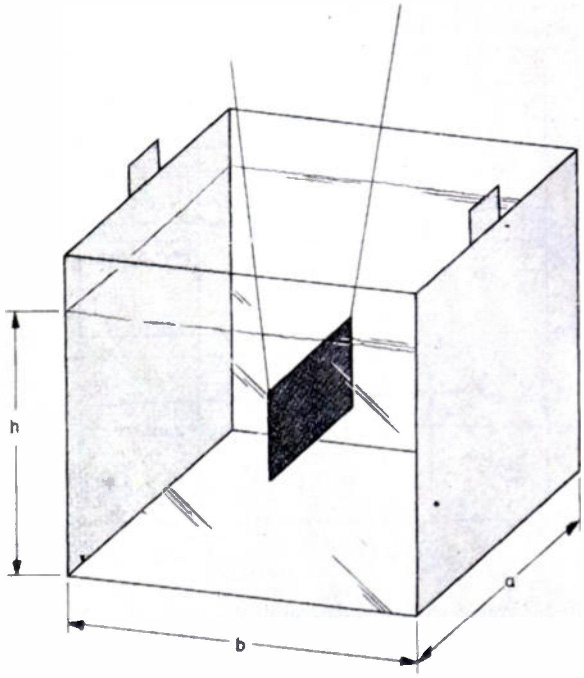

Fig. 1—Electrolytic cell containing a thin nonconducting obstacle.

to the conducting surfaces of the cell. The obstacles are cut from polystyrene sheet 0.005 inch thick, and are attached to the threads by a minimum quantity of cement. The electric polarizability is computed from the measured data by means of the following formula:

$$P = \frac{abh}{4}\left(\frac{R_1 - R_2}{R_1}\right), \tag{1}$$

where a, b, and h are the dimensions of the conducting solution in centimeters (Fig. 1), R_1 is the resistance of the cell with an obstacle in position, and R_2 is the resistance with the obstacle removed. The electric polarizability P has the units cm³, and applies to an aperture having the same shape and size as the obstacle. For further details on measurement technique, the author's previous paper should be consulted.[1]

Measurements were first made on a series of circular obstacles of various diameters in order to determine the maximum diameter for which the effect of proximity to the cell walls could be neglected. The theoretical value for an isolated circular aperture in an infinitely thin conducting wall is $P/d^3 = 1/12 = 0.08333 \ldots$ The experimental curve shown in Fig. 2 crosses this value at about $d = 2.75$ inches. For larger diameters the measured points are decreased by the proximity effect. For smaller diameters the values are increased by the finite obstacle

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† Sperry Gyroscope Co., Great Neck, Long Island, N. Y.

[1] S. B. Cohn, "Determination of aperture parameters by electrolytic-tank measurements," Proc. I.R.E., vol. 39, pp. 1416–1421; November, 1951. Also see vol. 40, p. 33; January, 1952.

[2] S. B. Cohn, "The electric and magnetic constants of metallic delay media," *Jour. Appl. Phys.*, vol. 22, pp. 628–634; May, 1951.

thickness. These effects are seen to be very small, however, since the measured curve is within 0.5 per cent of the correct value from $d=2.0$ to 3.2 inches. As an example of the effect of thickness, one obstacle 2.5 inches in diameter and 0.015-inch thick was tested and found to have a value of P/d^3, one per cent more than that obtained with the same diameter obstacle 0.005-inch thick.

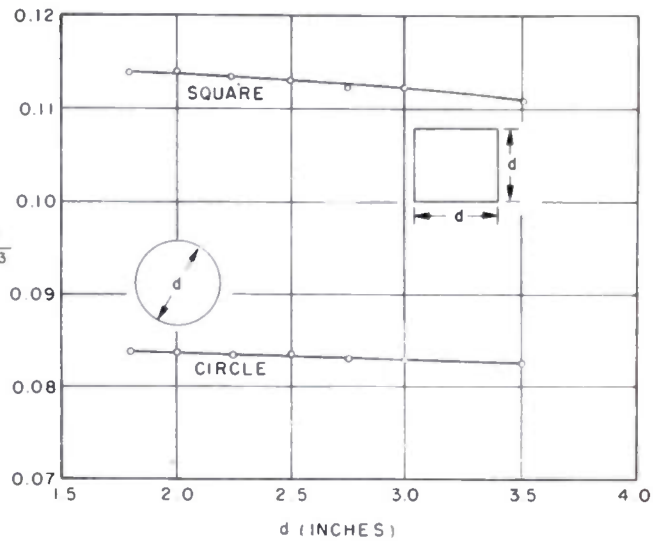

Fig. 2—Measured electric polarizability as a function of size for circular and square obstacles in a 6-inch cubical cell.

The curve for a series of square obstacles 0.005-inch thick is also given in Fig. 2. It is believed that this curve crosses the correct value, as in the case of the circles. Since the correct value is not known in advance, it will be assumed that the crossover point occurs for a square whose dimension d is such that the length 2.75 inches lies midway between d and the diagonal length $\sqrt{2}d$. This leads to a value $d=2.3$ inches and $P/d^3=0.1137$. The deviation from this value does not exceed 0.5 per cent for d between 1.8 and 2.6 inches.

For other shapes of obstacles it is assumed that correct results are obtained if the obstacles are made to fit the composite boundary formed by the superposition of a circle of diameter 2.75 inches and a square whose side is 2.3 inches. (This type of construction is illustrated in Fig. 5 of footnote reference 1.) It is believed that the error resulting from this assumption is very small and can be neglected.

The Electric Polarizability Data

The measured data are given in Figs. 3 and 4 for rectangular and rounded slots, crossed slots, rosettes, and dumbbells. The following theoretical formula for a *narrow* slot is also plotted in Fig. 3:

$$P = \frac{\pi}{16} lw^2, \qquad w/l \ll 1, \tag{2}$$

where w and l are dimensions defined in the figure. The agreement with the curves for rectangular and rounded slots is good for w/l up to 0.15, but the theoretical formula is increasingly in error for wider slots.

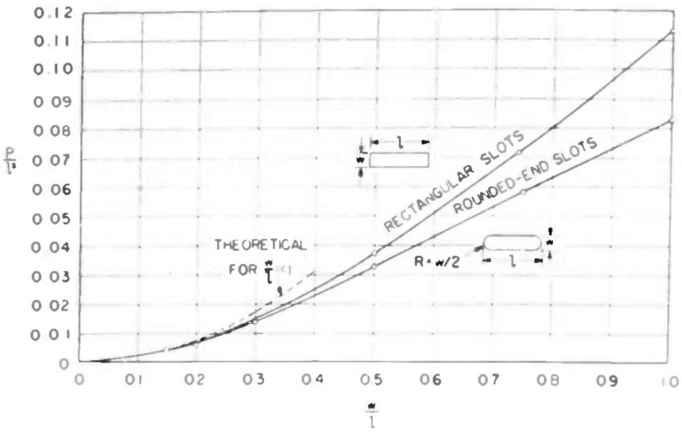

Fig. 3—Measured electric polarizability of rectangular and rounded slots.

It is of interest to note that for w/l small, P/l^3 for the crosses is approximately twice that for the rounded slot, while for the rosette, the ratio is approximately four. This is to be expected since each cross consists of two intersecting rounded slots while each rosette consists of four intersecting rounded slots. This correspondence holds very closely for the crosses for w/l up to at least 0.35, while for the rosettes it holds well only for w/l less than 0.1.

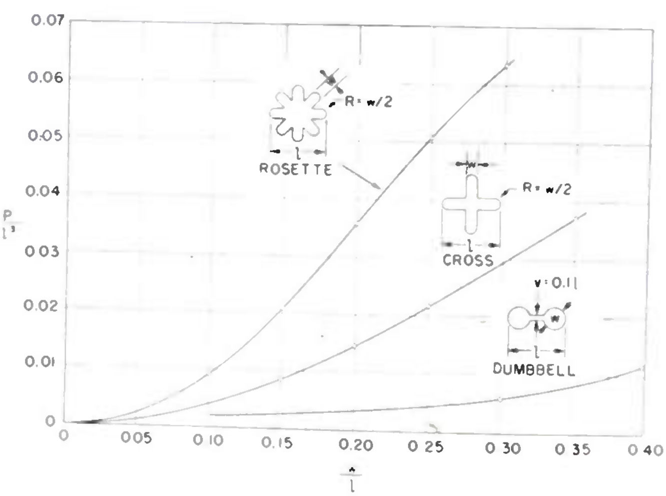

Fig. 4—Measured electric polarizability of cross, rosette, and dumbbell apertures.

The curve in Fig. 4 for the dumbbell aperture applies to a bar width equal to one-tenth of the total length, and would differ somewhat for any other width. One would expect the electric polarizability of the dumbbell to be approximately equal to the sum of the electric polarizabilities of two circular apertures plus that of a slot. The following empirical formula, which is based on this hypothesis, agrees within 2 per cent with the curve in Fig. 4, and should give good results for other bar widths.

$$P = \frac{w^3}{6} + \frac{\pi}{16}(l - w)v^2. \tag{3}$$

The dimensions l, w, and v are defined in Fig. 4.

Points taken from the curves are listed in Table I for the five aperture shapes tested. It is believed that these values are accurate within 1 or 2 per cent.

Conclusion

The electrolytic-tank method for measuring the electric polarizability of an aperture has been tested against the theoretical formulas for a circle and a narrow slot, and has been found to be capable of very high accuracy. The simplicity and economy of the method has made possible the accumulation in this paper of a large amount of data on practical aperture shapes not amenable to theoretical calculation. This data, together with the magnetic-polarizability data obtained previously by a similar method, should increase significantly the utility of Bethe's small-aperture coupling theory in the design of microwave devices.

TABLE I

Values of P/l^3

$w/l=$	0.1	0.15	0.2	0.25	0.3	0.4	0.5	0.75	1.0
Rectangle	0.0019	0.0041	0.0070		0.0147		0.0370	0.0731	0.1137
Rounded slot	0.0019	0.0041	0.0070		0.0143		0.0325	0.0585	0.0833
Cross	0.0039	0.0085	0.0144	0.0217	0.0293				
Rosette	0.0090	0.0209	0.0357	0.0508	0.0633				
Dumbbell*	0.0019				0.0058	0.0117			

*Width of bar $=0.1l$.

Multi-Element Directional Couplers*

S. E. MILLER†, MEMBER, IRE AND W. W. MUMFORD†, FELLOW, IRE

Summary—**It is shown that the backward wave in a directional coupler is related to the shape of the function describing the coupling between transmission lines by the Fourier transform. This facilitates the design of directional couplers for arbitrary directivities over any prescribed frequency band. Tightly coupled directional couplers are analyzed in simple terms, and it is shown that any desired loss ratio, including complete power transfer between lines, may be achieved. The theories are verified using waveguide models operating at 4,000, 24,000, and 48,000 mc, and it is indicated that the work is applicable to many types of electrical and acoustic transmission lines.**

THIS PAPER presents a theoretical approach which facilitates the handling of directional couplers using any number of coupling elements from two to an infinite number. In addition there is presented a simple theory accounting for some of the performance characteristics of tightly coupled directional couplers. Observations on a model having less than $\frac{3}{4}$-db coupling loss over an 18-per cent frequency band are to be given, as well as observations on models having higher losses.

Consider first the case of loose coupling. Fig. 1 shows diagrammatically two identical transmission lines each of which is parallel to the direction of propagation, along the x-axis. The region in which coupling exists between the two lines is confined to the interval $-L/2$ to $+L/2$. The variation of coupling between the lines is described by the function $\phi(x)$. The coupling may be either continuous or a series of discrete couplings. On the assumption of an exciting wave traveling to the right in line number two, the sum of all the forward current elements, referred to the plane $x=+L/2$, is given by

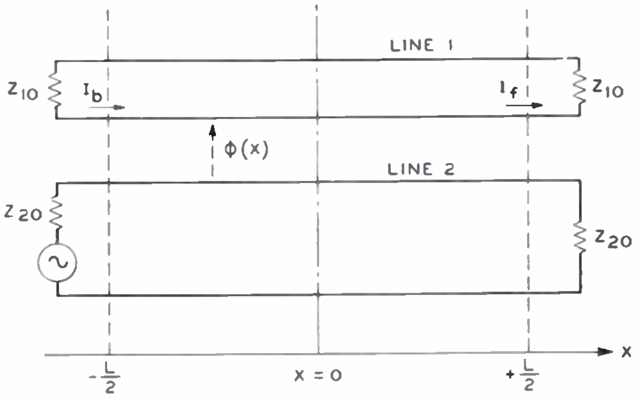

Fig. 1—Directivity derivation.

$$I_f = KF\int_{-L/2}^{+L/2}\phi(x)dx, \tag{1}$$

where

$$F = \frac{\epsilon^{-i(2\pi L/\lambda_g)}}{2Z_{10}}.$$

The factor K represents the fraction of the total induced current which travels forward in the undriven line. K is a measure of the directionality of the coupling on a differential length basis. The sum of all the backward current elements, referred to the plane $x=-L/2$, is given by

$$I_b = (1-K)F\int_{-L/2}^{+L/2}\phi(x)\epsilon^{-i(4\pi/\lambda_g)x}dx \cdots. \tag{2}$$

The ratio of the forward to the backward current is, of course, the directivity.

* Decimal classification: R310.4. Original manuscript received by the Institute, June 4, 1951; revised manuscript received May 11, 1952. Paper originally presented at the March, 1951 IRE National Convention.

† Bell Telephone Laboratories, Inc., Holmdel, N. J.

$$\text{Directivity} = \frac{I_f}{I_b}. \qquad (3)$$

As long as the phase of the coupling function $\phi(x)$ does not change between $-L/2$ and $L/2$, the forward current elements all add in phase. However, the backward current elements add in a form of destructive interference. The backward current expression has the form of the Fourier transform,[1] thus permitting the use of experience gained with the time and frequency domain relations in designing directivity characteristics. Another body of experience which bears on this problem is to be found in antenna design work. It turns out that the relation between the principal beam and the minor lobes of an antenna is related to the current excitation along the antenna in the same way that the directivity of a directional coupler is related to the shape of the coupling function. Of particular note in this regard is the work reported by Dolph[2] which can be interpreted to provide the optimum taper of coupling for minimizing the required coupling length in order to achieve a given amount of directivity over a broad band.

Consider a familiar example. Suppose the coupling between the two transmission lines is uniform over the interval L, as in Fig. 2. Then the directivity is given by

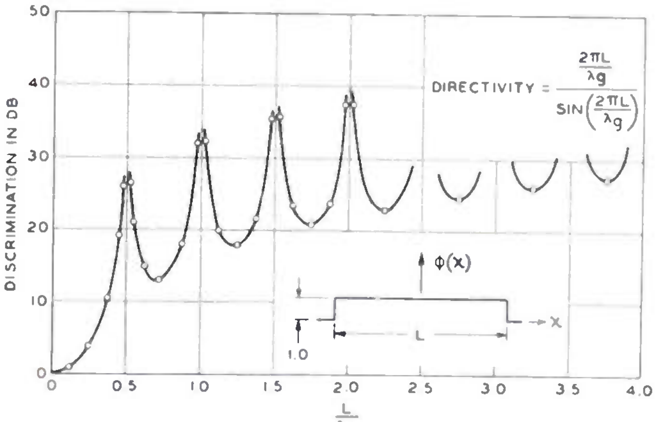

Fig. 2—Computed directivity for constant amplitude and phase of coupling.

the inverse of the familiar $\sin u/u$ function with $u = 2\pi L/\lambda_g$. The directivity is perfect for $L/\lambda_g = \frac{1}{2}$, 1, 1.5, and so on. However, the minima in directivity fall off as L/λ_g and a coupling interval of approximately three wavelengths is required in order to get broad-band directivity on the order of 25 db. There are a number of ways in which higher directivity can be obtained over a broad-band in a shorter length interval, two of which will be used as illustrations.

[1] The theoretical Fourier transform relation for the backward current was pointed out (after the authors had completed their work) by Folke Bolinder in a letter to the editor, Proc. I.R.E., vol. 39, p. 291; March, 1951. Also, the theoretical capabilities of distributed coupling in improving directivity characteristics were considered in some unpublished work of the late Arnold E. Bowen.

[2] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," Proc. I.R.E., vol. 34, pp. 335–348; June, 1946.

Suppose the coupling function is a linear taper as shown in Fig. 3. Then the locus of minimum directivity points falls off as $(L/\lambda_g)^2$ and a coupling interval on the order of one wavelength long produces broad-band directivity in excess of 25 db. At a length slightly greater than two wavelengths, 35-db directivity can be maintained over a very broad-band.

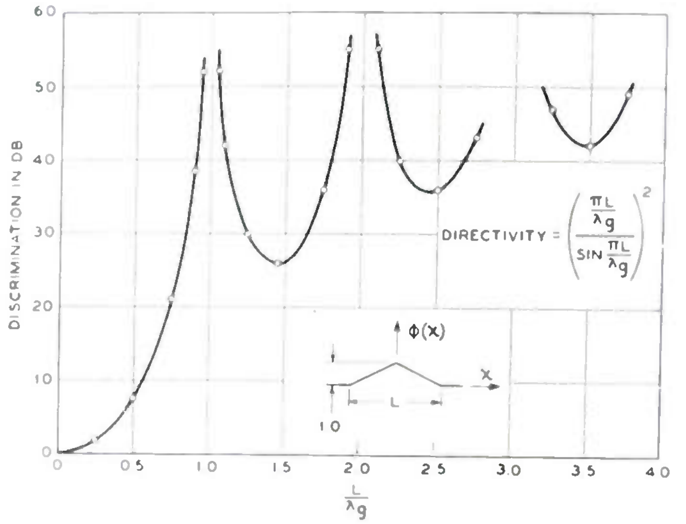

Fig. 3—Computed directivity for linear taper form of coupling.

Fig. 4 illustrates another way of achieving high directivity in a relatively short-length interval which is applicable when the bandwidth of interest is on the order of 40 per cent or less. This is frequently the case in waveguide work. This complex function, $\phi(x)$, may be

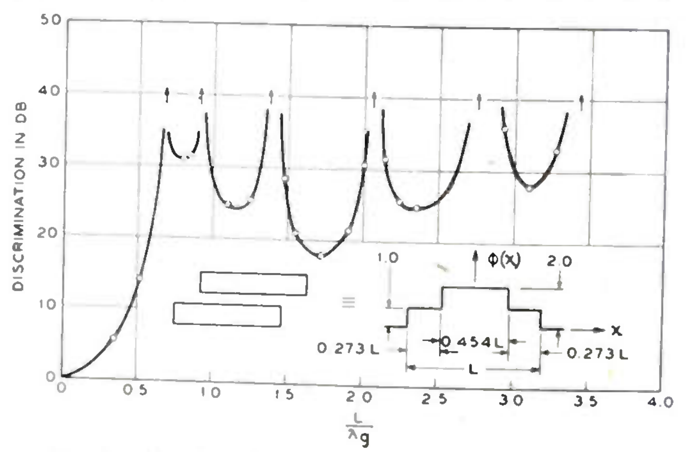

Fig. 4—Computed directivity for two constant-amplitude couplings superposed.

thought of as resulting from the linear superposition of two identical constant-amplitude coupling functions of equal length but displaced from each other along the x-axis. The net result is that there are two frequencies where infinite directivity should be observed and can be chosen independently. In this figure a choice has been made to place the infinite directivity points at $L/\lambda_g = 0.67$ and 0.87. As a result, the directivity is greater than 30 db over approximately ± 10-per cent band. The mean length of the coupling interval is approximately 0.8 wavelength.

Let us now examine some results of observations. Fig. 5 shows a jig used for measuring the directivity of a number of coupling arrangements in the frequency region near 4,000 mc. Coupling is achieved through holes in the small side of a rectangular waveguide. The common wall of the adjacent waveguides is made in the form of an insert which can be easily replaced during the course of the experiment.

In an attempt to realize uniform coupling between waveguides, one is tempted to use the rectangular slot as shown in Fig. 6. Without careful thought, one is likely to expect this slot to provide coupling between the lines on a differential length basis. The resulting directivity should then be the $\sin u/u$ form of directivity curve as previously shown. This is not the way the slot actually works. What actually happens is that the slot itself acts as a transmission line more or less independently of the adjacent waveguides. The wave runs along the slot from end to end with a high coefficient of reflection at the slot ends. Even when the slot is made several wavelengths long, where the theoretical directivity exceeds 20 db, the observed directivity is of the order of ± 5 db and the arrangement is characterized by high standing waves presented to the exciting wave.

In the coupling arrangement illustrated in Fig. 7, wires have been soldered at equal intervals along the length of the slot. The traveling waves which tend to form in the slot are localized between wires and, in effect, discrete couplings located at the center point between wires are produced. The theoretical and observed results associated with this insert are given in Fig. 8.

The solid curve marked by x's shows the theoretical directivity for the slot shown when a very large number of wires is placed at equally spaced intervals. This is a

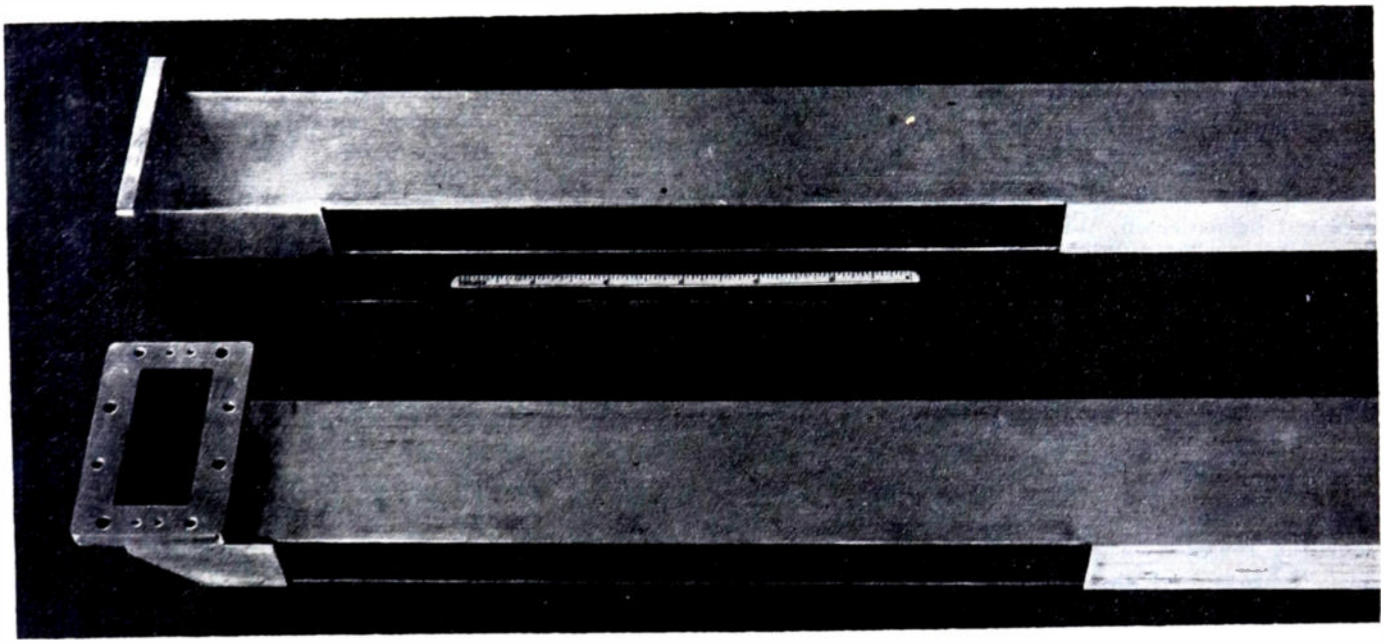

Fig. 5—4,000-mc directional coupler test jig.

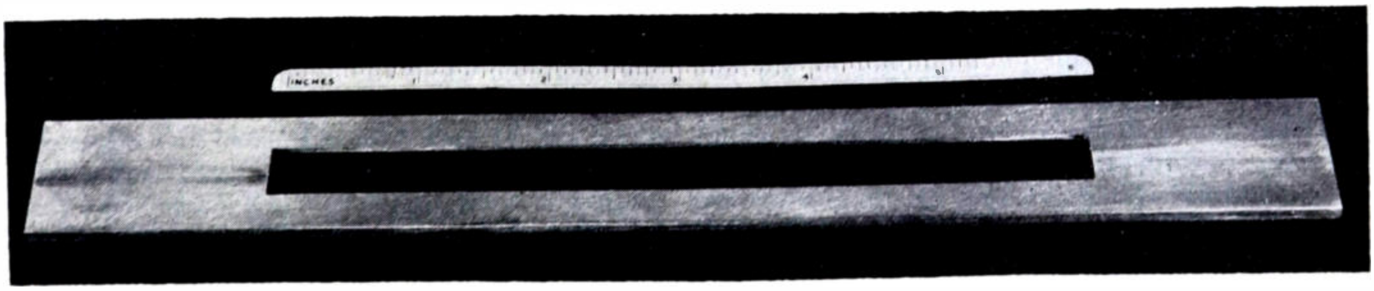

Fig. 6—Open slot coupling insert.

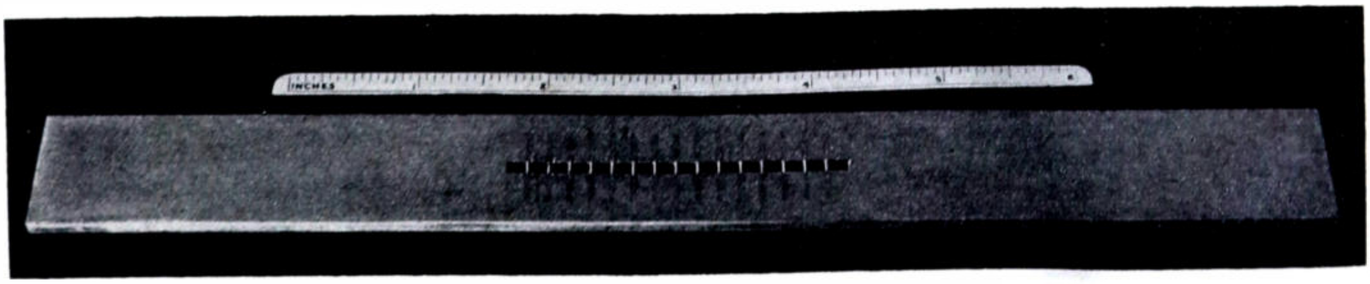

Fig. 7—Constant amplitude coupling insert.

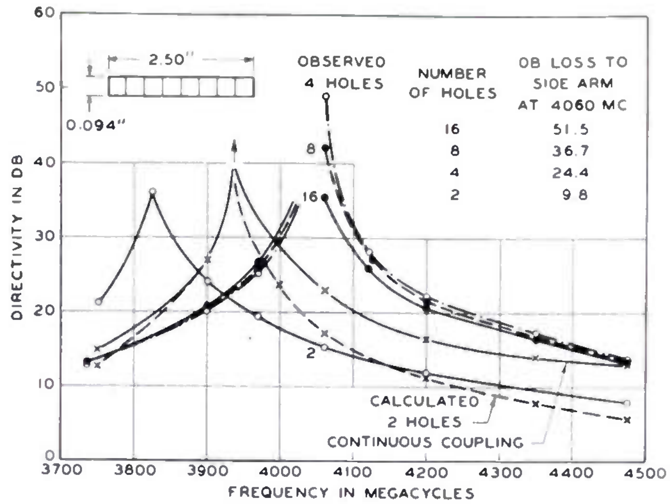

Fig. 8—Observed versus computed directivity for constant amplitude coupling. Number of holes varied from 2 to 16.

case of continuous coupling, and this curve is the portion of the $\sin u/u$ curve in the vicinity of 3,930 mc, where $L/\lambda_g = \frac{1}{2}$. When fewer wires are used, the theoretical directivity changes very little, and for as few as 3 wires, giving 4 equally spaced holes, the theoretical curves lie so close to the case of continuous coupling that they have not been drawn. With only one wire in the slot, there results 2 quarter-wave long holes with a center-to-center spacing of approximately one-quarter wavelength. The theoretical directivity in this case is the familiar cosine function and is drawn with x's and a dashed line. Consider now the observational results. With 15 equally spaced wires (16 holes) the resulting directivity is shown by the solid line at the right. The peak of directivity is displaced from the theoretical peak by approximately 7 per cent in guide wavelengths. After the removal of every other wire, producing 8 holes, the observed directivity is given by the dashed line. After the removal of every other wire again, producing 4 holes, the measured directivity is shown by the curve identified by circles. We observe that the directivity is essentially independent of the number of holes in the range of 8 or more holes per wavelength. For this range of 4 to 16 holes, the coupling loss is observed to be between 25 and 50 db and the loose coupling assumption under which the theoretical directivity is derived is actually justified. However, when alternate wires are again removed, leaving only 2 holes, the coupling loss is approximately 10 db—definitely not loose coupling. The observed peak in directivity is then shifted approximately 8 per cent in the longer wavelength direction from the theoretical peak. The shape, however, is very similar to the theoretical one.

Fig. 9 shows the observed directivity when the length of the slot and the number of holes are maintained constant, but the height of the slot is varied. Directivity should be independent of the amplitude of coupling and therefore independent of the height of the slot. The data given in Fig. 9 confirm this. The approximate coupling loss for each slot height is shown at the right. It is found that the current coupling is approximately proportional to the slot height, as one might expect. This characteristic makes it possible to build directional couplers using moderately complex coupling arrays without experimentation regarding hole size versus coupling loss.

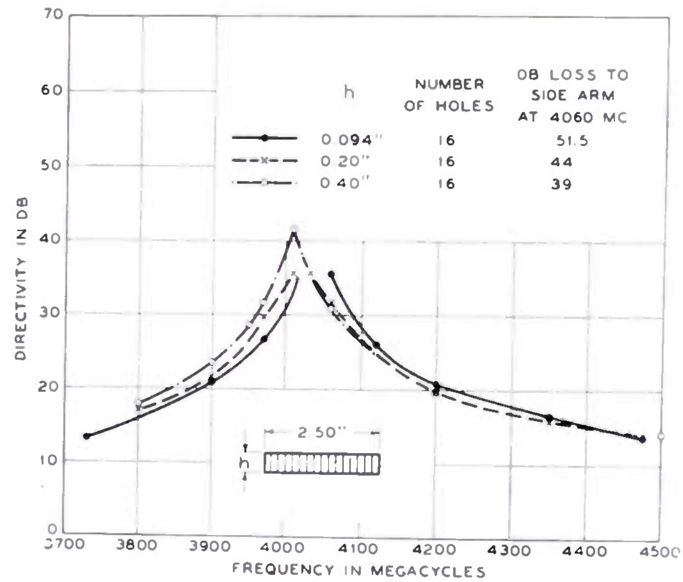

Fig. 9—Observed directivity for constant amplitude coupling when height of slot is changed.

Fig. 10 shows a coupling arrangement composed of two uniform couplings of the same length displaced with respect to each other along the longitudinal axis. Instead of using separate slots, however, a single slot has been used where the height is proportional to the total amplitude of coupling desired at the particular point. Fig. 11 shows the calculated and observed results. The calculated directivity based on continuous coupling is shown by the solid line marked with circles. For the model shown in Fig. 10, in which 22 holes were used, the observed coupling loss is 42 db and the observed directivity is as shown with the solid points.

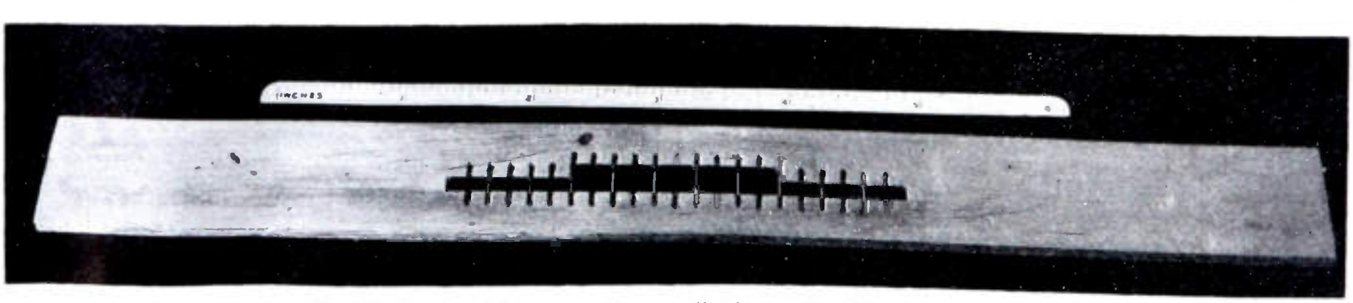

Fig. 10—Insert with two constant amplitude coupling functions superposed.

When alternate wires are removed, the loss drops to 28.6 db and the directivity observed is given by the crosses and dashed curve. The agreement between the observed and calculated directivity curves seems good evidence that the current coupling loss is proportional to the height of the slot. Other experimental evidence indicates that this is very nearly true as long as the coupling per hole is weak.

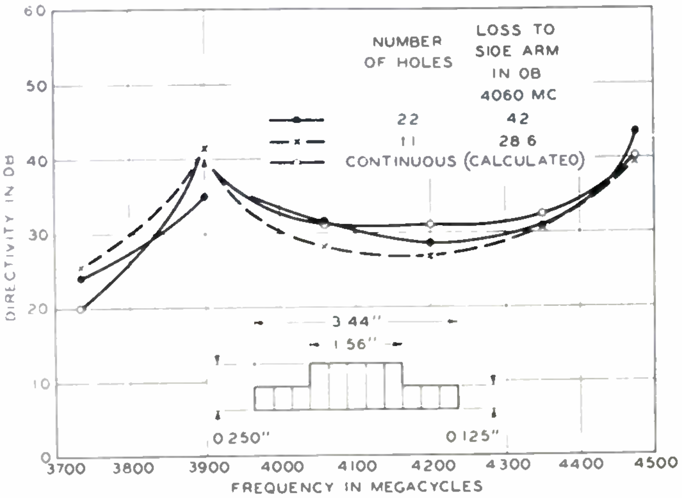

Fig. 11—Observed versus computed directivity for insert of Fig. 10.

Fig. 12 shows the linear taper form of coupling array, and the experimental and theoretical results are given in Fig. 13. The computed directivity based on continuous coupling is given by the dot-dash line with circles. The two peaks of directivity are the result of employing the linear taper plus a constant amplitude over the entire length, as shown in the sketch. The observed loss using 56 holes is 30.5 db and the observed directivity under this condition is given by the solid line with solid points. Except for a shift in frequency, the observed characteristic is quite similar to the computed one and directivity in excess of 35 db is observed over a frequency band of approximately 15 per cent. When alternate wires in this array are removed, leaving 28 holes, the coupling loss is 15.7 db and the directivity is given by the crosses and the dashed line. When alternate wires are again removed, leaving 14 holes, the observed loss is 5.6 db and the observed directivity is given by the solid line with circles. In this condition, there are about 7 holes per wavelength and a loss which is quite low.

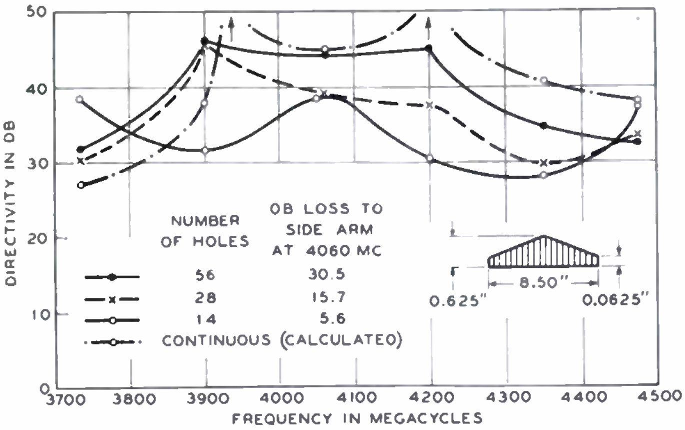

Fig. 13—Observed versus computed directivity for insert of Fig. 12.

Therefore, the departure from the theoretical directivity is not surprising. It is, however, interesting to note that the directivity is in excess of 28 db over a band exceeding 20 per cent.

In designing the models just described for use at 4,000 mc, no information on the particular waveguide was used except the guide wavelength which is calculable precisely from the dimensions of the guide. Therefore,

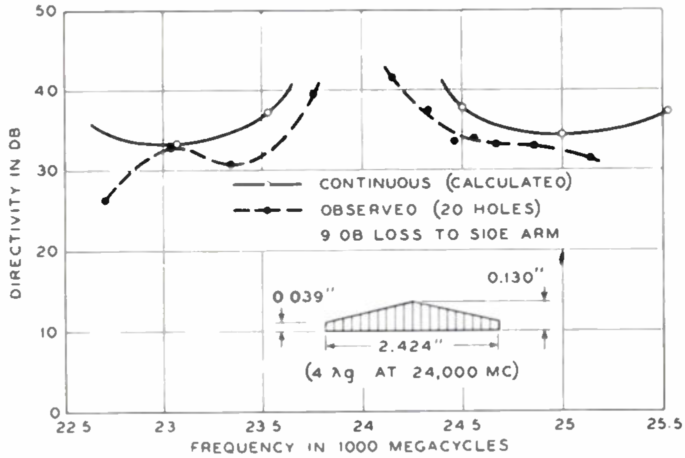

Fig. 14—Observed versus computed directivity for a 24,000-mc directional coupler.

the approach should be applicable directly to other waveguide sizes and other frequency ranges. This has been illustrated by building directional couplers at 24,000 and 48,000 mc.

Fig. 14 shows the observational results on a 24,000-

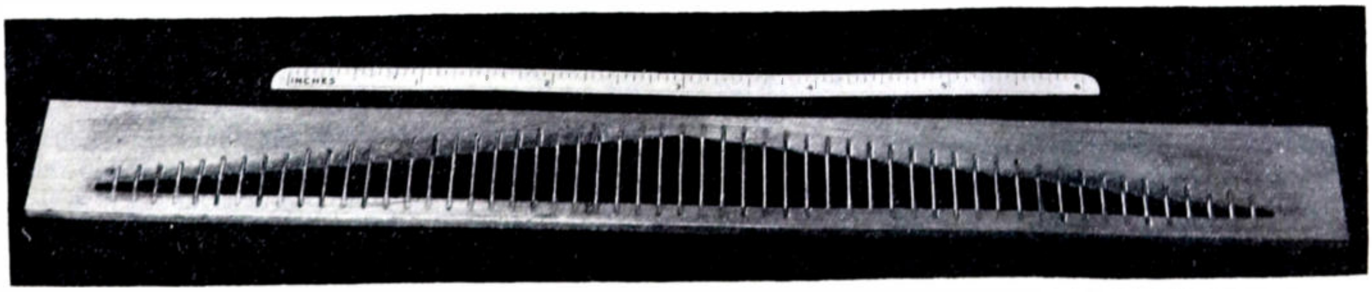

Fig. 12—Insert with linear taper plus constant amplitude coupling.

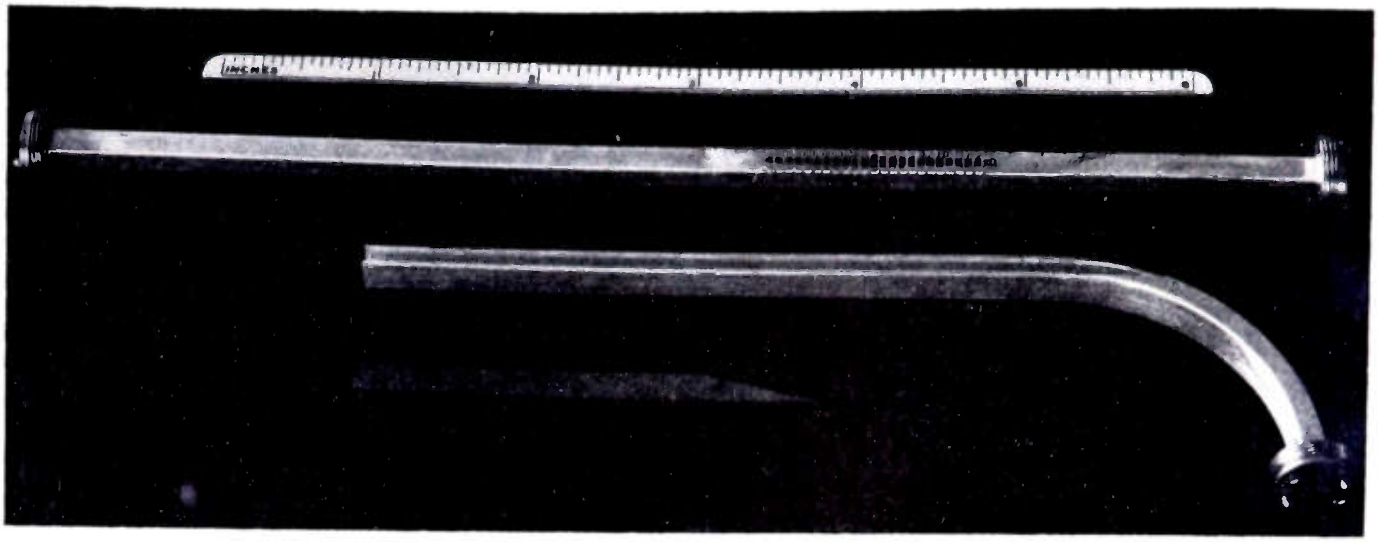

Fig. 15—48,000-mc directional coupler.

mc directional coupler wherein the coupling array was a linear taper superimposed on a constant amplitude of coupling. The total coupling interval is approximately 4 guide wavelengths and the observed loss for 20 holes was 9 db. The calculated directivity based on continuous coupling is given by the circles and the solid line. The observed directivity is given by the solid points and dashed line.

Fig. 15 shows a directional coupler made for operation at 48,000 mc. The inside dimensions of the waveguide are 0.094 by 0.188 inch. The length of the coupling array is 1.32 inches, approximately 4 guide wavelengths. The coupling array used is again a linear taper superimposed on a constant amplitude of coupling. The calculated directivity based on continuous coupling is given in Fig. 16 by the solid line and circles. The observed directivity using 24 holes is given by the solid points and dashed line. The coupling loss is 14 db.[3]

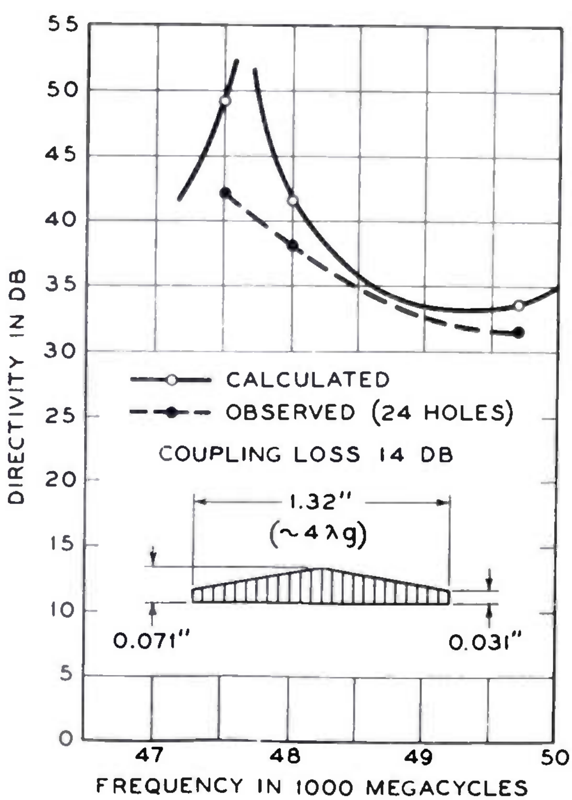

Fig. 16—Observed versus computed directivity for directional coupler of Fig. 15.

The results given so far have shown what can be done with equally spaced discrete couplings using amplitude tapers. Fig. 17 shows a coupling array which is easier to

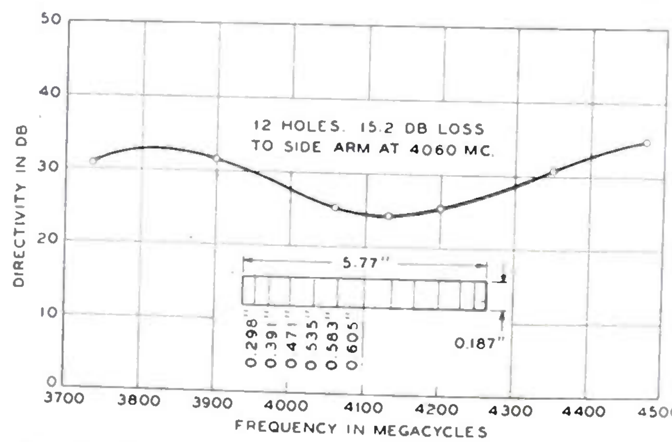

Fig. 17—Observed directivity for tapered amplitude and spacing of coupling elements.

build when the dimensions become small, as they do in waveguides above 50,000 mc. The amplitude of the coupling and the spacing between these couplings are both tapered. This is done by tapering the spacing between wires along a constant height slot. The observed directivity is given by the solid line. The coupling loss in this particular case is 15.2 db; similar couplers have been built with losses as low as 3 db.

The theoretical treatment of directivity already given was based on loose coupling, and the observational results have shown that using the loose coupling theory as a guide good directivities can be achieved with losses as small as 5 db. Consider now the tight-coupling case.

[3] Measurements on this model were made possible through the co-operation of A. G. Fox, Bell Telephone Laboratories, Holmdel, N. J.

Fig. 18 illustrates two identical lossless transmission lines symmetrically located about a means of continuous coupling. This coupling is assumed directional in the forward direction within a length interval in which negligible power is transferred between the lines. The effect of the coupling on the traveling waves in the two lines is given by the relations

$$\frac{dE_1}{dx} = -\alpha E_1 + \alpha E_2; \tag{4}$$

$$\frac{dE_2}{dx} = +\alpha E_1 - \alpha E_2. \tag{5}$$

The solution for the case where an input signal of magnitude unity is impressed on line 1 and no input is applied to line 2 is given by the relations

$$E_1 = \tfrac{1}{2}(1 + \epsilon^{-2\alpha x}) \tag{6}$$

$$E_2 = \tfrac{1}{2}(1 - \epsilon^{-2\alpha x}). \tag{7}$$

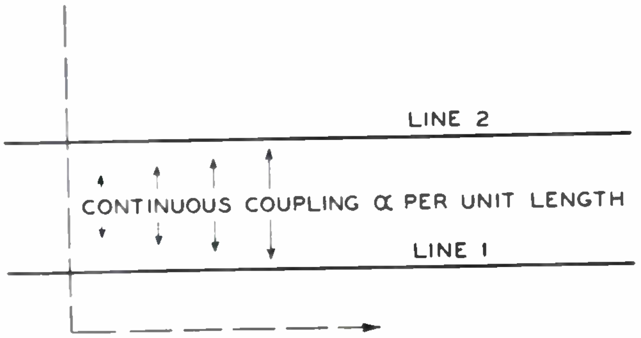

Fig. 18—Tight coupling relations.

The magnitude and relative phase of the waves on the two lines are shown in Fig. 19. The ordinate of the lower chart represents the magnitude of the wave on

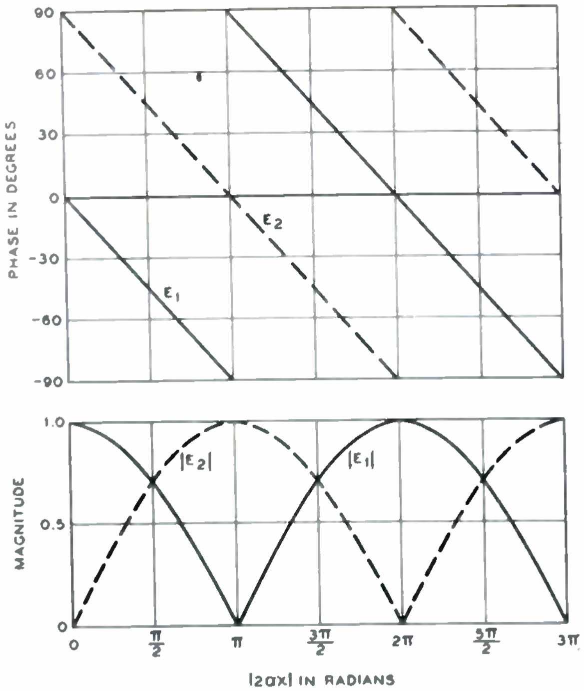

Fig. 19—Amplitude and phase of waves in a tightly coupled directional coupler.

either line, and the abscissa represents an integrated coupling magnitude: the product of coupling per unit length α times the distance x over which coupling is maintained. The driven-line wave magnitude declines co-sinusoidally and the undriven-line wave magnitude increases sinusoidally as the coupling is increased. Complete power transfer between the lines takes place,[4] and repeats cyclically as long as coupling is maintained. The coupling may be broken at any point where the waves in the two lines have a relation which it is desired to preserve. Thus, hybrids with any desired loss ratio may be readily formed.

The upper chart shows the phase of the wave in each line relative to that which would prevail in a wave traveling in a similar transmission line without coupling. The driving wave experiences a phase advance, whereas the wave in the side line is delayed for small couplings. This delay in the side line wave goes to zero at the point where complete power transfer occurs. Note that there is always a 90° phase difference between the two lines.

A physical picture of the power transfer is obtained from Fig. 20, which represents two transmission lines

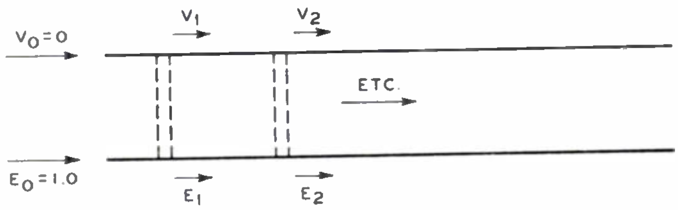

Fig. 20—Diagram for illustrating discrete coupling relations in tightly coupled directional couplers.

and two discrete couplings. Energy transferred from the lower line to the upper line at the first coupling experiences a 90° phase delay. This energy travels along the upper line to the second coupling and part of this energy returns to the lower line, with a further phase delay of 90°. Thus, energy which goes from the lower line to the upper line, and back to the lower line at a later coupling point, arrives in the lower line out of phase with the energy which traveled straight through in the lower line. A summation of such components eventually results in cancellation of the wave in the lower line.

Quantitative relations may be written for the sum of the forward wave components after an arbitrary number of discrete couplings. These relations help to answer the question, "How many couplings of a certain specified loss each are required to obtain the desired loss to the side arm?" The relations answering this question are plotted in Fig. 21. The abscissa is a number of coupling units, the ordinate is loss per coupling unit, and the parameter along the curves is over-all net loss to the side-arm output. For 10 coupling units, a 3-db net loss to the side arm is produced by making the loss per coupling unit equal to about 22 db. Experimentally, very low losses have been observed.

Fig. 22 shows three inserts used in the 4,000-mc test

[4] R. L. Kyhl, M.I.T. Radiation Laboratory Series, vol. 11, Chapt. 14, p. 887.

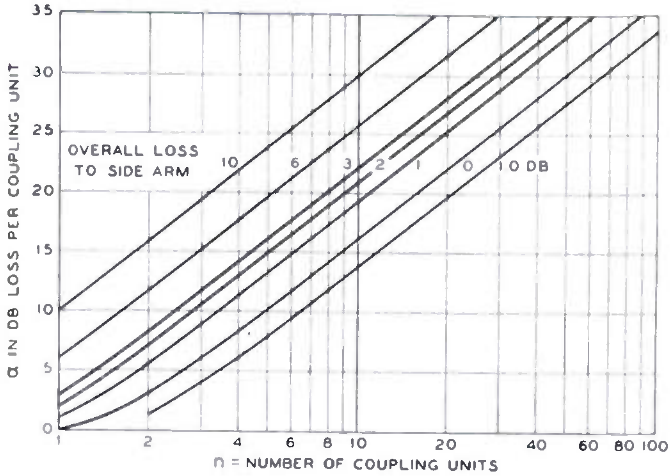

Fig. 21—Chart showing the relation between number of coupling units, loss per coupling unit, and over-all net loss to the side arm in directional couplers.

jig to realize low losses. The simplest array is the series of uniform-sized holes uniformly spaced at the top of the figure. The taper used is the same one previously demonstrated in connection with loose coupling work; the lower insert is the result of placing in tandem two identical arrays, each of which resulted from the combination of several uniform coupling functions. All three of these inserts gave less than three quarters of a db loss to the side arm over an 18-per cent frequency band, and over 15-db directivity. The directivity and loss for the linear taper are given in Fig. 23. The loss from input to side arm was not centered in this band, being 0.75 db at 3,737 mc and 0 db at 4,475 mc. The directivity was more than 20 db at any point in the frequency band.

To summarize, theoretically any predetermined bandwidth and arbitrarily large directivity can be achieved, using the approach outlined, without previous knowledge of coupling versus hole size. Coupling losses less than 0.1 db are achievable and the technique appears applicable to frequencies in excess of 50,000 mc.

Whereas the theory has been illustrated using waveguides, it is apparent that it is applicable to open-wire,

NUMBER OF HOLES	LOSS IN DB 3735 MC	4060 MC	4475 MC
7	0.75	0.1	0

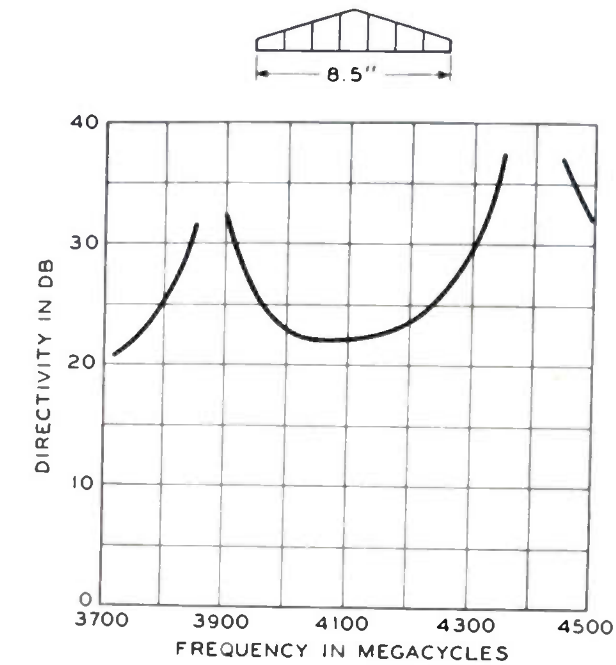

Fig. 23—Observed directivity for middle insert of Fig. 22.

coaxial, dielectric, lumped element electrical, or acoustic transmission lines.

In conclusion, the authors would like to express appreciation for the valuable assistance given in this work by Mr. E. L. Chinnock.

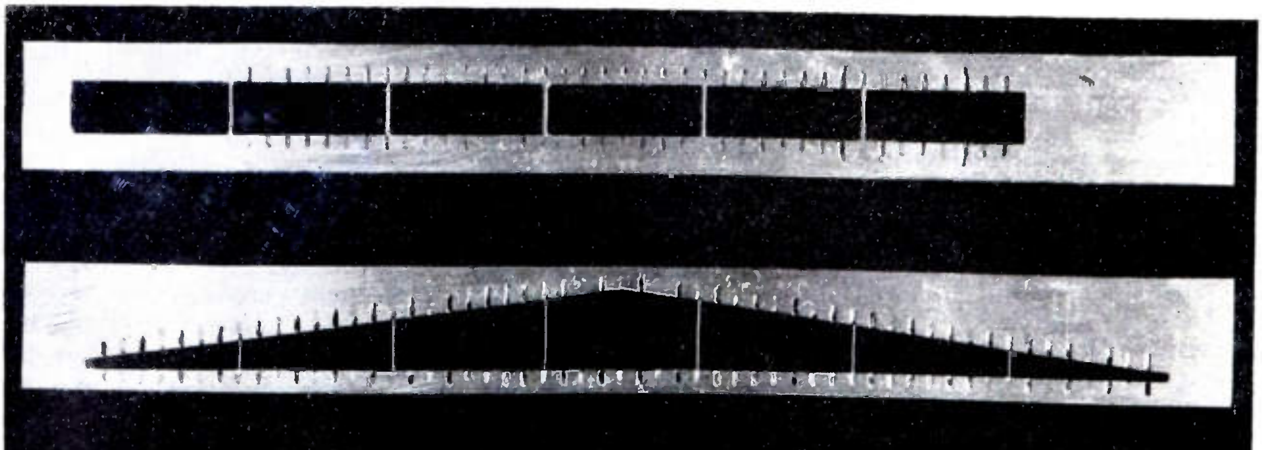

Fig. 22—Inserts used to demonstrate complete power transfer between lines.

Nonsynchronous Time Division with Holding and with Random Sampling*

J. R. PIERCE†, FELLOW, IRE AND A. L. HOPPER†, SENIOR MEMBER, IRE

Summary—There is a general type of system in which an indefinitely large number of transmitters can have access to any of an indefinitely large number of receivers over a medium of limited bandwidth. In these systems, signal-to-noise ratio goes down as more transmitters are used simultaneously. This paper describes a particular system which sends samples by means of coded pulse groups sent at random times. The signal-to-noise ratio is good in the absence of interference and the effect of interference is minimized by holding the previous sample if a sample is lost. An experimental system worked satisfactorily and gave close to the predicted signal-to-noise ratio. Such a system might be used to provide communication and automatic switching in rural telephony, or for other applications.

Introduction

THIS PAPER deals with a particular type of nonsynchronous time-division multiplex communication system which has been devised and tested by the authors. The first part of the paper discusses the advantages of systems of this general class and explains why a particular system was chosen for investigation. The second part describes the system in detail and presents experimental results and compares them with the simple theory. The third part discusses a possible application of this type of system.

Part I—Background and Description of System

Some years ago Shannon pointed out in unpublished work that conventional multiplex communication systems distinguish among channels by sending, in each channel, only signals which are orthogonal to any signals which may be sent in any other channels. This is true in frequency-division multiplex, because signals in nonoverlapping bands of frequencies are truly orthogonal functions of time in the sense that the integral of their product over an infinite time is zero; therefore, the channels carried by such signals can be completely separated. The pulses used in an ordinary time-division system are truly orthogonal functions, and the channels carried by different sets of pulses can be completely separated.

The difficulty with truly orthogonal functions is that a channel of given capacity can be divided into a limited number only of channels of a given lesser capacity if the signals in any channel are to be orthogonal to all signals in all other channels. For instance, if one has a frequency band of 40 kc, he can assign 10 specific 4-kc channels to 10 different talkers. There is nothing new left to assign to an eleventh talker, and an eleventh talker cannot use the same band of frequencies without switching, that is, without destroying the access to the channel of one of the 10 existing assignments. The same thing is true of ordinary pulse systems.

Shannon then made the suggestion that there might be assigned to channels not functions from a truly orthogonal set but functions from an approximately orthogonal set.[1] He pointed out as an example that there is an infinite number of noise signals in a given frequency range which over a long period of time are approximately orthogonal.

The use of approximately rather than truly orthogonal functions would of course necessarily result in some cross talk or "noise," but there would be compensating advantages. By use of approximately orthogonal functions, one could make a communication system with the general properties described in connection with Fig. 1.

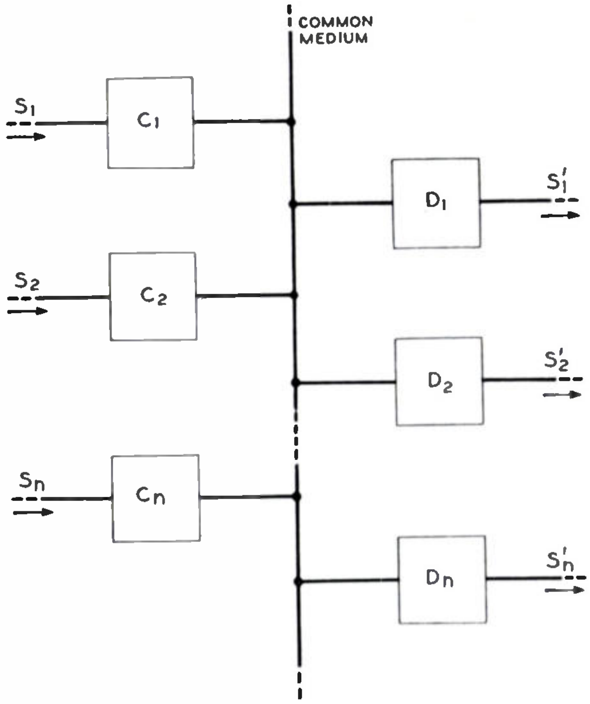

Fig. 1—System in which channels are assigned approximately orthogonal functions of time.

In the sort of system shown in Fig. 1, a number of speech or other signal channels $S_1, S_2 \cdots S_n$ are acted upon or modulated by coders $C_1, C_2 \cdots C_n$. These coders need not be synchronized in any fashion. The outputs of the coders go to a common medium, shown in Fig. 1 as a line. This common medium might, however, be a radio-frequency channel, either directional or

* Decimal classification: R460. Original manuscript received by the Institute, November 15, 1951; revised manuscript received June 12, 1952.
† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] The general relation of this to Zator coding should be noted. Zator coding is dealt with in various publications of the Zator Co., 79 Milk St., Boston 9, Mass.

nondirectional, a wire line, a waveguide, or any other transmission medium. The signals from the coders may be injected into the medium at one location, as at a multiplex terminal at one end of a transmission line, or at a number of locations, as from airplanes, automobiles, military vehicles, and the like. Decoders $D_1, D_2 \cdots D_n$ are connected to the common medium. Decoder D_1, for instance, can be adjusted to respond only to signals from coder C_1, and the like, giving an output S_1', proportional to the input S_1.

An important feature of such a system is that although the channel capacity of the common medium may be such as to allow *simultaneous* use of only a limited number of the channels $S_1 \cdots S_n$, for instance 50, yet many more distinct coders and decoders, for instance 1,000, can have uninterrupted access to the medium. Thus, without any switching, the common medium is at all times available to transmit any 50 out of the 1,000 possible channels. These figures are of course chosen only for the purpose of illustration. As an increasing number of channels are placed in simultaneous use over the common medium, there is a gradual degradation of quality.

There are a number of ways in which this sort of performance could be achieved. One way has been mentioned: the use of random or noise waveforms as carriers. This necessitates the transmission to or reproduction at the receiver of the carrier required for demodulation. Besides this, the signal-to-noise ratio in such a system is poor even in the absence of interference unless the bandwidth used is many times the channel width. The ratio of signal power to noise power in the absence of interference is approximately equal to the ratio of total bandwidth to channel bandwidth. Other systems using pulses have been described in the literature.[2,3]

In the system discussed here, the signal to be sent is sampled at somewhat irregular intervals, the irregularity being introduced by means of a statistical or "random" source. The amplitude of each of the samples is conveyed by a group of pulses, which also carries information as to which transmitter sent the group of pulses. A receiver can be adjusted to respond to pulse groups from one transmitter and to reject pulse groups from other transmitters.[4] When a pulse group is accepted, the amplitude of the sample which it carries is stored as a voltage on a capacitor. The voltage across the capacitor is the input to the output filter. When another pulse group is accepted, the voltage on the capacitor is changed to conform to the amplitude of the new sample, and the voltage is then held constant until another pulse group is accepted, and so on.

[2] E. Labin *et al.*, U. S. Patents Nos. 2,408,079; 1946; 2,410,350 (1946), 2,425,066; 1947.

[3] W. D. White, "Theoretical aspects of asynchronous multiplexing," Proc. I.R.E., vol. 38, pp. 270–275; March, 1950.

[4] Codes and coincidence circuits in receivers have been used in radar beacons. See, A. Roberts, "Radar Beacons," M.I.T. Radiation Laboratory Series, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 3, chapt. 5; 1947. The coincidence circuits there described are less powerful than the discriminating circuits we have used.

If a pulse group from one transmitter sufficiently overlaps a pulse group from the desired transmitter, the combined "distorted" pulse group is rejected, and at the receiver a sample is lost. When a sample is so lost, voltage across the capacitor will be in error by the difference between two successive samples.

If the overlap is not sufficient to cause loss of the sample, another sort of error may be introduced by distortion of the sample amplitude. A third kind of error can result from the production of a code group assigned to one transmitter by an accidental combination of signals from other transmitters. These latter two kinds of error can be reduced at the expense of increasing the number of lost samples by making the requirements for the acceptance of a code group very precise.

The same average sampling rate is used for all transmitters. If the sampling intervals were equal, overlapping of pulse groups from two transmitters would be followed by further successive overlappings until the sampling times drifted out of phase. This would result in the loss of a series of successive samples. By sampling the signal at somewhat irregular intervals, which we may call *random sampling*, loss of groups of successive samples is avoided. Listening tests show that a random loss of samples is less objectionable than a periodic loss of series of successive samples.

The advantages of this particular system are as follows:

(1) The signal-to-noise ratio is good in the absence of interference.

(2) In the presence of interference the noise is proportional to signal.

(3) Interference between transmitters is unintelligible noise.

(4) Holding of the previous sample in case a sample is lost reduces the noise due to interference for some signals, including speech.

(5) There is no limit to the number of assignments which can be made. The use of the same mean sampling rate for each transmitter is necessary to achieve this.

Part II—Experimental System

As built for these experiments the system uses pairs of equal and opposite video pulses, amplitude modulated, to carry and identify samples. By time separation of the two pulses of the pair, the transmitter introduces a code which permits recognition by the proper receiver. The receiver accepts or rejects this code by comparing the amplitudes at four taps along a delay line. Acceptance enables a gate which transfers the pulse amplitude to a storage capacitor and following filter to reform the original audio wave. When another transmitter happens to interfere with the correct one, the gate is disabled and no sample is transferred.

General Description

A simplified block diagram of the two transmitters and the receiver used in these experiments is shown in

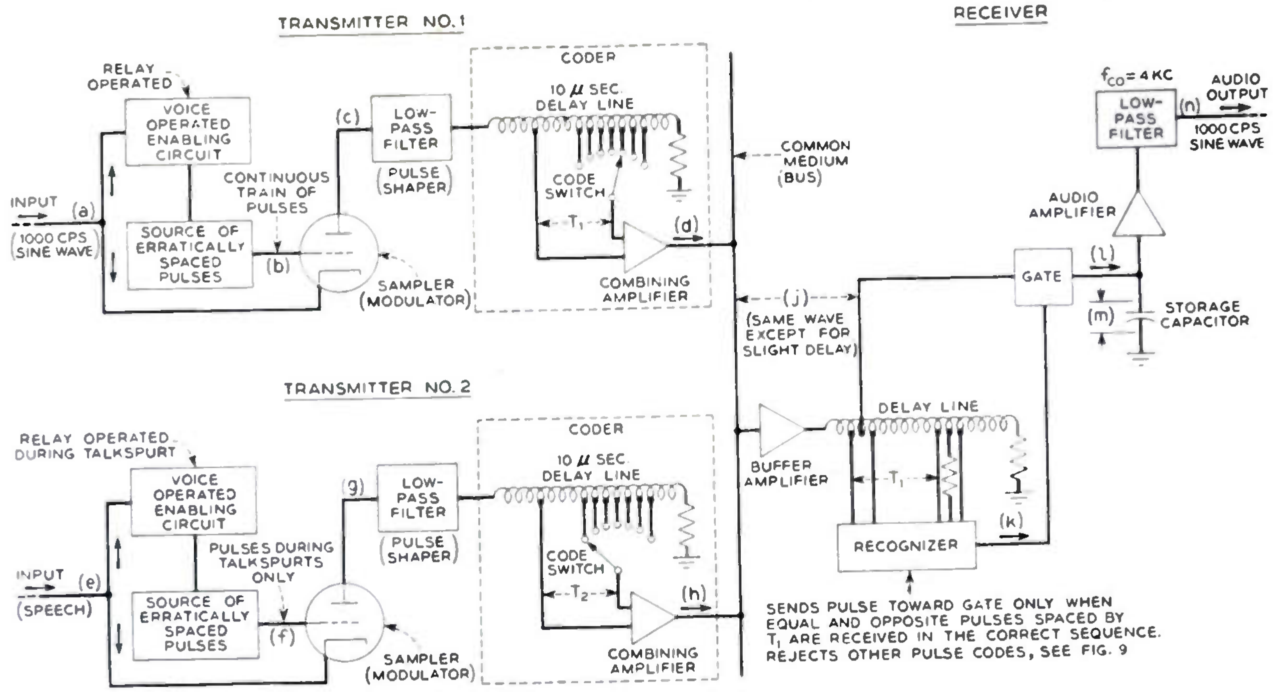

Fig. 2—Simplified schematic.

Fig. 2 and a simplified picture of the corresponding wave shapes is given in Fig. 3 to show briefly the operating principles of the system. It is assumed that transmitter #1 has its code switch set for a relatively long code, such as T_1 in Fig. 3(d), while transmitter #2 is switched to a relatively short code, such as T_2 in Fig. 3(h). The receiver is set to accept the T_1 code by the spacing of taps on its delay line and to reject all others.

To illustrate the principle of operation with the aid of Figs. 2 and 3, it is assumed that transmitter #1 is operating continuously with a 1,000-cps sine-wave input which holds the associated voice relay operated. Hence the input wave (a) is randomly sampled by pulses (b) to produce modulated pulses (c). These pulses are then paired by the delay line and made equal and opposite by the combining amplifier, resulting in the transmission of modulated and coded pulses (d) to the common medium. A low pass filter is used between the modulator and delay line for pulse shaping.

As shown at (e), transmitter #2 is assumed to be inactive at first because of lack of speech input. Hence only pulses coded by T_1 are present on the common medium (j). Since the receiver is set for this code, it correctly recognizes each pulse pair operating the gate by pulses (k) to produce samples of wave (j) as at (l). The voltage across the storage capacitor neglecting the dc component is a step wave as shown at (m). As shown at (n) the resulting filtered output at this time is practically undistorted.

However, at time (e') the speech input to transmitter #2 has built up to a point where its voice relay operates (time of operation assumed small for illustrative purposes). This causes wave (e) to produce modulated pulses (g) similar to those of transmitter #1. However,

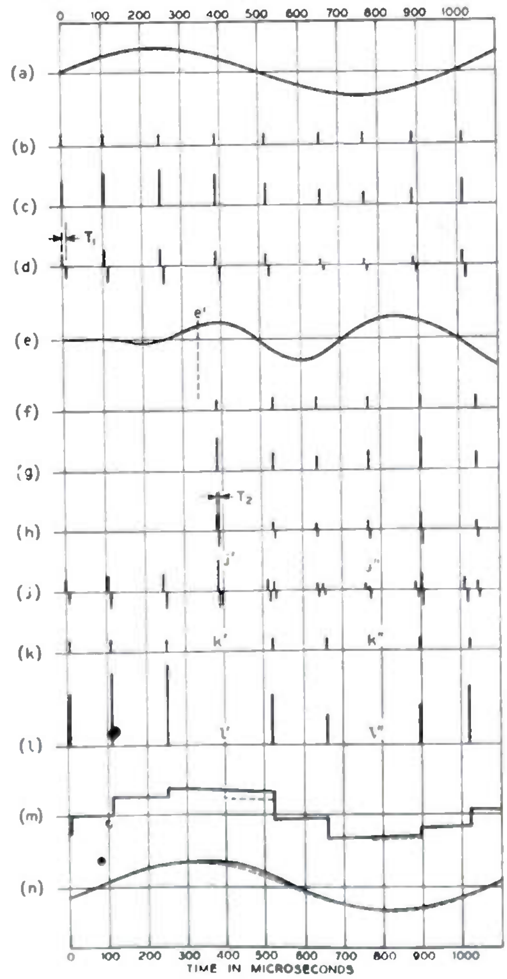

Fig. 3—Typical wave shapes.

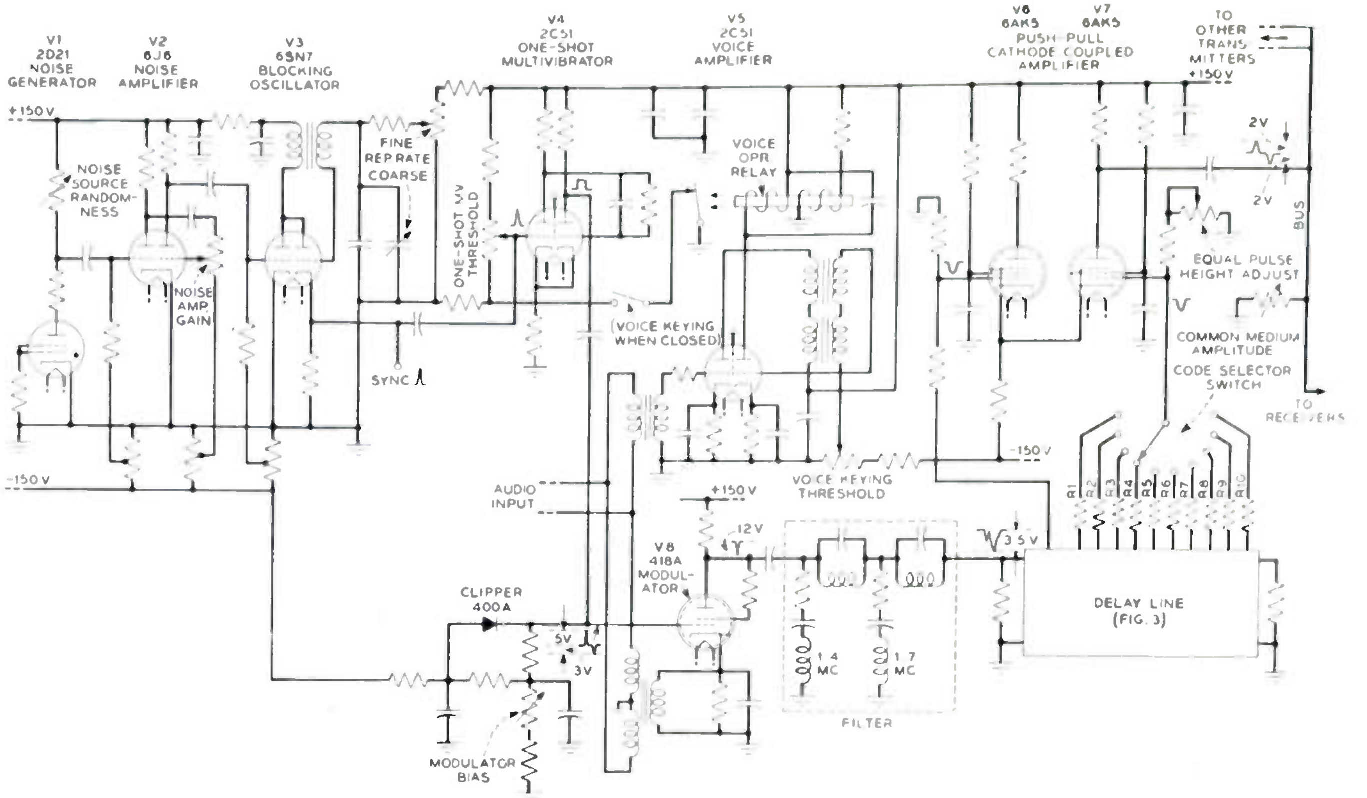

Fig. 4—Transmitter.

the code switch in this case is set for T_2 so that pulse trains (d) and (h) can be distinguished by their respective codes, T_1 and T_2.

Both transmitter outputs (d) and (h) appear superimposed on the common medium and at the gate input as shown at (j). Because of the different sampling rates and modulation of the two transmitters their outputs may combine to produce a wide variety of interferences. Two simple examples of interference are shown at j' and j''. At j' the positive pulses of the two transmitters coincide and add in phase. At j'' the negative pulse of transmitter #1 is subtracted from the positive pulse of transmitter #2.

The recognizer continuously monitors wave (j) as it passes down the receiver delay line and decodes it by means of five taps properly spaced along the line. Each time equal and opposite pulses spaced by T_1 (and positive before negative) are detected, a pulse (k) is transmitted to operate the gate. However, because of the interferences noted above no gate enabling pulses are sent at (k') and (k''). The gate output (l) is the same as before, except samples are now missing at (l') and (l''). Likewise, the step wave (m) and filtered output (n) are distorted because of the missing samples. The undistorted waves are shown by the dashed lines.

Delay Line

Identical delay lines are used in both transmitter and receiver. The lines have a total delay of approximately 10 μsec and a characteristic impedance of about 250 ohms. Taps are located $\frac{1}{4}$ μsec apart.

Transmitter

The transmitter is shown in Fig. 4. The source of erratically timed pulses is shown in the left half of this figure. A 2D21 thyratron V1 was found to be fairly satisfactory although some selection was necessary and the plate-load resistor had to be adjusted for "best-looking noise" as observed on an oscilloscope. Two out of five samples would give up to 1-volt peak-to-peak noise output with a relatively small output of discrete frequencies. Twin triode V2 increases the noise voltage to about 8-volts peak to peak as required by the following circuit.

The primary pulse source is blocking oscillator V3 which is triggered at a random rate by the noise source previously described. Since its output varies considerably with the repetition rate, for random sampling it is necessary to add one-shot multivibrator V4. The output of this stage when differentiated and clipped provides erratically timed pulses of constant amplitude to the grid of modulator V8.

The audio input is connected to the cathode of modulator V8 and also through amplifier V5 to a voice-operated relay. This relay operates during talk spurts to decrease the bias on one-shot multivibrator V4, allowing it to produce pulses. In the absence of audio input, V4 is biased off and no pulses are transmitted.

It is important that the pulses delivered to the common medium be shaped so as to be as nearly noninterfering as possible, that is, of zero amplitude except during the pulse interval. Referring to Fig. 5, (a) shows a short impulse about 0.4 μsec long at the base which appears at the plate of the modulator tube and is applied to the input of the low pass filter. This filter consists of two sections designed on the basis of a constant resistance of 250 ohms. The first section resonates at 1.4 mc while the second resonates at 1.7 mc. Fig. 5(b) shows the output of the filter which is a jagged wave due to the transmission of frequencies above the frequency of maximum attenuation. However, a few sections of the delay line effectively suppress these frequencies and give an over-all characteristic which closely approximates a Gaussian cut-off. The corresponding desirable wave form, shown in Fig. 5(c), is relatively free of undershoots and overshoots. Nevertheless, these undershoots and overshoots do become worse as the wave travels down the delay line. The wave shape at the far end of the line is shown in Fig. 5(d).

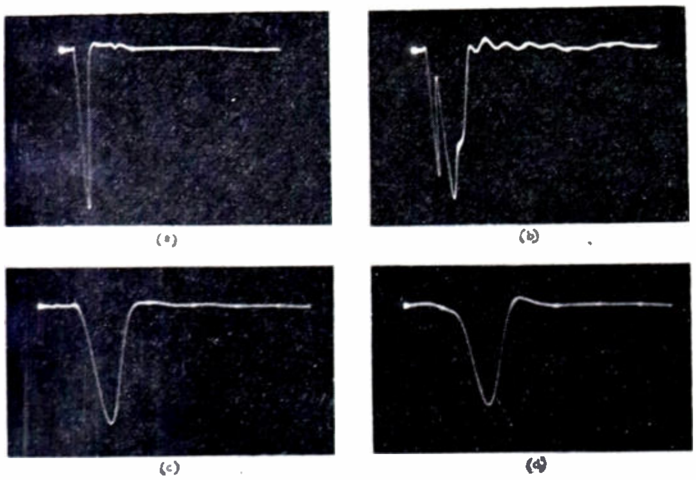

Fig. 5—Transmitter wave shapes—1μ sec timing dots. (a) Plate of modulator tube. (b) Output of filter. (c) Delay-line terminal 4. (d) Delay-line terminal 40.

As shown in Fig. 4, the combining amplifier is of the push-pull cathode-coupled type using pentodes V6 and V7. A control is provided at the grid of V7 to permit adjusting the pulses of a code pair to be equal in amplitude. To avoid having to adjust this control each time the code selector switch is operated, resistors R1 to R10 are chosen to compensate for the attenuation of the delay line. A common medium amplitude adjustment is also provided to maintain the proper amplitude for optimum recognition margins in the receiver.

Receiver

The basic problem is to enable a gate each time a desired pulse pair is received unless it is interfered with by pulses from another transmitter. For example, we might recognize the desired pair whenever the pulses were equal and opposite, and each had zero slope as observed at two properly spaced taps on the delay line. This method suffers because of the distortions in practical differentiating circuits and the timing uncertainty when the gradual slope near the pulse crests is the criterion. This led to the idea of checking at four points along the delay line as shown in Fig. 6. In effect, this recognizer uses the delay line for a kind of distortionless differentiation, and by operating on the steeply sloping sides of the pulses allows a high degree of timing discrimination. It also has the advantage of producing a desirably short pulse at the mixer output for operating the gate.

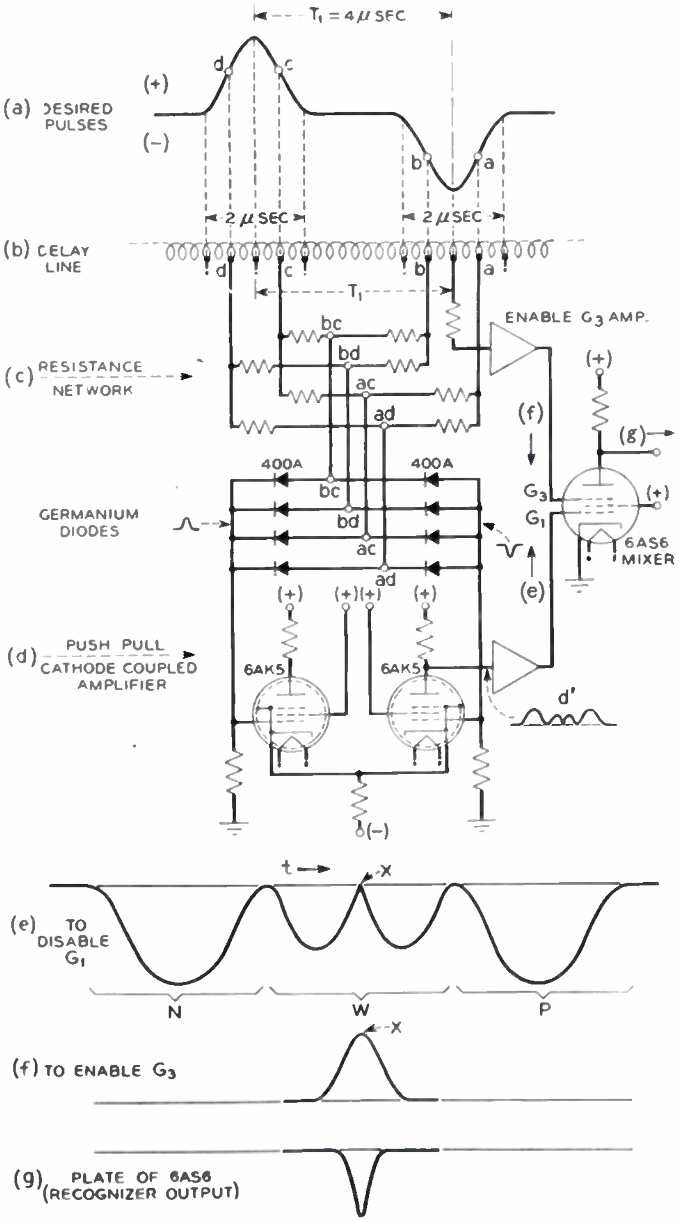

Fig. 6—Recognizer.

The action of the recognizer may be understood by referring to Fig. 6. The desired pulse pair is shown by the wave at (a), with $T_1 = 4$ μsec. Let us assume that this wave has traveled down the delay line shown at (b). Then at the instant shown by the relation of (a) to (b) the voltages measurable at the various taps on the delay line will be proportional to the amplitude of the wave above. At this instant the voltages at taps a, b, c, and d are related as

$$v_a = v_b = -v_c = -v_d. \tag{1}$$

This relationship may be established by three simultaneous equalities, such as $v_a = -v_c$, $v_a = -v_d$ and $v_b = -v_c$. A fourth equality, $v_b = -v_d$, may be added for symmetry. This is done by checking for zero voltage at the midpoint of resistances connecting a-c, a-d, b-c, and b-d as shown at (c). These four midpoints are checked for zero voltage by connecting them through diodes to a push-pull cathode-coupled amplifier as shown at (d). This causes a positive voltage output at d' at any time when any of the four points deviate from zero voltage. If we assume that the wave shown at (a) travels down the delay line shown at (b), the 6AK5 output will appear as shown at d'. This wave when amplified and inverted for disabling the control grid of the 6AS6 mixer is shown at (e). The center portion of this wave resembles the letter "W." The zero output at "X" in the "W" wave indicates the time coincidence shown between (a) and (b). At slightly earlier and later times the "W" wave has amplitude due to unbalance of the four midpoint voltages ac, ad, and so on. Four microseconds earlier pulse N was caused by the negative pulse of wave (a) crossing taps d and c. Four microseconds later pulse P is caused by the positive pulse crossing taps a and b. Thus the zero at "X" in the "W" wave is a unique recognition indication of the desired pulse pair, that is, when the interpulse spacing T_1 equals the intertap delay T_1. For all other pulse pairs there is no zero point "X" in the "W" wave.

To avoid false recognition when there are no pulse voltages present on any of the four taps, the suppressor grid of the 6AS6 mixer is enabled by a pulse derived from an additional tap on the delay line. This tap may be located to correspond to the first (that is the negative) pulse of the desired pair as shown in Fig. 6(b). When amplified and inverted, the wave presented to the suppressor grid is as shown at (f). When G_3 is enabled and G_1 is not disabled, as at time "X," a pulse appears at the plate of the 6AS6 as shown at (g).

Suppose that a pulse from another transmitter interferes with the wave in Fig. 6(a). It will obviously distort the wave at one or more of the four sampling points so that equality (1) is not satisfied. This will prevent a zero at "X" in the "W" wave so that the gate will not be enabled. However, if the interfering pulse happens to be equal and opposite to one of the pulses at (a) and spaced by T_1, there will be a zero in the "W" wave. But there will be no corresponding enable pulse as there was at (f) because the first pulse of this unwanted pair is positive instead of negative.

Of course *two* or more interfering transmitters might occasionally produce a pulse pair which is indistinguishable from the desired code shown at (a). However, an evaluation of this type of interference is beyond the scope of the present experiments.

The actual arrangement of the receiver circuit is shown in Fig. 7. For best signal-to-noise performance, both pulses of the pair are sampled for feeding the gate input. The negative pulse is taken directly from the delay line while the positive pulse is inverted by V_{12} and then mixed at the input to cathode follower V4. By staggering the two sample taps slightly with respect to the desired code, a fairly flat-topped wave is available at V4 as the gate input. This minimizes the noise due to variations in the exact time at which the gate operates due to variations in the amplitude of the wave in Fig. 6(g). For best recognition of pulse pairs in the presence of modulation the amplitude of pulses on the receiver delay line should be of the order of 10 volts. To obtain this, a 418A tetrode is used as buffer amplifier V1 with a step-down transformer feeding the delay line. Shunt feed to the transformer primary minimizes pulse distortion.

The performance of the recognizer is greatly improved by adding adjustable capacitors at points b-c and a-d as shown in Fig. 7. In this way phase shifts are introduced to compensate for the distortion of pulses during transmission. (Some distortion is due to a change in pulse length, amplitude, and preceding and succeeding undershoot and overshoot as the pulses travel down the transmitting and receiving delay lines.)

Pentodes V2 and V9 are used to raise the level of the enable and disable pulses to suitable levels for application to the grids of the mixer. The cathode follower half of V3 permits driving the 6AS6 suppressor from a low impedance source. The other half of V3 inverts the mixer output so as to present a positive pulse to trigger one-shot multivibrator V_{11}. The latter's function is to insure an "all-or-nothing" gating pulse in the presence of modulation and interference. The multivibrator output is differentiated and applied to the gate transformer by the driver half of V4. The other half of V4 is a cathode follower which applies the sample pulse at low impedance to the input of the gate.

The gate is operated for only $\frac{1}{4}$ μsec each time a sample is taken, so to charge the storage capacitor fully the gate resistance must be very low. For example, if the total charging resistance is 500 ohms and the capacitance is 300 mmf, the time constant will be 0.15 μsec. When the gate is unoperated, however, its resistance must be very high because the time between samples varies between 100 and 250 μsec (allowing for randomness and a missed sample) so that the last sample of the step wave must be held until the next sample is received to minimize distortion. If the resistance is 10 megohms, the discharge time constant is 3,000 μsec, which results in a moderate slope of the steps appearing across the storage capacitor shown in Fig. 3(m). As the storage capacitor is increased, the steps become flatter, but the high-frequency response of the system becomes poorer because of the longer charging time constant. Based on listening tests with one interfering transmitter, 300 mmf appears to be a good compromise value. The steps may be made to slope either up or down (or all toward the

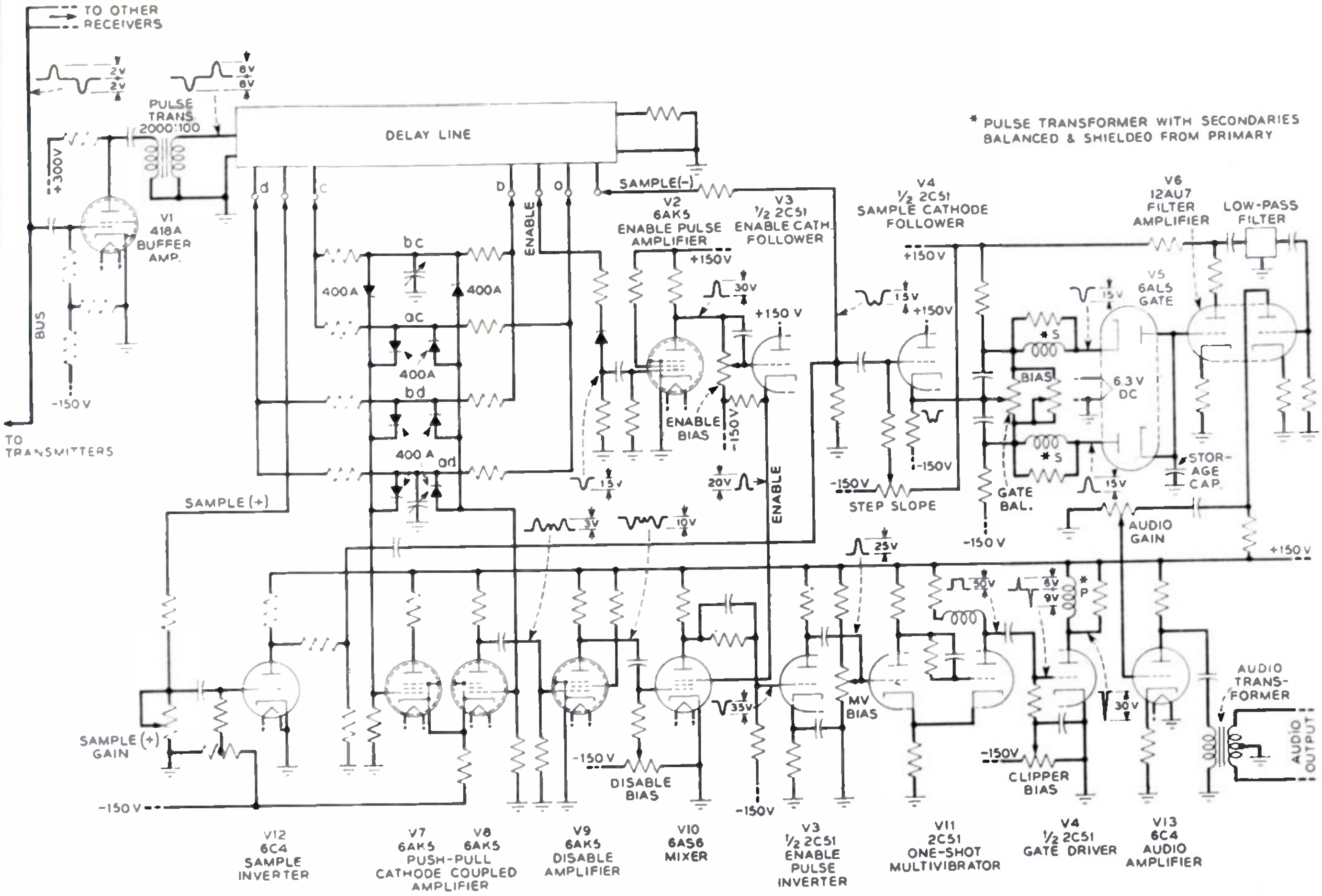

Fig. 7—Receiver.

axis as shown in Fig. 3(m)) by means of the STEP SLOPE adjustment since this establishes the unoperated gate input potential to which the stored charge leaks.

Two other adjustments are provided for optimizing the gate operation. One sets the total dc bias while the other adjusts the differential bias on the two diodes. A special pulse transformer is used in which the two secondaries are balanced and shielded from the primary. Thermionic rather than germanium diodes are needed because of the importance of high back resistance. In addition, direct coupling is used between the storage capacitor and the filter amplifier input to minimize leakage. A low pass filter eliminates the sampling frequencies from the audio output. The attenuation of this filter is about 3 db at 4,000 cps. The over-all audio response of the system is down 3 db at 3,000 cps.

Signal-to-Noise Ratio

The results of 1,000-cps signal-to-noise measurements at the output of the system are shown in Fig. 8, in which noise plus distortion is plotted as a function of total output. Tests were made both with and without an interfering transmitter using a 2B noise-measuring set with "F1A" line weighting.[6] To obtain the noise plus distortion readings, the 1,000-cycle signal was fed around the complete system from input to output with the proper attenuation and phase shift to cancel the signal at the output.

It is interesting to note how the noise and distortion increase along with the signal when both transmitters are operating. This sort of behavior is to be expected

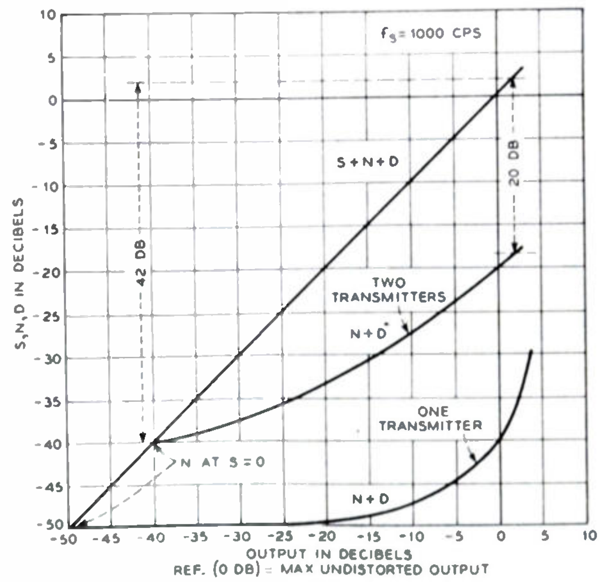

Fig. 8—Signal, noise, and distortion relative to maximum undistorted output. Output in decibels. Ref (0 db) = max undistorted output.

[6] This weighting takes into account the frequency characteristic of present telephone equipment. It is standard for measurement of noise on message circuits.

when the effect of missing samples is considered. As shown in Fig. 8, the maximum signal to background noise ratio is about 42 db. However, the maximum signal to (*maximum signal noise-plus-distortion*) ratio is only about 20 db. This is in fair agreement with a calculated value of 21.8 db obtained in the Appendix.

A rough idea of the interfering effects of a number of transmitters is obtained by driving the interfering one at rates which are various multiples of 8 kc while the signaling transmitter operates with jitter as before. Each simulated added transmitter degrades the signal to background noise ratio by roughly 3 db. Thus with five simulated interfering transmitters this ratio drops from 42 db to about 30 db. Ideally, interfering transmitters should not produce noise in the absence of signal unless overlapping pulse groups combined accidentally to form the receiver's code. As this cannot have occurred in the test described, the increase in background noise described above indicates imperfect functioning of the receiver.

Listening tests were conducted in which each transmitter was fed from its own tape recorder while the receiver was arranged to accept signals from either one or the other. In general the operation of the second transmitter did not interfere with intelligibility at normal talking levels although the transmission was judged to be below toll quality. The "loss-of-sample" distortion causes a certain "rasping" quality of speech. When more interfering transmitters are simulated as described above, the distortion becomes progressively worse as expected. It appears that speech intelligibility would still be tolerable with somewhere between five and eight active interfering transmitters.

Part III—Possible Application of System

For what use might a system of the type described be particularly suited? Simply as an illustration, let us consider how it might be used in connection with rural telephony.

Fig. 9 indicates the form a local exchange might take. It consists of a number of subscriber stations with directive antennas pointed at a central omnidirectional repeater station. Each subscriber transmits on a common transmitting frequency f_t and receives on a common receiving frequency f_r. Subscriber transmitter powers are adjusted so that the omnidirectional repeater receives signals from all subscribers at approximately the same level. It amplifies the received signal pulses, changes frequency from f_t to f_r, and reradiates omnidirectionally.

The system inherently provides both for talking and for automatic switching. Each subscriber is assigned a specific *number* or pulse code group, as, for instance, 2, or, perhaps preferably, 3 equal amplitude pulses spaced by times T_1 and T_2.[6] In the case of 3 pulses in a code group, each subscriber will be provided with two dials by which the T_1, T_2, that is, the code group or number of his transmitter and receiver, can be set simultaneously to any allowable number (T_1, T_2; or code group). Each subscriber is assigned his own number, to which his transmitter and receiver revert when his hook is down.

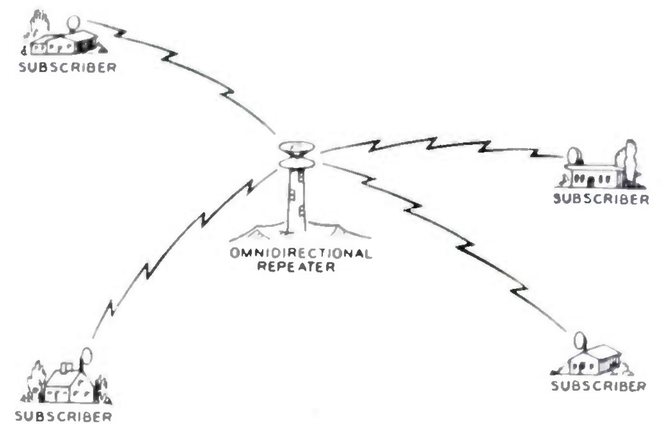

Fig. 9—Use of nonsynchronous system in rural telephony.

Each subscriber's receiver is on all the time. With the hook down, the bell rings when the receiver receives pulses corresponding to the subscriber's number. The subscriber's transmitter emits pulses when (1) the hook is up and (2) either a ring button is pushed, or the subscriber is talking.

To make a call, subscriber A raises his hook and dials subscriber B's number. He then presses a ring button which causes his transmitter to emit pulse groups corresponding to subscriber B's number. Subscriber B's bell rings. Subscriber B raises his receiver. A and B can now talk, both using the number of the called party, that is, subscriber B.

Among the "subscribers" there may be one or more operators or apparatuses at distant exchanges, to provide for communication between local exchanges.

We will note that a third subscriber C can talk to both the called and calling parties by dialing the called party's number. Subscriber C cannot reach a calling party by dialing the calling party's number. This might tend to reduce eavesdropping somewhat; but eavesdropping is still possible, as indeed it is now in rural party lines. We should also note that there is no busy signal, and that when the called party is talking on another code he simply does not hear or respond to a calling party's signal.

There are many other possible uses of such a system, civilian or military, either with or without the use of a repeater. For instance, such a system might be used to communicate with or between moving vehicles or ships. For such service, synchronous time-division multiplex seems to be ruled out because a number of paths of changing delays are involved. At the same time, frequency division may be made difficult by the excessive linearity requirements which it imposes on repeaters

[6] For T_1+T_2 ranging from 2 units to 15 units, this provides 105 separate code groups. Thus, a delay line 15 pulse lengths long could be used to generate or select any of 105 code groups.

that must amplify simultaneously signals of very different amplitudes.

In the rural telephony example, all pulses received come from a common transmitter (the central omnidirectional frequency-changing repeater). The outputs of all transmitters can be adjusted to give the same signal level at the repeater, and hence pulses from all transmitters have about the same amplitude. In the experimental system described, pulses from both transmitters had the same amplitude.

If there is no central repeater, or if transmitters move with respect to the central repeater, pulses from one transmitter may be hugely greater than those from another transmitter. Under these circumstances, information could be conveyed by frequency modulation of the radio frequency of the pulses. In this case the pulses could be limited somewhere in the receiver so that the strong and the weak pulses would come to have comparable amplitudes. Through this limiting, strong pulses would be lengthened with respect to weak pulses. This means that strong signals would tend to cause more interference than weak signals because the pulses of the strong signals would last longer than the pulses of the weak signals. In comparing pulses in a group, equality of pulse length rather than equality of pulse amplitude could be used in deciding whether or not several pulses belong to one code group.

Acknowledgment

The authors wish to acknowledge gratefully the advice and encouragement of A. A. Roetken and the assistance of J. L. Wenger in the construction and testing of the system.

Appendix

A. *Fraction of Samples Lost*

When a pulse from an interfering transmitter falls "on" a pulse of the desired transmitter, a desired pulse group will be lost, and this will cause the loss of one or more samples, depending on how many samples the pulse group carries.

What does "on" mean? Mathematically, a pulse of limited bandwidth never quite decays to zero. Furthermore, the nearness of spacing of pulses which will cause rejection depends on the sensitivity of the comparison circuits. Very sensitive comparison circuits will reject many samples whose amplitudes are only a little in error because of interfering pulses, but will allow few distorted samples to pass; less sensitive comparison circuits will accept more of the samples, but in doing so they will pass some distorted and hence "noisy" samples. Just losing samples causes noise; accepting badly distorted samples causes noise. There is some best compromise.

Consider a video pulse. Suppose that B is the bandwidth, say, to the 6-db down point. Experience shows that the length τ of the pulse at the "base" (from where it is near zero to where it is near zero again) is about

$$\tau = \frac{1}{B}. \tag{A1}$$

For radio-frequency pulses, of course, we will have

$$\tau = \frac{2}{B_r}, \tag{A2}$$

where B_r is the radio-frequency bandwith. We will assume that one pulse interferes with another if its center falls within the time interval τ.

Now, consider N transmitters operating simultaneously, each emitting 2 W/m pulse groups a second. Here W is half the sampling rate (the limiting audio bandwidth) and m is the number of samples per pulse group. Let n be the number of pulses in a pulse group.

The total time per second occupied by one transmitter is

$$\text{occupied time per second} = \frac{2Wn}{m}\tau. \tag{A3}$$

The number of pulses emitted per second by interfering transmitters is

$$\text{interfering pulses per second} = \frac{2Wn}{m}(N-1). \tag{A4}$$

Hence, the number of lost samples per transmitter (m times the number of lost pulse groups) is

lost samples per second, for one transmitter

$$= \frac{4W^2n^2}{m}(N-1)\tau. \tag{A5}$$

As the number of samples per second is $2W$, the fraction α of samples lost is

$$\alpha = \frac{2Wn^2}{m}(N-1)\tau. \tag{A6}$$

Let us assume a radio-frequency system, and use relation (A2)

$$\alpha = 4\frac{Wn^2}{B_r m}(N-1). \tag{A7}$$

It is of course optimistic to regard W as the audio-frequency bandwidth; the audio-frequency bandwidth will be appreciably less than W in any practical system.

In the case of the experimental system, approximately

$$B = 600{,}000 \text{ sec}^{-1} \text{ (video band)}$$
$$W = 5{,}000 \text{ sec}^{-1}$$
$$n = 2$$
$$m = 1$$
$$N = 2.$$

We should use (since B is the video band)

$$\alpha = 2\frac{Wn^2}{Bm}(N-1),$$

whence

$$\alpha = 0.067.$$

B. Noise Caused by Lost Samples

In random sampling systems the previous sample is held if a sample is lost. For a given fraction of samples lost, the signal-to-noise ratio depends on the nature of the signal. In the noise measurements made, the signal was a sine wave. In treating this case let us assume that

(1) successive samples are never lost;

(2) the received samples are $\sin 2\pi Wt/2\pi Wt$ waves, which have a uniform frequency spectrum up to a frequency W and no frequencies above this;

(3) the filter through which the samples pass has an amplitude response $F(f)$ which is zero above $f=W$;

(4) the signal is a sine wave of peak amplitude V;

(5) the error signals (missing pulses) constitute a noise of flat frequency distribution.

According to (1), if we lose a sample we utilize, instead, a sample in error by the difference between two successive samples, which we will call a. Let A be the amplitude of the signal. Then

$$\overline{A^2} = \frac{V^2}{2} \tag{B1}$$

and

$$\overline{a^2} = V^2 f \int_0^{1/f} [\sin 2\pi ft - \sin 2\pi f(t-T)]^2 dt$$

$$\overline{a^2} = 2V^2 \sin^2 \pi f T$$

$$\overline{a^2} = 2V^2 \sin^2 (\pi f/2W). \tag{B2}$$

Here T is the sampling interval which is $\frac{1}{2}\omega$.

The samples constitute a signal of a single frequency f, and give a mean-squared output

$$\overline{A^2}(F(f))^2.$$

By assumption (5) the samples give a mean-squared output

$$\overline{\alpha^2}\langle (F(f))^2\rangle,$$

where

$$\langle (F(f))^2\rangle = \frac{1}{W}\int_0^{W} (F(f))^2 df. \tag{B3}$$

Hence, the noise-to-signal ratio will be

$$N/S = 4\alpha \sin^2 (\pi f/2W) \frac{\langle (F(f))^2\rangle}{(F(f))^2}. \tag{B4}$$

We should now observe that this result is independent of sample shape if we assume $F(f)$ to be the over-all audio response including effects of both sample shape and filter, since we can regard the audio filter as merely changing the shape of the sample in a linear manner.

We should note that quite different results would be obtained for signals with different statistics. Thus, if we retain assumption (1) and replace the rest by an assumption that all received samples add on a power basis and if we assume that any sample value lying between $+V$ and $-V$ is equally likely, without regard for previous samples, we find that

$$N/S = 2\alpha \tag{B5}$$

instead of (B4).

Let us make a comparison between (B4) and (B5). Suppose that in (B4) we assume $f=1{,}000$ and $W=5{,}000$, and assume a flat response from 0 to 5,000. Then (B4) gives

$$N/S = 0.38\alpha$$

while (B5) gives

$$N/S = 2\alpha.$$

This comparison merely serves to show that some signals suffer less than others when a missing sample is replaced by the preceding one. In the case of a sine wave of period large compared with the sampling period, the amplitude does not change much in a sampling period and the distortion or "noise" is small. In any case, we should note that the noise to signal ratio is dependent only on the fraction of missing samples and not on the signal level.

The values $f=1{,}000$ and $W=5{,}000$, assumed above, are typical of the experimental system. If we assumed a flat audio response from 0 to 5,000 cycles, we would have

$$\frac{\langle (F(f))^2\rangle}{(F(f))^2} = 1.$$

In the first part of the Appendix, α was estimated as 0.067. Hence, for this flat band we estimate the noise-to-signal power ratio as

$$(0.38)(0.067) = 0.025,$$

corresponding to 16 db.

Actually, an output filter narrower than 5,000 cycles was used in the experimental system, and furthermore, noise was measured with a 2B noise set using F1A line weighting. Using the over-all frequency characteristic of the system filter and the line weighting (filter),

$$\frac{\langle (F(f))^2\rangle}{(F(f))^2} = 0.26.$$

Using this value, the noise-to-signal power ratio is

$$(0.38)(0.067)(0.26) = 0.0066,$$

corresponding to a signal-to-noise ratio of 21.8 db. The large-signal signal-to-noise ratio measured for the case above was 20 db.

A Compact Broad-Band Microwave Quarter-Wave Plate*

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Summary—**Differential phase shift between two orthogonal TE_{11} waves in a circular hollow waveguide is achieved with a reflectionless array of capacitive pins. Using transmission-line theory, an analysis of such a structure is made and, under the assumption that the pin susceptance varies with frequency as $j\omega C$, a broad-band 3-pin array acting as a quarter-wave plate may be designed. Such an array, which is only one inch long at X-band, has been tested. A voltage ellipticity ratio of less than 1.1 and vswr less than 1.2 is maintained over a 12-per cent band.**

Introduction

Various methods of obtaining in round waveguide a 90° phase difference between two modes in space quadrature have been described in the literature.[1,2] The method used here makes use of a cascade of three capacitive pins which load the waveguide for one mode. Such a cascade will be frequency sensitive because of the element spacing, and the variation of susceptance with frequency. However, these two effects may be made to compensate each other over a frequency band.

Analysis

Consider the waveguide to be for one mode a loaded line as shown in Fig. 1. (It is assumed that the screws act as shunt elements if their diameter is small compared to a wavelength.) Using network theory a transmission matrix may be found expressing V_i and I_i as linear functions of V_0 and I_0.[3]

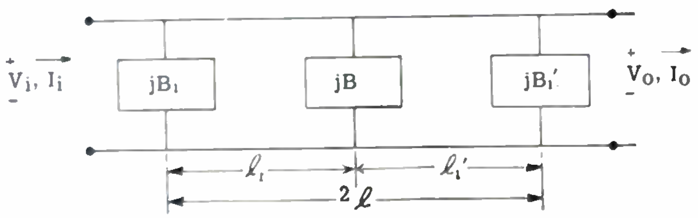

Fig. 1—Susceptance-loaded transmission line.

$$\begin{bmatrix} V_i \\ I_i \end{bmatrix} = \begin{bmatrix} t_{11} & t_{12} \\ t_{21} & t_{22} \end{bmatrix} \begin{bmatrix} V_0 \\ I_0 \end{bmatrix}, \tag{1}$$

where

$$t_{11} = \cos 2\beta l - \frac{B_1'}{Y_0}\sin 2\beta l - \frac{B}{Y_0}\sin\beta l_1 \cos\beta l_1' \tag{2a}$$

$$t_{12} = jZ_0\left(\sin 2\beta l - \frac{B}{Y_0}\sin\beta l_1 \sin\beta l_1'\right). \tag{2b}$$

β is the propagation constant and Z_0 and Y_0 are the characteristic impedance and admittance of the unloaded line.

For arbitrary phase shift θ, and a matched line, we want the loaded section of line of length l to have an electrical length $2\beta l+\theta$ with the characteristic impedance unchanged. This is equivalent to requiring

$$t_{11} = \cos(2\beta l + \theta) \tag{3a}$$

$$t_{12} = jZ_0 \sin(2\beta l + \theta). \tag{3b}$$

Solving (2) and (3) gives us

$$\frac{B}{Y_0} = \frac{\sin 2\beta l - \sin(2\beta l + \theta)}{\sin\beta l_1 \sin\beta l_1'} \tag{4a}$$

$$\frac{B_1'}{Y_0} = \frac{\sin\theta\cos\beta l_1 - (1-\cos\theta)\sin\beta l_1}{\sin\beta l_1' \sin(2\beta l + \theta)}. \tag{4b}$$

An equation for B_1/Y_0 may be obtained by exchanging l_1 and l_1' and B_1 and B_1' in (4b). Now let $\theta = 90°$, $l_1 = l_1' = l$. Then $B_1 = B_1'$. A plot of (4) under these conditions is shown in Fig. 2. Note that in the region $70° > \beta l > 45°$ both curves have positive slope for increasing values of λ (decreasing βl).

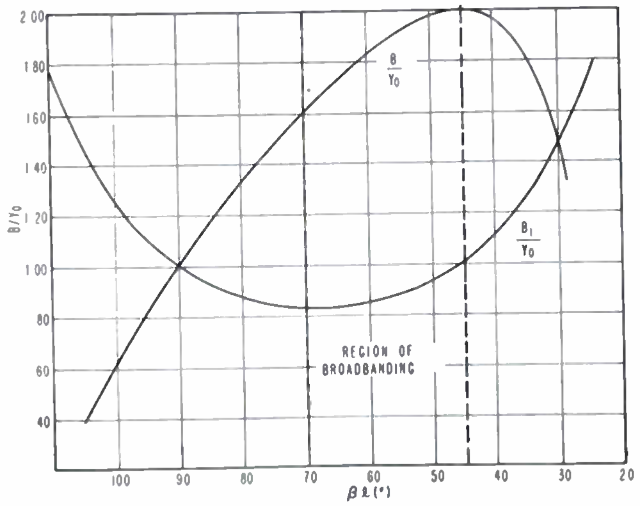

Fig. 2—Plot of (4a) and (4b) for $\theta = +90°$ and $l_1 = l_1' = l$.

The curves in Fig. 2 indicate the desired variation of B/Y_0 and B_1/Y_0 with frequency. We next assume that each pin approximates a lumped capacity. Then

* Decimal classification: R118.2×R310. Original manuscript received by the Institute, June 8, 1951; revised manuscript received, April 28, 1952.

† Antenna Research Branch, Radio Division I, Naval Research Laboratory, Washington 25, D. C.

[1] G. L. Ragan, "Microwave Transmission Circuits," Rad. Lab. Series, McGraw-Hill Book Co., Inc., vol. 9, pp. 369–378; 1948.

[2] A. G. Fox, "An adjustable waveguide phase changer," Proc. I.R.E., vol. 35, pp. 1489–1498; December, 1947.

[3] C. G. Montgomery, R. H. Dicke, and E. M. Purcell, "Principles of microwave circuits," Rad. Lab. Series, McGraw-Hill Book Co., Inc., vol. 8, p. 103; 1948.

$$\frac{B}{Y_0} = K\lambda_c \frac{\lambda_g}{\lambda_0^2} = K \frac{\lambda_c}{\lambda_0} \sqrt{\frac{1}{1 - \frac{\lambda_0^2}{\lambda_c^2}}}\,[4] \qquad (5a)$$

$$\frac{B_1}{Y_0} = K_1\lambda_c \frac{\lambda_g}{\lambda_0^2} = K_1 \frac{\lambda_c}{\lambda_0} \sqrt{\frac{1}{1 - \frac{\lambda_0^2}{\lambda^2}}}, \qquad (5b)$$

where K and K_1 are constants of proportionality and λ_c is the waveguide cutoff wavelength, a normalizing constant.

A plot of the factor $\lambda_g\lambda_c/\lambda_0^2$ versus λ_0/λ_c shows that for a region $0.75 < \lambda_0/\lambda_c < 1$ this assumed variation of B/Y_0 and B_1/Y_0 has positive slope with increasing wavelength. Thus, the variation of B/Y_0 and B_1/Y_0 required in Fig. 2 can be approximately met over a range of frequencies.

To find the best value of l, equate the assumed variation of B/Y_0 and B_1/Y_0 ((5)) with the desired variation ((4)), which gives

$$K = \frac{\lambda_0^2}{\lambda_c\lambda_g} \frac{\sin 2\beta l - \cos 2\beta l}{\sin^2 \beta l}, \qquad (6a)$$

$$K_1 = \frac{\lambda_0^2}{\lambda_c\lambda_g} \frac{1}{\sin \beta l(\cos \beta l + \sin \beta l)}. \qquad (6b)$$

It is desired to find the value of l which will make K and K_1 most nearly constant over a range of values of λ_0. To do this, the right-hand members of (6a) and (6b) may be plotted versus λ_0/λ_c for various values of l/λ_c as a parameter. It can then be seen that both K and K_1 have stationary values for a value of $l/\lambda_c = 0.25$ and for values of λ_0/λ_c between 0.80 and 0.90. These values determine the broad-band design.

Experimental Procedure

For an experimental check, a section was built as in Fig. 3. Each shunt element was made up of a pair of adjustable screws entering from opposite sides of the waveguide. To find the proper insertion a simple experiment was performed in which the standing-wave ratio introduced into a matched line by a single pair of opposed screws was measured. The susceptance versus insertion depth was then calculated.

The section was placed with the plane of the screws oriented at 45° to an incident, linearly polarized wave and the ver (voltage ellipticity ratio) measured beyond the screw section by means of a rotatable probe.

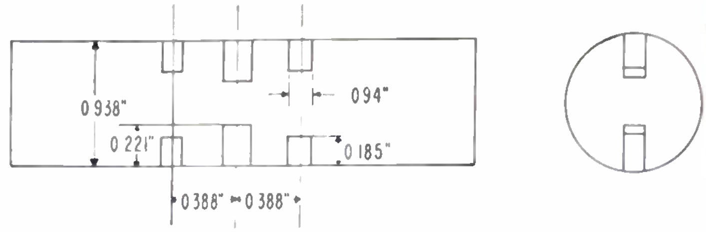

Fig. 3—Waveguide section; schematic, showing calculated dimensions.

Results are shown in Fig. 4. It was found that if each screw insertion was increased approximately 0.004 inch a somewhat flatter ellipticity characteristic was obtained. This adjustment was needed, it was thought, to compensate for effects neglected in the above theory, such as the series reactance of the pins, the coupling between adjacent pairs, and slight deformations of the pipe from perfect circularity.

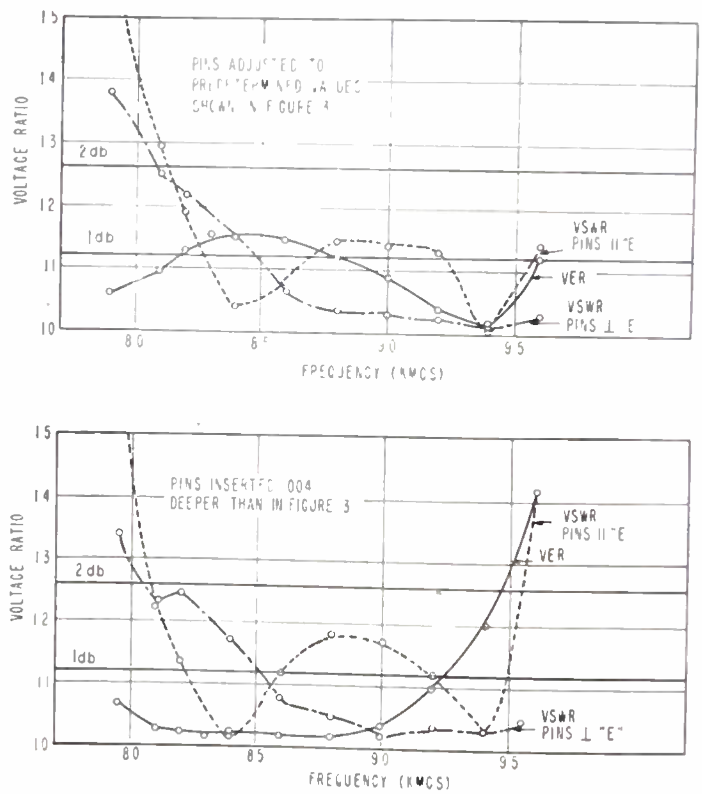

Fig. 4(a) and (b)—Voltage ellipticity and voltage standing-wave ratios versus frequency.

[4] *Ibid.*, p. 169

Fundamental Aspects of Linear Multiplexing*

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Summary—A linear multiplex system is defined as one in which the separation of signals belonging to different channels is achieved by the use of linear, time-variant or time-invariant, filters. It is shown that a fundamental property of such systems is that the sets of signals associated with their respective channels are linear and disjoint. Conversely, signals that belong to linear and disjoint sets can be transmitted simultaneously and separated at the receiving end by means of linear, generally time-variant, filters. It is shown that frequency-band compression cannot be achieved with a linear system.

In geometrical terms, the extraction of signals belonging to a specified channel may be regarded as the projection of the signal space on the manifold corresponding to the channel in question along a complementary manifold. The filtering process is formulated in analytical terms via the λ-domain technique. Methods of synthesizing linear multiplex systems of other than the conventional frequency- or time-division types are indicated.

I. Introduction

Multiplex communication systems may be divided into two basic categories, linear systems and nonlinear systems. A multiplex system is *linear* or *nonlinear* according as the separation of signals belonging to different channels is effected by linear or nonlinear filters. It is *synchronous* or *asynchronous*, depending on whether the signal-separating filters are time-variant or time-invariant.

The conventional time- and frequency-division multiplex systems, as well as the system based on the use of orthogonal functions,[1] fall into the category of linear systems. On the other hand, most of the code multiplex and asynchronous[2] systems fall into the category of nonlinear systems.

The purpose of this paper is to examine some of the basic theoretical aspects of linear multiplexing in the light of function space representation of signals. No attempt will be made to describe or analyze particular multiplex systems in detail, or to compare their performance in regard to bandwidth requirements, cross talk and interchannel interference, and other factors of practical importance. Thus, the present paper constitutes a preliminary investigation of some of the fundamental aspects of linear multiplexing, not a detailed study of various types of linear multiplex systems.

II. Basic Theory

For simplicity of analysis, it will be assumed that the system has only two channels, which will be designated by I and II. This restriction is not an essential one, and it does not detract from the generality of the results.

A functional diagram of an idealized multiplex system is shown in Fig. 1; $u_1(t)$ and $u_2(t)$ represent two possible messages, the set of all such messages constituting the message set (space) $\mathfrak{M}$. The messages $u_1(t)$ and $u_2(t)$ are operated upon by the modulators M_1 and M_2, respectively, resulting in the signals $v_1(t)$ and $v_2(t)$. These signals are added, and their sum $v(t)=v_1(t)+v_2(t)$ is transmitted through a common channel to the receiver.

Disregarding the noise, distortion, and delay, the received signal is identical with $v(t)$. At the receiving end, $v(t)$ is processed by two filters F_1 and F_2, which, ideally, yield the transmitted signals $v_1(t)$ and $v_2(t)$. These signals are operated upon by the demodulators D_1 and D_2, whose function is to transform the signals $v_1(t)$ and $v_2(t)$ into the transmitted messages $u_1(t)$ and $u_2(t)$, respectively. Needless to say, in practice $u_1(t)$ and $u_2(t)$ are obtained after some time delay and, in general, with some distortion.

The modulator M_1 transforms the message set $\mathfrak{M}$ into

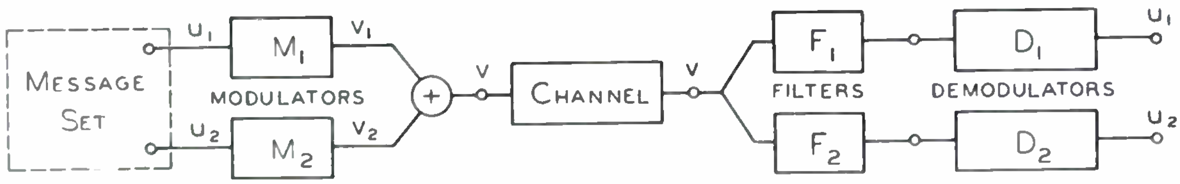

Fig. 1—Functional diagram of a typical two-channel multiplex system.

a set S_1, which consists of all possible signals $v_1(t)$ in channel I. Similarly, M_2 transforms $\mathfrak{M}$ into a set S_2, which consists of all possible signals $v_2(t)$ in channel II. These two sets give rise, collectively, to a set S, which consists of all possible signals $v(t)$, $v(t)=v_1(t)+v_2(t)$, at the receiving end of the system. Clearly, the sets S_1 and S_2 are subsets of S.

The sets in question assume a familiar form in the case of a frequency-division multiplex system. Here the set S_1 consists of all signals which do not contain frequencies outside of a band, say from f_1 to f_2 cps, while S_2 comprises those signals which do not contain frequencies outside, say, f_3 to f_4 cps, with $f_3>f_2$. The set S,

* Decimal classification: R460. Original manuscript received by the Institute, October 11, 1951.

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[1] N. Marchand and H. R. Holloway, "Multiplexing by Orthogonal Functions," IRE Conference on Airborne Electronics, Dayton, Ohio; May 23, 1951. See also, "Analysis of dot-sequential color television," Proc. I.R.E., vol. 39, pp. 1280–1287; October, 1951.

[2] W. D. White, "Theoretical aspects of asynchronous multiplexing," Proc. I.R.E., vol. 38, pp. 270–275; March, 1950.

then, consists of signals whose frequency content is confined to the bands f_1 to f_2 cps and f_3 to f_4 cps.

In general, the signals $v_1(t)$ and $v_2(t)$ are of a random nature and the sets S_1 and S_2 are ensembles, i.e., sets with a probability measure. However, the statistical structures of S_1 and S_2 are seldom well defined and are usually lacking in stability. For this reason, it is expedient to treat S_1 and S_2 simply as sets, and not as ensembles.

The principal components of a multiplex system are the filters F_1 and F_2, whose function is to extract the transmitted signals $v_1(t)$ and $v_2(t)$ from the received signal $v(t)$, $v(t)=v_1(t)+v_2(t)$. The establishment of some of the basic properties of these filters is the main concern of the following discussion:

Since the filters F_1 and F_2 have similar functions, it will be sufficient to consider just one of them, say F_1. Using the symbolic notation

$$y = Fx \tag{1}$$

to indicate that $y(t)$ is the response of a filter F (at rest) to a signal $x(t)$, the operation performed by F_1 may be written

$$F_1(v_1 + v_2) = v_1. \tag{2}$$

This relation expresses the fact that, ideally, the filter F_1 yields the signal $v_1(t)$, without any distortion or delay, by operating on the sum of $v_1(t)$ and $v_2(t)$.

It is evident that (2) should hold for any signal $v_1(t)$ belonging to S_1, and any signal $v_2(t)$ belonging to S_2. Thus, a more complete statement of the function of the filter F_1 is

$$F_1(v_1 + v_2) = v_1 \tag{3}$$

for all v_1 in S_1 and all v_2 in S_2.

It is quite clear that (3) cannot possibly be satisfied unless the sets S_1 and S_2 possess some rather special properties. Two sets of signals, S_1 and S_2, will be called *linearly separable* if there exists a linear filter F_1 satisfying (3). Clearly, the sets of signals associated with the channels of a linear multiplex system must be linearly separable.

A question of basic importance is: Under what conditions are two sets of signals linearly separable? An answer to this question is formulated in the sequel.

First, it will be shown that S_1 and S_2 must be *linear* sets. In mathematical terminology, a set is called *linear* if it contains every finite linear combination of its elements. Thus, if v' and v'' are any two elements of a set S, then S is a linear set if $av'+bv''$, where a and b are arbitrary constants, is an element of S.

A simple example of a linear set is the set of signals which do not contain frequencies higher than, say, f_0 cps. Clearly, if $v'(t)$ and $v''(t)$ are two such signals, then $av'(t)+bv''(t)$ is also a signal of the same type.

The fact that S_1 and S_2 must be linear sets is easily established. Thus, (3) may equivalently be written in the form

$$F_1v_1 = v_1 \quad \text{for all } v_1 \text{ in } S_1 \tag{4}$$

$$F_1v_2 = 0 \quad \text{for all } v_2 \text{ in } S_2. \tag{5}$$

If (4) holds for v_1' and v_1'' which are members of S_1, then it also holds for $av_1'+bv_1''$, where a and b are arbitrary constants. Consequently, $av_1'+bv_1''$ is a member of S_1, and hence S_1 is a linear set. The same reasoning applies to S_2. Thus, a necessary condition for linear separability is that S_1 and S_2 should be linear sets.

Next, it will be shown that S_1 and S_2 must be *disjoint* sets, that is, should not have any element in common except the null signal $v(t)\equiv 0$. Suppose that a signal $v(t)\neq 0$ belongs to both S_1 and S_2; then, according to (4) the response of F_1 to this signal is $v(t)$. But, according to (5) the response is zero. In consequence of this contradiction, a nonzero signal $v(t)$ cannot belong to both S_1 and S_2—which implies that the sets S_1 and S_2 are disjoint.

To summarize, linearly separable sets of signals have two fundamental properties: (a) linearity and (b) disjointness.

An important question which arises at this point is: Are these two properties sufficient? In other words, are two sets of signals S_1 and S_2 linearly separable—and hence usable in a linear multiplex system—provided only that they are linear and disjoint? It will be shown in the next two sections that the answer to this question is in the affirmative, with one generally unimportant qualification concerning the physical realizability of the filters F_1 and F_2. Thus, it appears that, in so far as linear multiplex systems are concerned, the only essential properties that S_1 and S_2 need possess are those of linearity and disjointness.

The practical significance of this conclusion stems from the fact that one can readily construct a large variety of sets of signals which are linear and disjoint, and which, consequently, can be employed for synthesizing various types of linear multiplex systems other than those of the conventional time- and frequency-division types. Thus, the designer of a multiplex system finds at his disposal a number of different types of systems of which he can choose one best suited for his purposes. It is well to emphasize, however, that of all possible linear multiplex systems, those employing frequency- and time-division, although not necessarily the best in all important respects, are certainly the simplest in both conception and design.

III. Geometrical Interpretation

Many aspects of multiplexing are greatly clarified when considered in the light of geometrical representation of signals. In applying the geometrical approach, a signal $v(t)$ is represented as a vector v in a vector space (signal space) Σ. Since the use of the geometrical approach in communication theory is of rather recent origin,[3] it might be helpful to precede the application of

[3] C. E. Shannon, "Communication in the presence of noise," Proc. I.R.E., vol. 37, pp. 10–21; January, 1949.

this approach to linear multiplex systems with a few words of introduction. Detailed treatments of function spaces can be found in the literature of mathematics[4,5] and physics.[6,7]

Consider a signal $v(t)$ defined over the infinite interval $-\infty<t<\infty$. When $v(t)$ is represented as a vector v in a signal space Σ, the co-ordinates of v are, roughly speaking, the coefficients of a suitable resolution of $v(t)$ into a set of component signals. In particular, when $v(t)$ is a band-limited signal of band f_0, the co-ordinates of v may be taken as the values of $v(t)$ at regularly spaced sampling instants $1/2f_0$ seconds apart.

A linear set of signals, such as S_1, corresponds in the signal space to a *linear manifold* Σ_1, which is a linear subspace of Σ. Familiar examples of linear manifolds are straight lines and planes in the three-dimensional space. Naturally, one cannot visualize a multidimensional linear manifold. However, thinking in "three-dimensional" terms is very helpful since many of the properties of multidimensional linear manifolds can be inferred from those of straight lines and planes.

It was shown in the preceding section that linearly separable sets of signals are (a) linear and (b) disjoint. If S_1 and S_2 are two such sets, the corresponding manifolds Σ_1 and Σ_2 in the signal space Σ are likewise linear and disjoint. In geometrical terms, this means that Σ_1 and Σ_2 are linear manifolds which do not intersect each other except at the origin. A simple example of such manifolds is furnished by two straight lines passing through the origin. Another example is that of a plane passing through the origin and a straight line (also passing through the origin) which does not lie in this plane.

For purposes of visualization, it will suffice to use a two-dimensional diagram of the signal space Σ, as shown in Fig. 2, with the tacit understanding that Σ actually is infinite-dimensional. In a two-dimensional diagram, the manifolds Σ_1 and Σ_2 assume the form of two straight lines emanating from the origin. The signals $v_1(t)$ and $v_2(t)$, belonging to S_1 and S_2, respectively, correspond to the vectors v_1 and v_2 in Σ_1 and Σ_2. The sum signal $v(t)=v_1(t)+v_2(t)$ at the receiving end of the multiplex system corresponds to the vector $v=v_1+v_2$, the resultant of v_1 and v_2.

At the receiving end, the filter F_1 is supposed to extract the signal $v_1(t)$ from the received signal $v(t)=v_1(t)+v_2(t)$. Geometrically, the problem is that of finding v_1, given v and knowing that v_1 and v_2 lie in Σ_1 and Σ_2, respectively. The solution of this problem is clear; v_1 is obtained by projecting v on Σ_1 along Σ_2. It can readily be shown that the same solution applies in infinite-dimensional signal spaces. Thus, the operation performed by the filter F_1 may be interpreted in geometrical terms as the projection of the signal space Σ on the manifold Σ_1 (representing the set of signals S_1 in channel I) along the manifold Σ_2 (representing the set of signals S_2 in channel II).

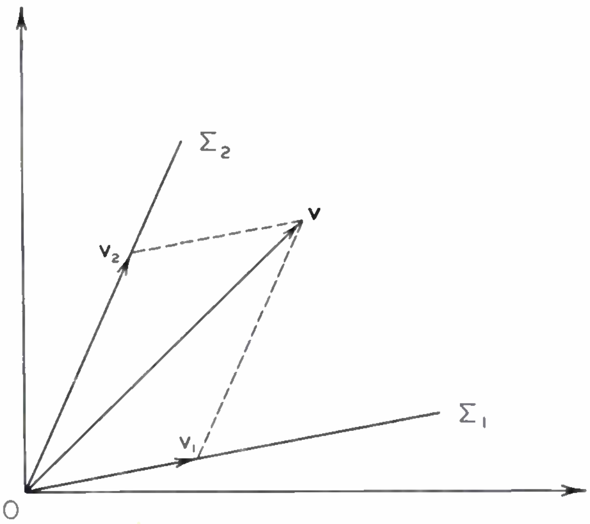

Fig. 2—Geometrical representation of the filtering process for the case where Σ_1 and Σ_2 are linear manifolds.

This geometrical interpretation is helpful in several respects. In particular, it suggests a simple way of characterizing the filter F_1. As a convenient illustration, consider a four-dimensional signal space in which the co-ordinates of a point represent the values of the corresponding signal at four equispaced sampling instants, and suppose that the manifolds Σ_1 and Σ_2 are two planes passing through the origin. Assume that the plane Σ_1 is specified in terms of two vectors with components $(k_{11}, k_{21}, k_{31}, k_{41})$ and $(k_{12}, k_{22}, k_{32}, k_{42})$ which it contains, and that Σ_2 is similarly specified by two vectors whose components are $(k_{13}, k_{23}, k_{33}, k_{43})$ and $(k_{14}, k_{24}, k_{34}, k_{44})$. With these vectors, construct the matrix

$$k = \begin{Vmatrix} k_{11} & k_{12} & k_{13} & k_{14} \\ k_{21} & k_{22} & k_{23} & k_{24} \\ k_{31} & k_{32} & k_{33} & k_{34} \\ k_{41} & k_{42} & k_{43} & k_{44} \end{Vmatrix}, \tag{6}$$

whose columns are the vectors in question.

In geometrical terms, the filter F_1 takes the received signal $v(t)$—represented by a four-dimensional vector v—and projects it on the plane Σ_1 along Σ_2, thus yielding the signal $v_1(t)$ in channel I. Regarding v_1 and v_2 as column matrices, the transformation of v into v_1 may be written in a matrix form

$$v_1 = W_1 v, \tag{7}$$

where the 4×4 matrix W_1 constitutes, in effect, a matrix representation of the impulsive response of the filter F_1. The elements of W_1 can readily be found by making use of the fact that, geometrically, W_1 represents the projection of Σ on Σ_1 along Σ_2. Thus, one obtains the following expression for the general element of W_1:

[4] M. Frechet, "Les Espaces Abstraits," Gauthier-Villars, Paris; 1928.
[5] F. J. Murray, "Linear Transformations in Hilbert Space," Princeton University Press, Princeton, N. J.; 1941.
[6] J. von Neumann, "Mathematische Grundlagen der Quantenmechanik," Dover Publications, Inc., New York, N. Y.; 1943.
[7] H. Weyl, "Theory of Groups and Quantum Mechanics," Dover Publications, Inc., New York, N. Y.; 1949.

$$W_{1(ij)} = \sum_{\lambda=1}^{2} k_{i\lambda} k_{\lambda j}^{-1}. \tag{8}$$

Here k_{ij} is the general element of the matrix k, and k_{ij}^{-1} is that of the inverse of k. It can easily be verified that, given the vector $v = v_1 + v_2$, where v_1 and v_2 are vectors in Σ_1 and Σ_2, one obtains the vector v_1 by performing the matrix operation

$$v_1 = W_1 v. \tag{9}$$

This is equivalent to operating on the sum signal v with a filter F_1 whose impulsive response (in matrix form) is W_1.

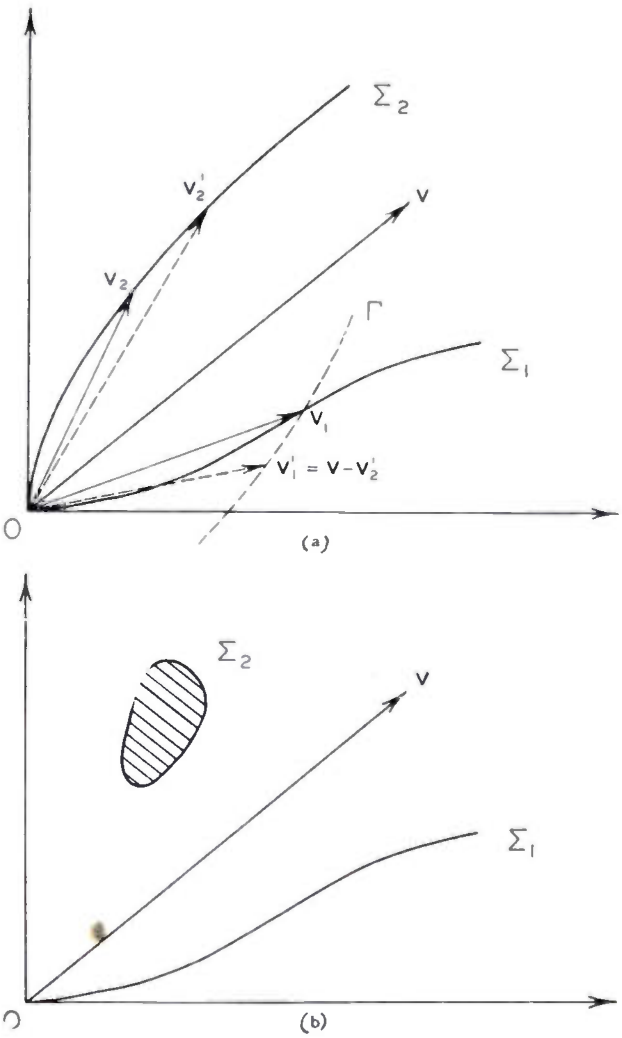

Fig. 3—(a) Case where v_1 can be extracted from v_1+v_2. (b) Case where v cannot be extracted from v_1+v_2.

At this point it is worth while to digress briefly and consider the case where the manifolds Σ_1 and Σ_2 are nonlinear.[8] Suppose that the two-dimensional diagram of the manifolds in question is of the form shown in Fig. 3(a). Here Σ_1 and Σ_2 represent the loci of the tips of v_1 and v_2. It is clear that, given the sum $v = v_1 + v_2$ and the manifolds Σ_1 and Σ_2, one can determine v_1 by plotting the locus of $v - v_2'$, where v_2' is a vector in Σ_2, and finding the intersection of this locus, Γ, with Σ_1. Thus, the extraction of v_1 is achieved essentially by solving a system of simultaneous equations defining Σ_1 and Γ for the co-ordinates of v_1.

In the case under consideration, the filtering operation

$$F_1 v = v_1, \tag{10}$$

which maps Σ onto Σ_1, is nonlinear. In this connection, an important observation is that v_1 can, in principle, be extracted from the sum $v_1 + v_2$ (although in some cases the result may not be unique) provided the sum of the dimensions of Σ_1 and Σ_2 does not exceed that of the signal space Σ. (In the example under consideration, Σ_1 and Σ_2 are one-dimensional and Σ is two-dimensional.)

Now suppose that Σ_1 has the same form as in Fig. 3(a), but Σ_2 is two-dimensional. This case is illustrated in Fig. 3(b). Here, the shaded area represents the possible locations of the tip of v_2. In this case, it is quite evident that, given the sum $v = v_1 + v_2$, one cannot extract v_1 by any operation, linear or nonlinear. More generally, it appears that if the sum of the dimensions of Σ_1 and Σ_2 exceeds that of Σ, a signal v_1 cannot be extracted from the sum $v_1 + v_2$ by the use of a continuous (linear or nonlinear) operation.

One important conclusion which stems from the above discussion is that it is impossible to achieve bandwidth compression with a linear multiplex system. More specifically, consider a message set which consists of all signals of bandwidth not exceeding f_0 and defined over a long time interval T. The dimension of the message space is $2f_0T$, and correspondingly the dimension of the manifolds Σ_1 and Σ_2 in the signal space Σ is likewise $2f_0T$. Since the dimension of Σ should not be less than the sum of the dimensions of Σ_1 and Σ_2, the dimension of Σ should be greater than or equal to $4f_0T$. This implies that the bandwidth of the common channel cannot be less than $2f_0$, which is the sum of the bandwidths of component channels. Consequently, the common channel bandwidth cannot be less than twice the bandwidth of message set, so long as the system is designed to transmit any possible message in the message set, i.e., any signal of bandwidth not exceeding f_0.

IV. Analytical Formulation

The filtering process discussed in the preceding two sections can be formulated in analytical terms by employing the technique of resolution of signals into a suitable set of component signals.[9]

Briefly, the conventional frequency analysis of a signal $v(t)$ is based on the resolution of $v(t)$ into a set of com-

[8] A somewhat more detailed discussion of this case is given in L. A. Zadeh, "Some basic problems in communication of information," *Trans. N. Y. Acad. Sci.*, vol. 14, pp. 201–204; March, 1952.

[9] L. A. Zadeh, "A general theory of linear signal transmission systems," *Jour. Frank. Inst.*, vol. 253, pp. 293–312; April, 1952.

plex exponential component signals. The resolution is expressed by the Laplace integral

$$v(t) = \frac{1}{2\pi j}\int_C e^{\lambda t}V(\lambda)d\lambda, \qquad (11)$$

where $\{e^{\lambda t}\}$ represents[10] a set of exponential component signals; λ is the complex frequency; C is a straight line parallel to the imaginary axis; and $V(\lambda)$ plays the role of a weighting function. The function $V(\lambda)$ is formally given by the (bilateral) Laplace transform of $v(t)$

$$V(\lambda) = \int_{-\infty}^{\infty} e^{-\lambda t}v(t)dt, \qquad (12)$$

where $e^{-\lambda t}$ is the kernel of the transformation.

More generally, $v(t)$ may be resolved into a set of component signals $\{k(t;\lambda)\}$ of some suitable but otherwise arbitrary form. Correspondingly, the expression for $v(t)$ in terms of the component signals reads

$$v(t) = \int_C k(t;\lambda)V(\lambda)d\lambda, \qquad (13)$$

where λ plays the same role as in (11). The weighting function $V(\lambda)$ is called the *spectral* function of $v(t)$ relative to $k(t;\lambda)$. The set of functions $\{k(t;\lambda)\}$ defines a so-called λ *domain*, of which the time and frequency domains are special cases. Thus, when $k(t;\lambda)=\delta(t-\lambda)$, where $\delta(t-\lambda)$ is a unit impulse, λ = time, the associated λ domain is the time domain, and $V(\lambda)$ is identical with $v(\lambda)$. On the other hand, when $k(t;\lambda)=e^{\lambda t}/2\pi j$, λ = complex frequency, the associated λ domain is the frequency domain, and $V(\lambda)$ is the Laplace transform of $v(t)$.

Returning to the consideration of linear multiplex systems, it will be noted that the conventional time- and frequency-division methods of multiplexing are particular cases of what might be referred to as "λ domain division." The term "division" implies that a set of signals $\{k(t;\lambda)\}$ corresponding to some set A_1 of values of λ is assigned to channel I, while a disjoint set of component signals corresponding to some nonoverlapping set A_2 of values of λ, is assigned to channel II. In the particular case of frequency division, the sets A_1 and A_2 consist of two nonoverlapping frequency bands, and the component signals are of the form $e^{\lambda t}/2\pi j$. On the other hand, in the case of time division, A_1 and A_2 consist of nonoverlapping time intervals, and the component signals are unit impulse functions $\delta(t-\lambda)$.

By using the notion of λ domain, the linearity and disjointness of two sets of signals can be conveniently formulated in terms of the spectral functions of the signals in question. Thus, if A_1 and A_2 are two nonoverlapping sets of values of λ (on the contour C), then two sets of signals S_1 and S_2 are linear and disjoint if the spectral function of any signal in S_1 vanishes for λ not in A_1, while that of any signal in S_2 vanishes for λ not in A_2. In other words, the spectra of these signals do not overlap in the associated λ domain. This mode of characterization of linear and disjoint sets of signals is analogous to that commonly used in the case of the frequency domain.

The sets A_1 and A_2 may be regarded as the "bands" occupied by channels I and II, respectively, in the associated λ domain. If B_1 is the "bandwidth" of channel I, i.e., Lebesgue measure of A_1, and B_2 is that of channel II, then the total "bandwidth" is the sum of B_1 and B_2; that is,

$$\text{total "bandwidth"} = B = B_1 + B_2. \qquad (14)$$

In the case of frequency division, (14) reduces to the statement that the total bandwidth of the system (in the conventional sense of bandwidth) is the sum of the bandwidths of component channels. Similarly, in the case of time division, (14) reduces to another obvious statement, namely, that the total transmission time is the sum of transmission times for individual channels. More generally, (14) implies that, in a multiplex system based on division in some domain, the total "bandwidth" occupied by the system in this domain is equal to the sum of "bandwidths" of component channels. Since any linear multiplex system is, in one way or another, based on division in a λ domain, it can be concluded that the "bandwidths" of component channels combine additively in some λ domain. In particular, in the case of a frequency-division multiplex system, the bandwidths combine additively in the frequency domain; while in the case of a time-division system, the "bandwidths," i.e., transmission times, combine additively in the time domain.

In the preceding section, the filtering operation performed by F_1 was characterized in geometrical terms as the projection of the signal space Σ on Σ_1 along Σ_2. In what follows, the filtering process is treated from a different point of view based on the resolution of signals into a set of component signals $\{k(t;\lambda)\}$.

A linear system such as F_1 can be characterized in a variety of ways. One convenient mode of characterization involves a so-called *characteristic function*[9] $K_1(t;\lambda)$, which is defined as the response of F_1 to $k(t;\lambda)$—both regarded as functions of time involving λ as a parameter. For example, in the time domain $k(t;\lambda)=\delta(t-\lambda)$ and the associated characteristic function is the response of F_1 to $\delta(t-\lambda)$, that is, the impulsive response of F_1. Thus in the time domain $K_1(t;\lambda)=W_1(t,\lambda)$, where $W_1(t,\lambda)$ denotes the impulsive response of F_1.

Consider now a multiplex system in which the "band" occupied by channel I in some specified λ domain, defined by $\{k(t;\lambda)\}$, is a set A_1 of values of λ (on the contour C). The filter F_1 should pass, without distortion or delay, all component signals $k(t;\lambda)$ in which λ belongs to A_1, and reject all those in which λ (on C) does not belong to A_1. From this it follows at once that the characteristic function $K_1(t;\lambda)$, which is the response of F_1 to $k(t;\lambda)$, is given by

[10] The braces are used here and elsewhere to denote a set of functions. Thus $\{e^{\lambda t}\}$ represents a set of functions $e^{\lambda t}$, each function corresponding to a value of λ on the contour C.

$$K_1(t;\lambda) = \begin{Bmatrix} k(t;\lambda), & \text{for } \lambda \text{ in } A_1 \\ 0, & \text{for } \lambda \text{ not in } A_1 \end{Bmatrix}. \tag{15}$$

It is more desirable to characterize the filter F_1 in terms of its impulsive response $W_1(t, \xi)$, i.e., the response to a unit impulse $\delta(t-\xi)$. It can readily be shown[9] that the relation between the impulsive response and a characteristic function is

$$W_1(t, \xi) = \int_C K_1(t;\lambda)k^{-1}(\lambda;\xi)d\lambda, \tag{16}$$

where $k^{-1}(\lambda; t)$ is the inverse of $k(t; \lambda)$. [The inverse of $k(t; \lambda)$, $k^{-1}(\lambda; t)$, is related to $k(t; \lambda)$ by the equation

$$\int_C k(t;\lambda)k^{-1}(\lambda;\xi)d\lambda = \delta(t-\xi). \tag{17}$$

The inverse of $k(t;\lambda)=e^{\lambda t}/2\pi j$ is $k^{-1}(\lambda; t)=e^{-\lambda t}$. Similarly, the inverse of $k(t;\lambda)=\delta(t-\lambda)$ is $k^{-1}(\lambda;t)=\delta(t-\lambda)$.]

To obtain the expression for the impulsive response of F_1, it is sufficient to substitute the expression for $K_1(t; \lambda)$ (15) into (16). This yields

$$W_1(t, \xi) = \int_{A_1} k(t;\lambda)k^{-1}(\lambda;\xi)d\lambda, \tag{18}$$

where the integral is taken over the set A_1, i.e., the "band" of channel 1. This result provides the desired expression for the impulsive response of F_1. Once $W_1(t, \xi)$ has been obtained, the filter F_1 may be synthesized directly in the form of a tapped delay-line filter with time-varying weights.

The familiar cases of the time and frequency domains will serve to illustrate the calculation of $W_1(t, \xi)$ from the general expression (18) derived above. First, consider a frequency-division system in which the band A_1 is the frequency interval $-\omega_0 \leqq \omega \leqq \omega_0$. In this case,

$$k(t;\lambda) = \frac{e^{j\omega t}}{2\pi j}, \qquad \lambda = j\omega \tag{19}$$

and

$$k^{-1}(\lambda; t) = e^{-j\omega t}. \tag{20}$$

Substituting these expressions in (18) yields

$$W_1(t,\xi) = \frac{1}{2\pi j}\int_{-j\omega_0}^{j\omega_0} e^{j\omega t}e^{-j\omega\xi}d(j\omega)$$
$$= \frac{\sin \omega_0(t-\xi)}{\pi(t-\xi)}, \tag{21}$$

which will be recognized as the impulsive response of a conventional low-pass filter with zero phase shift in the pass band.

Next consider a time-division system in which the band A_1 consists of intervals of length t_0, which are uniformly distributed with period T_0 on the t-axis. In other words, A_1 consists of values of λ satisfying the inequalities

$$nT_0 < \lambda < nT_0 + t_0, \quad n = 0, \pm 1, \pm 2, \cdots \tag{22}$$

In the case under consideration, $k(t; \lambda)$ and $k^{-1}(\lambda; t)$ are expressed by

$$k(t;\lambda) = \delta(t-\lambda) \tag{23}$$

and

$$k^{-1}(\lambda; t) = \delta(t-\lambda). \tag{24}$$

Substituting these in (18) gives

$$W_1(t,\xi) = \int_{A_1} \delta(t-\lambda)\delta(\lambda-\xi)d\lambda, \tag{25}$$

which simplifies to

$$W_1(t,\xi) = \delta(t-\xi), \quad \text{for } \xi \text{ in } A_1, \tag{26}$$
$$= 0, \quad \text{for } \xi \text{ not in } A_1.$$

Physically, this represents the impulsive response of a switch which is closed periodically every T_0 seconds for t_0 seconds. Needless to say, this is the usual form of the filter F_1 in the case of a time-division system.

It will be noted that the impulsive response of the filter F_1 in the case of a frequency-division system does not vanish for $t<\xi$. This implies that F_1 is not physically realizable since the impulsive response of any physical system is zero for $t<\xi$. However, $W_1(t-\beta, \xi)$, where β is a constant, may be made as small as desired for $t<\xi$ by making β sufficiently large. Thus, one can approximately realize a filter F_1^* whose impulsive response is $W_1(t-\beta, \xi)$. The filter F_1^* would yield the same output as F_1, but with a time-delay β.

The same problem is usually encountered in connection with other types of linear multiplex systems. Thus, in general, $W_1(t, \xi)$ as given by (18) does not vanish for $t<\xi$. Consequently, it is necessary to introduce a sufficiently long time delay β and realize approximately a filter F_1^* whose impulsive response is $W_1(t-\beta, \xi)$. This can always be done provided that $W_1(t, \xi)$ approaches zero uniformly as $t-\xi$ approaches minus infinity. With this generally unimportant qualification, one can state that the filter F_1 is physically realizable to within a constant time delay, with as small an error as desired.

It should be noted that the general expression for $W_1(t, \xi)$, obtained above, includes that expressed by (8) as a special case. Thus, (8) may be regarded as a particular case of (18), in which t and λ assume the discrete values identified by the subscripts i and j. The matrices k and k^{-1} are discrete forms of $k(t; \lambda)$ and its inverse $k^{-1}(\lambda; t)$, respectively. Finally, in (8), the summation over the subscripts associated with the manifold on which the projection takes place, corresponds to the integration over the set A_1 in (18).

V. Synthesis of a Linear Multiplex System

The conventional multiplex systems utilize essentially two types of linearly separable signals: (a) signals which occupy nonoverlapping "bands" in the time domain (time division) and (b) signals which occupy nonoverlapping bands in the frequency domain (frequency division). A question which naturally arises is: How can one synthesize, at least in theory, a system in which the signals occupy nonoverlapping bands in some domain other than the time or frequency domains?

The preceding discussion suggests two approaches to this problem, one analytical, the other geometrical. In the analytical approach, one chooses a suitable (continuous or discrete) set of component signals generated by a function $k(t; \lambda)$, and then assigns a set A_1 of values of λ to channel I and a nonoverlapping set A_2 to channel II. Correspondingly, the impulsive response of the filter F_1 (at the receiving end of channel I) is obtained from (18), while that of the filter F_2 (at the receiving end of channel II) is obtained from the same expression, except A_1 is replaced by A_2. Finally, the filters F_1 and F_2 are approximated to within a constant time delay by physically realizable filters F_1^* and F_2^*.

The chief difficulty in this approach is that, at present, the inverse functions $k^{-1}(\lambda; t)$ are known for only a relatively small number of $k(t; \lambda)$ functions, and of these not all are of practical interest. The removal of this difficulty requires an expanded catalogue of various types of $k(t; \lambda)$ functions and their respective inverses.

The geometrical approach, although similar in principle to the analytical approach, is better adapted than the latter for numerical work, and has the advantage of furnishing a visual picture (in three-dimensional space) of the multiplex process. In employing this approach, one selects some suitable (linear and disjoint) manifolds Σ_1 and Σ_2 in the signal space Σ, and assigns them to channels I and II, respectively. A vector v_1 in Σ_1 represents a signal $v_1(t)$ which may be transmitted over channel I. Similarly, a vector v_2 in Σ_2 corresponds to a signal $v_2(t)$ in channel II. At the receiving end, the separation of signals is accomplished by projecting the signal space Σ on the manifold Σ_1 along Σ_2, which yields the signals belonging to channel I, and by projecting Σ on Σ_2 along Σ_1, which gives the signals in channel II.

In practice, in employing the geometrical approach it is expedient to divide the infinite time interval $-\infty < t < \infty$ into a succession of reasonably short time intervals T, each of which corresponds approximately to a finite-dimensional signal subspace. To illustrate this procedure, suppose that each such subspace is approximated by a four-dimensional subspace of Σ. Assume that the manifolds Σ_1 and Σ_2 in this subspace are two nonintersecting planes, of which Σ_1 is defined by two vectors $k_1 = (k_{11}, k_{21}, k_{31}, k_{41})$ and $k_2 = (k_{12}, k_{22}, k_{32}, k_{42})$, while Σ_2 is similarly defined by $k_3 = (k_{13}, k_{23}, k_{33}, k_{43})$ and $k_4 = (k_{14}, k_{24}, k_{34}, k_{44})$.

A signal $v_1(t)$ in Σ_1 may be represented as a vector v_1 of the form

$$v_1 = \alpha_1 k_1 + \alpha_2 k_2, \tag{27}$$

where α_1 and α_2 are arbitrary constants. Similarly, v_2 may be written as

$$v_2 = \alpha_3 k_3 + \alpha_4 k_4, \tag{28}$$

where α_3 and α_4 are likewise arbitrary constants. By assigning numerical values to α_1, α_2, α_3, and α_4, one obtains two four-point signals v_1 and v_2. Repeating this process, not necessarily with the same k vectors, one obtains two trains of four-point signals which, collectively, constitute a pair of linearly separable signals $v_1(t)$ and $v_2(t)$ defined over an arbitrarily long interval.

The separation of these signals is achieved by projecting each four-dimensional subspace on Σ_1 along Σ_2 by the use of the matrix equation (9). By repeating this operation for each four-point interval, one can extract a composite signal $v_1(t)$ (consisting of a train of four-point signals), from the sum of $v_1(t)$ and $v_2(t)$. The same procedure (with the roles of Σ_1 and Σ_2 interchanged) is followed for the extraction of $v_2(t)$ from the sum of $v_1(t)$ and $v_2(t)$.

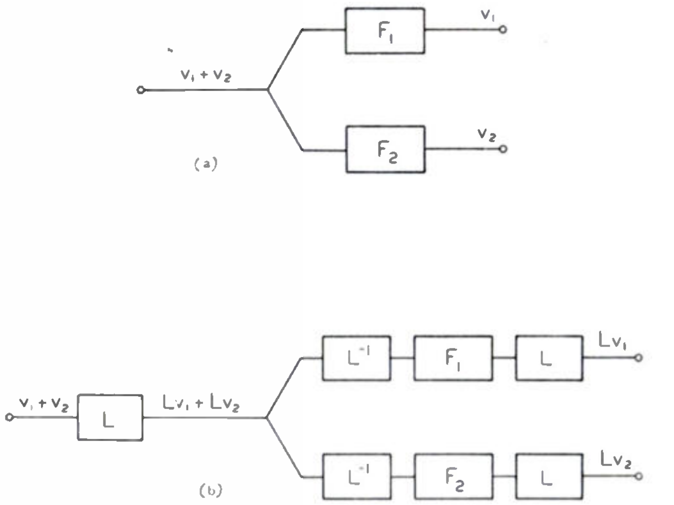

Fig. 4—Linear transformation of a multiplex system.

It will be noted that many different types of multiplex systems can be obtained from a given prototype system through a process of linear transformation[11] which is illustrated in Fig. 4. The prototype system is shown in Fig. 4(a), while the derived system is shown in Fig. 4(b). In the latter system, L represents a linear network, generally of a time-variant type, and L^{-1} stands for the inverse system. It is seen that, if the sets of signals associated with channels I and II in the prototype system are S_1 and S_2, the corresponding sets in the transformed system are $L(S_1)$ and $L(S_2)$, where $L(S)$ represents the set of signals resulting from operating with L on the elements of S.

It should be remarked that this method of derivation is applicable regardless of whether the prototype multiplex system is linear or not. It is essential, however, that L be a linear network.

[11] L. A. Zadeh and K. S. Miller, "Generalized ideal filters," *Jour. Appl. Phys.*, vol. 23, pp. 223-228; February, 1952.

A Method for the Construction of Minimum-Redundancy Codes*

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Summary—An optimum method of coding an ensemble of messages consisting of a finite number of members is developed. A minimum-redundancy code is one constructed in such a way that the average number of coding digits per message is minimized.

Introduction

ONE IMPORTANT METHOD of transmitting messages is to transmit in their place sequences of symbols. If there are more messages which might be sent than there are kinds of symbols available, then some of the messages must use more than one symbol. If it is assumed that each symbol requires the same time for transmission, then the time for transmission (length) of a message is directly proportional to the number of symbols associated with it. In this paper, the symbol or sequence of symbols associated with a given message will be called the "message code." The entire number of messages which might be transmitted will be called the "message ensemble." The mutual agreement between the transmitter and the receiver about the meaning of the code for each message of the ensemble will be called the "ensemble code."

Probably the most familiar ensemble code was stated in the phrase "one if by land and two if by sea." In this case, the message ensemble consisted of the two individual messages "by land" and "by sea", and the message codes were "one" and "two."

In order to formalize the requirements of an ensemble code, the coding symbols will be represented by numbers. Thus, if there are D different types of symbols to be used in coding, they will be represented by the digits $0, 1, 2, \cdots, (D-1)$. For example, a ternary code will be constructed using the three digits 0, 1, and 2 as coding symbols.

The number of messages in the ensemble will be called N. Let $P(i)$ be the probability of the ith message. Then

$$\sum_{i=1}^{N} P(i) = 1. \tag{1}$$

The length of a message, $L(i)$, is the number of coding digits assigned to it. Therefore, the average message length is

$$L_{av} = \sum_{i=1}^{N} P(i)L(i). \tag{2}$$

The term "redundancy" has been defined by Shannon[1] as a property of codes. A "minimum-redundancy code" will be defined here as an ensemble code which, for a message ensemble consisting of a finite number of members, N, and for a given number of coding digits, D, yields the lowest possible average message length. In order to avoid the use of the lengthy term "minimum-redundancy," this term will be replaced here by "optimum." It will be understood then that, in this paper, "optimum code" means "minimum-redundancy code."

The following basic restrictions will be imposed on an ensemble code:

(a) No two messages will consist of identical arrangements of coding digits.

(b) The message codes will be constructed in such a way that no additional indication is necessary to specify where a message code begins and ends once the starting point of a sequence of messages is known.

Restriction (b) necessitates that no message be coded in such a way that its code appears, digit for digit, as the first part of any message code of greater length. Thus, 01, 102, 111, and 202 are valid message codes for an ensemble of four members. For instance, a sequence of these messages 1111022020101111102 can be broken up into the individual messages 111-102-202-01-01-111-102. All the receiver need know is the ensemble code. However, if the ensemble has individual message codes including 11, 111, 102, and 02, then when a message sequence starts with the digits 11, it is not immediately certain whether the message 11 has been received or whether it is only the first two digits of the message 111. Moreover, even if the sequence turns out to be 11102, it is still not certain whether 111-02 or 11-102 was transmitted. In this example, change of one of the two message codes 111 or 11 is indicated.

C. E. Shannon[1] and R. M. Fano[2] have developed ensemble coding procedures for the purpose of proving that the average number of binary digits required per message approaches from above the average amount of information per message. Their coding procedures are not optimum, but approach the optimum behavior when N approaches infinity. Some work has been done by Kraft[3] toward deriving a coding method which gives an average code length as close as possible to the ideal when the ensemble contains a finite number of members. However, up to the present time, no definite procedure has been suggested for the construction of such a code

* Decimal classification: R531.1. Original manuscript received by the Institute, December 6, 1951.

† Massachusetts Institute of Technology, Cambridge, Mass.

[1] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. 27, pp. 398–403; July, 1948.

[2] R. M. Fano, "The Transmission of Information," Technical Report No. 65, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; 1949.

[3] L. G. Kraft, "A Device for Quantizing, Grouping, and Coding Amplitude-modulated Pulses," Electrical Engineering Thesis, M.I.T., Cambridge, Mass.; 1949.

to the knowledge of the author. It is the purpose of this paper to derive such a procedure.

Derived Coding Requirements

For an optimum code, the length of a given message code can never be less than the length of a more probable message code. If this requirement were not met, then a reduction in average message length could be obtained by interchanging the codes for the two messages in question in such a way that the shorter code becomes associated with the more probable message. Also, if there are several messages with the same probability, then it is possible that the codes for these messages may differ in length. However, the codes for these messages may be interchanged in any way without affecting the average code length for the message ensemble. Therefore, it may be assumed that the messages in the ensemble have been ordered in a fashion such that

$$P(1) \geqq P(2) \geqq \cdots \geqq P(N-1) \geqq P(N) \qquad (3)$$

and that, in addition, for an optimum code, the condition

$$L(1) \leqq L(2) \leqq \cdots \leqq L(N-1) \leqq L(N) \qquad (4)$$

holds. This requirement is assumed to be satisfied throughout the following discussion.

It might be imagined that an ensemble code could assign q more digits to the Nth message than to the $(N-1)$st message. However, the first $L(N-1)$ digits of the Nth message must not be used as the code for any other message. Thus the additional q digits would serve no useful purpose and would unnecessarily increase L_{av}. Therefore, for an optimum code it is necessary that $L(N)$ be equal to $L(N-1)$.

The kth prefix of a message code will be defined as the first k digits of that message code. Basic restriction (b) could then be restated as: No message shall be coded in such a way that its code is a prefix of any other message, or that any of its prefixes are used elsewhere as a message code.

Imagine an optimum code in which no two of the messages coded with length $L(N)$ have identical prefixes of order $L(N)-1$. Since an optimum code has been assumed, then none of these messages of length $L(N)$ can have codes or prefixes of any order which correspond to other codes. It would then be possible to drop the last digit of all of this group of messages and thereby reduce the value of L_{av}. Therefore, in an optimum code, it is necessary that at least two (and no more than D) of the codes with length $L(N)$ have identical prefixes of order $L(N)-1$.

One additional requirement can be made for an optimum code. Assume that there exists a combination of the D different types of coding digits which is less than $L(N)$ digits in length and which is not used as a message code or which is not a prefix of a message code. Then this combination of digits could be used to replace the code for the Nth message with a consequent reduction of L_{av}. Therefore, all possible sequences of $L(N)-1$ digits must be used either as message codes, or must have one of their prefixes used as message codes.

The derived restrictions for an optimum code are summarized in condensed form below and considered in addition to restrictions (a) and (b) given in the first part of this paper:

(c) $L(1) \leqq L(2) \leqq \cdots \leqq L(N-1) = L(N)$. (5)

(d) At least two and not more than D of the messages with code length $L(N)$ have codes which are alike except for their final digits.

(e) Each possible sequence of $L(N)-1$ digits must be used either as a message code or must have one of its prefixes used as a message code.

Optimum Binary Code

For ease of development of the optimum coding procedure, let us now restrict ourselves to the problem of binary coding. Later this procedure will be extended to the general case of D digits.

Restriction (c) makes it necessary that the two least probable messages have codes of equal length. Restriction (d) places the requirement that, for D equal to two, there be only two of the messages with coded length $L(N)$ which are identical except for their last digits. The final digits of these two codes will be one of the two binary digits, 0 and 1. It will be necessary to assign these two message codes to the Nth and the $(N-1)$st messages since at this point it is not known whether or not other codes of length $L(N)$ exist. Once this has been done, these two messages are equivalent to a single composite message. Its code (as yet undetermined) will be the common prefixes of order $L(N)-1$ of these two messages. Its probability will be the sum of the probabilities of the two messages from which it was created. The ensemble containing this composite message in the place of its two component messages will be called the first auxiliary message ensemble.

This newly created ensemble contains one less message than the original. Its members should be rearranged if necessary so that the messages are again ordered according to their probabilities. It may be considered exactly as the original ensemble was. The codes for each of the two least probable messages in this new ensemble are required to be identical except in their final digits; 0 and 1 are assigned as these digits, one for each of the two messages. Each new auxiliary ensemble contains one less message than the preceding ensemble. Each auxiliary ensemble represents the original ensemble with full use made of the accumulated necessary coding requirements.

The procedure is applied again and again until the number of members in the most recently formed auxiliary message ensemble is reduced to two. One of each of the binary digits is assigned to each of these two composite messages. These messages are then combined to form a single composite message with probability unity, and the coding is complete.

TABLE I

Optimum Binary Coding Procedure

Message Probabilities												
Original Message Ensemble	*Auxiliary Message Ensembles*											
	1	2	3	4	5	6	7	8	9	10	11	12
												1.00
											0.60	
										0.40	0.40	
									0.36	0.36		
								0.24	0.24	0.24		
0.20	0.20	0.20	0.20	0.20	0.20	0.20	0.20	0.20	0.20			
							0.20	0.20	0.20			
0.18	0.18	0.18	0.18	0.18	0.18	0.18	0.18	0.18				
						0.18	0.18	0.18				
					0.14	0.14	0.14					
0.10	0.10	0.10	0.10	0.10	0.10	0.10	0.10					
0.10	0.10	0.10	0.10	0.10	0.10	0.10						
0.10	0.10	0.10	0.10	0.10	0.10	0.10						
				0.10	0.10							
		0.08	0.08	0.08	0.08							
			0.08	0.08								
0.06	0.06	0.06	0.06	0.06								
0.06	0.06	0.06	0.06									
0.04	0.04	0.04	0.04									
*0.04	0.04	0.04										
0.04	0.04	0.04										
0.04	0.04											
	0.04											
0.03												
0.01												

Now let us examine Table I. The left-hand column contains the ordered message probabilities of the ensemble to be coded. N is equal to 13. Since each combination of two messages (indicated by a bracket) is accompanied by the assigning of a new digit to each, then the total number of digits which should be assigned to each original message is the same as the number of combinations indicated for that message. For example, the message marked *, or a composite of which it is a part, is combined with others five times, and therefore should be assigned a code length of five digits.

When there is no alternative in choosing the two least probable messages, then it is clear that the requirements, established as necessary, are also sufficient for deriving an optimum code. There may arise situations in which a choice may be made between two or more groupings of least likely messages. Such a case arises, for example, in the fourth auxiliary ensemble of Table I. Either of the messages of probability 0.08 could have been combined with that of probability 0.06. However, it is possible to rearrange codes in any manner among equally likely messages without affecting the average code length, and so a choice of either of the alternatives could have been made. Therefore, the procedure given is always sufficient to establish an optimum binary code.

The lengths of all the encoded messages derived from Table I are given in Table II.

Having now determined proper lengths of code for each message, the problem of specifying the actual digits remains. Many alternatives exist. Since the combining of messages into their composites is similar to the successive confluences of trickles, rivulets, brooks, and creeks into a final large river, the procedure thus far described might be considered analogous to the placing of signs by a water-borne insect at each of these junctions as he journeys downstream. It should be remembered that the code which we desire is that one which the insect must remember in order to work his way back upstream. Since the placing of the signs need not follow the same rule, such as "zero-right-returning," at each junction, it can be seen that there are at least 2^{12} different ways of assigning code digits for our example.

TABLE II

Results of Optimum Binary Coding Procedure

i	$P(i)$	$L(i)$	$P(i)L(i)$	Code
1	0.20	2	0.40	10
2	0.18	3	0.54	000
3	0.10	3	0.30	011
4	0.10	3	0.30	110
5	0.10	3	0.30	111
6	0.06	4	0.24	0101
7	0.06	5	0.30	00100
8	0.04	5	0.20	00101
9	0.04	5	0.20	01000
10	0.04	5	0.20	01001
11	0.04	5	0.20	00110
12	0.03	6	0.18	001110
13	0.01	6	0.06	001111
			$L_{av}=3.42$	

The code in Table II was obtained by using the digit 0 for the upper message and the digit 1 for the lower message of any bracket. It is important to note in Table I that coding restriction (e) is automatically met as long as two messages (and not one) are placed in each bracket.

Generalization of the Method

Optimum coding of an ensemble of messages using three or more types of digits is similar to the binary coding procedure. A table of auxiliary message ensembles similar to Table I will be used. Brackets indicating messages combined to form composite messages will be used in the same way as was done in Table I. However, in order to satisfy restriction (e), it will be required that all these brackets, with the possible exception of one combining the least probable messages of the original ensemble, always combine a number of messages equal to D.

It will be noted that the terminating auxiliary ensemble always has one unity probability message. Each preceding ensemble is increased in number by $D-1$ until the first auxiliary ensemble is reached. Therefore, if N_1 is the number of messages in the first auxiliary ensemble, then $(N_1-1)/(D-1)$ must be an integer. However $N_1 = N - n_0 + 1$, where n_0 is the number of the least probable messages combined in a bracket in the original ensemble. Therefore, n_0 (which, of course, is at least two and no more than D) must be of such a value that $(N-n_0)/(D-1)$ is an integer.

In Table III an example is considered using an ensemble of eight messages which is to be coded with four digits; n_0 is found to be 2. The code listed in the table is obtained by assigning the four digits 0, 1, 2, and 3, in order, to each of the brackets.

TABLE III

Optimum Coding Procedure for $D=4$

Message Probabilities: Original Message Ensemble	Auxiliary Ensembles			$L(i)$	Code
			→1.00		
		→0.40			
0.22	0.22	0.22		1	1
0.20	0.20	0.20		1	2
0.18	0.18	0.18		1	3
0.15	0.15			2	00
0.10	0.10			2	01
0.08	0.08			2	02
	→0.07				
0.05				3	030
0.02				3	031

Acknowledgments

The author is indebted to Dr. W. K. Linvill and Dr. R. M. Fano, both of the Massachusetts Institute of Technology, for their helpful criticism of this paper.

Coding with Linear Systems*

JOHN P. COSTAS†, ASSOCIATE, IRE

Summary—**Message transmission over a noisy channel is considered. Two linear networks are to be designed: one being used to treat the message before transmission and the second to filter the treated message plus channel noise at the receiving end. The mean-square error between the actual transmission circuit output and the delayed message is minimized for a given allowable average signal power by proper network design. Numerical examples are given and discussed.**

I. Introduction

THE PROBLEM to be considered here is that of message transmission over a noisy channel. As shown in Fig. 1, a message function, $f_m(t)$, is to be sent down a channel into which a noise function, $f_n(t)$, is introduced additively. The resultant system output is represented by $f_0(t)$. In most communication systems, the opportunity exists to code the message before its introduction into the transmission channel. Recently, Wiener, Shannon, and others have considered coding processes of a rather complex nature wherein the message function is sampled, quantized, and the resulting sample values converted into a pulse code for transmission. Although this technique may be quite useful in many instances, its application will be restricted by the complexity of the terminal equipment required. In this discussion, the coding and decoding systems will be limited to linear networks. In Fig. 1, network $H(\omega)$ will be used to code the message before transmission and network $G(\omega)$ will perform the necessary decoding. Network $H(\omega)$ must be designed so that the message is predistorted or coded in such a way as to enable the decoding or filtering network $G(\omega)$ to give a better system output than would have been possible had the message itself been sent without modification.

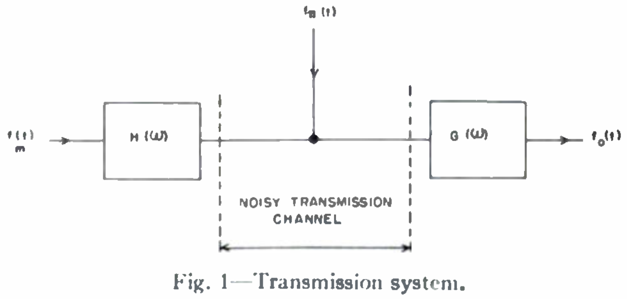

Fig. 1—Transmission system.

Before going further, a criterion of performance must be chosen for the transmission system. That is, some measurable quantity must be decided upon to enable us to determine whether one particular network pair $H(\omega)-G(\omega)$ is more satisfactory than some other pair. No single performance criterion can be expected to apply in all situations and no such claim is made for the mean-square error measure of performance which is to be used

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here. Wiener[1] and Lee[2] have used the mean-square criterion extensively in the past although Lee has shown in an unpublished memorandum that many other error criteria can be handled mathematically. The mean-square error, $\mathcal{E}$, of the system of Fig. 1 is defined by

$$\mathcal{E} = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{T} [f_0(t) - f_m(t-\alpha)]^2 dt. \tag{1}$$

For a given transmission delay α, networks $H(\omega)$ and $G(\omega)$ must be designed so that the mean-square difference between the actual system output, $f_0(t)$, and the delayed message, $f_m(t-\alpha)$, is minimized. The complete derivation for the design of the optimizing networks will appear in a forthcoming report[3] of the Research Laboratory of Electronics, M.I.T., and only the final results of this report will be discussed here.

II. Optimum Network Pair

If, in Fig. 1, the optimum network pair $H(\omega)-G(\omega)$ is used, the mean-square error will be the minimum possible for the specified delay α, and this error will be denoted by $\mathcal{E}_{\min}$. If the delay is increased and the new minimum error is computed, it will be found to be smaller than before. That is, the minimum error will be smaller the larger the allowable system delay. In the limit when the delay time becomes infinite, the smallest possible error obtainable using linear networks will result. This error is called the irremovable error and is indicated by $\mathcal{E}_{\text{irr}}$. That is,

$$\mathcal{E}_{\text{irr}} = \mathcal{E}_{\min}(\alpha) \quad \text{for} \quad \alpha \to \infty. \tag{2}$$

Thus, an optimum network design based on a long (theoretically infinite) delay will result in the best system performance. The irremovable error is an important item since it represents the ultimate possible performance of the system of Fig. 1.

Before proceeding to the optimum-design equations for the coding and decoding filters, certain assumptions and constraints must be introduced. The power-density spectra of the message and noise are assumed to be known and will be indicated by

$\Phi_{mm}(\omega)$ = power-density spectrum of message function $f_m(t)$

$\Phi_{nn}(\omega)$ = power-density spectrum of noise function $f_n(t)$.

The message and noise in this discussion are assumed to be independent (uncorrelated) variables. Finally, a constraint must be placed on the average signal power at the input to the transmission channel. Since $|H(\omega)|^2\Phi_{mm}(\omega)$ represents the power-density spectrum of the transmitted signal, the average signal power, $P_{\text{ave.}}$, will be given by

$$\int_{-\infty}^{\infty} |H(\omega)|^2 \Phi_{mm}(\omega) d\omega = P_{\text{ave.}} \tag{3}$$

Consider now that some arbitrary network is chosen for $H(\omega)$. Under the condition of a fixed (but not necessarily optimum) coding network, the best decoding network, $G(\omega)$, will be given by

$$G(\omega) = \frac{H(-\omega)\Phi_{mm}(\omega)e^{-j\alpha\omega}}{|H(\omega)|^2\Phi_{mm}(\omega) + \Phi_{nn}(\omega)}, \quad \alpha \to \infty. \tag{4}$$

The irremovable system error which will result when an arbitrary $H(\omega)$ is used in conjunction with a $G(\omega)$ designed according to (4) is given by

$$\mathcal{E}_{\text{irr}}^{(H\ \text{fixed})} = \int_{-\infty}^{\infty} \frac{\Phi_{mm}(\omega)\Phi_{nn}(\omega)}{|H(\omega)|^2\Phi_{mm}(\omega) + \Phi_{nn}(\omega)} d\omega. \tag{5}$$

The optimum coding network may now be found by solving for that $H(\omega)$ function which minimizes (5) while satisfying the power constraint (3). This optimum $H(\omega)$ function can be shown to be given by

$$|H(\omega)|^2 = [\gamma\sqrt{\Phi_{mm}(\omega)\Phi_{nn}(\omega)} - \Phi_{nn}(\omega)]/\Phi_{mm}(\omega) \tag{6a}$$

and

$$|H(\omega)|^2 = 0. \tag{6b}$$

Equation (6a) holds for all values of ω which make the right-hand side positive; for all other ω, (6b) must be used. The constant γ is adjusted so that the power constraint (3) is satisfied. The irremovable error which will result from optimum linear coding may be found by substitution of (6a) and (6b) into (5). The irremovable error for the case where no coding but only optimum filtering at the receiver is used may be found by setting $H(\omega)$ equal to a constant in (5). This constant must be chosen to satisfy (3).

Note that only the magnitude of the transfer function of the coding network is involved in (5) and (6). The phase contribution of $H(\omega)$ is not important since the decoding network, $G(\omega)$, as given by (4), provides the necessary phase correction.

III. Numerical Examples

To illustrate the use of the above equations, two examples will be given and discussed. A white noise spectrum shall be assumed where

$$\Phi_{nn}(\omega) = a^2. \tag{7}$$

The message function, $f_m(t)$, will be assumed to have a power-density spectrum given by

$$\Phi_{mm}(\omega) = \frac{\beta/\pi}{\omega^2 + \beta^2}, \tag{8}$$

and the average power, $P_{\text{ave.}}$ of (3) shall be taken as unity. Two cases are to be considered:

[1] N. Wiener, "The Extrapolation, Interpolation, and Smoothing of Stationary Time Series with Engineering Applications," John Wiley and Sons, Inc., New York, N. Y.; 1949.

[2] Y-W Lee, "Application of Statistical Methods to Communication Problems," Technical Report No. 181, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; September, 1950.

[3] J. P. Costas, "Coding with Linear Systems." Technical Report No. 226, Research Laboratory of Electronics, M.I.T., Cambridge Mass.; February 20, 1952.

Case I $a^2=1/10\beta\pi$;
Case II $a^2=1/2\beta\pi$.

For Case I, (6a) holds for all ω out to $\omega=8.45\beta$, and (6b) must be used for all frequencies higher than this cutoff value. The irremovable error without coding is found to be 0.302, and with optimum coding it is reduced to 0.285. The spectra involved are plotted in Fig. 2. Note that the coding network attenuates both the

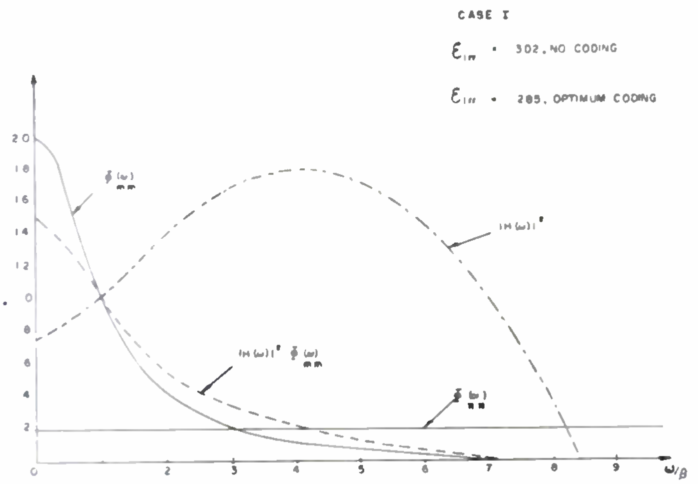

Fig. 2—Power spectra and coding network transfer function; Case I.

frequencies below $\omega=\beta$ and above about $\omega=7\beta$. The power saved in these bands is used to boost the middle range of frequencies.

Case II differs from the above in that the noise level has been raised by a factor of 5. Fig. 3 shows the changes

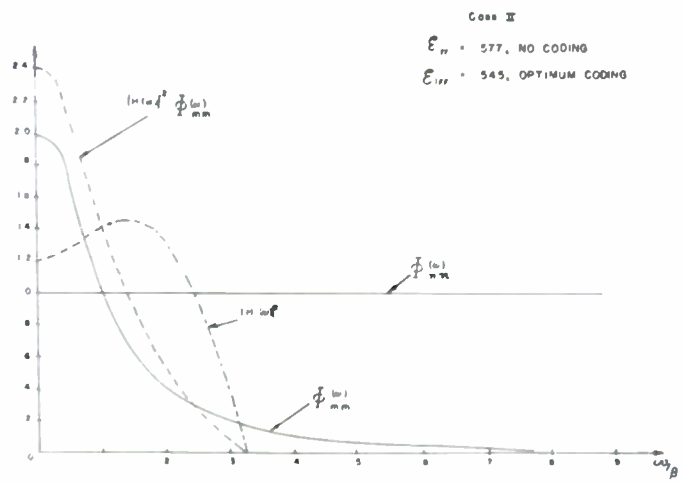

Fig. 3 Power spectra and coding network transfer function; Case II.

in the coding network which are required to combat the severe channel noise. Note that the cutoff frequency has dropped to about $\omega=3.25\beta$ and that a boost is given to all frequencies below about $\omega=2.5\beta$. The improvement in irremovable error is rather small, dropping from 0.577 with no coding to 0.545 with optimum coding. A decibel gain versus log ω plot of the optimum coding networks for the two cases is shown in Fig. 4 for comparison.

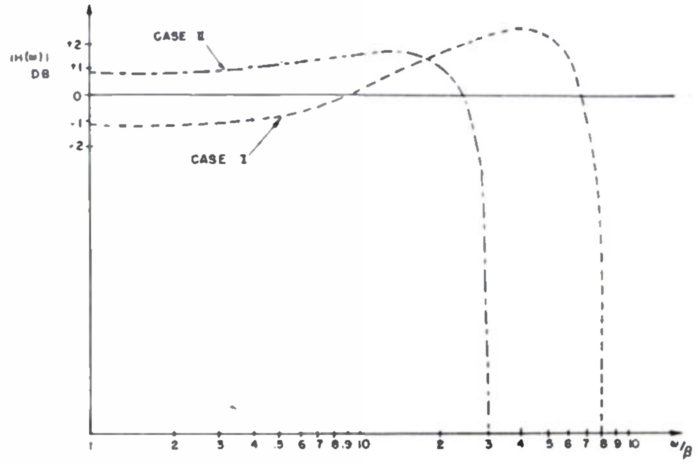

Fig. 4—Frequency plot of $|H(\omega)|$ in decibels for Cases I and II.

IV. Conclusions

The rather moderate improvement in system performance resulting in the above examples is due in part to the use of linear coding and decoding networks. In addition, the particular noise and message spectra assumed do not demonstrate fully the advantages to be gained by optimum linear coding. Less well-behaved spectrum functions would have resulted in a greater improvement in system performance in the case of optimum coding as compared to straight message transmission. In most cases, a coding network which is a fair approximation of the optimum network as given by (6a) and (6b) will usually yield a performance sufficiently close to optimum for all practical purposes. The performance of any given coding network can be checked by a substitution of its system function [after normalization with respect to the power constraint (3)] into (5).

V. Acknowledgment

The author wishes to thank Professor Y-W Lee, Dr. C. A. Stutt, and Mr. C. A. Desoer of the Research Laboratory of Electronics, M.I.T., for their many helpful comments and suggestions.

Notes on Methods of Transmitting the Circular Electric Wave Around Bends*

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Summary—**The tendency for energy to be converted out of the circular electric wave in bent round pipe may be avoided by one of three general approaches: (1) by removing the degeneracy between TE_{01} and TM_{11}, (2) by converting to a normal mode of the bent guide at both ends of the bend, and (3) by utilizing dissipation in the unwanted modes to prevent power transfer to them. All three approaches are discussed. Normal attenuation in round pipe should be effective in moderating straightness requirements. Elliptical guide and special waveguide structures may be used to negotiate intentional bends; bending radii in the range one to 1,000 feet appear acceptable at 50,000 mc for waveguides 3/8-inch to 2 inches in diameter, respectively.**

Introduction

RECENTLY PUBLISHED DATA[1] indicate that transmission losses on the order of 3 db per mile have been observed using the circular electric wave (TE_{01}) in straight round pipes. The question immediately arises, what can be done about bends? The theoretical effects of propagating the circular electric wave in curved round pipes have been studied.[2,3,4,5] This paper summarizes the problem and describes alternate solutions.

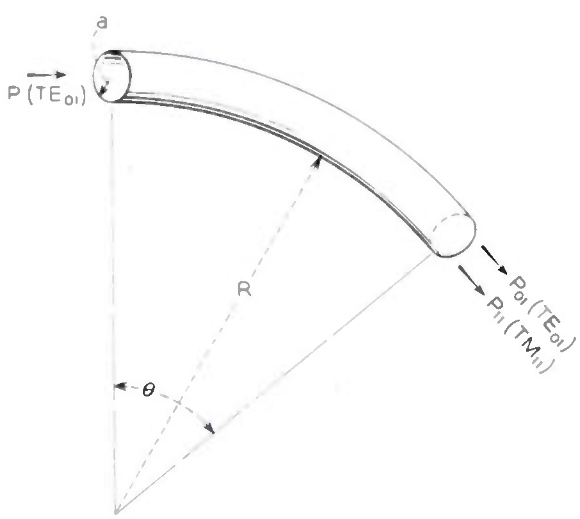

Fig. 1—Round waveguide bent in one plane.

Characteristics of a Bent Round Guide with TE_{01} Excitation

When a pure TE_{01} wave in round guide is propagated into a curved section (Fig. 1), theory which is based on

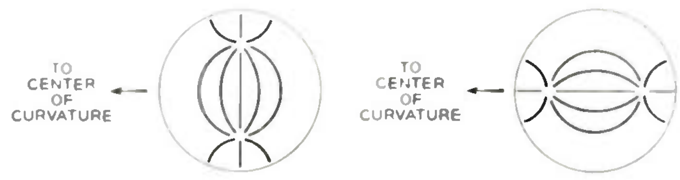

Fig. 2—Electric field directions for the TM_{11}'' and TM_{11}' modes.

no dissipation[2,4] shows that energy is transferred from the TE_{01} mode to the TM_{11} mode. At an angle θ_c,

$$\theta_c = \frac{\pi}{2.32}\,\frac{\lambda_0}{a} \text{ (radians)}, \tag{1}$$

where a is the waveguide radius and λ_0 is the free-space wavelength, the power emerging from the bend is entirely in the TM_{11}'' mode, with orientation as in Fig. 2(a). At other bend angles, θ, the amplitudes of the TE_{01} and TM_{11}'' waves emerging from the bend into straight pipe may be expressed (for input normalized to unity) as follows:

$$TE_{01} \text{ amplitude} = \left|\cos\left(\frac{\pi}{2}\,\frac{\theta}{\theta_c}\right)\right| \tag{2}$$

$$TM_{11}'' \text{ amplitude} = \left|\sin\left(\frac{\pi}{2}\,\frac{\theta}{\theta_c}\right)\right|. \tag{3}$$

This behavior is sketched in Fig. 3. There is a 90° time phase difference between transverse magnetic intensities of TE_{01} and TM_{11}'' components at the bend output.

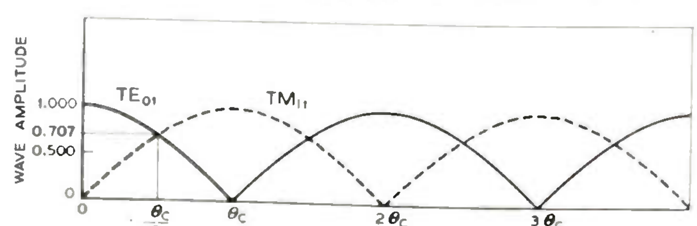

Fig. 3—TE_{01} and TM_{11}'' wave amplitudes versus bend angle.

The behavior of a section of curved line in terms of its input and output waves, as given above, is independent of the bending radius for gradual bends. The reason lies

* Decimal classification: R118.2. Original manuscript received by the Institute, November 27, 1951; revised manuscript received, April 24, 1952.

† Bell Telephone Laboratories, Holmdel Radio Laboratory, Holmdel, N. J.

[1] S. E. Miller and A. C. Beck, "Low loss waveguide transmission," given orally at the March, 1951, IRE National Convention. To be submitted to the PROCEEDINGS.

[2] M. Jouguet, "Effects of the curvature on the propagation of electromagnetic waves in guides of circular cross-section," *Cables and Trans.* (Paris), vol. 1, no. 2, pp. 133–153; July, 1947.

[3] W. J. Albersheim, "Propagation of TE_{01} waves in curved waveguides," *Bell Sys. Tech. Jour.*, vol. 28, no. 1, pp. 1–32; January, 1949.

[4] S. O. Rice, Unpublished Memorandum.

[5] M. Jouguet, "Wave propagation in nearly circular waveguides: transmission-over-bends devices for H_0 waves," *Cables and Trans.* (Paris), vol. 2, no. 4, pp. 257–284; October, 1948.

in the fact that the TE_{01} wave is degenerate with (and therefore has the same phase velocity as) TM_{11}, one of the modes which is coupled to the TE_{01} mode by the bend. This may be stated more quantitatively in terms of the coupled transmission-line equivalent of the bent round waveguide, Fig. 4. When the coupled lines are

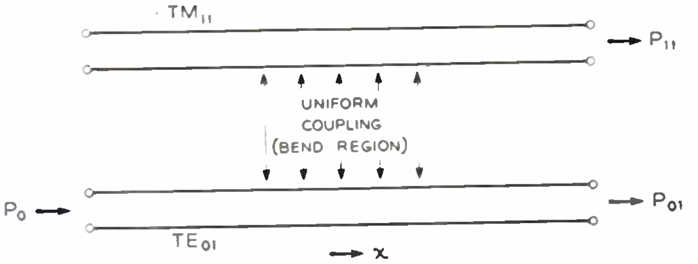

Fig. 4—Coupled transmission-line equivalent of the bent round waveguide.

dissipationless and have the same phase velocity, it may be shown that

$$\frac{P_{11}}{P_0} = \sin^2 (cx) \tag{4}$$

$$\frac{P_{01}}{P_0} = \cos^2 (cx), \tag{5}$$

where c is the coupling per unit length and x is the length co-ordinate. Thus the bend output is uniquely determined by the product of path length and coupling per unit length. The path length varies as the bending radius, and the coupling per unit length varies inversely as the bending radius for gradual bends; hence the product of coupling path length times coupling per unit length is dependent only on the total bend angle.

An alternative description of the wave propagation in the bend region itself may be given in terms of modes which are orthogonal in the bend region. The TE_{01} and TM_{11}'' modes of straight circular pipe are not normal modes of the curved region. When a pure TE_{01} wave is impressed on the input of the curved region, the energy divides equally into two of the normal modes of the curved region. Because familiar mathematical functions do not describe simply the propagation effects in the curved region, the solution for these effects is obtained by perturbation theory and the curved region's normal modes are described in terms of *combinations* of the normal modes of straight circular pipe.

The two normal modes of the curved region which are excited equally by pure TE_{01} bend input are[4]

Mode	*Propagation Constant*	
$(TE_{01} + TM_{11}'')$	$i\dfrac{2\pi}{\lambda_0}\left[(1-\nu^2)^{1/2} + \dfrac{a}{\sqrt{2}\ 3.83R}\right]$	(6)
$(TE_{01} - RM_{11}'')$	$i\dfrac{2\pi}{\lambda_0}\left[(1-\nu^2)^{1/2} - \dfrac{a}{\sqrt{2}\ 3.83R}\right],$	(7)

in which ν is the ratio of free-space wavelength to cutoff wavelength in straight circular pipe. Each of these normal modes contains equal amounts of energy in the TE_{01} and TM_{11}'' field distributions; the time phase difference between the TE_{01} and TM_{11}'' transverse magnetic intensities is either 0° or 180°.

After traveling in curved pipe through an angle θ, the phase difference between the $(TE_{01}+TM_{11}'')$ and $(TE_{01}-TM_{11}'')$ modes is

$$\frac{2\pi}{\lambda_0}\,\frac{\sqrt{2}\,a\theta}{3.83} = \frac{2.32a\theta}{\lambda_0}\ \text{(radians.)} \tag{8}$$

The mode TM_{11}', whose orientation is given in Fig. 2(b), is normal in both straight and bent circular pipe (to the second order of approximation), and has a propagation constant of

$$i\frac{2\pi}{\lambda_0}(1-\nu^2)^{1/2}$$

in both curved and straight pipe.

Discrete Angle-Bend Solution

The most elementary solution to the problem of taking the circular electric wave around bends is to use a total bend angle of $2\theta_c$, $4\theta_c$ or $n\theta_c$, n an even integer.

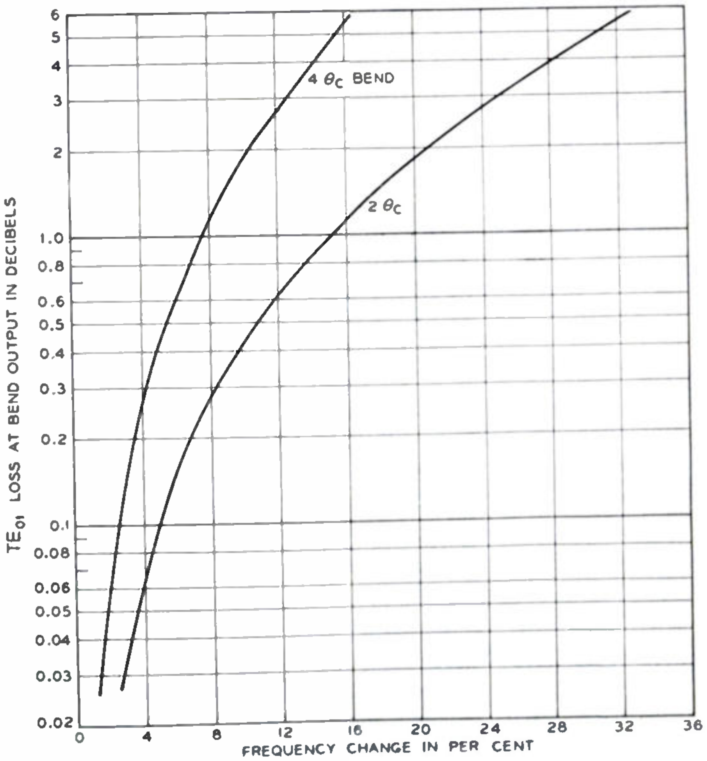

Fig. 5—Theoretical bandwidth of discrete angle bends.

As shown in Fig. 3, at these angles the energy is back in the TE_{01} wave. The limitations of this approach are twofold: (1) Only certain specific bend angles are allowed, certainly a severe restriction in most practical installations, and (2) the frequency bandwidth of the solution is limited since θ_c (1) is a direct function of λ_0. Fig. 5 shows the energy loss for the TE_{01} wave in bends of $2\theta_c$ and $4\theta_c$ as a function of departure from midband frequency. For bandwidths of ±5 per cent or less, the loss may be tolerable, but at ±25 per cent or more bandwidth (which the waveguide itself is certainly capable of handling) excessive bend losses occur.

Solution by Degeneracy Removal

In most types of transmission lines, geometric changes of the type associated with bends, impedance level changes, or even mode-type transformations may be made without undesired effects provided the transition per wavelength along the axis of propagation is not too

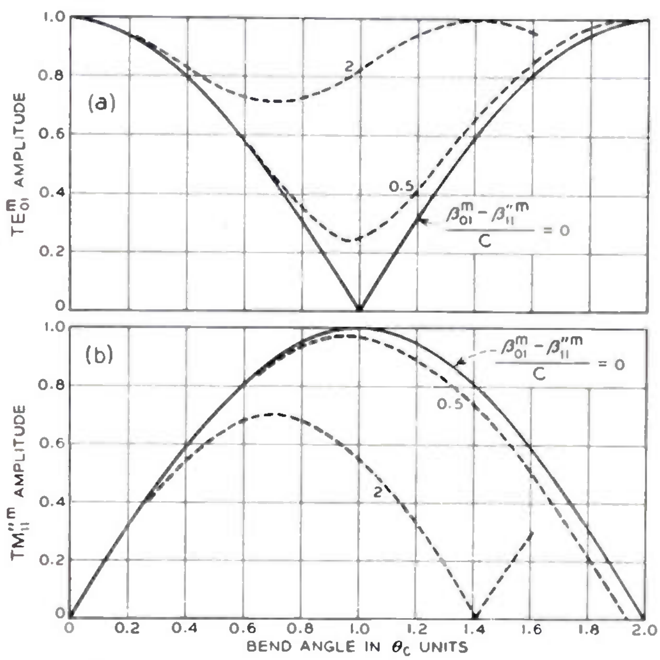

Fig. 6—The effects of degeneracy removal on transmission through bends. (a) TE_{01}^{m} bend output. (b) $TM_{11}''^{m}$ bend output.

large. This is the familiar tapered transmission-line approach. Why, then, is it that a very gradual bend is as bad as a more rapid bend in causing TE_{01} loss? The answer lies in the fact that the bend couples TE_{01} to a mode TM_{11} which has the *identical* phase velocity. If we remove this degeneracy, we create a situation in which a gradual bend causes less loss than a more rapid bend because the components of energy transferred from TE_{01} to TM_{11}'' at different locations along the axis of propagation no longer add in phase.

The change in transmission effects, which result when the degeneracy is removed, can be analyzed on a coupled transmission-line basis as sketched in Fig. 4. We consider only the TE_{01} to TM_{11} coupling in a bend and assume the coupling coefficient between these modes is not altered by the structural modification which removes the degeneracy. This value of coupling coefficient is

$$c = \frac{1.16a}{R\lambda_0}, \tag{9}$$

which, when used in (4) and (5), will give the same amplitude for TE_{01} and TM_{11}'' as (2) and (3) for the output of curved circular pipe.

In the altered guide in which the degeneracy has been removed, the propagation constant of at least one of the modes must be other than the circular pipe value, and the modes should be given new designations. We shall designate the modified TE_{01} wave as TE_{01}^{m} and the modified TM_{11}'' wave as $TM_{11}''^{m}$.

Considering the dissipationless case only for the present, it is found that an index of transmission performance in the curved modified line is

$$\frac{\beta_{01}^{m} - \beta_{11}''^{m}}{c} \tag{10}$$

where β_{01}^{m} and $\beta_{11}''^{m}$ are the phase constants of the TE_{01}^{m} and $TM_{11}''^{m}$ waves, respectively, in straight modified line. The amplitudes of TE_{01}^{m} and $TM_{11}''^{m}$ at the end of a curved section of modified line are plotted in Fig. 6 (a) and (b) for several values of the parameter (10). The abscissa is the bend angle, expressed as multiples of the extinction angle θ_c for circular guide of the same radius.

Fig. 6 shows that the energy exchange between modes is still periodic as a function of bend angle when the degeneracy is broken, but the maximum energy lost from TE_{01} is reduced as the ratio (10) is made large.

Fig. 7 shows the amplitudes of $TM_{11}''^{m}$ and TE_{01}^{m} at the bend angle where maximum energy transfer has taken place, as a function of the ratio (10).

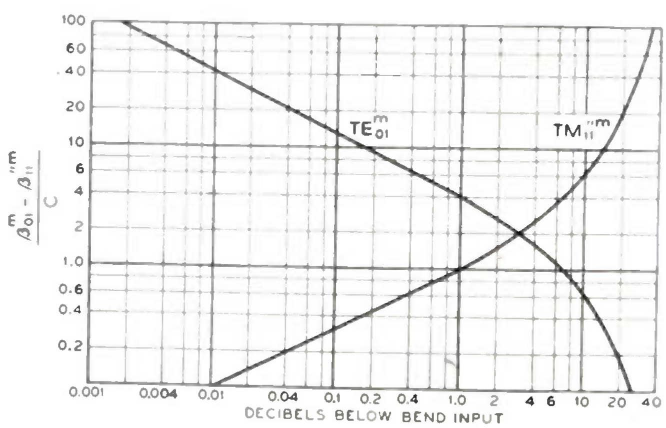

Fig. 7—Bend-output wave amplitudes at the bend angle of maximum conversion versus $\beta_{01}^{m}-\beta_{11}''^{m}/c$.

In order to relate the ratio (10) to physical quantities, it is convenient to rewrite (10) as

$$\frac{2\pi R}{1.16a}\left(\sqrt{1-(\nu_{01}{}^{m})^2} - \sqrt{1-(\nu_{11}''^{m})^2}\right), \tag{11}$$

which gives the ordinate for Fig. 7 in terms of the ν's, which are the free space to cutoff wavelength ratios for the TE_{01}^{m} and $TM_{11}''^{m}$ modes in straight modified pipe.

When the TE_{01} and TM_{11} modes are far from cutoff, as is the case in low-loss waveguides, $\nu^2 \ll 1$ and (11) becomes

$$\frac{\pi R}{1.16a}\left((\nu_{11}''^{m})^2 - (\nu_{01}{}^{m})^2\right). \tag{12}$$

Consider an example to place order of magnitude. For a 50,000-mc wave in 2-inch diameter pipe, a 0.5-per cent difference in cutoff wavelength between TE_{01} and TM_{11} would yield a ratio (12) of 20 for a bending radius of 2,960 feet, corresponding to less than 0.05-db maximum loss to TE_{01} (at worst bend angle); a pipe length of 4,650 feet would be required to negotiate a 90° bend.

For guides far from cutoff, as in the case chosen, the allowable bending radius (1) varies inversely with the $TE_{01}-TM_{11}$ cutoff wavelength difference, (2) varies inversely as the ratio $\lambda_0^2/_a^3$, and (3) decreases as the acceptable TE_{01} bend loss increases (see Fig. 7, noting $(\beta_{01}{}^m-\beta_{11}{}''^m)/c$ varies directly as the bending radius R).

In principle, the problem of transmitting the circular electric wave around bends can be solved by breaking the $TE_{01}-TM_{11}$ degeneracy; the question is, how much increase in $TE_{01}{}^m$ versus TE_{01} heat loss will occur when the waveguide has been altered to remove the degeneracy? To answer this we must consider a specific structure.

Elliptic Waveguide Solution

One way of removing the $TE_{01}-TM_{11}$ degeneracy is to deform the walls of the circular waveguide. An elliptic guide is an example of such a deformation, on which there is available some information in the literature. We are interested here in small amounts of eccentricity, and need to know cutoff wavelengths and attenuation constant very accurately. The work of Chu[6] did not provide the desired information, and hence the computations reported herein are based on new derivations made by Gray of the Bell Telephone Laboratories.

For the $TE_{01}{}^m$ mode in a slightly elliptic guide, Gray determines that the cutoff constant is

$$k_{01}{}^m = k\left(1+\frac{1}{4}e^2+\frac{k^2+10}{64}e^4\right)\cdots, \qquad (13)$$

where $k=3.8317$ and e is the eccentricity. The TM_{11}'' mode of round guide may divide into two modes in elliptic guide, depending on the location of the major and minor axes of the cross section relative to the bending radius. For the "even"[6] wave Gray finds

$$k_{11}{}''^m = k\left(1+\frac{1}{8}e^2+\frac{k^2+14}{256}e^4\right)\cdots \qquad (14)$$

and for the "odd" wave

$$k_{11}{}''^m = k\left(1+\frac{3}{8}e^2+\frac{k^2+62}{256}e^4\right)\cdots. \qquad (15)$$

The $TE_{01}{}^m-TM_{11}{}''^m$ cutoff wavelength difference is plotted in Fig. 8, based on (13) and (14). For an eccentricity of about 0.3 there is a 1 per cent cutoff wavelength difference. Fig. 9 shows that an eccentricity of 0.3 corresponds to an approximately 5 per cent difference between the major and minor axes of the ellipse.

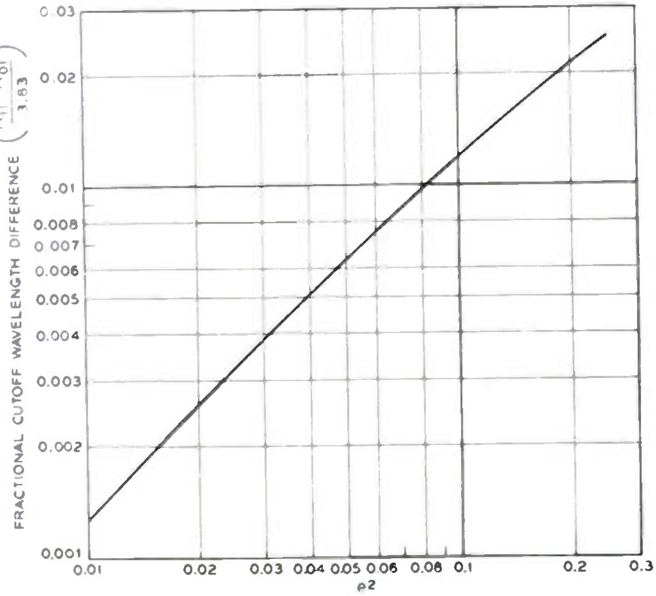

Fig. 8—The effect of eccentricity on the difference between the cutoff wavelengths of TE_{01} and TM_{11} in elliptic waveguide.

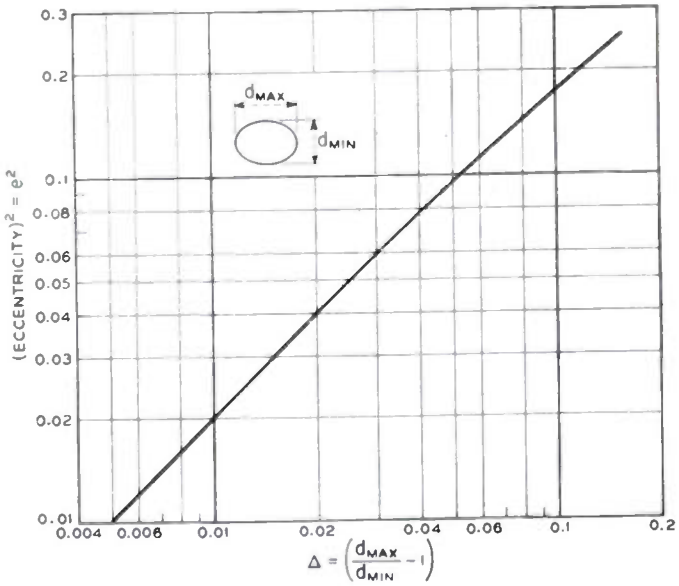

Fig. 9—Elliptic waveguide eccentricity versus the major-to-minor axis ratio.

It is important that no new degeneracy with TE_{01} be created by going to the elliptical cross section, and Gray finds that is the case.

The heat loss for the $TE_{01}{}^m$ wave is given by Gray as[7]

$$\alpha_{01}{}^m = \frac{R}{\eta a}\left(\frac{\nu^2}{\sqrt{1-\nu^2}}\left[\left(1+\frac{1}{4}e^2\right.\right.\right.$$

$$\left.\left.+\frac{k^2e^4}{16J_0{}^2(k)}\log\frac{1+\sqrt{1-e^2}}{e}\right]+\frac{k^2e^4}{32}\sqrt{1-\nu^2}\right). \qquad (16)$$

[6] Lan Jen Chu, "Electromagnetic waves in elliptic hollow pipes of metal," *Jour. Appl. Phys.*, vol. 9, pp. 583; September, 1938.

[7] Terminology is that of S. A. Schelkunoff, "Electromagnetic waves," D. Van Nostrand Co., New York, N. Y., 1943.

The relation between eccentricity and $TE_{01}{}^m$ loss in straight elliptic guide is plotted in Fig. 10. An eccentricity of 0.3 results in 25 to 35 per cent more heat loss than in circular guide, with little dependence on proximity to cutoff.

The allowable bending radius may now be calculated for a preselected maximum bend loss as a function of increased heat loss due to eccentricity. The steps are as follows: The ratio $(\beta_{01}{}^m - \beta_{11}{}''^m)/c$ is obtained from Fig. 7 for a preselected $TE_{01}{}^m$ loss due to bending at the angle of maximum conversion; for $(\beta_{01}{}^m - \beta_{11}{}''^m)/c$ equal to 20, this bend loss is 0.043 db. Then for the selected condition, the ratio (11) is determined, from which the bending radius may be calculated as a function of eccentricity. The increased heat loss due to eccentricity is also known ((16) and Fig. 10), so the bending radius

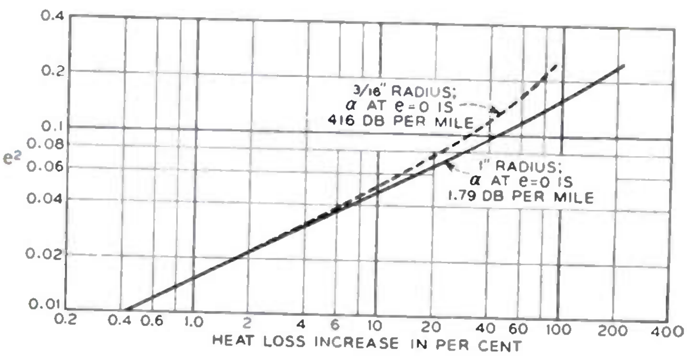

Fig. 10—The increase in TE_{01} heat loss due to waveguide ellipticity for 1 inch or $\frac{3}{16}$ inch guide radius and a frequency of 50,000 mc.

may be plotted directly as a function of increased heat loss due to eccentricity. The results are given in Figs. 11 and 12 for guides of 2-inch and $\frac{3}{8}$-inch diameter operated at 50,000 mc.

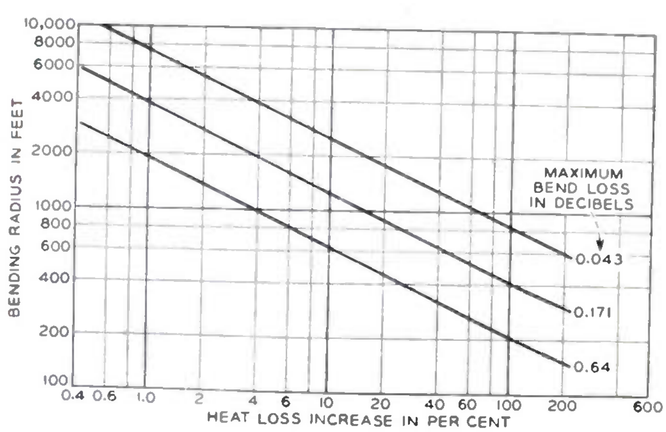

Fig. 11—The allowable bending radius versus the associated increase in TE_{01} heat loss in elliptic guide, with maximum bend loss as a parameter, for a 2-inch diameter guide at 50,000 mc.

In the 2-inch diameter low-loss guide, bending radii on the order of 250 to 1,000 feet may be tolerated, depending on maximum bend loss accepted, at a penalty of a 50-per cent increase in heat loss above the value for a circular cross section. In the $\frac{3}{8}$-inch diameter, which might be used in short runs, a bending radius as low as one foot may be tolerable.

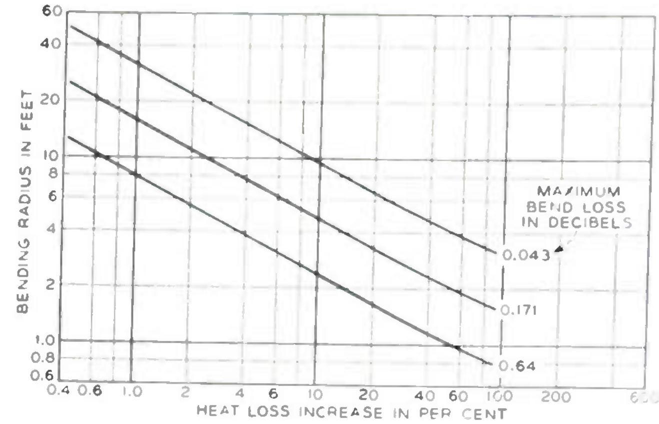

Fig. 12—The allowable bending radius versus the associated increase in TE_{01} heat loss in elliptic guide, with maximum bend loss as a parameter, for a $\frac{3}{8}$-inch diameter guide at 50,000 mc.

Fig. 13 shows allowable bending radius versus waveguide diameter for 50,000-mc operation and for an eccentricity of 0.3, with maximum bend loss as parameter.

Dissipation in the guide walls, which will be discussed in a subsequent paragraph, alters the elliptic guide bend performance for very large bending radii, but does not detract from its usefulness in avoiding bend losses.

The above analysis shows that elliptic waveguides present one solution to the bend problem. To avoid losses due to accidental deviations from straightness, the long line may be given some ellipticity. If more rapid bends must be negotiated, a guide of greater ellipticity might be employed in the bend region only, with suitable tapers in straight sections adjacent to the bend.

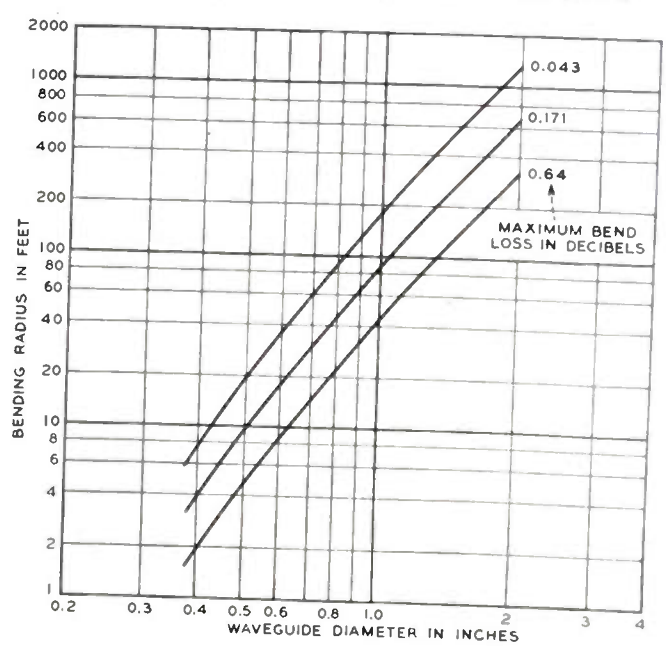

Fig. 13—The allowable bending radius versus guide diameter, with maximum bend loss as a parameter, for elliptic guides of eccentricity $e=0.3$ and a frequency of 50,000 mc.

Alternate Methods of Removing the Degeneracy

In general, any alteration in the circular guide which affects the TE_{01} and TM_{11} waves differently will remove the degeneracy and thus become a potential solution to the bend problem. One such alteration is to put circular corrugations in the wall transverse to the axis of propagation, forming a structure similar to the familiar sylphon bellows. This corrugated guide has circular symmetry and a cross section as sketched in Fig. 14.

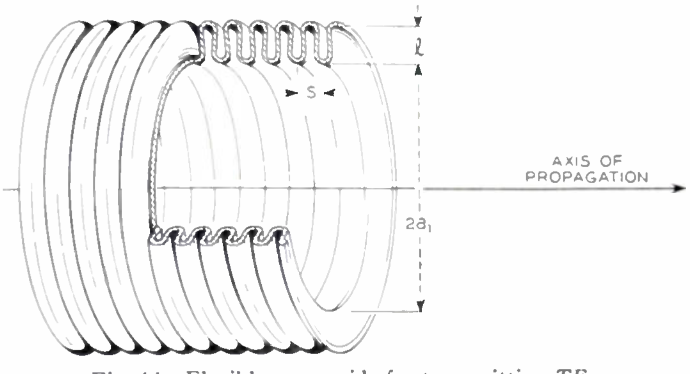

Fig. 14—Flexible waveguide for transmitting TE_{01} around bends (due to King).

The remarkable TE_{01} transmission characteristics of this structure were first discovered by King of the Bell Telephone Laboratories. He finds that the circular electric wave undergoes bend losses of 0.1 db or less for bends as large as the critical angle. The spacing "S" (Fig. 14) is made a small fraction of a wavelength so that TE_{01} propagates very nearly as though in a solid pipe of radius "a," whereas the TM_{11} wave experiences additional loading due to the radial grooves of length "l." Thus the degeneracy is removed. The mechanical flexibility of this structure, combined with its ability to transmit TE_{01} in bends, make it very attractive in certain applications.

Normal Mode Solution

Another approach to the bend problem is to utilize one of the normal modes of curved round guide. There are many such modes, but the ones most closely related to TE_{01} are TM_{11}' (Fig. 2(b)) and $TE_{01} \pm TM_{11}''$. The

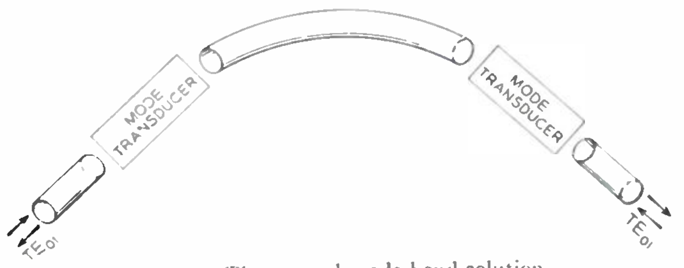

Fig. 15—The normal-mode bend solution.

general layout would be as sketched in Fig. 15. A mode transducer would be placed at both ends of the curved region to transform TE_{01} to one of the curved region's normal modes. Thus bends of arbitrary length could be negotiated. The question arises whether it is possible to perform the mode transformation.

The TM_{11}'' mode may be formed from the TE_{01} wave using a section of bent round pipe of total angle θ_c as given by (1). The TM_{11}' mode may, in turn, be formed by rotating the polarization of the TM_{11}'' wave in a section of straight elliptic guide whose major and minor axes are inclined at 45° to the initial polarization of the TM_{11}'' wave.[5] A combination of a θ_c-angle bend and a suitable length of elliptic guide therefore constitute a TE_{01} to TM_{11}' mode transformer for use in the layout of Fig. 15. All elements of this arrangement lie in one plane.

As an alternative, the section of elliptic guide may be eliminated from the mode transducer by making the θ_c-angle bend in a plane perpendicular to the plane of the arbitrary bend, thereby presenting the arbitrary bend with an input wave of TM_{11}' which is a normal mode of the curved region. This reduces the mode transducer to a simple θ_c-angle bend, but puts the elements of the bend (the mode transducers and arbitrary bend) in a three-dimensional arrangement between the two straight guides which it is the objective to join.[5] In practice this method would be awkward.

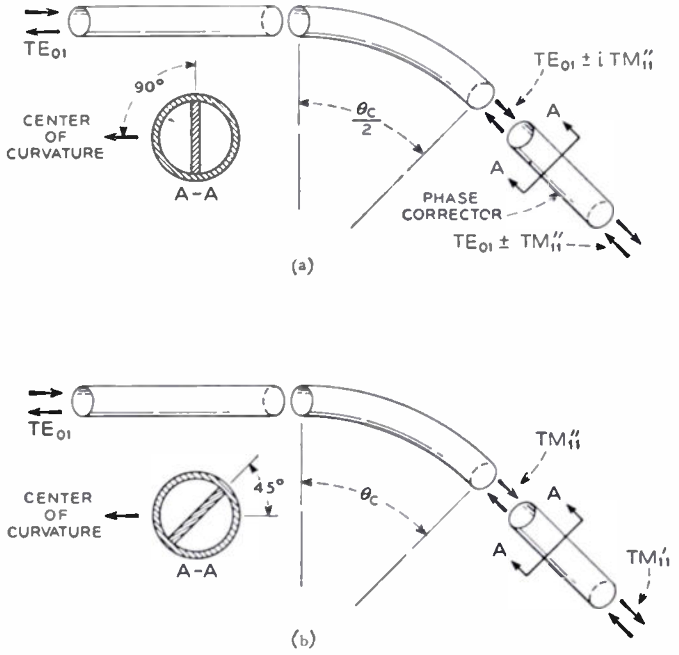

Fig. 16—Mode transducers applicable in the configuration of Fig. 15. (a) TE_{01} to $(TE_{01} \pm TM_{11}'')$ mode transducer. (b) TE_{01} to TM_{11}' mode transducer.

Fig. 16 shows two transducers from TE_{01} to normal modes of the bend region wherein all elements are in a single plane. In Fig. 16(b), the first transformation is from TE_{01} to TM_{11}'' in a θ_c-angle bend, followed by a rotation of the polarization by means of a longitudinal diametral dielectric sheet to convert TM_{11}'' to TM_{11}'. The function of the dielectric sheet is the same as that of the section of elliptic pipe mentioned above, but may be more easily controlled in practice.

Fig. 16(a) shows a TE_{01} to $(TE_{01} \pm TM_{11}'')$ transducer. The $\theta_c/2$ bend divides the input TE_{01} wave into equal powers in TE_{01} and TM_{11}'' waves; but this is not the normal mode of the bend region because there is a 90° time phase difference between the TE_{01} and TM_{11}'' transverse magnetic intensities, instead of the required 0° or 180°. This is indicated by the designation $TE_{01} \pm iTM_{11}''$ on the sketch. However, by introducing a 90°

delay difference between the TE_{01} and TM_{11}'' components in a section of straight pipe, the $(TE_{01} \pm TM_{11}'')$ wave is created. This 90° phase correction is accomplished in the Fig. 16(a) layout by a longitudinal diametral dielectric sheet, which should be many wavelengths long to avoid mode conversion effects, and should be of such dielectric constant and thickness to avoid reflection at the ends.

A very attractive TE_{01} to TM_{11}' mode transducer has been designed by Morgan and evaluated experimentally

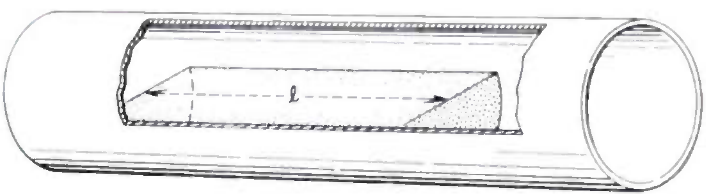

Fig. 17—A TE_{01} to TM_{11} mode transducer (due to Morgan).

by King, both of the Bell Telephone Laboratories. The structure, sketched in Fig. 17, consists of a half-circular cylinder of low-loss dielectric material whose dielectric constant relative to that for free space (i.e., ϵ_r) is very nearly unity. For a length "l" of this half-circular cylinder such that

$$l = \frac{2.073\lambda_0}{\epsilon_r - 1} \cdots, \tag{17}$$

where λ_0 is the free-space wavelength, Morgan determines that a TE_{01} incident wave is completely converted to TM_{11}. The orientation may obviously be made so that the transducer output is TM_{11}' at the start of the arbitrary bend. Morgan has also found it possible to make the transformation TE_{01} to $TE_{01} \pm TM_{11}''$ in a structure similar to Fig. 17 under certain conditions.

The structure of Fig. 17 may be employed in place of the $\theta_c/2$-angle bend in Fig. 16(a), the length required being one-half that given by (17).

One disadvantage of all of the "normal-mode" solutions described above is that the mode conversions necessary at the ends of the bend are frequency sensitive. Bandwidths on the order of those shown in Fig. 5 are about what might be expected of this general approach.

The Dissipation Solution

The TE_{01} to TM_{11}'' mode conversion which takes place in a curved round guide has the form sketched in Fig. 3 only in the limit of zero dissipation. We seek to describe here the effect of dissipation.

One way of showing the effects of dissipation is to consider how ideal mode filters alter bend performance. Suppose in the illustration of Fig. 18 the bend angle β were equal to θ_c and no mode filters were used. Then, as described in the first section of this paper, the bend output in TE_{01} would be zero—complete loss of signal. Suppose we now insert at the half angle $\beta/2$ a mode filter with no loss to TE_{01} and complete absorption of TM_{11}

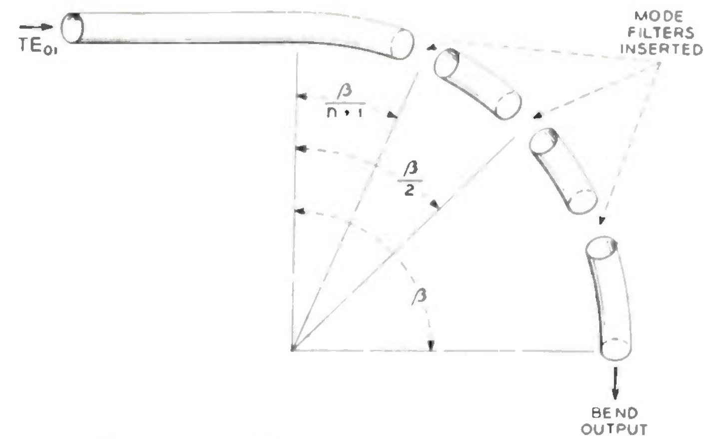

Fig. 18—Sketch for illustrating the effects of mode filters in a bend.

(such a filter has been approximated in practice).[1] Then the TE_{01} output of the first $\beta/2$-angle bend plus mode filter would be (2)

$$\cos\left(\frac{1}{2}\,\frac{\pi}{2}\right) = 0.707 \tag{18}$$

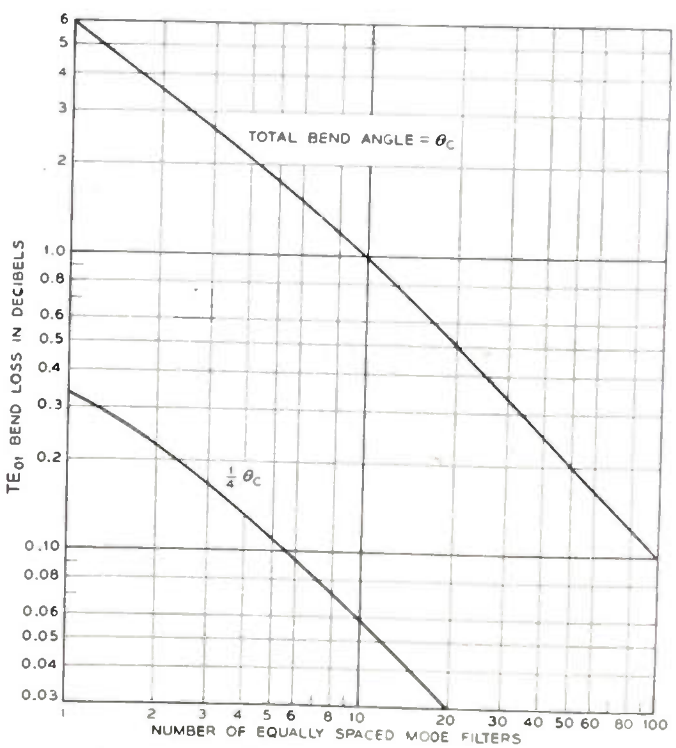

Fig. 19—The effect on TE_{01} bend loss due to mode filters in the configuration of Fig. 18.

for unit TE_{01} wave as the bend input, and there would be no TM_{11} output from the mode filter. At the output of the second half of the bend, the TE_{01} amplitude would be

$$\cos^2\left(\frac{\pi}{4}\right) = 0.5. \tag{19}$$

Thus by adding dissipation to TM_{11} only, the θ_c-angle bend loss has been decreased from infinite loss to 6-db loss. For the condition of "n" TM_{11} mode absorbers inserted at equal intervals along the θ_c-angle bend, it may be shown that the TE_{01} bend output is

$$\cos^{n+1}\left(\frac{\pi}{(n+1)2}\right). \tag{20}$$

This function has been plotted in Fig. 19. As the number of TM_{11} mode absorbers is increased without limit, the TE_{01} bend loss approaches zero. For an arbitrary bend of angle "β" and "n" equally spaced mode filters, the amplitude of the TE_{01} bend output is

$$\cos^{n+1}\left(\frac{\beta\pi}{(n+1)\theta_c 2}\right). \tag{21}$$

Fig. 19 also shows this relation for the bend angle $\beta=\theta_c/4$.

There are structures which make use of dissipation to transmit the circular electric waves around bends with low loss, a few of which are sketched in Fig. 20. The characteristic which these structures have in common is very high transmission loss for the modes to which energy tends to be transferred and very low transmission loss to the circular electric waves. The ability of the structure of Fig. 20(c) to transmit TE_{01} around bends was first observed by Fox at Holmdel. King has also shown experimentally that structures of the form of Fig. 20(a) may be used to avoid bend losses.

Having found that dissipation may be used to avoid TE_{01} bend losses in special structures, we may inquire whether or not dissipation in ordinary circular pipe will have an effect in reducing TE_{01} bend losses. The attenuation constant for TM_{11} in 2-inch diameter pipe at 50,000 mc is nearly 50 times the attenuation constant for TE_{01} in the same pipe; thus the structure inherently has the general property needed to avoid bend loss by means of dissipation.

The effects of dissipation in smooth lines have been determined using the coupled transmission-line analogy of Fig. 4. It has been determined that the amplitude of the TE_{01} output of a bend is[8]

$$E_{TE_{01}} = \left[\frac{1}{2} - \frac{(\gamma_1-\gamma_2)}{2\sqrt{(\gamma_1-\gamma_2)^2-4c^2}}\right]\epsilon^{r_1 x} + \left[\frac{1}{2} + \frac{(\gamma_1-\gamma_2)}{2\sqrt{(\gamma_1-\gamma_2)^2-4c^2}}\right]\epsilon^{r_2 x} \tag{22}$$

where

c = coupling per unit length along bend
x = length of pipe in the bend
$\gamma_1=\alpha_{01}+i\beta_{01}$
$= TE_{01}$ propagation constant
$\gamma_2=\alpha_{11}+i\beta_{11}$
$= TM_{11}$ propagation constant
$r_1=-\frac{1}{2}(2ic+\gamma_1+\gamma_2)+\frac{1}{2}\sqrt{(\gamma_1-\gamma_2)^2-4c_2}$
$r_2=-\frac{1}{2}(2ic+\gamma_1+\gamma_2)-\frac{1}{2}\sqrt{(\gamma_1-\gamma_2)^2-4c^2}$.

[8] The author has in preparation another paper in which the coupled transmission-line theory will be given in more detail, in a form applicable to many other problems as well as this one.

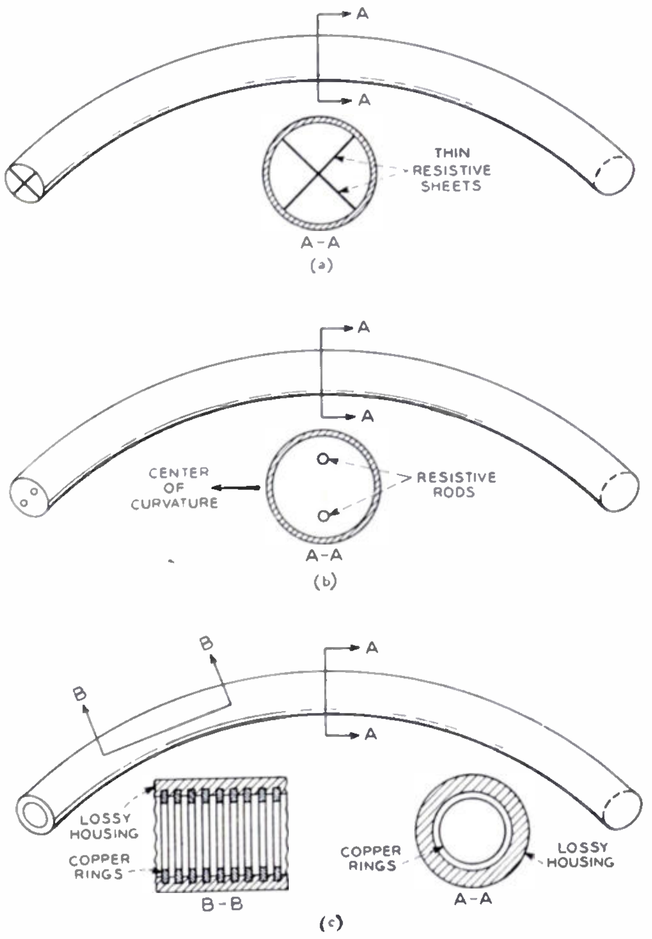

Fig. 20—Structures for negotiating bends using dissipation to prevent TE_{01} loss. (a) Longitudinal diametral resistive sheets. (b) Longitudinal resistive rods. (c) Closely spaced copper rings in a Lossy housing.

The TE_{01} and TM_{11} propagation constants γ_1 and γ_2 are for straight circular guide. It is assumed that the coupling "c" is unchanged by the presence of dissipation and is given by (9). Because of the degeneracy, the imaginary part of the propagation constant is the same for TM_{11} and TE_{01}, and

$$(\gamma_1-\gamma_2) = (\alpha_{01}-\alpha_{11}). \tag{23}$$

When $(\alpha_{01}-\alpha_{11})^2$ is very large compared to $4c^2$ so that

$$\sqrt{(\gamma_1-\gamma_2)^2-4c^2} \simeq |\alpha_{01}-\alpha_{11}|, \tag{24}$$

the TE_{01} amplitude given by (22) approaches

$$E_{TE_{01}} \simeq \epsilon^{r_1 x}$$
$$r_1 \simeq -\alpha_{01} - i\left(c+\frac{\beta_{01}+\beta_{11}}{2}\right). \tag{25}$$

Thus, the principal effect of a bend is to modify the imaginary portion of the propagation constant, provided that the ratio

$$\left|\frac{\alpha_{01}-\alpha_{11}}{c}\right| \tag{26}$$

is suitably large. For any given value of $(\alpha_{01}-\alpha_{11})$, (26) may be made as large as desired by making the value of c small, i.e., the bending radius large. (See (9).)

Fig. 21 shows the TE_{01} bend loss versus bend angle with the ratio (26) as a parameter; the bend loss is considered to be the actual TE_{01} bend output compared to

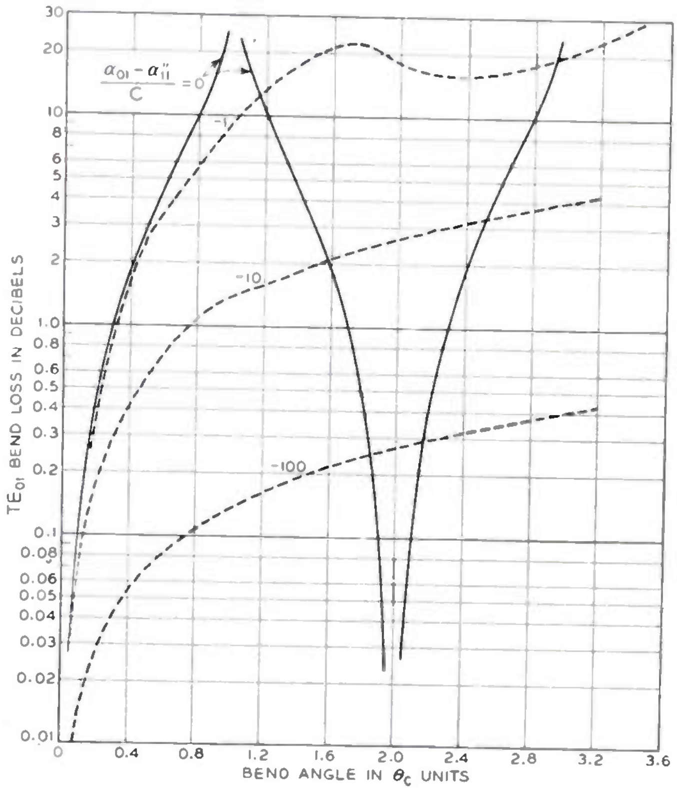

Fig. 21—TE_{01} bend loss versus bend angle, with $\alpha_{01}-\alpha_{11}''$ as a parameter.

the TE_{01} output from the same length of straight pipe. The curve for $(\alpha_{01}-\alpha_{11})/c$ equal to zero is the dissipationless case given by previous authors. For $(\alpha_{01}-\alpha_{11})/c$ greater than one, the bend loss becomes appreciably less than predicted by theory based on no dissipation; for $(\alpha_{01}-\alpha_{11})/c=100$, the loss in a critical angle bend is under 0.15 db instead of the infinite loss predicted by dissipationless theory. In practice, smaller ratios (26) are likely to be encountered, however. In Figs. 22 and 23 the bend loss is plotted versus bending radius for some waveguides of interest at a frequency of 50,000 mc. Note that the total bend angles for which these curves apply is different, depending on the pipe diameter. Reducing the guide size reduces the bend loss at a rapid rate, because (1) the bend loss is less for guides nearer cutoff at the same fraction of a critical angle bend (as shown in Figs. 22 and 23) and (2) the critical angle becomes larger as the guide size is reduced.

For a critical angle (18.3°) bend in a 2-inch diameter guide, a bending radius of 35,000 feet is required to reduce the bend loss to 1 db. In a 1-inch diameter guide, this 18° bend with a bending radius of 35,000 feet would produce under 0.15-db bend loss (Fig. 22).

It may not be too difficult to maintain unintentional deviations from straightness within 5 feet in 500 feet, in the form of an arc of a circle, and this condition corresponds to a bending radius on the order of 50,000 feet. This corresponds to a value of $(\alpha_{01}-\alpha_{11})/c$ of almost 20 in a 2-inch diameter pipe at 50,000 mc. It would appear,

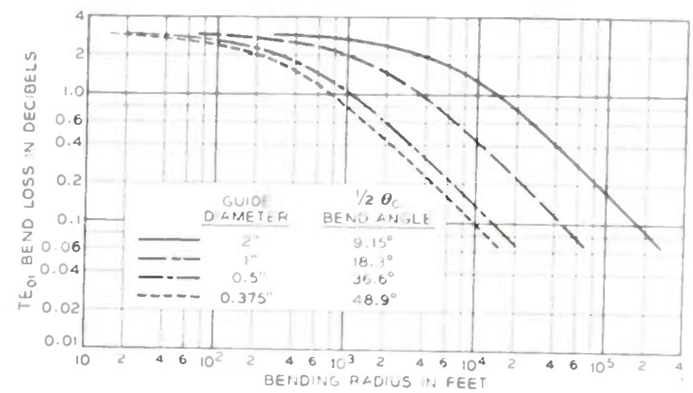

Fig. 22—TE_{01} bend loss versus bending radius for a $\frac{1}{2}$ θ_c bend in several round guides operated at 50,000 mc. The reduction in bend loss for large bending radii is due to dissipation in the walls.

therefore, that the losses due to deviations from straightness will be significantly reduced (compared to the predictions of dissipationless theory) because of inherent loss in the circular guide walls.

Intentional bends might be made in a radius on the order of 5,000 feet, which approximates the curves used on a high-speed railroad. This corresponds to a value of $(\alpha_{01}-\alpha_{11})/c$ of about 2 in a 2-inch diameter pipe, and the bend losses become undesirably large. One alternative is to go to a 1-inch diameter pipe, thereby raising θ_c

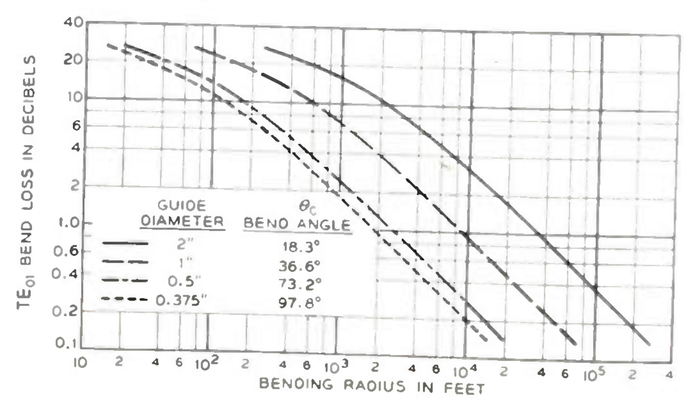

Fig. 23—TE_{01} bend loss versus bending radius for a θ_c-angle bend in several round guides operated at 50,000 mc.

from 18.3° to 36.6° and also raising $(\alpha_{01}-\alpha_{11})/c$ from about 2 to 7.2. This would reduce the bend loss to around one db for a 20° bend. Another alternative involves the use of mode filters in the manner described in connection with Figs. 18 and 19. If an ideal mode filter were added every 100 feet along a bend of radius 5,000 feet, the bend would be divided into segments $\theta_c/6$ each (for a 2-inch diameter pipe), resulting in a bend loss of about 0.04 db per 1.15° of bend. The ultimate attractiveness of this approach, compared to the "normal-mode" or "degeneracy-removal" approach, depends on how closely an ideal mode filter can be approached in practice; TE_{01} loss in the mode filter will, of course, limit the number of filters which can be added profitably.

Conclusion

The tendency for energy to be converted out of the circular electric wave in bent round pipe may be avoided by one of three general approaches: (1) by removing the $TE_{01}-TM_{11}$ degeneracy, (2) by converting to a normal mode of the bent guide at both ends of the bend, and (3) by utilizing dissipation in the unwanted modes to prevent power transfer to them. Methods (1) and (3) may be used singly or in combination for avoiding extreme straightness requirements on normally straight sections of line and for negotiating intentional bends.

Normal dissipation in solid round copper guide should be effective in moderating straightness requirements, but does not appear to make possible an attractive bending radius for intentional bends. Other waveguide structures, such as those of Fig. 20, may enable the dissipation approach to solve the intentional bend problem.

Removing the degeneracy by making the pipe elliptical increases the normal heat loss for the modified TE_{01} wave, and the tolerable bending radius is a compromise with this heat-loss increase. For a heat-loss increase of about 50 per cent, the tolerable bending radius is on the order of 300 to 1,000 feet (depending on the bend loss tolerated) for a 2-inch diameter guide operated at 50,000 mc. A transition from a circular guide in straight runs to a slightly elliptic guide for intentional bends is one way of avoiding the increased heat loss of the elliptic guide for the majority of line mileage.

Removal of the degeneracy and the dissipation approach offer no serious limitation of bandwidth.

A number of methods of converting to a normal mode of the bend region appear to be available, all of which appear to be limited to bandwidths on the order of 10 per cent.

An Improved Theory of the Receiving Antenna*

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Summary—**The theory of the center-loaded receiving antenna is improved by introducing the expansion parameter of King and Middleton, and generalized to take account of a load consisting of a two-wire line with finite spacing. First-order formulas for the distribution of current are obtained, together with approximate second-order formulas for the complex effective length of the antenna. Theoretical results are compared with experiment.**

Introduction

The integral equation for the cylindrical, center-loaded receiving antenna in a linearly polarized electric field of arbitrary orientation was formulated by Hallén[1] and reformulated and extended to the elliptically polarized field by Harrison and King.[2] In these analyses the integral equation is solved by iteration using an expansion parameter introduced by L. V. King[3] and independently by Hallén, viz., $\Omega=2\ln(2h/a)$ where h is the half-length and a the radius of the antenna. The numerical evaluation of the distribution of current and of the complex effective length by King and Harrison is based on a first-order formula in which terms of the order $1/\Omega$ are retained and the load at the center of the antenna is lumped.

Since the definition of the effective length involves the simultaneous definition of impedance in a manner to make this identically the impedance of the same antenna when center-driven, the analysis of the receiving antenna depends on the determination of the impedance of the driven antenna. It has been shown consistently in experimental investigations by D. D. King, Conley, Tomiyasu, and especially by Hartig and by Hartig, King, Morita, and Wilson[4] that the measured impedance of a cylindrical antenna is in good agreement with the Hallén theory provided this is carried out at

* Decimal classification: R120. Original manuscript received by the Institute, November 28, 1951; revised manuscript received April 21, 1952.

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[1] E. Hallén, "Theoretical investigations into transmitting and receiving antennae," *Nova Acta (Uppsala)* series 4, 11; pp. 1-44; November, 1938.

[2] C. W. Harrison, Jr., and R. King, "The receiving antenna in a plane polarized field of arbitrary orientation," Proc. I.R.E., vol. 32, p. 35; January, 1944.

[3] L. V. King, "On the radiation field of a perfectly conducting base-insulated cylindrical antenna over a perfectly conducting plane earth," *Phil. Trans.* series A, vol. 236, pp. 381–422; November, 1937.

[4] D. D. King, "The measured impedance of cylindrical dipoles," *Jour. Appl. Phys.*, vol. 17, p. 844; October, 1946. Cruft Laboratory Technical Report No. 2, Harvard University, Cambridge, Mass.; October, 1946.

P. Conley, "Antennas and open-wire lines," pt. III, *ibid.*, vol. 20, pp. 1022; November, 1949. Cruft Laboratory Technical Report No. 35; March, 1948.

K. Tomiyasu, "Antennas and open-wire lines," pt. II, *ibid.*, vol. 20, p. 892; October, 1949. Cruft Laboratory Technical Report No. 49; June, 1948.

E. O. Hartig, "Circular Apertures and their Effects on Half-Dipole Impedances," Cruft Laboratory Technical Report No. 107; June, 1950.

E. O. Hartig, R. King, T. Morita, and D. G. Wilson, "Measurement of antenna impedance using a receiving antenna," Proc. I.R.E., vol. 11, pp. 1458–1460; November, 1951. Cruft Laboratory Technical Report No. 94; December, 1949.

least to a second-order solution using the expansion parameter Ψ introduced by King and Middleton.[5,6] A first-order solution using the parameter Ω is not adequate quantitatively. Results obtained from first-order and approximate second-order analyses of the center-loaded receiving antenna following the procedure of King and Middleton have been reported,[7] but without a description of the basic theory.[8]

The theoretical[9,10] and experimental investigations of the impedance of a cylindrical antenna center-driven from an open-wire line (or base-driven from a coaxial transmission line) with a finite spacing of the conductors have demonstrated the importance of transmission-line end-effects and of coupling of the transmission line to the antenna near their junction. The ideal theoretical impedance Z_0 of an antenna center-driven by a discontinuity in scalar potential may be interpreted operationally as the extrapolation to zero line spacing of the *measured*, apparent impedance Z_{sa} on a line with a finite spacing that is reduced progressively. The apparent impedance Z_{sa} terminating a line that is end-loaded by an antenna is obtained from the theoretical impedance Z_δ of the antenna (as determined theoretically with a separation 2δ between the adjacent ends), in conjunction with a suitable terminal-zone network of lumped elements.[9] Since the *same* apparent impedance Z_{sa} is involved in the analysis of the receiving antenna when loaded by a transmission line with finite line spacing, it is evident that Z_δ and the appropriate terminal-zone network must be known if the current or the power in the receiver or other load at the end of the transmission line is to be determined.

Outline of the Theory

The general analysis for the current in a cylindrical antenna follows that in the earlier analyses[2] except that a general parameter Ψ (defined as proportional to the approximately constant ratio of vector potential to current on the surface of the antenna) is used instead of $\Omega = 2\ln(2h/a)$, and integrals are extended from $-h$ to $-\delta$ and δ to h instead of from $-h$ to h. The resulting expression for the even current at a point z along a receiving antenna with an impedance Z_δ and an effective load $Z_{L\delta}$ (which includes end and coupling effects) is

$$I_z = U\left\{u_\delta(z) - v_\delta(z)\frac{Z_\delta}{Z_\delta + Z_{L\delta}}\left[u_\delta(\delta)Z_{L\delta} - \frac{2q_0}{\beta_0}\sin q_0\delta\right]\right\}, \quad (1)$$

where

$$u_\delta(z) = \frac{j4\pi}{\zeta_0\Psi}\left\{\frac{\cos q_0h\cos\beta_0(z-\delta) - \cos q_0z\cos\beta_0(h-\delta)}{\cos\beta_0(h-\delta) + A_{1\delta}/\Psi + \cdots} + \frac{\frac{1}{\Psi}\left[m_{1\delta}(z)\cos\beta_0\delta + p_{1\delta}(z)\sin\beta_0\delta\right] + \cdots}{\cos\beta_0(h-\delta) + A_{1\delta}/\Psi + \cdots}\right\} \quad (2)$$

$$v_\delta(z) = \frac{j2\pi}{\zeta_0\Psi}\left\{\frac{\sin\beta_0(h-z) + M_{1\delta}(z)/\Psi + \cdots}{\cos\beta_0(h-\delta) + A_{1\delta}/\Psi + \cdots}\right\} \quad (3)$$

$$U = -E\cos\psi/\beta_0\sin\theta_2 \quad (4)$$

$$\beta_0 = 2\pi/\lambda; \qquad q_0 = \beta_0\cos\theta_2. \quad (5)$$

The first-order functions $A_{1\delta}$, $M_{1\delta}(z)$, $m_{1\delta}(z)$, and $p_{1\delta}(z)$, and corresponding higher-order functions with subscripts 2, 3, and so on, are integrals obtained in the iteration. The first-order functions may be expressed in terms of tabulated generalized sine and cosine integrals. The angles Ψ, θ_2, and $\bar{\theta}_2 = \pi/2 - \theta_2$ are illustrated in Fig. 1.

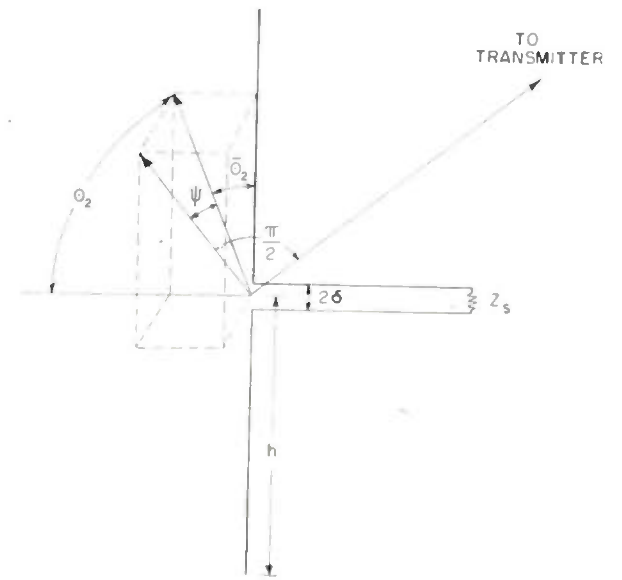


Fig. 1—Center-loaded receiving antenna in linearly polarized field.

The real expansion parameter Ψ is the magnitude at $z = z_r$ of

$$\Psi(z) = \left(\int_{-h}^{-\delta} + \int_{\delta}^{h}\right) g(z, z')K(z, z')dz', \quad (6)$$

where z_r is an appropriate reference point. For $\beta_0h < 2\pi$ $z_r = 0$.[11] The kernel in (6) is

[5] R. King, and D. Middleton, "The cylindrical antenna, current and impedance," *Quart. Appl. Math.*, vol. 3, pp. 302–305; January, 1946; vol. 4, p. 199; July, 1946.

[6] *Ibid.*, "The thin cylindrical antenna; a comparison of theories," *Jour. Appl. Phys.*, vol. 17, p. 273; April, 1946.

[7] R. King, "Graphical Representation of the Characteristics of Cylindrical Antennas," Cruft Laboratory Technical Report No. 20; October, 1947.

T. Morita, "Measurement of current and charge distributions on cylindrical antennas," Proc. I.R.E., vol. 38, p. 898; August, 1950. Cruft Laboratory Technical Report No. 66; August, 1950.

[8] This theory is developed in mimeographed form in Chapter IV, "Notes on Antennas," Cruft Laboratory, 1949. It has been reproduced and extended by S. H. Dike in Technical Report No. 14, Radiation Laboratory, Johns Hopkins University (June, 1951) under the title "Difficulties with Present Solutions of the Hallén Integral Equation." This report is discussed critically in the reference given in footnote 12.

[9] R. King, "Antennas and open-wire lines," *Jour. Appl. Phys.*, pt. I, vol. 20, p. 832; September, 1949.

[10] R. King, and K. Tomiyasu, "Terminal impedance and generalized two-wire line theory," Proc. I.R.E., vol. 37, p. 1134; October 1949. Cruft Laboratory Technical Report No. 74; April, 1949.

[11] It is shown in the reference given in footnote 12 that for electrically short antennas the expansion parameter is best defined by $\Psi(0) - \Psi(h)$ instead of $\Psi(0)$.

$$K(z, z') = \frac{e^{-j\beta_0 R}}{R}. \tag{7}$$

$$R = \sqrt{(z - z')^2 + a^2} \tag{8}$$

is the distance from the point on the surface where the vector potential is defined to the element of integration dz' at z' along the axis. The function $g(z, z')$ is the ratio of the current at z' to the current at z. It is given by

$$g(z, z') = \frac{u_\delta(z') - Sv_\delta(z')}{u_\delta(z) - Sv_\delta(z)}. \tag{9}$$

where S is the factor of $v_\delta(z)$ in (1). If the same method were followed as in the analysis of the center-driven antenna,[5] (6) would be integrated using (9) with the zeroth order values of $u_\delta(z)$ and $v_\delta(z)$ as obtained from (2) and (3). However, the distribution function (9) obtained in this manner is a function not only of the ratio h/a and of $\beta_0 h$, but of the angle θ and the load impedance $Z_{L\delta}$. Obviously, it is not practicable to use a different distribution function for the same antenna when its orientation or its load is changed. Since the expansion parameters determined from (6) in extreme cases differ relatively little, and since the over-all accuracy of a given order of solution is not sensitive to small changes in the expansion parameter, it is satisfactory to use the *same expansion parameter* for a given antenna for reception and transmission. Accordingly, the parameter[5,6] Ψ as defined for the transmitting antenna is used.

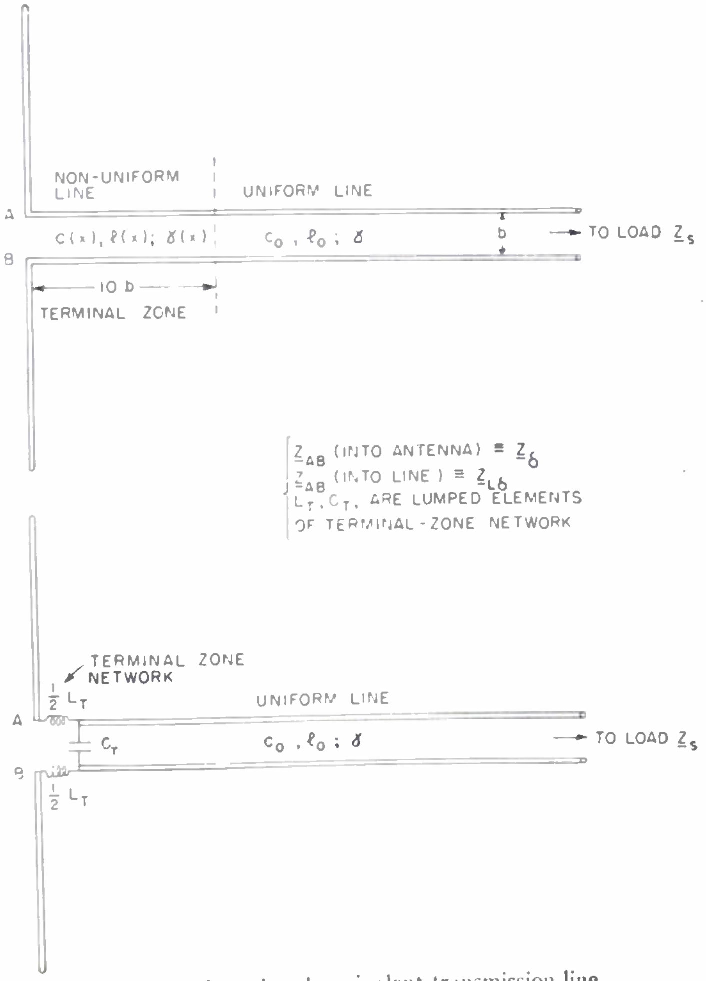


Fig. 2—Actual and equivalent transmission line.

First-order distributions of current in receiving antennas under a variety of load conditions have been computed from (1), with $\delta = 0$ and represented graphically.[7] An extensive comparison of the theoretical curves with measured values by Morita[10] shows good agreement.

Since the impedance of the center-loaded receiving antenna in the equivalent series circuit is also the impedance of the same antenna when center-driven, the *same* terminal-zone networks may be used to take account of end and coupling effects.[4,7] A typical circuit is in Fig. 2.

The Complex Effective Length

By application of Thévenin's theorem at the junction of the antenna and the line or other load, the current I_δ into the load $Z_{L\delta}$ is given by

$$I_\delta = V_\delta(Z_{L\delta} = \infty)/(Z_\delta + Z_{L\delta}), \tag{10}$$

where $V_\delta(Z_{L\delta} = \infty)$ is the open-circuit voltage maintained by the external field across the terminals of the antenna when the load is disconnected so that $Z_{L\delta} = \infty$, and Z_δ is the impedance of the antenna looking into these terminals. This voltage is

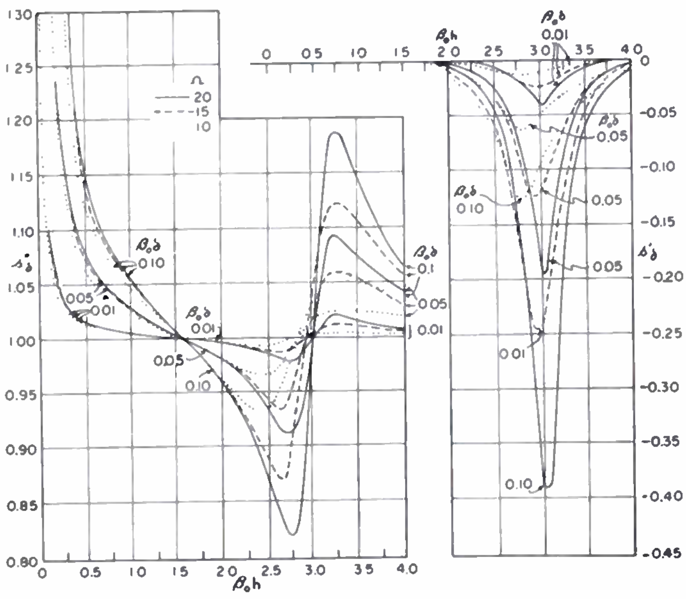


Fig. 3—The function $s_\delta = s_\delta'' + js_\delta'$.

$$V_\delta(Z_{L\delta} = \infty) = -E \cos\psi \cdot 2h_{e\delta}(\theta_2), \tag{11}$$

where the complex effective length of an antenna of length $2h$ is defined by

$$2h_{e\delta}(\theta_2) \equiv \left[u_\delta(\delta)Z_\delta + \frac{2q_0}{\beta_0}\sin q_0\delta\right]/\beta_0 \sin\theta_2. \tag{12}$$

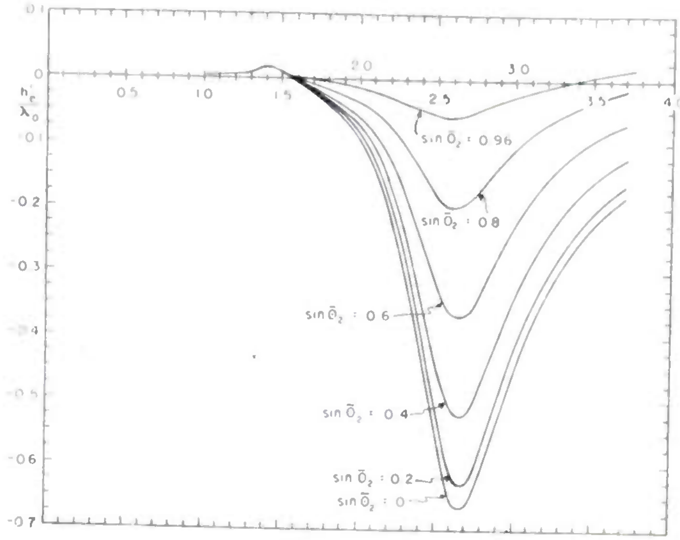


Fig. 4(a)—Imaginary part of complex effective length per wavelength; $\Omega=10$.

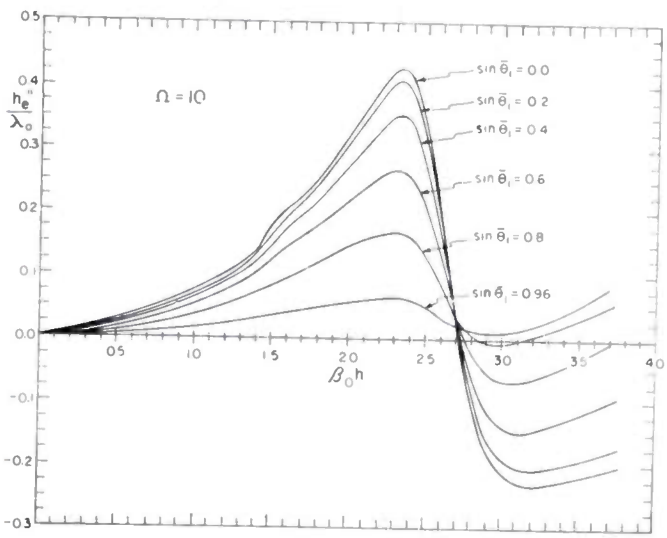


Fig. 4(b)—Real part of complex effective length per wavelength; $\Omega=10$.

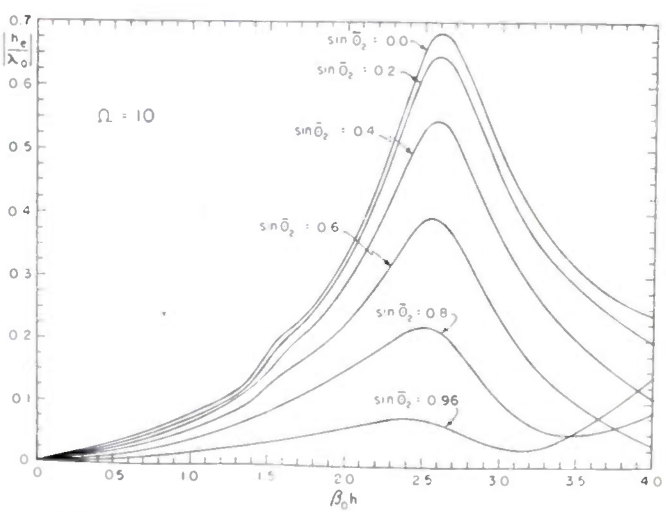


Fig. 4(c)—Magnitude of complex effective length per wavelength; $\Omega=10$.

For $\delta=0$,

$$2\beta_0 h_e(\theta_2) = Z_0 u_0(0)/\sin\theta_2. \tag{13}$$

It can be shown that (12) may be expressed as follows:

$$h_{2\delta}(\theta_2) = h_e(\theta_2)s_\delta - \delta\sin\theta_2, \tag{14}$$

where s_δ is a complex function which is represented graphically in Fig. 3 (see page 1115). Note that the term $\delta\sin\theta_2$, which is subtracted on the right in (14), is the effective half-length of a short, end-loaded antenna of actual length 2δ, that is, of the section of conductor which is missing at the center of the antenna owing to the finite spacing of the transmission line or the finite physical length of the load.

In (13), $u_0(0)$ is proportional to the current at the center of an unloaded receiving antenna and Z_0 is the impedance of a center-driven antenna. Since it has been shown[10] that the first-order distribution of current in an unloaded receiving antenna is a good approximation, and since $u_0(0)$ and Z_0 are by definition completely independent quantities, a combination of first-order $u_0(0)$ and second-order Z_0 should lead to an effective length that is comparable in accuracy with the second-order impedance. The complex effective length

$$h_e(\theta_2) \equiv h_e = h_e'' + jh_e' \tag{15}$$

is plotted in the form h_e'/λ_0, h_e''/λ_0, and h_e/λ_0, with $\beta_0 h$ as variable and $\cos\theta_2$ as parameter for $\Omega=2\ln(2h/a)=10$ in Figs. 4(a), 4(b), and 4(c). It is plotted in the complex plane with $\beta_0 h$ as running variable and $\cos\theta_2$ as parameter in Fig. 5.

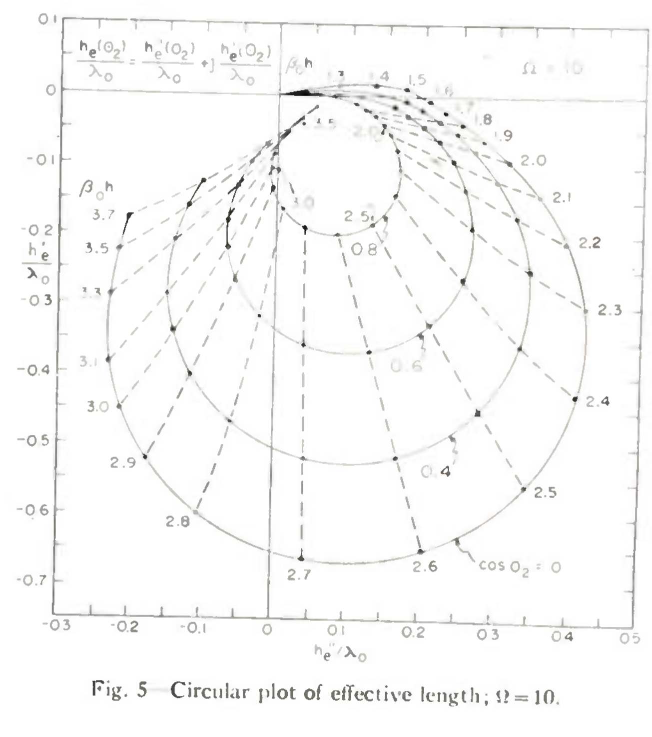
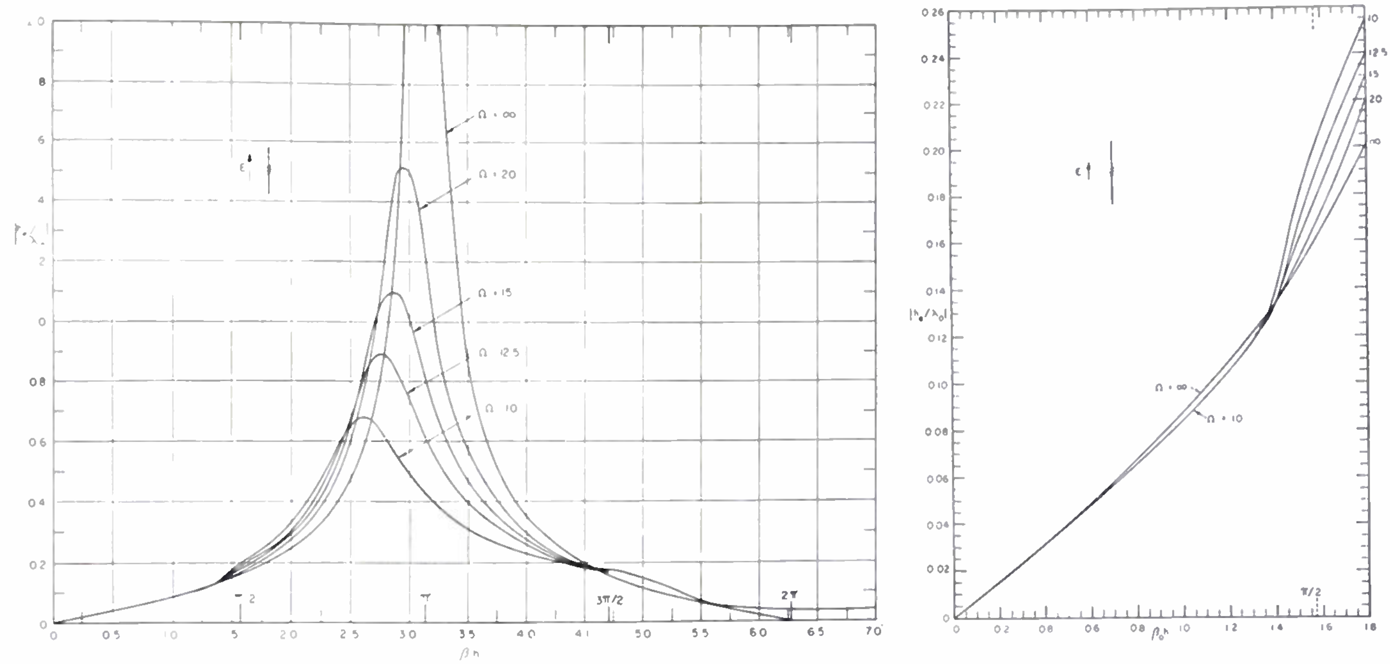


Fig. 5—Circular plot of effective length; $\Omega=10$.



(a) (b)

Fig. 6 (a) and (b)—Magnitude of effective length per wavelength for several values of Ω.

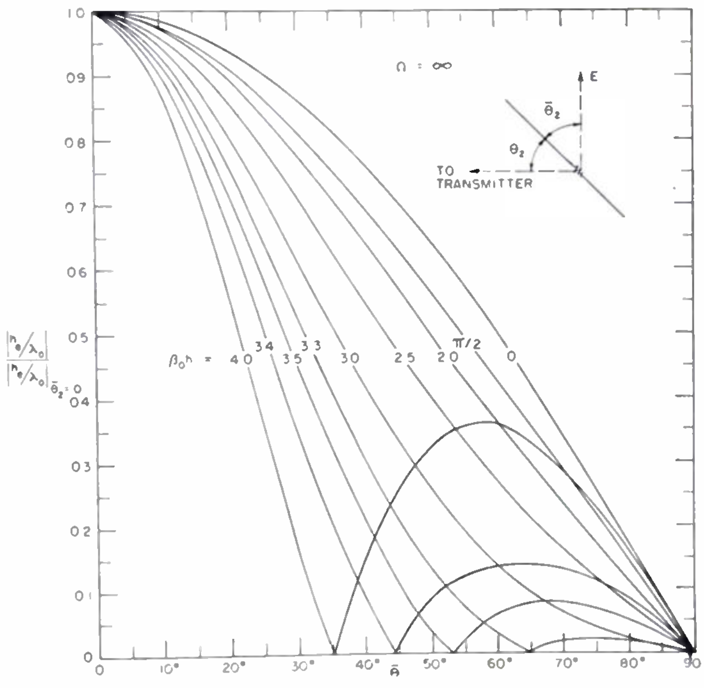


Fig. 7—Normalized zeroth-order effective length; $\Omega=\infty$.

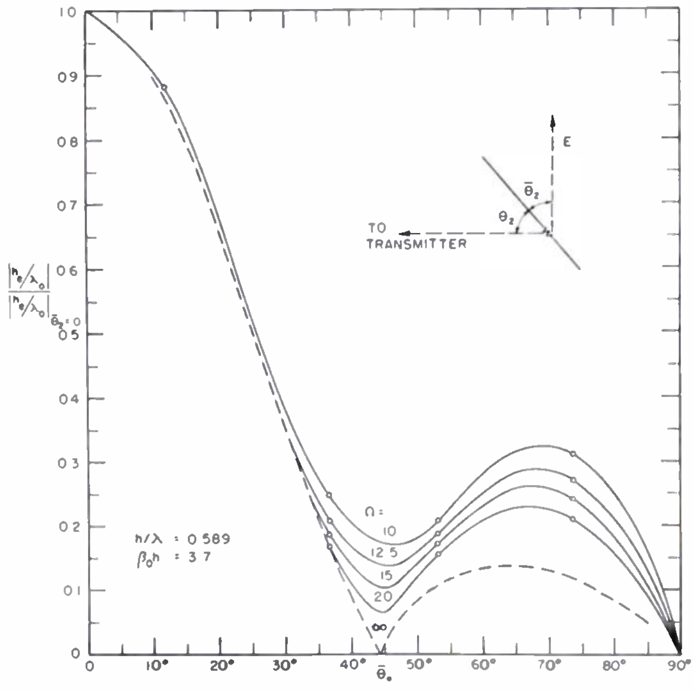


Fig. 8—Normalized effective length for $h/\lambda_0=0.589$.

With numerical values of the complex effective lengths available, the voltage $V_\delta(Z_{L\delta}=\infty)$ of the generator in the equivalent series circuit for the receiving antenna may be determined in both magnitude and phase referred to the electric field of the distant transmitter.[2] The magnitude of the effective length with the electric field parallel to the antenna ($\theta_2=\pi/2$) is shown in Fig. 6(a) and (b) for values of Ω corresponding to $h/a=75$, 260, 900, 11,000, and ∞. The zeroth-order curve ($\Omega=\infty$) is a fair approximation for thin antennas that are not too long. For electrically short antennas ($\beta_0 h<<1$), the first-order effective length in Fig. 6(b) was verified using an accurate formula.[12]

The directional properties of the receiving antenna are illustrated in Fig. 7 where the zeroth-order magni-

[12] R. King, "Theory of Electrically Short Transmitting and Receiving Antennas," Cruft Laboratory Technical Report No. 141; March, 1952. Accepted for publication in the *Jour. Appl. Phys.*, but the Appendix with the critical discussion referred to in footnote 8 is omitted.

tude of the effective length is plotted as a function of $\bar{\theta}_2=\theta_2-90°$ with $\beta_0 h$ as parameter. These zeroth-order curves are good approximations even for quite thick antennas if $\beta_0 h$ does not approach π. When $\beta_0 h$ exceeds π, minor lobes occur and sharp nulls are replaced by minima as illustrated for $\beta_0 h=3.7$ in Fig. 8.

Note that according to the reciprocal theorem the effective length of a receiving antenna is the same as the vertical field factor for the same antenna when driven.

Experimental Measurement of the Effective Length

An experimental determination of the magnitude of the effective length $|h_e(\theta_2)|$ was carried out by Morita and Taylor. The apparatus consisted of a receiving antenna erected vertically over a large, highly conducting screen and base-loaded by a coaxial line terminated in its characteristic impedance R_c. A constant electric field E parallel to the antenna was maintained by a distant transmitter, and the relative power to the load was measured as a function of the length h of the antenna.

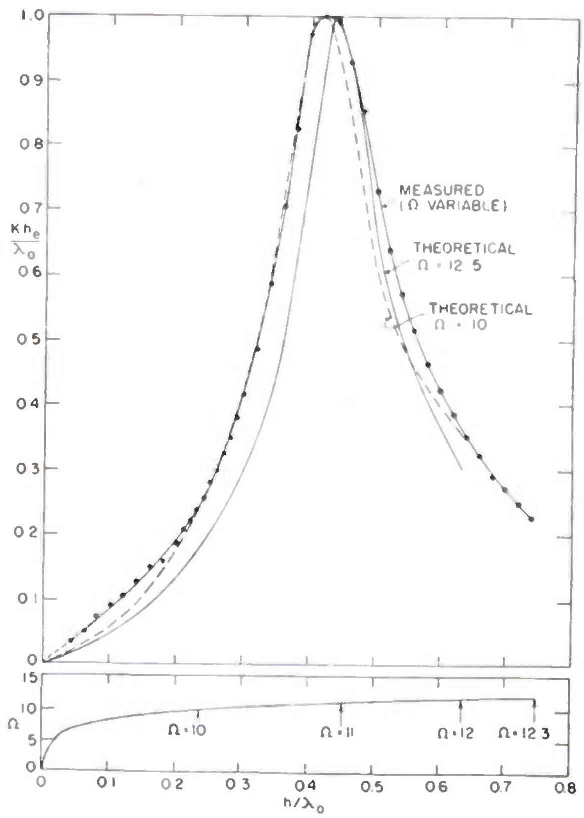


Fig. 9—Normalized effective length of receiving antenna as measured by Morita and Taylor as a function of h/λ_0. Corresponding values of $\Omega=2ln(2h/a)$ are shown together with theoretical curves for the normalized effective length for two values of Ω.

Using (10) and (11), it follows that

$$|h_{e\delta}(\theta)| \sim |I_\delta(Z_{L\delta}+Z_\delta)|, \tag{16}$$

where $Z_\delta=R_\delta+jX_\delta$ is the impedance of the antenna. Since the current in the matched line, and hence in the load, is given by

$$|I_\delta| = \sqrt{P_L/R_c}, \tag{17}$$

it follows that the magnitude of the effective length per wavelength is

$$|h_{e\delta}(\theta_2)/\lambda_0| \sim \sqrt{(P_L/R_c)[(R_\delta+R_c)^2+X_\delta^2]}. \tag{18}$$

Since R_c is known and P_L as a function of h/λ_0 has been measured, the relative value of $|h_{e\delta}(\theta_2)/\lambda_0|$ as a function of h/λ_0 can be determined from (18) using the measured value of the apparent impedance. The radius b of the sheath of the coaxial line is sufficiently small so that $Z_{\delta a} \doteq Z_\delta$. Using measured values of P_L, R_δ, and X_δ and with $R_c=65.9$ ohms, the normalized effective length of the antenna was determined and plotted in Fig. 9. Correspondingly normalized theoretical curves taken from Fig. 6(a) and (b) are also shown. In the experimental determination, h/a for the antenna varied as the length was decreased; the theoretical curves are computed for constant h/a. It follows that a continuous direct comparison is not possible. However, using the value of $\Omega=2ln(2h/a)$ shown at the bottom of Fig. 9, it is seen that the general agreement for corresponding values of Ω is quite good.

Power in the Load, Directivity, Effective Cross Section

The power transferred to the load of a receiving antenna is

$$P_L = \tfrac{1}{2}I_L^2 R_{L\delta} = \tfrac{1}{2}I_\delta^2 R_{L\delta}, \tag{19}$$

where $I_\delta=I_L$ is given by (10) with (11). This power is maximized in so far as adjustment of the load is concerned when $Z_{L\delta}$ is the complex conjugate of the antenna impedance Z_δ. In this case

$$P_{L\max} = \frac{|h_{e\delta}(\theta_2)E\cos\psi|^2}{2R_\delta}. \tag{20}$$

This may be expressed as follows:

$$P_{L\max} = \frac{|\lambda_0 E\cos\psi|^2}{8\pi\zeta_0} D(\theta_2, \beta_0 h), \tag{21}$$

where

$$D(\theta_2, \beta_0 h) \equiv \frac{\zeta_0}{\pi}\frac{|\beta_0 h_{e\delta}(\theta_2)|^2}{R_\delta}. \tag{22}$$

The dimensionless directivity[13] or gain defined in (22) is plotted in Fig. 10 as a function of $\beta_0 h$ with $\theta_2=\pi/2$, $\delta=0$, using $|he(\theta_2)/\lambda_0|$ as given in Fig. 6 and second-

[13] Note that the effective dissipation cross section of the antenna with a *conjugate-matched load* is equal to the reradiating or scattering cross section and is given by $\sigma_{\text{dis}}=\sigma_{\text{rad}}=(\lambda_0^2/4\pi)D(\theta_2, \beta_0 h)$.

It is evident that (22) may be expressed in the form $h_e=k\lambda_0\sqrt{DR_L}$, where $k=1\text{-}\sqrt{4\pi\zeta_0}$ and the load resistance R_L is substituted for the *equal* antenna resistance. This formula does *not* mean that the effective length of a receiving antenna is a function of its load. It is seen from (12) or (13) that h_e depends only on the dimensions and orientation of the antenna. It is the power to the load that depends on the load, not the effective length.

order resistances for $R_\delta \doteq R_0$. For $\beta_0 h < \pi/2$ the corrected resistances as determined from the exact analysis for the electrically short antenna[12] are used. Curves for a

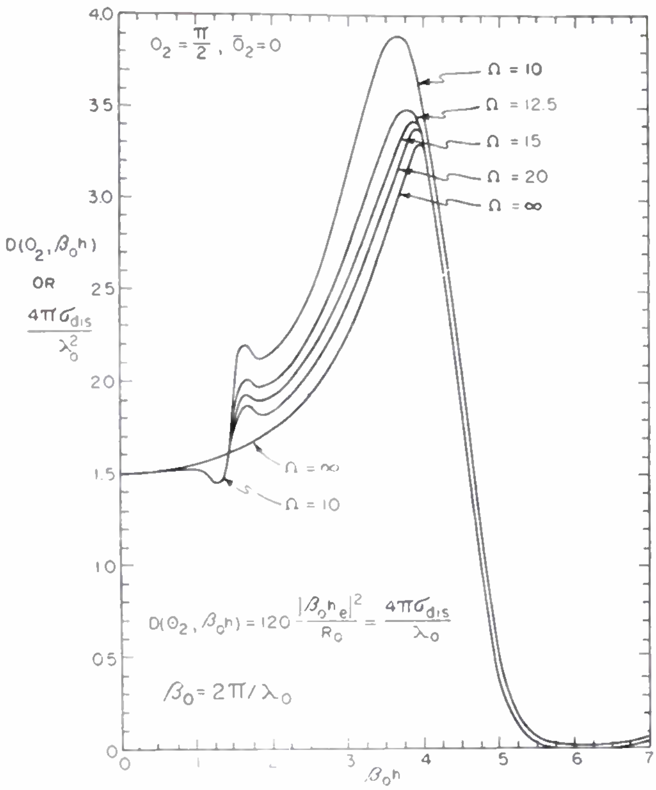


Fig. 10—Gain or directivity of receiving antenna parallel to the incident electric field.

range of values of θ_2 are in Fig. 11. The maximum value of $D(\theta_2, \beta_0 h)$ (which gives $P_{L\,\max\max}$ occurs when $\beta_0 h$ is near 4 and when $\theta_2 = \pi/2$ or $\bar{\theta}_2 = 0$. The magnitude of this

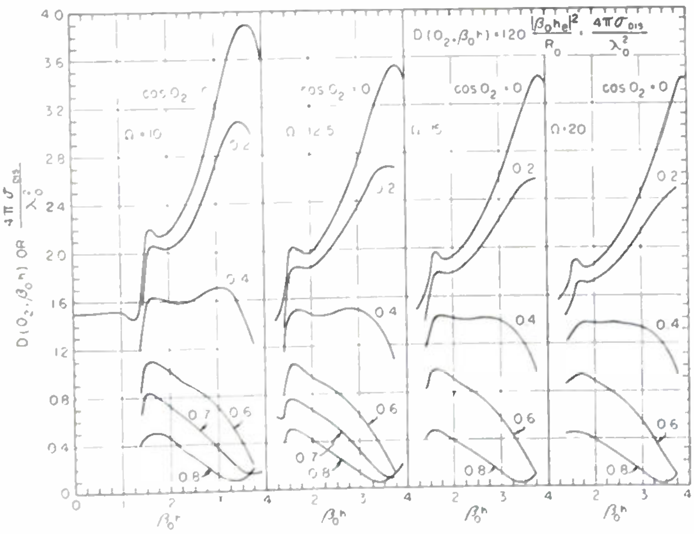


Fig. 11—Gain or directivity of receiving antenna with its orientation angle θ_2 (Fig. 1) as parameter.

power is greater for thicker antennas. Note that the power to a matched load is almost independent of h for $\beta_0 h \leq 0.5$. However, owing to the high capacitive reactance and very low resistance of short antennas, a conjugate match is difficult to obtain for them, and losses in the matching network may exceed the power to the load. The minor extremes and the general behavior of the curves in Fig. 10 near resonance are noteworthy. Since small changes well within the possible error of second-order resistances and first-order effective lengths can modify greatly the detailed structure of the curves near resonance, the minor maximum and minimum may be much less significant; in particular, the decrease slightly below 1.5 in the latter appears questionable.

Experimental Determination of the Power in the Load

An experimental determination of the power in a conjugate matched load ideally requires the direct measurement of the power dissipated in an impedance $Z_{L\delta} = Z_\delta^*$. Since it is difficult to adjust $Z_{L\delta}$ accurately for each length of the antenna, an essentially equivalent and more convenient and accurate procedure is to make use of a load given by $Z_{L\delta} = R_c$ for the transmission line. By determining the relative effective length of the antenna in terms of the power to R_c and substituting the values of $h_{e\delta}(\theta_2)$ so determined in (22), a quantity proportional to the power transferred to a load that is the complex conjugate of the impedance of the antenna may

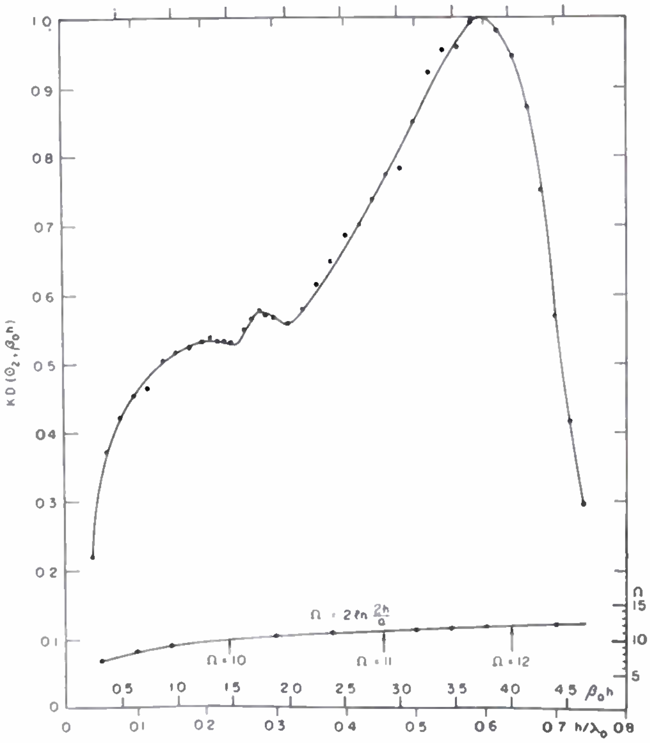


Fig. 12—Measured normalized gain of receiving antenna. Data of Morita and Taylor.

be computed. Using the experimentally determined values of the normalized effective length given in Fig. 9, the normalized relative directivity $KD(\theta_2, \beta_0 h)$ as determined by Morita and Taylor are in Fig. 12.

If account is taken of the fact that h/a is not constant, the general shape of the curve in Fig. 12 is in good agreement with the curves in Fig. 10 except for small values of $\beta_0 h$ where the experimental method fails. Note that the location of the maximum value of $D(\theta_2, \beta_0 h)$ agrees with theory. Moreover, and in agreement with theoretical predictions, a definite irregularity in the form of a minor maximum occurs near resonance. The amplitude of this oscillation in the otherwise smooth experimental curve is smaller than for the corresponding value of h/a in the theoretical curves, but the general behavior is correctly given. Using the method of conjugate match, similar results were obtained by Dike and King[14], except that their results do not confirm the minor maximum near resonance.

[14] S. H. Dike, and D. D. King, "The Cylindrical Dipole Receiving Antenna," Technical Report No. 12, Radiation Laboratory, Johns Hopkins University, Silver Springs, Md.; May, 1951. A Cruft Laboratory Technical Report discussing Dr. Dike's measurements critically is in preparation.

Conclusion

The theory of the center-loaded receiving antenna (or base-loaded antenna over a conducting plane) has been improved by obtaining an approximate second-order result using the expansion parameter of King and Middleton. Account is taken of the finite separation of a transmission-line load. Theoretical curves for effective length and for maximum power to the load (directivity) are compared with measured results with satisfactory agreement.

Acknowledgments

The co-operation of Dr. T. Morita in making available measurements made by him and Dr. J. Taylor is gratefully acknowledged.

Cross Polarization of Scattered Radio Waves*

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Summary—The polarization of the signal reaching a receiving antenna by the scattering mechanism proposed by Booker and Gordon[1] is investigated. Equations are presented which give the response of dipole antennas oriented horizontally, vertically, and axially,[2] relative to a linear polarized source. The relative response of the three antennas is calculated for selected values of the scattering parameters and a comparison made with the measured response of similar antennas to a 102.9-mc signal arriving over a path length of 147 miles.

I. Introduction

IN A PREVIOUS PAPER,[3] the author developed a method for computing the total radio energy arriving at a receiving point by the scattering mechanism proposed by Booker and Gordon.[1] In the previous work, no attempt was made to analyze the polarization of the total received energy. The present paper extends the analysis to include a study of the polarization of the scattered radio energy reaching the receiver. The responses of antennas with horizontal, vertical, and axial orientation are presented for a signal originating at a transmitter with linear polarization.

Numerical integration is performed for a selected set of values of the scattering parameters to give a measure of the cross polarization which may be expected. These results are compared with similar measurements made on a horizontally polarized 102.9-mc signal arriving over a path length of 147 miles.

II. Discussion of Assumptions

The equations presented in this paper are based on a number of assumptions which are made to simplify the solution as much as possible and which still approximate actual conditions. The assumptions involved are as follows:

(1) Refraction is taken as standard and straight-line propagation over a smooth earth of 4/3 radius used.

(2) Secondary scattering is negligible and no loss of energy in the incident beam occurs as the result of scattering.[4]

(3) The scale-of-turbulence and the mean-square deviation of the index-of-refraction are the same throughout all regions of the sky.

(4) The sky is evenly illuminated over the important scattering region by a transmitter with a linear polarization. The receiving antennas are dipoles with normal radiation characteristics.

(5) Only direct radiation from the scattering centers is considered. The effect of ground reflections can be included, if desired, by considering the radiation pattern of the receiving antenna and its image as a receiving unit. This, of course, assumes that the distance be-

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† Electrical Engineering Research Laboratory, University of Texas, Box F, University Sta., Austin 12, Tex.

[1] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," Proc. I.R.E., vol. 38, pp. 401–412; April, 1950.

[2] See Section IV for detailed description of these orientations.

[3] A. H. LaGrone, "Volume integration of scattered radio waves," Proc. I.R.E., vol. 40, p. 54; January, 1952.

[4] A. H. LaGrone, W. H. Benson, Jr., and A. W. Straiton, "Attenuation of radio signals caused by scattering," *Jour. Appl. Phys.*, vol. 22, pp. 672–674; May, 1951.

tween the antenna and its image is small as compared to the distance to the scattering center so that θ and X in (1) are approximately the same for the two antennas.

(6) In the numerical example, which is compared with the field-measured data, conditions are assumed which approximate those of the field measurements.

III. Magnitude of the Scattered Signal

The magnitude of the power scattered per unit solid angle, per unit incident power density, and per unit macroscopic element of volume is deduced by Booker and Gordon[1] to be

$$\sigma(\theta, x) = \frac{\overline{(\Delta\epsilon/\epsilon)^2}(2\pi l/\lambda)^3}{\lambda[1+\{(4\pi l/\lambda)\sin\tfrac{1}{2}\theta\}^2]^2}\sin^2 x = G\sin^2 x, \quad (1)$$

where ϵ is the average permittivity, $\Delta\epsilon$ is the departure of the permittivity from its average value, l is the scale-of-turbulence, λ is the wavelength, θ is the angle between the direction of incidence and the direction of scattering, and X is the angle between the direction of the electric field vector and the direction of scattering.

IV. Scattered Power Received from a Unit Scattering Volume

Let three identical dipoles at R (Fig. 1) be oriented as follows: (a) one horizontal and normal to line TR, (b) one in the vertical plane and normal to line TR, and (c) one lying along the line TR. These will henceforth be referred to as the horizontal, vertical, and axial dipoles, respectively. The subscripts h, v, and a will be used to identify them in equations.

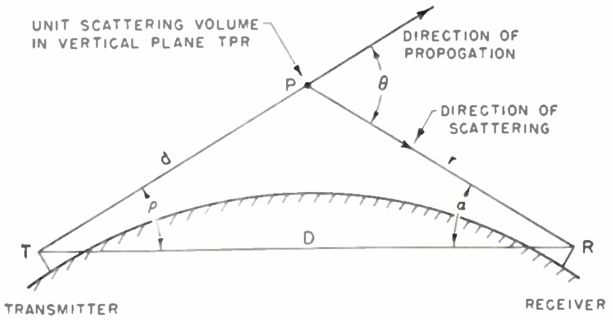

Fig. 1—Scattering geometry with unit scattering volume in a vertical plane passing through the transmitter and receiver.

The elementary dipole induced in the unit scattering volume will, in general, be so oriented as to produce a field at the receiver which will have horizontal, vertical, and axial components as these are defined for the dipoles above. With this system of dipoles, then, it can be shown that the power received by the dipoles will be

$$W_h = \frac{AG\phi_h}{r^2}[C_1\cos^2\theta\sin^2\delta + C_2\cos^2\delta + 2\sqrt{C_1C_2}\cos\theta\sin\delta\cos\delta], \quad (2)$$

$$W_v = \frac{AG\phi_v}{r_2}[C_1\cos^2\theta\cos^2\delta + C_2\sin^2\delta - 2\sqrt{C_1C_2}\cos\theta\sin\delta\cos\delta], \quad (3)$$

$$W_a = \frac{AG\phi_a}{r^2}[C_1\cos^2\theta], \quad (4)$$

where

A = effective area of dipole,
ϕ_h, ϕ_v, and ϕ_a = dipole radiation characteristics,
δ = arc tan $[\tan\beta\cos\alpha]$,
α = elevation angle at receiver (Fig. 1),
β = angle of tilt of the plane TPR from the vertical plane (Fig. 2),
C_1 = power-density component associated with the electric-field-intensity component in the plane TPR (Fig. 2) incident on the unit scattering volume at P,
C_2 = power-density component associated with the electric-field-intensity component normal to the plane TPR (Fig. 2) incident on the unit scattering volume at P,
r = distance from the unit scattering volume to the receiver (Fig. 1).

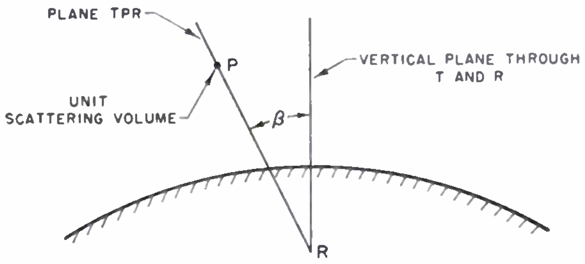

Fig. 2—Scattering geometry with unit scattering volume in plane TPR tilted at the angle β with respect to the vertical plane passing through transmitter and receiver.

V. Numerical Example

An example of cross polarization is computed by numerical integration for $(l/\lambda)=4$. The source is assumed to radiate a signal which is polarized normal to the plane TPR (Fig. 1) and to evenly illuminate the important scattering region of the sky. Several approximations are possible for $(l/\lambda)=4$ which do not seriously affect the final solution, but because of space limitations, cannot be given in detail here. The final equations which were used to compute the total scattered power received by the dipoles are given below with the results tabulated in Table I.

$$P_h = P_s\int_{\beta=-\pi/2}^{\pi/2}\int_{\theta_m}^{\theta}(\theta-\theta_m)\frac{\sigma_h}{D}\,d\theta d\beta \quad (5)$$

$$P_v = P_s \int_{\beta=-\pi/2}^{\pi/2} \int_{\theta_m}^{\theta} (\theta - \theta_m) \frac{\sigma_v}{D} d\theta d\beta \qquad (6)$$

$$P_a = P_s \int_{\beta=-\pi/2}^{\pi/2} \int_{\rho_m}^{\rho} \int_{\alpha_m}^{\alpha} \frac{\sigma_a}{D} d\alpha d\rho d\beta \qquad (7)$$

P_s = power radiated per unit solid angle by the source (P_s was set equal to unity in the numerical example).

$$\sigma_h = G[\cos^2\theta \sin^4\beta + \cos^4\beta + 2\cos\theta \sin^2\beta \cos^2\beta]$$

$$\sigma_v = G[\cos^2\theta \cos^2\beta \sin^2\beta + \cos^2\beta \sin^2\beta - 2\cos\theta \sin^2\beta \cos^2\beta]$$

$$\sigma_a = G[\cos^2\theta \sin^2\beta \sin^2\alpha].$$

The subscript m denotes minimum value of the angle (value at grazing, Fig. 1).

TABLE I

RELATIVE SIGNAL STRENGTHS FOR SCATTERING WITH $(l/\lambda) = 4$

	Signal Level in db Relative to Horizontal Dipole Signal at Various Distances from Transmitter				
Dipole	25 miles	45 miles	75 miles	125 miles	205 miles
Horizontal	0.0	0.0	0.0	0.0	0.0
Vertical	−37.2	−36.1	−34.6	−32.6	−29.7
Axial	−28.4	−27.7	−26.7	−25.4	−23.6

VI. FIELD MEASUREMENT OF CROSS-POLARIZED SIGNAL

To test this theoretical analysis, field-strength measurements were made on radio station KPRC-FM, Houston, Texas. Three identical dipoles and measuring equipments were set up to record simultaneously and continuously the power received on horizontal, vertical, and axial dipoles as these are defined in Section IV. KPRC-FM broadcasts a horizontally polarized signal of 102.9 mc from an antenna located on top of a tower 342 feet above local terrain. The receiving dipoles were 32 feet above local terrain and 147 miles from the transmitting antenna.

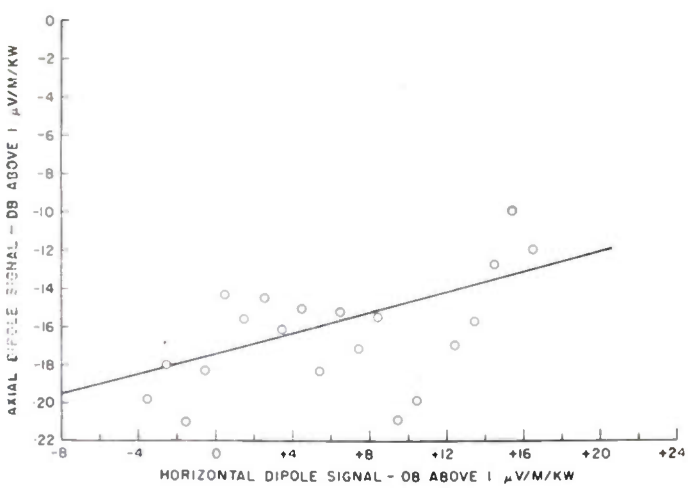

Fig. 3—Axial dipole component of scattered wave.

The field-measured data are shown in Figs. 3 and 4 for the period November 13 through November 28, 1950. The abscissas of the points represent hourly median values of the horizontal signal and the ordinates composite hourly median values of the axial or vertical signal. The straight lines are the weighted least-square lines. The weight given each point was the number of hours of data represented by the point.

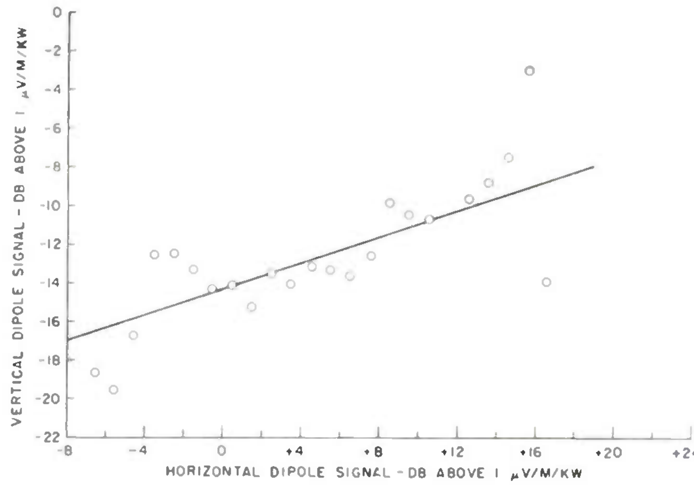

Fig. 4—Vertical dipole component of scattered wave.

Fig. 5 is a plot of the hourly median signal for a single day. Each point represents a single hourly median signal.

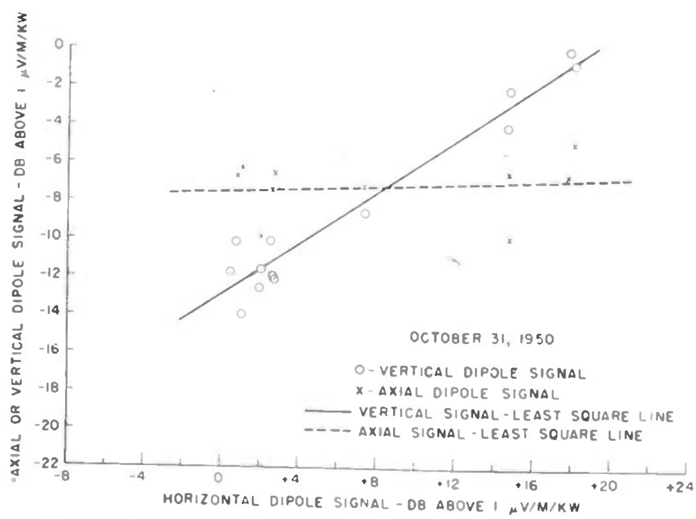

Fig. 5—Axial and vertical dipole components of scattered wave.

VII. COMPARISON OF THEORETICAL AND FIELD-MEASURED DATA

The numerical example in Section V was computed for $(l/\lambda) = 4$. This confined the major scattered field to small values of θ and meant that the scattered signals were coming from near the horizon. Under these conditions, and with the receiving dipole characteristics as assumed, the axial and vertical signals were found to be approximately 25 and 32 db, respectively, below the horizontal signal.

The field measurements shown in Figs. 3 and 4 show the axial and vertical dipole signals varying from 29 and 25 db, respectively, below the horizontal signal during strong signal periods to 13 and 10 db, respectively, below the horizontal signal during weak signal periods.

A comparison of the field measurements with the numerical example reveals that the axial and vertical dipole signals are interchanged in their relative magnitudes. On a few days, such as shown in Fig. 5, the vertical signal did drop below the axial signal for long periods of time. The fact that the relative magnitudes are interchanged is not too surprising, however, as a small vertical component is known to be propagated in the direction of Austin by KPRC-FM. Such a component would contribute materially, by the scattering process, to the signal received on the vertical dipole while producing relatively little effect on the horizontal and axial dipole signals. This could account for the difference noted in the measured and computed signals. No axial component could be propagated; hence, no effect similar to this could be associated with the axial dipole signal.

Measurements made by this laboratory in December, 1949 and January, 1950[5] on KPRC-FM, Houston, Texas and WFAA-FM, Dallas, Texas show conclusively that, under strong signal conditions, the major part of the signal comes from the horizon and that, under weak signal conditions, a significant part of the signal does not come from the horizon. The measurements were made using a conventional dipole and a directive antenna (double dipole) with both horizontally polarized and pointed in the direction of the transmitter. In comparing the signals received by the two antennas, it was noted that the ratio of the signals received was a function of the signal level. For strong signals, the measured ratio was the normal gain of the directive antenna, indicating that the signals were arriving horizontally and from the direction of the transmitter. For weaker signals, however, the signal ratio in decibels decreased linearly with the strength of the signal received by the directive antenna, indicating that the signals were not all coming from the horizon in the direction of the transmitter, but were coming from a rather large area of sky.

The strong signals in the field-measured data would then appear to have come from near the horizon and should compare with the signals in the numerical example for $(l/\lambda)=4$. Figs. 3 and 4 show the axial and vertical signals to be at least 25 db below the horizontal signal under the strong signal conditions, which does indicate some agreement with the numerical integration values. Cross-feed in the system at these signal levels, however, prevented an accurate measurement of the axial and vertical dipole signals and made it impossible to use this criterion to distinguish between scattered signals and those due to internal reflections or reflections from elevated layers.

Rough calculations were made for $(l/\lambda)=0.08$. In this case, the significant scattering region is extended to large values of θ. These calculations show the axial and vertical dipole signals to be only 8 and 10 db, respectively, below the horizontal signal, Table II. This would correspond to the case of the weak signals which come from a rather large area of the sky. Under the weak signal conditions the field measurements show very good agreement with the numerical example as vertical and axial signals were found 10 and 13 db, respectively, below the horizontal signal.

TABLE II

RELATIVE SIGNAL STRENGTH FOR SCATTERING WITH $(l/\lambda)=0.08$

	Signal Level in db Relative to Horizontal Dipole Signal at Various Distances from Transmitter				
Dipole	25 miles	45 miles	75 miles	125 miles	205 miles
Horizontal	0.0	0.0	0.0	0.0	0.0
Vertical	−10.4	−10.3	−10.3	−10.2	−10.2
Axial	− 7.9	− 8.0	− 8.0	− 8.0	− 8.0

XII. CONCLUSION

1. A formula is presented for calculating the scattered power received on horizontally, vertically, and axially polarized dipoles in terms of the scattering parameters.

2. Cross polarization in the scattered wave is relatively unimportant for (l/λ) large, as shown for $l/\lambda=4$. In view of this, it is evident that no significant cross polarization in the scattered wave would be expected at microwave frequencies. As (l/λ) decreases, the extent of cross polarization in the scattered wave increases; hence, the extent of cross polarization is an indication of the magnitude of the scale-of-turbulence.

3. The stronger scattered signals come principally from near the horizon and indicate larger values for the scale-of-turbulence with negligible cross polarization.

4. Experimental results confirm the existence of field components at the receiver which were not present in the transmitted wave.

[5] A. W. Straiton, D. F. Metcalf, and C. W. Tolbert, "A study of tropospheric scattering of radio waves," PROC. I.R.E., vol. 39, pp. 643–648; June, 1951.

Correspondence

Sweep-Frequency Oblique-Incidence Ionosphere Measurements over a 1,150 km Path*

The National Bureau of Standards has been conducting a sweep-frequency time-delay-measurement experiment between Sterling, Virginia, and St. Louis, Missouri. Equipment has been installed to permit simultaneous pulse transmission and reception at both ends of the 1,150-km path as well as vertical-incidence virtual height-versus-frequency recording at the path midpoint, which is located near Batavia, Ohio.

Although it is too early to draw definite conclusions from the work, it is felt that the accompanying display of two undisturbed-day records may be of interest. Referring to Fig. 1, the upper print contains plots of equivalent path length versus frequency for pulsed signals received at Sterling from St. Louis. It also contains plots of virtual height-versus-frequency made at vertical incidence with the same equipment. Considering the oblique-incidence records, the following points are of interest: (A) E-layer transmission, with a maximum usable frequency of 12.7 mc; (B) (C) F_1-layer transmission, with an ordinary-wave maximum usable frequency (B) of 11.6 mc; (D) (E) F_2-layer transmission with an ordinary-wave maximum usable frequency of 10.3 mc; (F) two-hop F_1-layer transmission; (G) two-hop F_2-layer transmission; (H) sporadic-E-layer transmission. The traces at (J) are of local, vertical-incidence reflections.

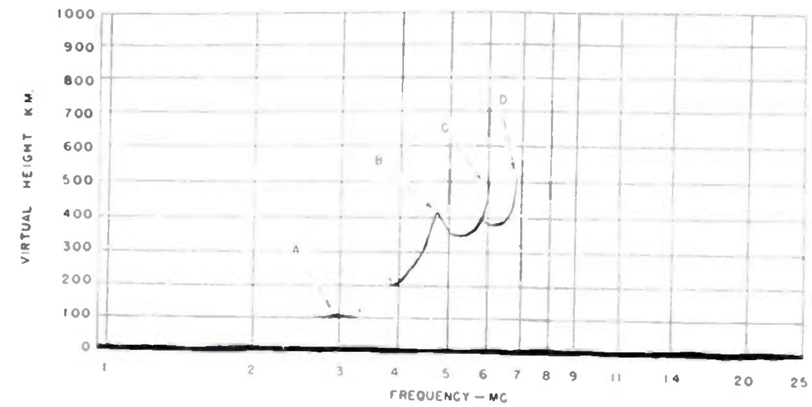

Fig. 2—Corresponding vertical-incidence recordings made at Batavia, Ohio. Both records made on September 4, 1951 at 11:30 A.M.

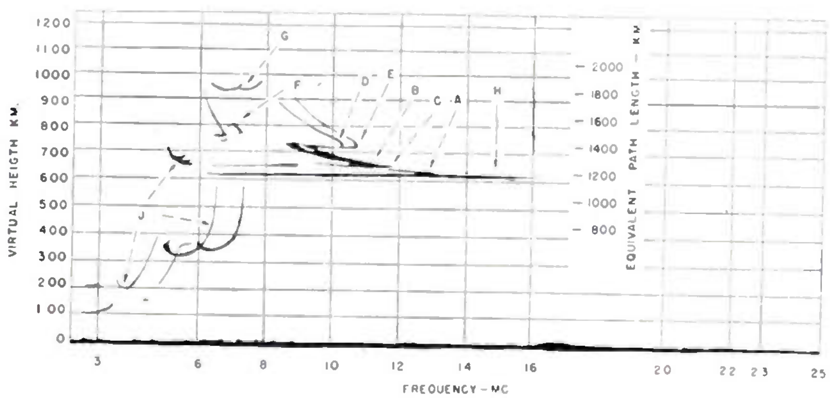

Fig. 1—Oblique-incidence and vertical-incidence recordings made at Sterling, Virginia.

Fig. 2, a conventional ionosphere recording of virtual height versus frequency made at vertical incidence at Batavia, Ohio, contains the following: (A) E-layer ordinary wave, with critical frequency of slightly less than 3 mc; (B) F_1 ordinary wave, with a critical frequency of 4.7 mc; (C) (D) F_2-layer reflections, with an ordinary-wave critical frequency (C) of about 6 mc.

The two records described above were made on September 4, 1951, at 11:30 a.m.

One purpose of the experiment is to check the accuracy of the transmission-curve method[1] of obtaining oblique-incidence maximum usable frequencies from vertical-incidence data. As an illustration, the following data have been scaled from the records shown:

TABLE I

Layer	*Maximum Usable Frequency Scaled from Oblique-Incidence Records*	*Maximum Usable Frequency from Midpoint Data and Transmission-Curve Calculation*
E ordinary	12.7	12.9
F_1 ordinary	11.6	11.0
F_2 ordinary	10.3	10.2

It will be noted that a fair agreement has been obtained between the observed and calculated maximum usable frequencies. The results are not to be considered conclusive because of possible height-scale errors in the vertical-incidence records.

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E. E. Ferguson
Central Radio Propagation Laboratory
National Bureau of Standards
Washington 25, D. C.

* Received by the Institute, February 18, 1952.

[1] N. Smith, "The relation of radio sky-wave transmission to ionosphere measurements," Proc. I.R.E., vol. 27, p. 332; May, 1939.

A Note on "A Precision Decade Oscillator"*

My attention has been drawn to an article by Edwards, entitled "A Precision Decade Oscillator."[1] It is of particular interest to me because my company, Muirhead and Company, Ltd., which in 1938 produced a commercial RC oscillator using a Wien bridge network, has been engaged in the manufacture of precision decade oscillators since 1940.

The original suggestion of a decade oscillator came from Wigan, who was then at the Ministry of Supply and was interested in variable-frequency oscillators having a frequency accuracy and stability of a few tenths of 1 per cent.

Our aim was to produce an oscillator covering the range 1 cps to 100 kc on 4 decade dials in 1 cps steps up to 10 kc and 10 cps steps up to 100 kc. The frequency accuracy achieved was 0.1 to 0.2 per cent over the major portion of this range.

As a result of preliminary work, it became obvious that when resistance decades were used over a range of 10,000 to 1 the effect of amplifier output impedance became significant. Wigan then introduced an additional resistance R_1 into the RC network, as shown in Fig. 1, where A represents an

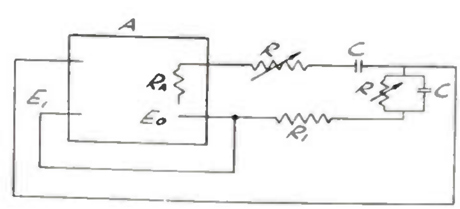

Fig. 1

amplifier with output impedance R_A and zero phase shift between input voltage E_i and open-circuit output voltage E_0. The two

* Received by the Institute, December 26, 1951.

[1] C. M. Edwards, "A precision decade oscillator for 20 cycles to 200 kilocycles," Proc. I.R.E., vol. 39, pp. 277–278; March, 1951.

Correspondence

R's are assumed equal to one another as are the two C's. It can easily be shown that the frequency of oscillation ω_0 with unity loop gain and R_1 equal to zero is given by

$$\omega_0 = \frac{1}{RC\sqrt{1+\frac{R_A}{R}}}.$$

Thus, if ω_0 is to be accurate to 0.2 per cent, either the factor R_A/R must not exceed 0.4 per cent or the series R must be reduced to offset the effect of R_A. Trimming the individual series resistances R in this way is practicable when the range of R is 10:1, as in the oscillator described by Edwards, but becomes awkward and cumbersome when a range of 10,000 to 1 is used.

If, however, the resistance R_1 is made equal to $R_A/2$, the frequency ω_0 is given by $\omega_0=1/RC$, which is the desired condition. Under these circumstances the network reduces to that shown in Fig. 2, and it is clear that E_1 is then in phase with E_0.

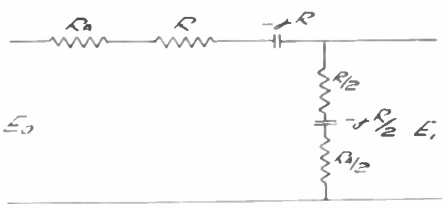

Fig. 2

The advantages derived from this additional resistance are considerable. It makes possible the use of comparatively low R and high C in the RC network, and avoids the necessity for separate trimming of the individual resistances or the stray capacitances associated with the various positions of the decade switches. Also, this resistance can be used to cancel out, to a limited but very useful extent, the effect of phase shift in the amplifier at the higher frequencies. For this purpose the resistance R_1 is set somewhat higher than its theoretical value. Since the influence of this resistance on the oscillation frequency increases as the main tuning resistances R are reduced, its effect is negligible at low frequencies but beneficial at the higher frequencies where amplifier phase shift may be of significance.

In the oscillator the main tuning resistances vary from 200 ohms to 2 megohms. They are 0.1 per cent wire-wound nonreactive types, except for the highest values which, of course, affect only the lowest frequencies. In practice, the resistance R_1 is adjusted until proper decading is obtained for the first decade dial, and thereafter it is only necessary to trim up the main tuning condensers to bring the actual frequencies in line with the indicated frequencies.

In addition to good frequency accuracy and stability, other advantages of the decade oscillator are its ready repeatability of setting and the availability of highly accurate incremental changes in frequency either in minute or large steps.

Apart from the 1 cps to 100 kc oscillator, to which reference has been made, various other instruments using the same principle have been manufactured during the last ten years. One of these is a decade oscillator covering the range 100 cps to 40 kc associated with a transmission measuring set for measuring gains and losses. This is a portable instrument arranged to work from normal ac mains or a 12-volt accumulator, and was designed in conjunction with the Ministry of Supply during World War II for the maintenance of Army carrier telephone circuits in the field. A later development is a new decade oscillator covering the range 0.1 cps to 20 kc.

Many hundreds of decade oscillators of various types have been manufactured by my company, and are in service in England and in other parts of the world. Experience gained with them has proved the high accuracy and stability of this type of instrument, and has shown that their field of use is extremely wide and varied.

J. A. B. Davidson
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Muirhead and Company, Ltd.
Elmers End
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England

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2. British Patent No. 524314; October 20, 1940; E. R. Wigan.
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4. E. R. Wigan and J. A. B. Davidson, "The Decade Oscillator," Muirhead Technique, vol. 1, pp. 20, 31; 1947.
5. H. D. Binyon, "Applications of the Decade Oscillator Type D-105-A," Muirhead Technique, vol. 4, p. 27; 1950.

A Stable Amplitude-Modulated Microwave Generator*

In the course of an experimental investigation at a frequency of 9,300 mc, it became desirable to have a low-power, amplitude-modulated, radio-frequency generator of good frequency stability. The most commonly used radio-frequency generator for such an application consists of a reflex-klystron oscillator that is amplitude-modulated by a square wave applied to the reflector. This system has a number of inherent disadvantages; in particular, a rather elaborate regulated-power supply is required, and it is usually necessary to allow about a two-hour warm-up period before the frequency is stable. These disadvantages may be avoided by the use of a somewhat different system.

The cavity-stabilized reflex-klystron system, developed by Pound,[1] has very good frequency-stability, and it does not require an elaborate power supply. It is not convenient to amplitude modulate the klystron. It is possible, however, to use the stabilized klystron to generate a cw signal and then to modulate this signal by means of a crystal detector (Type 1N23B). In this way a very stable generator was assembled. The system requires only a one-minute warm-up period for most applications. There is a slow drift in amplitude for about five minutes, but after that time the stability is sufficient for very accurate measurements. It should be mentioned that the crystal modulator absorbs some power and that the maximum modulation is about 50 per cent. The available power is accordingly somewhat less than that of the first-mentioned system.

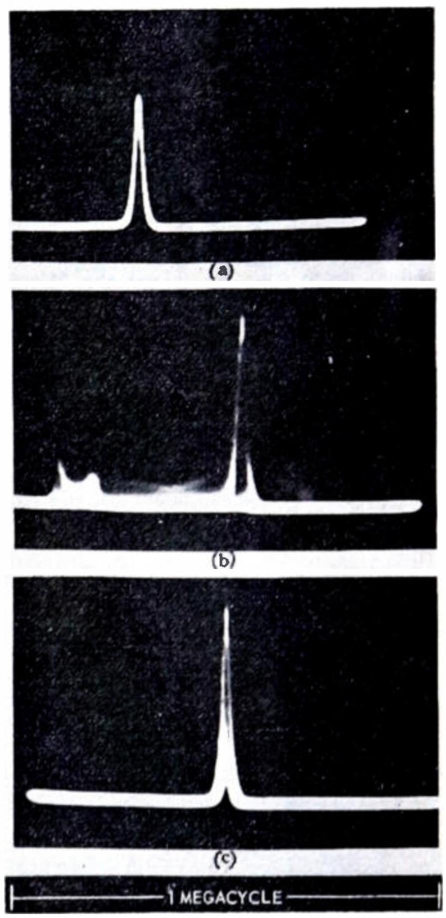

Fig. 1—Spectrograms showing the reduction of frequency pulling of a klystron 723/AB by use of a Pound stabilizer. (a) No crystal modulation; no stabilization. (b) 1,000-cps crystal modulation; no stabilization. (c) 1,000-cps crystal modulation; Pound stabilizer.

The modulated crystal tends to influence the frequency of the klystron oscillator. As a point of interest, some photographs were taken of the frequency spectrum with and without the stabilizing system. These photographs, with the pertinent data, are given in Fig. 1. It is clear from the data that the Pound stabilizer greatly reduces frequency-pulling.

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* Original manuscript received by the Institute, June 11, 1952. This work was done at Cruft Laboratory, Harvard University, Cambridge, Mass.

[1] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 35, pp. 1405–1415; 1947.

Contributors to Proceedings of the I.R.E.

For a photograph and biography of Dr. W. R. G. Baker, see page 99 of the January, 1952, issue of the Proceedings of the I.R.E.

Herbert J. Carlin (M'47–SM'50) was born in New York, N. Y. in 1917. He attended Columbia University, where he received the B.S. degree in 1938 and the M.S. degree in 1940. In 1947 he was awarded the D.E.E. degree from Brooklyn Polytechnic Institute.

H. J. Carlin

From 1940 to 1945 Dr. Carlin was associated with the Westinghouse Company as a design engineer in the power-system relay section of the meter division, and has written several papers on power-system protection. He joined the Microwave Research Institute of the Polytechnic Institute of Brooklyn in 1945, and has made contributions in the field of microwave networks and microwave power measurements. He holds the position of research supervisor, and lectures in the graduate school at the Institute.

Dr. Carlin is a member of the A.I.E.E., A.A.A.S., Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

Seymour B. Cohn (S'41–A'44–M'46–SM'51) was born in Stamford, Conn. on October 21, 1920. He received the B.E. degree in electrical engineering from Yale University in 1942; the M.S. degree in communication engineering in 1946, and the Ph.D. degree in engineering sciences and applied physics in 1948, both from Harvard University.

Seymour B. Cohn

From 1942 to 1945 Dr. Cohn was employed as a special research associate by the Radio Research Laboratory of Harvard University, also representing that Laboratory as a technical observer with the U. S. Army Air Force in the Mediterranean Theater of Operations.

Since March, 1948 Dr. Cohn has been employed by the Sperry Gyroscope Company, and now holds the position of research engineer in the Microwave Instruments and Components Department.

Dr. Cohn is a member of Tau Beta Pi and Sigma Xi, and is serving on the Papers Review Committee of the IRE.

John P. Costas (S'46–A'51) was born on September 16, 1923, in Wabash, Ind. He obtained from Purdue University the B.S. degree in electrical engineering in 1944. Two years were then spent in Naval Service as radar officer in which time Dr. Costas attended the Harvard and M.I.T. Radar Schools. He returned to Purdue and obtained the M.S. degree in electrical engineering in 1947. In 1951 he obtained the degree of D.Sc. from M.I.T.

John P. Costas

Dr. Costas is presently employed as a member of the Electronics Laboratory staff of the General Electric Company.

H. V. Cottony (M'45–SM'51) was born in Nizhni-Novgorod, Russia, on March 27, 1909. He received the B.S. degree in electrical engineering from Cooper Union Institute of Technology in 1932, the M.S. degree in electrical engineering from Columbia University's School of Engineering in 1933, and the E.E. degree from Cooper Union in 1946.

H. V. Cottony

From 1935 to 1937 Mr. Cottony was a research engineer for the Sonotone Corporation. From 1937 to 1941 and, again, from 1945 to the present he has been employed at the National Bureau of Standards as a physicist and radio engineer. His most recent work at the Bureau has involved radio-noise measurements and antenna studies.

During the period 1941 to 1945 Mr. Cottony was on military duty at the Office of the Chief Signal Officer and at the Signal Corps Laboratories, where he served successively as assistant officer-in-charge of the Aircraft Radio Section; project officer for radar AN/TPS-3; officer-in-charge of the Antenna and Mechanical Design Section; and chief of the Thermionics Branch.

Mr. Cottony is a member of AIEE, Tau Beta Pi, and Mu Alpha Omicron.

H. A. Hess (A'49) was born in Kirchheim-Teck, Wuerttemberg, Germany, on June 15, 1910. He attended the Stuttgart Technical University from 1930 to 1933 and the University of Jena from 1933 to 1935. The following two years he was a scientific assistant at the Heinrich Hertz Institute, Berlin, and received the Dr. Phil. Nat. degree from the Friedr. Schiller University of Jena in 1937.

H. A. Hess

During the period 1938 to 1940, Dr. Hess was employed in the research laboratories of the Telefunken Company, Berlin. In 1941 he became a technical assistant at the German Patent Office, Berlin, and in 1941 he was obliged to serve as a civilian employee of the Luftkriegsakademie, Berlin-Gatow. He did research in hf propagation in Denmark from 1942 to 1945. Since 1950, Dr. Hess has been employed at the Service de Prévision Ionosphérique Militaire (France) at Freiburg.

Philip N. Hess (S'50–A'51) was born February 18, 1926, in Minot, N. D.

After two years as a radio technician in the U. S. Navy during World War II Mr. Hess attended North Dakota State College, receiving the B.S. degree in electrical engineering in 1949. He then joined the staff of the electrical engineering college at Oklahoma Agricultural and Mechanical College as an instructor.

Philip N. Hess

During his work at Oklahoma A and M, Mr. Hess was associated with the tornado identification and tracking project and received the M.S. degree in 1950. He then became an instructor of electrical engineering at the University of Minnesota where he is currently engaged in research in the field of nonlinear mechanics.

Andrew L. Hopper (A'42–M'46–SM'51) was born on January 11, 1906 in Mahwah, N. J. He received the E.E. degree from Rensselaer Polytechnic Institute in 1928.

A. L. Hopper

Since that date Mr. Hopper has been engaged in various areas of communications research and development, mostly for Bell Telephone Laboratories. This work has been in such a variety of fields as machine switching, telegraph, radar, proximity fuses and, more recently, the transcontinental microwave radio-relay system. At present he is engaged in microwave repeater research.

Contributors to Proceedings of the I.R.E.

David A. Huffman (S'44–A'47) was born in Alliance, Ohio, on August 9, 1925. He received the B.E.E. and M.Sc. in E.E. degrees from the Ohio State University in 1944 and 1949, respectively.


David A. Huffman

From 1944 to 1946 Mr. Huffman served as a radar maintenance officer aboard the destroyer U.S.S. Duncan. From 1947 to 1950 he was an instructor in the department of electrical engineering at the Ohio State University, during which time he was also in charge of a classified electronics project with the O.S.U. Research Foundation. In 1950 he was associated with an Air Navigation Development Board psychological planning group, and in 1951 with the Physical Science Laboratory, State College, New Mexico.

At present Mr. Huffman holds the International Business Machines fellowship for work in automatic control systems at the Massachusetts Institute of Technology. He is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, Pi Mu Epsilon, and Sigma Pi Sigma.

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Joseph F. Hull (M'50) was born in Montello, Wis. on August 25, 1921. He received the B.S. degree in electrical engineering from the University of Wisconsin in June, 1943. He joined the U. S. Army Enlisted Reserve Corps in 1942, but was placed on inactive status during the war in order that he might carry on research at the General Electric Research Laboratory under the sponsorship of the Office of Scientific Research and Development. From 1943 to 1945, he worked on the development of high-power continuous-wave magnetrons for radar countermeasures at the General Electric Co.


Joseph F. Hull

In 1945, Mr. Hull was activated by the Army, and assigned to the thermionics branch of the Signal Corps Engineering Laboratories, Fort Monmouth, N. J. to carry on research in the field of microwaves. Since 1946 he has been employed as a civilian research engineer by the Signal Corps.

Mr. Hull received the M.S. degree in electrical engineering from Rutgers University in May, 1951. Concurrent with his work at the Signal Corps, he is presently engaged in graduate study at Polytechnic Institute of Brooklyn, N. Y.

Mr. Hull is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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J. R. Johler (A'47) was born in Scranton, Pa., on February 23, 1919. He received the B.A. degree from the American University in 1941 and the B.S. degree in engineering from the George Washington University in 1950, his major fields of study being physics and electrical engineering. He has also attended the Graduate School of the National Bureau of Standards.


J. R. Johler

From 1942 to 1944 Mr. Johler worked in the field of ballistic research at the Aberdeen Proving Ground, Aberdeen, Md. While on active duty in the Navy from 1944 to 1946 he attended the Radio Materiel School of the Naval Research Laboratory in Washington, D. C.

Since 1946 Mr. Johler has been employed by the National Bureau of Standards, Central Radio Propagation Laboratory. He has been principally concerned with instrumentation and measurement of radio wave-propagation parameters. At present he is assistant in charge of the Radio Noise Studies program, devoting special attention to the measurement of the absolute intensity of atmospheric radio noise on a world-wide basis.

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Herbert L. Jones (A'37–SM'46) was born on December 2, 1904, in Copperton, N. M. He received the B.A. degree in physics and mathematics at the University of Oregon in 1926, and the M.S. and Ph.D. degrees in physics, mathematics, and electrical engineering at Oregon State College, in 1934 and 1935, respectively.


Herbert L. Jones

Dr. Jones worked as telephone engineer for Pacific Telephone and Telegraph, in Portland, Ore., from 1926–1929, and as radio telephone engineer at the Bell Telephone Laboratories from 1929–1932. After receiving the Ph.D. degree, he taught electrical engineering at the University of New Mexico for ten years and has since been professor of electrical engineering at Oklahoma A & M College, Stillwater, Okla., in charge of graduate studies in communications. His study of the electrical characteristics of tornadoes has been in progress since 1947.

Dr. Jones is an active member of the National Society of Professional Engineers, and a member of Phi Beta Kappa, Eta Kappa Nu, Phi Kappa Phi, and Sigma Tau.

For a photograph and biography of Ronald King, see page 997 of the August, 1952, issue of the Proceedings of the I.R.E.

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A. H. LaGrone (M'48–SM'51) was born in Panola County, Texas on September 25, 1912. He received the B.S. degree in electrical engineering from the University of Texas in 1938. After four years as distribution engineer with the San Antonio Public Service Company, he was commissioned in the U. S. Naval Reserve and ordered to active duty in 1942. During this time Mr. LaGrone was instructor in radar at the Massachusetts Institute of Technology and later radar officer aboard the U.S.S. *Gillette*, D.E. 681, in the Atlantic.


A. H. LaGrone

At the conclusion of World War II Mr. LaGrone, then a lieutenant commander, was ordered to inactive duty and accepted the position of radio engineer with the Electrical Engineering Research Laboratory, the University of Texas. Mr. LaGrone was recipient in 1948 of the M.S. degree in electrical engineering from this university.

Mr. LaGrone is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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Alan A. Meyerhoff (S'46–A'48) was born in Baltimore, Md. on March 20, 1926. He pursued his professional education at Rutgers University where he received the degrees of B.S. in 1946 and M.S. in 1947.


A. A. Meyerhoff

From 1947 to 1948 he was employed by Philco Corporation in its research division where he worked on audio recording systems and crystal rectifier circuits. Since then, Mr. Meyerhoff has been associated with Coles Signal Laboratory of the Signal Corps Engineering Laboratories, where he is concerned with the development of radio relay sets and with the investigation of special problems associated therewith.

Mr. Meyerhoff is a member of Tau Beta Pi and Phi Beta Kappa.

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For a photograph and biography of Dr. Kenneth S. Miller, see page 998 of the August, 1952, issue of the Proceedings of the I.R.E.

Contributors to Proceedings of the I.R.E.

Stewart E. Miller (M'46) was born in Milwaukee, Wis., in 1918. He attended the University of Wisconsin, transferring to Massachusetts Institute of Technology to study communications engineering under a Bell System cooperative plan. He received the B.S. and M.S. degrees in electrical engineering in 1941.

S. E. MILLER

Mr. Miller then joined Bell Telephone Labs, engaging in design of centimeter-wave radar transmitter receivers and development of repeaters for the coaxial-cable carrier system from 1942 to 1949. He has since been in charge of a research group investigating communication possibilities of microwave guided-wave systems.

Mr. Miller is a member of Tau Beta Pi, and Eta Kappa Nu and an associate member of Sigma Xi.

❖

William W. Mumford (A'30-SM'46-F'52) was born on June 17, 1905, in Vancouver, Wash. He received the A.B. degree in math and physics from Willamette University, Salem, Ore., in 1930.

W. W. MUMFORD

Mr. Mumford was a radio operator in the U. S. Coast Guard in 1923. He joined the Western Union Telegraph Co. in 1924, becoming manager-operator at South Bend, Wash. in 1925. He was associated with the Oregon State Highway Testing Laboratory in Salem as a laboratory technician from 1928 to 1930. Since 1930 he has been on the technical staff of Bell Telephone Laboratories at Holmdel, N. J., where he has been engaged in high-frequency propagation and the development of microwave components.

Mr. Mumford is secretary of the administrative committee of the Professional Group on Microwave Electronics.

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J. R. Pierce (S'35-A'38-SM'46-F'48) was born in Des Moines, Iowa on March 27, 1910. He received the B.S. degree in electrical engineering in 1933 and the Ph.D. in 1936 from the California Institute of Technology.

J. R. PIERCE

Since 1936 Dr. Pierce has been at Bell Telephone Labs, working largely on vacuum tubes. He is now director of electronics research.

Dr. Pierce received the IRE Fellow award in 1948, the Eta Kappa Nu "Outstanding Young Electrical Engineer" award for 1942, and the IRE Morris Liebman Memorial Prize for 1947.

Dr. Pierce is a member of Tau Beta Pi and the British Interplanetary Society, and a fellow of A.P.S.

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R. H. Rhéaume (M'45) was born on August 30, 1909 in Stamford, Conn. He received the M.E. degree from Stevens Institute of Technology in 1930 and the M.S. degree in electrical engineering from Columbia University in 1941. He is a licensed professional engineer in the states of New York and Connecticut, a member of the National Society of Professional Engineers, the A.I.E.E., and Tau Beta Pi.

R. H. RHÉAUME

Mr. Rhéaume was with Bell Telephone Laboratories as an acoustical research engineer from 1930 to 1932, the Western Electric Co. (Kearny Works) as a manufacturing engineer from 1940 to 1945, and with Machlett Laboratories, as a design engineer to 1951. He is now executive engineer for the Hanovia Chemical and Manufacturing Co.

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Alan J. Simmons (A'47) was born in New York City on October 14, 1924. He received the B.S. degree in physics and chemistry from Harvard University in 1945 and the M.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1948. While in the U. S. Navy in 1944-1945, he attended the radar schools at Harvard and M.I.T.

A. J. SIMMONS

From 1946 to 1948 Mr. Simmons was a research assistant at the Research Laboratory for Electronics, M.I.T., and since that time has been working on microwave antennas and allied problems in the Antenna Research Branch at the Naval Research Laboratory, Washington, D. C.

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Max Sucher was born in Poland in 1913. He received the B.S. degree in physics from Brooklyn College in 1933 and the M.S. degree in physics from Brooklyn Polytechnic Institute in 1947.

From 1936 to 1938 Mr. Sucher was with the National Bureau of Standards and from 1940 to 1946 with the Bureau of Ships of the Navy Department. From 1946 to 1947 he served as a research fellow in the physics department of Brooklyn Polytechnic Institute, joining the staff of the Polytechnic Research and Development Company in 1947. Since 1950 he has been with the Microwave Research Institute of Brooklyn Polytechnic Institute as a research associate.

MAX SUCHER

Mr. Sucher is a member of the American Physical Society and Sigma XI.

❖

A. W. Warner (M'52) was born in Sewickley, Pa. in 1915. He received the B.A. degree, with a major in physics, from the University of Delaware in 1940 and the M.S. degree in physics from the University of Maryland in 1942.

A. W. WARNER

In the same year Mr. Warner was a member of the faculty of Lehigh University, leaving in July to join the Western Electric Company, where he worked on the development of crystal-unit test equipment. In 1943 Mr. Warner became a member of the technical staff of Bell Telephone Laboratories, where he has been engaged in the design of high-frequency plated crystal units.

Mr. Warner is a member of the Acoustical Society of America.

❖

Lotfi A. Zadeh (S'45-A'47-M'50) was born on February 4, 1921, in Baku, Russia. He attended Alborz College of Teheran, and received the B.S. degree in electrical engineering from the University of Teheran in 1942. He worked for a year as a technical contractor with the U. S. Army Forces in Iran, and came to the United States in 1944. He resumed his studies, receiving the M.S. degree from the Massachusetts Institute of Technology in 1946, and the Ph.D. degree from Columbia University in 1949. In 1946 he joined the staff of Columbia University, where he is now assistant professor of electrical engineering.

LOTFI A. ZADEH

Dr. Zadeh is an associate member of the American I.E.E., and a member of the American Physical Society, the American Mathematical Society, and Sigma Xi.

Institute News and Radio Notes

National Electronics Conference Set

The eighth annual National Electronics Conference is scheduled for the Sherman Hotel, Chicago, Ill., September 29–October 1.

Nearly 100 papers will cover a broad field of electronic research, development, and industrial application, supplemented by over 75 exhibits by manufacturers and institutions foremost in the electronics fields.

Highlighting the social program will be three luncheons featuring prominent speakers, an evening banquet, and a full three-day program for the ladies.

The conference is sponsored by the American Institute of Electrical Engineers, Illinois Institute of Technology, Institute of Radio Engineers, Northwestern University, University of Illinois, with Purdue University, University of Wisconsin, and the Society of Motion Picture and Television Engineers participating.

Advance registration may be made to National Electronics Conference, Inc., Karl Kramer, Executive Secretary, 852 East 83rd St., Chicago 19, Ill.

HF Measurements Conference Slated

Under the joint sponsorship of IRE, AIEE, and the National Bureau of Standards, the Third Conference on High-Frequency Measurements will be held on January 14–16, 1953, in Washington, D. C

The Conference will follow the pattern of similar meetings held in 1949 and 1951, and will be devoted exclusively to the techniques and problems of high-frequency measurements, with emphasis on new developments.

Annual Physics Prize Established

The Bell Telephone Laboratories and the American Physical Society have established the Oliver E. Buckley Solid State Physics Prize. The prize consists of an annual award of $1,000 to be given to the person adjudged to have made an important contribution to the advancement of knowledge in solid state physics within the five years immediately preceding the award.

The prize, endowed by a trust fund of $50,000, is provided by Bell Telephone Laboratories, and is named in honor of the Laboratories' former President and Board Chairman who retired September 1, 1952.

The prize, to be administered by the American Physical Society, will be available for each calendar year; the Society may delay an award to a subsequent year. A total of 25 prizes will be awarded during the 25-year life of the trust, and in 1978, the remaining funds are to be turned over to the American Physical Society for its uses and purposes.

The five committee members of the Society to select the first winner are: Harvey Brooks, Harvard University; J. B. Fisk, Bell Telephone Laboratories; J. C. Slater, Massachusetts Institute of Technology; C. S. Smith, University of Chicago; and J. H Van Vleck, Harvard University.

Calendar of COMING EVENTS

Annual Meeting of the Instrument Society of America, Cleveland, Ohio, September 8–12

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

Annual Meeting of the Optical Society of America, Hotel Statler, Boston, Mass., October 9–11

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

IAS-IRE-RTCA-ION Symposium on Electronics in Aviation, New York, N. Y., January 26–30

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

9th Joint Conference of RTMA of United States and Canada, Ambassador Hotel, Los Angeles, Calif., April 16–17

1953 National Conference on Airborne Electronics, Dayton, Ohio, May 11–13

Technical Committee Notes

The **Standards** Committee met on June 12, with Ernst Weber as acting chairman in the absence of A. G. Jensen. The Committee further considered the proposed Standards on Receivers: Definitions of Terms. The next item considered was a list of Radio Astronomy Definitions (51 IRE 24. PS1) proposed by the Wave Propagation Committee. Professor Weber informed those present that C. R. Burrows of the Wave Propagation Committee was a delegate to the URSI Australia meeting and planned to take the definitions to Australia with a view to correlating them with attending international representatives. Professor Weber mentioned the importance of the Standards Committee in letting Dr. Burrows know whether the definitions were reasonably acceptable or of fundamental differences of opinion regarding them. It was moved that the proposed definitions on Radio Astronomy be referred to the Wave Propagation Committee with a summary of the remarks made at the meeting. It was moved also that a letter be sent to Dr. Burrows with a summary of the Standards Committee remarks, pointing out that the comments in no way constitute approval of the definitions and that the list is being referred to the Wave Propagation Committee for further consideration. The Committee next considered a request from the American Standards Association for reaffirmation or revision of eight American Standards. Each standard was discussed and voted upon. The next item discussed was a letter ballot of ASA C16 covering consideration of eleven RTMA Standards. After the comments received from various members of the Standards Committee, it was moved that the list be approved with the exception of RTMA Standard TR-106 and TR-112, and that an affirmative vote be recorded for the new RTMA Standard TR-112-A, not included in the list. It was also decided that the IRE Representatives on C16 be instructed on how to vote on these standards. Chairman Weber asked for consideration of a letter from P. S. Christaldi suggesting that all members of IRE committees and subcommittees be supplied automatically with a copy of each new standard as it is issued. It was agreed that it would be sufficient to send one copy to each member of the Standards Committee and to each subcommittee chairman, with a form attached by means of which a request could be made for additional copies. R. R. Batcher, Chairman of the Annual Review Committee, announced that he hoped to have a statement to make at the next meeting of the Standards Committee on plans for the current year's Annual Review.

The **Audio Techniques** Committee met on June 26, under the Chairmanship of C. A. Cady. H. W. Augustadt pointed out that the Audio Techniques definitions to be considered had been under discussion for some time and although there had been preliminary discussions on the terms with the coordinator and his committee, the list in its original form (dated September 24, 1951) had never been presented as a proposed standard. There followed a discussion on the conflict

SECOND 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.

(Technical Committee Notes Cont.)

with certain definitions and those already published as standards, particularly in connection with the transducer and receiver groups. R. A. Miller confirmed that the list now under consideration did not include terms covered by the current Transducer Standards. It was the consensus of opinion that every effort should be made to clear the list for presentation to the Standards Committee before further conflicts arose with related groups. W. L. Black moved that each term be considered and voted upon separately. The motion was seconded. Many of the terms now under review are closely related to ASA C16.5-1942. Accordingly, these terms should be reviewed concurrently with this Standard at the next meeting. Mr. Augustadt was requested to begin preparation of a Master Index of Audio Definitions, including those terms which have been previously accepted through co-ordination with the Transducer Task Group.

Under the Chairmanship of F. J. Gaffney, the **Measurements and Instrumentation** Committee met on June 26, 1952. Chairman Gaffney reported that H. E. Dinger had agreed to accept the chairmanship of the Subcommittee on Interference Measurements. Mr. Dinger has indicated by letter that he is actively working on the formation of a Subcommittee and has already received three acceptances from well-qualified individuals in the field. The Chairman reported on the resignation of E. I. Green as Chairman of the Subcommittee on High-Frequency Measurements, and on the appointment of R. W. Lowman of Airborne Instruments Laboratory, as Chairman. This Subcommittee also includes C. J. Franks and B. M. Oliver. C. D. Owens, reporting for Subcommittee 25.3, on Magnetic Measurements, indicated that he had been in contact with the RTMA Subcommittee 21 C1 on High-Frequency Cores, and with the Subcommittee on Electronic Cores of the Metal Powders Association. Both of these groups will welcome IRE assistance in the matter of basic definitions and methods of measurement. Mr. Owens' Subcommittee has started a program to supply this need and has made some new appointments to the Subcommittee. This Subcommittee will generate definitions for such basic quantities as permeability, Q, and so on, for high-frequency cores and will later study the problem of methods of measurement of these characteristics. Reports were submitted by G. L. Fredendall, Chairman of Subcommittee 25.5; Arnold Peterson, Chairman of Subcommittee 25.4; and P. S. Christaldi of Subcommittee 25.10. Dr. Christaldi presented to the Committee for initial consideration and comment a number of definitions in the field of cathode-ray oscillography. Dr. Christaldi strongly urged that the Committee members review the proposed definitions and comment to him at their earliest convenience.

On June 19, the **Television Systems** Committee met under the Chairmanship of R. E. Shelby. The Committee scope was reviewed and revised. The next item was a report by M. W. Baldwin, Jr., of the Subcommittee on Color Television Definitions. Mr. Baldwin stated that this Subcommittee had been appointed by A. G. Jensen. The present membership was given. Mr. Baldwin submitted a list of proposed television definitions (52 IRE 22. PS1), dated March 19, 1952, stating that this present listing incorporates modifications made after the Subcommittee received comments from several Committee members, and when the first draft was circulated with letter ballot. The Subcommittee on Definitions asked its Chairman to recommend to the Television Systems Committee that the Standards Committee be requested to publish all previous color television definitions each time that new terms are approved for publication. There are currently ten to twenty more definitions being considered, and others may be added. It was moved and seconded that the question of republishing definitions, now before the Television Systems Committee, when a subsequent list is submitted, be referred to the Subcommittee on Definitions for recommendation. It was decided unanimously that the list of 43 definitions (52 IRE 22. PS1—Proposed Television Definitions—3/19/52) be approved and forwarded to the Standards Committee. An Ad Hoc Committee has been organized to define the term "picture element." The committee will undertake the task of preparing the interim definition and the study leading to a comprehensive definition. It was recognized by the Television Systems Committee that more than one definition may emerge. A request was made for color television to be considered in drawing the definition and that the desirability of a relationship to a resolution chart be kept in mind.

Under the Chairmanship of W. J. Poch, the **Video Techniques** Committee met on June 3. Activities of the Subcommittee on Video Systems and Components were discussed. G. L. Fredendall reported on progress made at the last meeting. Several definitions approved at the last meeting of the Standards Committee were discussed.

Over 1,000 Attend Southwestern Conference

An over-all success was acclaimed by the Fourth Southwestern IRE Conference and Radio Engineering Show, held May 16–17, 1952, at the Rice Hotel, Houston, Tex.

With over 1,000 individuals registering for the conference, crowded technical sessions, and the witnessing of one of the finest groups of exhibits, the Houston IRE Section made a record stride as sponsor and host to the conference.

W. C. Copp, advertising manager of the PROCEEDINGS, is shown at the IRE exhibit of the Radio Engineering Show.

General Conference Chairman G. K. Miller is shown opening the Conference banquet program. At the right is Donald B. Sinclair, IRE President.

The program was highlighted with the opening address presented by Donald B. Sinclair, IRE President, and the banquet address given by Commander T. A. M. Craven, prominent consulting engineer of Washington, D. C.

Notice to Authors

It is planned to publish a special issue of the **Proceedings** devoted primarily or entirely to the subject of Ultra-High Frequencies. This issue is tentatively scheduled to appear in January, 1953.

UHF topics to be covered will include tubes, transmitters, receivers, antennas, studio equipment, and propagation data.

Authors wishing to submit material for consideration for publication in the UHF issue must forward their papers to the Institute by October 1, 1952.

Professional Group News

Audio

After sponsoring two interesting sessions at the IRE Western Convention, the Audio Group is now planning its sessions at the National Electronics Symposium, Chicago, Ill., to be held September 29–October 1.

The TRANSACTIONS PGA-8 has been mailed to members of the Group.

Antennas and Propagation

The 268-page PGAP-3 TRANSACTIONS has been mailed to the Antennas and Propagation Group members. The Group's Institutional Listings Committee has been successful in obtaining ten advertising listings (at twenty-five dollars each) for the recently published TRANSACTIONS.

Airborne Electronics

The members of the Airborne Electronics Group were polled recently to determine the opinion on the enlargement of the Group's scope and a possible change in the Group's title. The results will be announced at an early date.

Other activities include the mailing of the August NEWSLETTER to the Group members, and scheduling of the 1953 National Conference on Airborne Electronics for May 11–13, Dayton, Ohio.

Broadcast and Television Receivers

Technical sessions will be sponsored by the Broadcast and Television Receivers Group at the IRE/RTMA Radio Fall Meeting, October 20–22, in Syracuse, N. Y. During this time, the Annual Meeting of the Administrative Committee of the Group will be held.

PGBTR-1 TRANSACTIONS has been mailed to members of the Group.

Broadcast Transmission Systems

Four technical papers and a panel discussion were sponsored by the Broadcast Transmission Systems Group at the IRE Western Convention in Long Beach, Calif. The Group now is planning for the Annual Broadcast Symposium in Philadelphia at the Franklin Institute early this fall.

Electron Devices

The first NEWSLETTER and a notice of a two-dollar assessment for publications have been sent to members of the Electron Devices Group.

The Group is making plans now toward the IRE/RTMA Radio Fall Meeting in Syracuse, October 20–22.

Electronic Computers

Members of the Electronic Computers Group have been mailed the proceedings of the Electronic Computer Symposium held April 30–May 1, at the University of California, Los Angeles, Calif.

The Group plans to publish in a TRANSACTIONS the papers from the technical sessions held at the IRE Western Convention.

Engineering Management

The PGEM-1 NEWSLETTER has been sent to the members of the Engineering Management Group. A notice of a one-dollar assessment for future publications was included with the NEWSLETTER.

Information Theory

A tentative program has been planned for the Information Theory Symposium to be held in late October. The program includes: Tutorial review of information theory, Chairman, M. J. DiToro; Tutorial review of statistics, Chairman, M. Schwartz; Advances, Chairman, A. G. Clavier; Miscellaneous applications of information theory, Chairman, W. White; Applications to communication systems, Chairman, W. G. Tuller.

Instrumentation

Papers from the Group's two technical sessions at the IRE Western Convention will be published in a TRANSACTIONS of the Instrumentation Group.

The Group is now making plans for its participation in three forthcoming meetings: The Joint Meeting on High-Frequency Measurements, January 14–16, 1953, Washington, D. C.; the Joint Meeting on Radio Meteorology, Austin, Tex., Fall, 1953; and, the West Coast Symposium on Quality Components, May, 1953.

Microwave Theory and Techniques

Papers are being solicited for the Symposium on Microwave Circuitry to be held November 7, 1952, New York, N. Y. Material for the symposium should be sent to the Technical Program Chairman: A. C. Beck, Bell Telephone Labs., Red Bank, N. J.

Nuclear Science

Plans for the Nuclear Science Group's Annual Conference are being made. An announcement covering the conference will be released soon.

Quality Control

The PGQC-1 TRANSACTIONS has been mailed to members of the Group. At the same time, a notice of a two-dollar assessment was sent to the members for the TRANSACTIONS and for future publications.

Radio Telemetry and Remote Control

The Group participated in a two-day Joint IRE/AIEE Telemetering Conference, held at the Lafayette Hotel, Long Beach, Calif., August 26–27. The meeting was held in conjunction with the IRE Western Convention.

J. A. Doremus, chief engineer of the control and carrier division, Motorola, Inc., Chicago, Ill., has been appointed to serve on the Administrative Committee of the Radio Telemetry and Remote Control Group. He will serve out the unexpired term of W. R. Thurston who unavoidably has had to resign from the Committee.

Vehicular Communications

Among the recent activities of the Vehicular Communications Group are the amendment of the Constitution and By-Laws and plans for the Annual Fall Meeting.

The PGVC-2 TRANSACTIONS has been mailed to the Group members.

(*Continued on page 1132*)

Transactions of IRE Professional Groups

The following issues of TRANSACTIONS have recently been published by IRE Professional Groups and additional copies are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the prices listed below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-members*
Antennas and Propagation	PGAP-3; URSI-IRE Meeting, April, 1952 (268 pages)	$4.80	$7.20	$14.40
Audio	PGA-8; July, 1952 issue (40 pages)	.80	1.20	2.40
Broadcast and Television Receivers	PGBTR-1; Symposium on "UHF Television Receiver Considerations" (12 pages)	.50	.75	1.50
Quality Control	PGQC-1; papers presented at 1951 Radio Fall Meeting and 1952 IRE National Convention (60 pages)	1.20	1.80	3.60
Vehicular Communications	PGVC-2; Symposium on "What's New in Mobile Radio" (32 pages)	1.20	1.80	3.60

* Public libraries and colleges can purchase copies at IRE Member rates.

Professional Group News, (cont.)

Professional Group Chapters

The IRE Executive Committee has approved the ruling of the IRE Professional Groups Committee that Professional Group Chapters active on or before May 31, 1952, are exempted from the regulation of submitting a petition for the formation of chapters. The ruling was formed in order that the chapters may receive the ten-dollar rebate allowed for chapter meetings.

For the formation of a Professional Group Chapter, a new chapter or existing inactive chapters must follow the procedure of submitting a petition signed by ten members of the relative IRE Section.

Los Angeles Group Chapters

A report on the IRE Professional Group Chapters functioning actively with the Los Angeles IRE Section has been submitted. The following is a list of the individual Professional Group Chapters, their chairmen or representatives, and their activities.

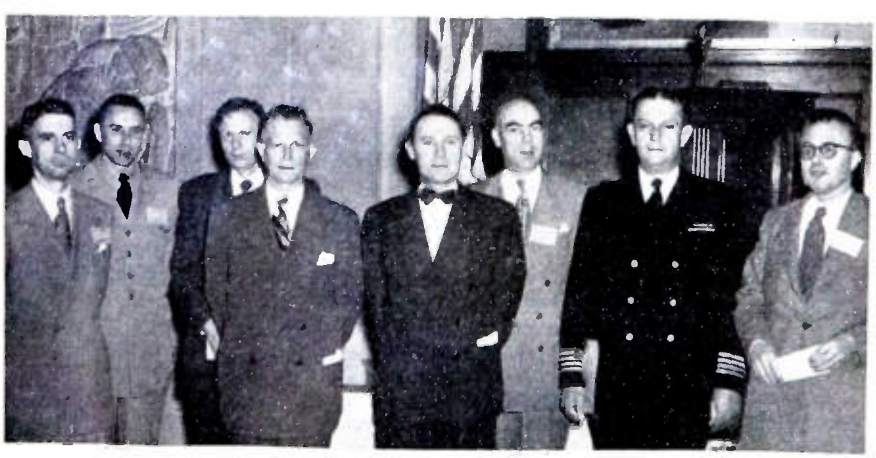
Airborne Electronics: Chairman, G. M. Greene. Meetings are planned for every two months with the scheduled program to include an introductory speaker and two technical papers. The chapter sponsored two sessions at the IRE Western Convention. **Antennas and Propagation**: Chairman, L. C. Van Atta. The chapter sponsored four sessions at the recent IRE Western Convention. **Audio**: Representative, J. K. Hilliard. An organization of periodic meetings will be arranged in the near future. The chapter sponsored two sessions at the IRE Western Convention. **Broadcast and Television Receivers**: Representative, H. E. Rice. Two sessions were sponsored by the Group chapter at the IRE Western Convention. **Broadcast Transmission Systems**: Chairman, P. G. Caldwell. The chapter holds monthly meetings and sponsored a session at the IRE Western Convention. **Circuit Theory**: Chairman, Louis Weinberg. The chapter holds bimonthly meetings and sponsored two sessions at the IRE Western Convention. **Electron Devices**: Representative, H. Q. North. Regular chapter meetings are being considered. Two sessions of the Group were held at the IRE Western Convention. **Electronic Computers**: Chairman, H. D. Huskey. The chapter holds monthly meetings and held two sessions at the IRE Western Convention. **Information Theory**: Chairman, R. B. Conn. Bimonthly meetings are held and one session was sponsored by the chapter at the IRE Western Convention. **Instrumentation**: Representative, W. D. Hershberger. Periodic chapter meetings are under consideration and the Group sponsored two sessions at the IRE Western Convention. **Radio Telemetry and Remote Control**: Representative, W. E. Lehan. Four sessions were held by the Group at the IRE Western Convention. **Vehicular Communications**: Chairman, Maurice Kennedy. The chapter plans to hold monthly meetings.

Chicago Group Chapters

There are ten active Professional Group Chapters in Chicago, Ill. One or more meetings have been held by each Group during the season, and 35 papers have been presented. The Chicago IRE Section's plan of coordination with Group Chapters is a shining example of Section initiative in encouraging Group Chapter activity and growth. The following is a list of the chairmen of the various Group Chapters Chicago Section.

Audio: R. E. Troxel, Shure Bros., Inc.; Antennas and Propagation: J. S. Brown, Andrew Corporation; Broadcast and Television Receivers: M. G. Beier, Zenith Radio Corp.; Broadcast Transmission Systems: E. W. Jacker, Station WAIT; Circuit Theory: A. A. Gerlach, Armour Research Foundation; Industrial Electronics: A. Crossley, Crossley and Associates; Instrumentation: J. F. Byrne, Motorola, Inc.; Nuclear Science: B. S. Schwartz, Manufacturer and Manufacturer's Representative; Quality Control: H. E. May, Motorola, Inc.; Vehicular Communications: R. V. Dondanville, Commonwealth Edison Company.

Components Symposium Speakers



Among the speakers at the Progress in Quality Electronic Components Conference, sponsored jointly by the IRE Instrumentation Group, AIEE, and RTMA, held in Washington, D. C., are: (left to right) J. G. Reid, Chairman of the Conference; Lt. Col. C. B. Lindstrand, USAF; A. V. Astin, Director, NBS; J. A. Milling, Chairman Electronics Production Board; Glen McDaniel, RTMA; E. A. Speakman, Research and Development Board; Capt. Rawson Bennett, USN Bureau of Ships; and, W. A. G. Dummer, Telecommunications Research Establishment, London, Eng.

NRL Changes Command

A change-of-command ceremony was held recently at the Naval Research Laboratory of the Office of Naval Research, Washington, D. C., to mark the shift in directors of the laboratory.

Captain Willis Henry Beltz became the director of NRL succeeding Captain Frederick Raymond Furth, who in turn relieved Captain Beltz as assistant chief of the Bureau of Ships for Electronics.

Serving in the United States Naval Reserve since 1925, Captain Beltz transferred to the regular Navy in 1946. He has served in various engineering capacities with RCA and with radio operations and the installment of various types of radio equipment for the Navy.

Captain Furth has served with the Navy since his graduation from the Naval Academy in 1924. He has been associated continuously with communications engineering as a radio officer and in the development and operational use of radio, radar, sonar, and other electronic equipment.

Radio Meteorology Conference Slated for 1953

The University of Texas will be host to a four-day conference on Radio Meteorology, scheduled for the second week in November, 1953, Austin, Tex. This meeting will be jointly sponsored by the American Meteorological Society, the Radar Weather Conference, the IRE Professional Group on Antennas and Propagation, National Commission II on Tropospheric Radio Propagation of URSI, and the Joint Commission on Radio Meteorology.

The sessions (which will not meet simultaneously) will include such topics as: tropospheric radio wave propagation mechanisms, radar storm detection and rainfall determination, cloud physics, turbulence, spherics, refractive index climatology and forecasting, and atmospheric reflections. Review papers submitted at the Group's invitation, will supplement individually submitted reports on specific research activities.

A call for papers will be issued during the first months of 1953 to permit the advance publication of 1,000 to 1,500-word abstracts. Further information may be obtained from any of the following members of the Steering Committee for the conference: L. G. Cumming, Institute of Radio Engineers, 1 East 79 St., New York, N. Y.; J. R. Gerhardt, University of Texas, Austin, Tex.; W. E. Gordon, Joint Commission Radio Meteorology, Cornell University, Ithaca, N. Y.; Martin Katzin (representing URSI National Commission II), Wave Propagation Research Branch, Naval Research Laboratory, Washington, D. C.; J. S. Marshall (representing the Radar Weather Conference), McGill University, Montreal, Canada; Newbern Smith (representing the IRE Professional Group on Antennas and Propagation), Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.; and, K. C. Spengler, American Meteorological Society, 3 Joy Street, Boston, Mass.

IRE/RTMA Radio Fall Meeting

HOTEL SYRACUSE, SYRACUSE, N. Y.—OCTOBER 20–22, 1952

Monday, 9:30 A.M., October 20

UHF SESSION

(Arranged by the IRE Professional Group on Broadcast and Television Receivers)

Chairman, L. M. Clement, Crosley Div., Avco Manufacturing Corp.

"Study of Noise Reduction by Feedback in Ultra-High Frequency Amplifiers," A. B. Glenn, General Electric Co.

"Connection of UHF and Color Adaptors to VHF Receivers," L. H. Horn, Underwriters' Laboratories, Inc.

"A UHF Grid Dip Meter," A. E. Hylas and Walter Tyminski, Allen B. DuMont Laboratories, Inc.

"UHF Tuning Devices," Norman Altman and Fred Barr, General Instrument Corp.

Monday, 2:00 P.M., October 20

ELECTRON TUBE QUALITY CONTROL SESSION

(Arranged by the IRE Professional Group on Quality Control)

Chairman, J. R. Steen, Sylvania Electric Products Inc.

"Performance Evaluation of Special Red Tubes," H. J. Prager, RCA.

"Development of Inspection Manual for Reliable Tubes," E. K. Wimpy, Hytron Radio and Electronics Co.

"Adaptation of Industry Proposals to Subminiature Tube Evaluation," A. J. Heitner, Sylvania Electric Products Inc.

"Latest Developments in Airinc Reliability Program," R. E. Moe, General Electric Co.

Monday, 8:00 P.M., October 20

MEETING OF SYRACUSE TECHNOLOGY CLUB AND SYRACUSE IRE SECTION

The program of this meeting will include several talks by representatives of the Kimball Glass Company.

Tuesday, 9:00 A.M., October 21

Symposium on NTSC Color Television Receiver Development

(Arranged by the IRE Professional Group on Broadcast and Television Receivers)

Chairman, D. B. Smith, Philco Corp.

"General Considerations in the Design of a Color Television Receiver," B. D. Loughlin and C. J. Hirsch, Hazeltine Electronics Corp.

"IF and Video Design Considerations as Applied to the Color Signal," B. S. Parmet, Motorola, Inc.

"Color Synchronization," W. E. Good, General Electric Co.

"Color Signal Demodulators," D. H. Pritchard and R. N. Rhodes, RCA.

"Color Phase Alternation," S. J. Klapman, Admiral Corp.

General Discussion of Preceding Papers

Tuesday, 2:00 P.M., October 21

COLOR TELEVISION SECTION

(Arranged by the IRE Professional Group on Broadcast and Television Receivers)

Chairman, E. W. Engstrom, RCA

"Principles of Colorimetry as Applied to Television," F. J. Bingley, Philco Corp.

"Colorimetric Analysis of Gamma Corrected Color Television Systems," D. C. Livingston, Sylvania Electric Products Inc.

"Instrumentation for Color Television Development," K. E. Farr, Westinghouse Electric Co.

Tuesday, 7:00 P.M., October 21

Fall Meeting Dinner

Toastmaster, J. W. McRae, Bell Telephone Laboratories, Inc.

Speaker and Subject to be announced later

Wednesday, 9:00 A.M., October 22

ELECTRONIC DEVICES SESSION

(Arranged by the IRE Professional Group on Electronic Devices)

Chairman, G. D. O'Neill, Sylvania Electric Products, Inc.

"Future Trends in Tube Design," R. R. Law, RCA.

"Mechanisms in Transistor Electronics," R. M. Ryder, Bell Telephone Laboratories, Inc.

"Operational Aspects of the 'Sleeping Sickness'," Irving Levy and F. M. Dukat, Raytheon Manufacturing Co.

"Properties of PNP Diffused Junction Transistors," J. S. Saby, General Electric Co.

"The Application of RCA Point-Contact Transistors," R. M. Cohen, RCA.

(Part of this Session may be extended into Wednesday evening)

Wednesday, 2:00 P.M., October 22

GENERAL TELEVISION SESSION

Chairman, A. V. Loughren, Hazeltine Electronics Corp.

"Problems of Television Interference," W. B. Smith, Canadian Department of Transport.

"AFC Circuit Design for Television," G. D. Doland, Philco Corp.

"Ninety-Degree Cathode-Ray Sweep System Consuming less than 'Fifty-Degree' Power," C. E. Torsch, General Electric Co.

"Design Considerations for Series Heater Strings in Television Receivers," M. B. Knight, RCA.

IRE People

John Milton Miller (A'17–F'20) has retired as deputy director of research of the Naval Research Laboratory, Office of Naval Research.

J. M. MILLER

A native of Hanover, Pa., Dr. Miller is a graduate of Yale University, having received the Ph.D. degree in physics in 1915.

From 1907–1919, Dr. Miller was a physicist with the National Bureau of Standards, and from 1919–1923, a radio engineer at the Radio Laboratory, Air Station, Navy Department, Anacostia, D. C. He then joined NRL as a radio engineer. During the period 1925–1936, he was in charge of radio receiver research at the Atwater Kent Manufacturing Company in Philadelphia, and from 1936–1940, he was assistant head of the research laboratory for the RCA Radiotron Company.

Returning to NRL in 1940, Dr. Miller subsequently became superintendent of Radio I Division in 1945, was named associate director of research in 1951, and then was appointed scientific research administrator in 1952. During this latter period he continued to act as superintendent of Radio I Division at the laboratory.

Dr. Miller has served as a patent expert with the government and has been issued more than 20 patents of his own in the radio field. His inventions include fundamental circuits for quartz crystal oscillators. He collaborated in the perfecting of crystals cut to have zero temperature coefficient, and the designing of the first high-powered crystal-controlled radio transmitter.

Dr. Miller was awarded the Distinguished Civilian Service Award in 1945 for "initiation of the development of a new flexible radio-frequency cable urgently needed in radio and radar equipment which solved a desperate material shortage in the United States during World War II, a well-deserved honor."

IRE People

Harold Goldberg (A'38–M'44–SM'44) has been appointed chief of the ordnance development Program C, Division 17, of the National Bureau of Standards. The ordnance program is concerned with research, development, and engineering of electronic ordnance devices.

H. Goldberg

Dr. Goldberg was born in Milwaukee, Wis., in 1914, and received the B.S. degree in electrical engineering in 1935, the M.S. degree in 1936, and the Ph.D. degree in 1937, from the University of Wisconsin. In 1941 he received a Ph.D. in physiology.

Dr. Goldberg served as senior engineer with the Stromberg-Carlson Company research department from 1941–1945 and as principal research engineer for Bendix Radio Division, Bendix Aviation Corporation, on problems of microwave research and development from 1945–1947. He joined the NBS as chief of the ordnance research section, ordnance development division and was appointed assistant chief of Division 13 in 1950. In 1951 he became chief of branch B of the Division.

Dr. Goldberg has been granted ten patents in the electronics field and some 50 others are pending. He is a member of the American Institute of Electrical Engineers, the American Physical Society, the American Association for the Advancement of Science, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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Lew H. Morse (A'46–M'49) has been appointed sales engineer of the Cyde H. Schryver Sales Company.

L. H. Morse

Mr. Morse was born in Methuen, Mass., on March 24, 1917. After attending Oberlin College, he joined the Army Air Force in 1941, where he taught and wrote on navigational subjects. He was later assigned to the Fifth Emergency Rescue Squadron at Biloxi, Miss.

Upon discharge from the Air Force, Mr. Morse joined the advertising department of the Aireon Corporation as a technical writer. Subsequent connections with Collins Radio Company, Central Radio and Television School, and Bendix Aviation Corporation have all involved working in technical writing, editing, and advertising.

Mr. Morse is the editor of "The Local Oscillator," a publication of the Kansas City IRE Section.

Samuel Heller (A'44) has been appointed chief engineer of the American Rectifier Corporation, New York, N. Y.

Mr. Heller received the E.E. degree in 1935 from Cooper Union College, and is the author of several books. He is an associate member of the American Institute of Electrical Engineers.

❖

Thomas B. Moseley (A'42–M'46) has been appointed sales representative of the television transmitter division of the Allen B. DuMont Laboratories, Inc. He will represent the company at the Southwest headquarters in Dallas, Tex.

Prior to his recent appointment, Mr. Moseley was the director of control orders, sales and service department of Collins Radio. He has been a representative for the long lines department for the American Telephone and Telegraph, and has been associated with the Mutual Broadcasting System. Mr. Moseley is a native of Del Rio, Texas.

Samuel S. Mackeown (M'29–F'40), professor of electrical engineering at the California Institute of Technology, Pasadena, Calif., died recently. He was 56 years old.

S. S. Mackeown

Dr. Mackeown was born in New York, N. Y., and received the B.A. and Ph.D. degrees from Cornell University in 1917 and 1923, respectively.

From 1917–1918, Dr. Mackeown was a laboratory assistant at the National Bureau of Standards, and from 1918–1919, he served as an officer with the United Signal Corps. Later, he was an assistant physicist at NBS doing research work on vacuum tubes, and then transferred to the Western Electric Company in New York, also to work on vacuum tubes.

From 1920–1923, Dr. Mackeown was a physics instructor at Cornell University and consequently became a National Research Fellow in physics at the California Institute of Technology. In 1926 he became an assistant professor of electrical engineering at the Institute conducting courses on vacuum tubes and radio communications. He remained with the Institute until his death.

Dr. Mackeown served on the IRE Board of Editors from 1941–1951, and was the IRE Representative at the California Institute of Technology from 1941–1950.

Alfred L. Green (A'28–F'38), died in Australia last year, according to a notice recently received at The Institute of Radio Engineers.

Dr. Green was born on February 3, 1905, in Brockley, London, Eng. He received the B.S. degree in 1926, the M.S. degree in 1929, the Ph.D. degree in 1934, and the D.Sc. degree in 1944, from King's College University, London.

Dr. Green's professional career included work as a mathematics instructor, and investigator to radio research boards and councils for scientific and industrial research in Great Britain and Australia.

At the time of his death, Dr. Green was officer in charge, Ionospheric Predication Service of the Commonwealth Observatory, Mount Stromlo, Canberra, A.C.T., Australia.

Everard M. Williams (S'36–A'41–SM'44) has been appointed head of the department of electrical engineering at Carnegie Institute of Technology.

E. M. Williams

Designer of the oscillator and deflector for the Carnegie Tech two-and-a-half million dollar synchro-cyclotron, Dr. Williams has been a member of the Carnegie faculty since 1945. He was made professor of electrical engineering in 1949

Dr. Williams was born in New Haven, Conn., in 1915, received the B.E.E. degree in 1936, and the Ph.D. degree in 1939 at Yale University, where he won the Yale Engineering Association Scholarship three times and a Charles A. Coffin Fellow for the year 1938–1939.

From 1939–1942, Dr. Williams served as an instructor in electrical engineering at Pennsylvania State College. During World War II, he served as branch engineer with the Development Branch Special Project Laboratory at Wright Field. He was presented the President's Certificate of Merit for "outstanding fidelity and meritorious conduct in aid of the war effort."

Dr. Williams is the author of technical engineering papers, is a licensed professional engineer in Pennsylvania, and serves as a consultant on electronic techniques for a number of large industrial firms. He is a member of Sigma Xi, Pi Mu Epsilon, Tau Beta Pi, Eta Kappa Nu, the American Institute of Electrical Engineers, and the American Society for Engineering Education. He is also a member of the Committee on Guided Missiles, Research and Development Board.

Dr. Williams served as Chairman for the IRE Pittsburgh Section in 1947–1948.

Books

Electrical Communications Experiments by Henry R. Reed, T. C. G. Wagner, and George F. Corcoran

Published (1952) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 451 pages+6-page index+viii pages. 228 figures. 5¼×9. $6.75.

Henry R. Reed, T. C. G. Wagner, and George F. Corcoran are members of the electrical engineering department, University of Maryland, Baltimore, Md.

The authors of this book have compiled a progressive series of experiments intended for undergraduate students primarily interested in principles of communications. The textbook is divided into four parts: dc fundamentals; ac fundamentals; engineering electronics; and radio engineering. Each part contains fifteen experiments, and the approximate frequency range is dc to 5 mc.

The introductory material presents a clear discussion of the uncertainties in electrical measurements and the treatment of experimental data. Various types of report formats are outlined, followed by the experiments which are methodically written, with emphasis on "apparatus," "discussion," "laboratory exercises," and "report suggestions." Under the heading of "discussions" the theories to be investigated are thoroughly presented, and the experimental procedure is outlined under "laboratory exercises." Finally, through "report suggestions," the data are graphed and quantitative comparisons made with the theories. Precautions are noted and methods compared wherever it is necessary.

Most of the experiments are well documented and the schematic diagrams are neatly presented. Particular attention is given to principles of operation rather than details of the apparatus itself. Only standard measuring equipment and components are utilized.

The authors have attempted to cover a broad field within the 451 pages; however, certain topics have been insufficiently treated. For example, very little mention is made of experimental techniques for the study of transients, filters, and pulsing and synchronizing circuits, although the interested reader will discover that the methods described by the authors can be readily extended to cover such topics. It is a very useful book and one that both teachers and students should find most helpful.

ANTHONY B. GIORDANO
Polytechnic Institute of Brooklyn
Brooklyn 2, N. Y.

Television Engineering by Donald G. Fink

Published (1952) by McGraw-Hill Book Company, Inc., 330 West 42 St., New York 36, N. Y. 690 pages +12-page appendix+19-page index+xiv pages. 512 figures. 6×9. $8.50.

Donald G. Fink is the editor of *Electronics*, McGraw-Hill Book Company, Inc., New York, N. Y.

During the past few years, the rapid growth of knowledge in the field of television engineering has created the need for an up-to-date reference on the subject. This new volume fulfils that need. It covers the basic theory and fundamental principles of television as seen in the light of recent technical advances, and places the latest-developments in transmitting and receiving techniques at the readers fingertips.

The text is divided into eleven main sections: the television system; analysis and synthesis of images; cameras and picture tubes; scanning and synchronization; transmission of the video signal; video amplification; carrier transmission of picture and sound signals; color fundamentals; color television systems; television broadcasting equipment; and television receiving equipment. In covering such an extensive subject as television engineering, the author necessarily has had to limit his discussion to the high points of most items. Nevertheless, the value of the book has been greatly enhanced by the inclusion of numerous footnotes and bibliographical references with each chapter.

The section on analysis and synthesis of images has been presented with special care. This is gratifying because it is this aspect of the subject that determines the maximum possibilities and limitations of television picture transmission and reproductions. Similarly, the review of color fundamentals is noteworthy, as it forms the background upon which color television systems are based. The sections on color systems gives current information on sequential and simultaneous methods, but will no doubt be expanded in future editions as the color television art develops.

At the end of the book is an appendix listing definitions and standards extracted from the FCC "Standards of Good Engineering Practice Concerning Television Broadcasting Stations." The figures referred to are included in earlier pages of the book, but for the reader's convenience, they should have been repeated in the appendix proper. In this reader's opinion, it would have been well worth while to have reproduced the entire FCC document, as the "Standards" have been found to be out of print.

This volume can be recommended as a textbook for the technical student as well as a valuable addition to the experienced engineer's library. In general, it is clearly written, copiously illustrated, and fills the need for a modern comprehensive reference in the broad field of television engineering.

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Mandl's Television Servicing by Matthew Mandl

Published (1952) by The Macmillan Company, 60 Fifth Ave., New York 11, N. Y. 413 pages+3-page index+3-page appendix+xiii pages. 294 figures. 5½×9. $5.50.

Matthew Mandl is the director of electronics and television courses at Temple University, Philadelphia, Pa.

This book, which is intended for the serviceman, could also be useful to the advanced student of radio, and to those training for television servicing. In this respect, the book fulfills the promise of its title, and is probably the most comprehensive and complete book of this type available today. Design engineers also may find some useful reading in this book.

The field of the serviceman is covered in eighteen chapters. The last chapter deals with color television, a subject presently curtailed by the National Production Authority. However, inclusion of this topic broadens the scope of the book.

The author begins by discussing receiver fundamentals and presents a tabulation of qualifications which a competent technician should possess. This is a worth while deviation since much previous training is required to understand the material. Parallels are drawn between servicing radio receivers and servicing television receivers, and a reasonably complete tabulation of common television troubles is included. The amount of uhf information at the end of the book should be sufficient to render to the reader some of the problems which will exist when uhf television comes into general use.

A useful feature of the book is the division of it into two parts, the larger of which discusses all types of receiver servicing; much information on projection systems is included. The smaller section, dealing with color television, uhf, and the use of test equipment, is particularly praiseworthy since skillful and intelligent use of equipment can speed up a servicing job. This book not only discusses certain operations to be performed but also tells how it should be done and how the equipment should be connected.

This first edition is up to date as much as possible, allowing for the ever present hiatus between the author's submission of a final manuscript and the appearance of the work in print. It is recommended reading for those who face problems in television servicing.

JOHN H. BATTISON
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1952 Edition of The Radio Amateurs Handbook

Published (1952) by the American Radio Relay League, West Hartford 7, Conn. 771 pages including catalogue section+13-page index. 1202 illustrations including charts and tables. 6½×9½. $3.00.

This 29th edition of a standard manual on radio, although an annual publication, presents a progressive coverage of the subjects. The sections on theory and design fundamentals have, in the present volume, been extensively rewritten and rearranged. The chapter on vacuum tube data with its comprehensive source on tube information, includes recently announced tube types. The treatment of the various subjects presented is clear and understandable.

Although the contained material is addressed to radio amateurs, the handbook's scope and circulation have expanded into numerous fields of practical radio. Text matter covering expansions of radio into engineering practice includes chapters on mobile radio and measurement equipment. Doubtless, there are traditional reasons for continuing the word "amateur" in the title of this useful manual, but it is significant that copies of the book are found in the book racks of many practicing and operating radio and television engineers.

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Books

Frequenzmodulation by Paul Guettinger

Published (1951) by Verlag Leemann, Zuerich, Switzerland. 172 pages+18-page bibliography+3-page index. 101 figures. Price: 29 Swiss francs.

Paul Guettinger is a research engineer, Brown Boveri, Ltd. Baden, Switzerland.

This book, first published in 1947 and reviewed in the June, 1948, issue of the PROCEEDINGS, was written to aid students as well as design engineers in understanding and applying both the theoretical and practical problems of frequency modulation. Although the book cannot cover all the detail problems of the field, the author has tried to form every chapter into a complete unit, making the book of value as a reference even to specialists. The results are shown with precise mathematical equations presented in a clear manner which can readily be applied. Care has been taken to keep all symbols uniform throughout the book.

The opening chapter, "General Theory of Frequency and Phase Modulation," explains the nature of frequency modulation followed with the difference and the relation between phase and frequency modulations; a method of determining the PM and FM from the spectrum is given. A second chapter on "Distortion" discusses effects of modulation curves, filters and circuits, multiple transmission paths, and demodulation on distortion. This is followed by a chapter on "Influence of Noise on FM," which treats noise, frequency modulated and pulse signals, as well as crossmodulation and effects of nonlinear discriminators. The next section on "FM Transmitters" gives the general design of FM and PM transmitters, and shows how reactance tubes work and ways of stabilizing the carrier frequency. The last chapter on "FM Receivers" shows a complete receiver diagram including all values of the circuit elements.

The various stages are discussed, such as, RF amplifier, mixer and oscillator, IF-stage, limiter, and discriminator. The material on discriminators is completely re-written giving the general discriminator equation for various Q's, the Foster-Seeley discriminator, the ratio detector, and the Philips discriminator.

A mathematical appendix which covers Bessel functions and complex integration is included, plus 405 references.

HANS K. JENNY
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Radar and Electronic Navigation by G. J. Sonnenberg

Published (1952) by D. Van Nostrand Company, Inc., 250 Fourth Avenue, New York 3, N.Y. 265 pages+7-page index+vii pages. 196 figures. 5½×8½. $6.00.

In a manner appropriate to the instruction of marine navigators, the author lucidly treats the principles, operational practice, and limitations of currently used electronic aids to ship navigation.

There is a chapter devoted to each of: decca, loran, consol, echo sounders, and radar. Direction finding, azimuth finding, and radio range, early electronic aids still of importance in navigation, are omitted. Air navigators will find no reference to the ICAO standard aids to navigation and landing omnirange, DME, ILS, etc.

In the opening chapter, fundamentals of electronics and the geometry of navigation systems are discussed in elementary terms Although some liberties have been taken with theory, the simple and necessarily sketchy treatment of radiation, resonance, cathode-ray tubes and their associated circuits, hyperbolic co-ordinate systems, and propagation is probably adequate for the navigator.

The chapters covering the actual navigation systems are in good detail. Principles of the shore transmitters used in decca, loran, and consol are touched briefly. The operation of modern shipboard electronic navigation equipment is described, and critical circuits of the equipment are explained. Also, pertinent pictures of commercial apparatus are included. Accuracies, ranges, limitations, and dangers associated with use of the various navigation aids are covered not only from the semitechnical standpoint, but are illustrated by quotations from navigator's reports and governmental advice to mariners.

The title of the book seems misleading, in that one infers much more comprehensive content than the book actually has. The book, intended expressly for the sea navigator, is of limited value to the air navigator. Because the book is somewhat nontechnical and gives no hint of future developments, and does not cover many important aids to navigation, the radio engineer will probably prefer to use other and more complete texts in this field.

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Radio Astronomy by Bernard Lovell and J. A. Clegg

Published (1951) by Chapman and Hall Ltd., 37 Essex St., London, W. C. 2, Eng. 227 pages+7-page index+4-page appendix. 120 figures. 5¾×7¾. 16s.

Bernard Lovell and J. A. Clegg are members of the staff of the University of Manchester, Eng.

In presenting the first complete book on radio astronomy, the authors have recognized the birth of this new science from the alliance of astronomy, astrophysics, physics, and electronics. Although experts in such fields will consider some or all of the first four introductory chapters to be somewhat elementary, the student of radio astronomy will find them necessary.

Some readers may be surprised by the range of subjects discussed in the category of "radio astronomy," which include meteor studies, aurora borealis, and solar flares with their terrestrial effects. Eight chapters represent a comprehensive review of the existing knowledge of meteor studies, and this may impress some readers as being out of balance with the amount of space devoted to the combined subjects of solar and galactic radio waves. Nevertheless, those who seek additional information in specific fields will be aided by the substantial references at the end of each chapter.

A book of this type serves as a true and useful function of collecting and presenting the current level of scientific achievement in its field. However, the extremely rapid growth of scientific achievement, especially in the fields of solar and galactic radio-frequency radiations, is rolling back the frontiers of this branch of science. Its advancement will require the serious student of radio astronomy to use the book of Lovell and Clegg as a stepping stone to the latest technical literature.

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Washington, D. C.

Transient Electric Currents by Hugh Hildreth Skilling

Published (1952) by McGraw-Hill Publishing Co., Inc., 330 W. 42 St., New York 36, N.Y. 356 pages+5-page index +vii pages. 105 figures. 6×9. $6.00.

Hugh Hildreth Skilling is professor of electrica engineering, Stanford University, Stanford, Calif.

This book, written especially as a text for college upper classmen, presupposes some knowledge of elementary calculus and a first course in physics.

The treatment is based mainly on the classical method of solution of differential equations. Only the simpler concepts are borrowed from the operational calculus.

The author proceeds systematically from the simplest circuits to more complicated networks: from circuits where the driver is a steady emf or where no emf supply is acting, to cases where a response to an alternating emf is sought. Chapters on coupled resonant circuits and circuits with variable parameters carry the general method of attack to more difficult cases Considerable attention is devoted to traveling waves on long lines and a study of reflection at the ends with various terminations.

Throughout the book, the author stresses the generality of the method of solution and shows how to check the work. His discussion of the physics of the phenomena, brought out by the mathematical results, will appeal to the teacher as well as the pupil. Problems solved in the text are chosen to cover a variety of cases and the problems which accompany each chapter furnish a searching test of the ability of the student to apply basic principles to new problems.

This second edition of the work differs principally from the previous one in the inclusion of a chapter on the Laplace Transformation. This is introduced as being an extension of ideas encountered in a study of Fourier series. Although the treatment is compressed into a compass of only 50 pages, the main concepts are developed and a number of problems are solved which have been solved earlier by the classical method.

In the opinion of this reviewer, the author has produced a textbook outstanding for clarity, consistency, and readability. A future textbook by the same author on the Laplace Transformation alone would be welcome.

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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 I.R.E.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 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

534.321.9 **2099**

The Measurement of Ultrasonic Attenuation in Solids by the Pulse Technique and some Results in Steel—R. L. Roderick and J. Truell. (*Jour. Appl. Phys.*, vol. 23, pp. 267–279; February, 1952.)

534.756 **2100**

Signal Translation in Hearing—L. O. Schott. (*Bell. Lab. Rec.*, vol. 30, pp. 2–8; January, 1952.) Account of investigations of the process by which sound waves are converted to electrical pulses in the ear, using a circuit analogue for the hair-cell pulse generators.

534.771 **2101**

Audiometer Measurement of Hearing Loss—F. J. Meister. (*Arch. Tech. Messen*, no. 192, pp. 15–18; January, 1952.) Two audiometers are briefly described, one, of the heterodyne oscillator type, providing a continuous frequency range from 20 cps to 20 kc, and the other, a RC generator, providing 10 frequencies between 64 and 11,584 cps. Practical details are discussed.

534.839:[621-1+629.13 **2102**

Sonic and Ultrasonic Noise in Engineering and in Aviation—P. Bugard, M. Guennec and J. Selz. (*Ann. Télécommun.*, vol. 7, pp. 47–55; January, 1952.) Report of measurements made with a piezoelectric microphone sensitive up to 80 kc, (*a*) using a frequency analyzer of range 1–100 kc and a pen recorder, (*b*) using a recording cro.

534.843:534.7 **2103**

Room Acoustics from the Point of View of the Singer and Speaker—R. Husson. (*Ann. Télécommun.*, vol. 7, pp. 58–74; February, 1952.) Analysis of physiological processes of articulation and the effects of room acoustics on the vocal system. A short reverberation time suitable for an audience may cause rapid fatique in the performer.

534.846 **2104**

Acoustical Researches in the Municipal Theatre of Budapest—T. H. Tarnóczy. (*Acta. Tech. Acad. Sci. Hungaricae*, vol. 2, nos. 2–4, pp. 285–301; 1952. In English.) Report of investigations of reverberation, echo paths, and attenuation at different locations, with a view to structural modifications to improve the acoustical characteristics of the theater.

534.85/.86:534.32 **2105**

Critical Study of the Concept of High Fidelity in the Response of Electroacoustic Systems—A. Moles. (*Onde élect.*, vol. 32, pp. 11–25; January, 1952.) Three criteria of acoustic quality are examined: (*a*) the minimum change in intensity perceptible by the ear at different frequencies and sound levels; this leads to a definition of an acoustic information unit; (*b*) the inability of the ear to detect a change in quality between two positions not far apart in a good concert hall; (*c*) the maximum variation that can be introduced unnoticed by an audience in the reproduction of a piece of music heard normally just previously. Experimental investigations of (*b*) and (*c*) are described. In practice a system has "perfect" fidelity if its response characteristic is level to within ±2 db over the af range.

621.317.7.018.78.029.4 **2106**

A Distortion Analyzer for the Audio-Frequency Range—G. Hoffmann. (*Fernmeldetech. Z.*, vol. 5, pp. 31–38; January, 1952.) An explanation of the basic principles of analyzers suitable for measurement of all harmonics, combination tones, etc., within a range of about 30 cps–30 kc, and an outline description of a possible type of equipment.

621.395.61/.62 **2107**

Coupled Electroacoustic Transducers—F. A. Fischer. (*Arch. elekt. Übertragung*, vol. 6, pp. 35–36; January, 1952.) Correction to paper noted in 571 of April.

621.395.623.7 **2108**

A Step towards the High-Fidelity Reproducer: the Focusing Baffle—P. Forestier. (*TSF et TV*, vol. 28, pp. 66–68; February, 1952.) Description of a commercially available loudspeaker unit comprising a baffle formed of part of an ellipsoid, effective for the middle and upper frequencies, together with a spherical resonator effective for the lower frequencies.

621.395.625.3 **2109**

Explanation of the Inner Mechanism of Magnetic-Tape Recording—K. Schwarz. (*Frequenz*, vol. 6, pp. 37–44; February, 1952.) Longitudinal and transverse magnetization of a permanent-magnet material are distinguished and the frequency characteristic in each case is discussed. Transverse magnetization of the tape is predominant in normal recording techniques. Causes of nonlinear distortion in recording and reproduction are noted.

621.395.625.3 **2110**

New Magnetic-Recording Head—M. Camras. (*Jour. Soc. Mot. Pic. Telev. Eng.*, vol. 58, pp. 61–66; January, 1952.) Description of a type of recording head with a third magnetic pole mounted above the air gap between the other two. It is claimed that such heads produce a more uniform field throughout the thickness of the magnetic layer, with a more rapid field decay at the trailing edge.

621.395.92 **2111**

The Medresco Hearing Aid—C. J. Cameron and E. W. Ayers. (*P.O. Elec. Engrs' Jour.*, vol. 44, part 4, pp. 153–158; January, 1952.) A broad outline is given of the work done by the Post Office in redesigning the hearing aid developed for the Medical Research Council (see 1245 and 1246 of 1948), incorporating a magnetic receiver and a new microphone, with modification of the amplifier circuit. A description is given of typical Medresco aids and their principal component parts. Production and performance of these aids is briefly mentioned and possible future improvements in design are noted.

ANTENNAS AND TRANSMISSION LINES

621.392.015.3:517.432 **2112**

A Particular Application of Operational Calculus to Transients on Transmission Lines—R. Codelupi. (*Alta Frequenza*, vol. 21, pp. 27–38; February, 1952.) The quadripole equations of a uniform line with negligible attenuation and phase distortion are put into a form suitable for application of operational calculus. As examples, the expression for the far-end voltage of a line is transformed to one which shows the reflected waves, and the wave form corresponding to a unit step voltage input is determined (*a*) for the input end of an infinitely long coaxial cable, (*b*) for the output from a 300-m length of 75-Ω coaxial cable terminated by a 75-Ω resistor in parallel with a 0.01-μF capacitor. For (*b*) the results are in good agreement with measurements.

621.392.21 **2113**

Very Flexible Two-Wire Line with Rapid Radial Diminution of Field, for U.H.F. Transmission—H. Kleinwächter and H. Weiss. (*Onde élect.*, vol. 32, pp. 46–50; February, 1952.) Investigation of a transmission line consisting of a twisted pair of very thin insulated wires, for use as a low-power feeder for cm waves. Optimum wire diameter and insulation thickness are determined, taking flexibility and attenuation into account. The radial diminution of the radiation field is calculated approximately. Results of measurements on a line consisting of 1-mm copper wires with polythene coating of outer diameter 2 mm agree well with calculations, the attenuation being of the order of 0.7 db/m. For a line using 0.05-mm wires the calculated attenuation is about 5 db/m.

621.392.26 2114

Compilation of the Propagation Constants of an Inhomogeneously Filled Waveguide—L. G. Chambers. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 19–21; January, 1952.) Application to the waveguide problem of the Rayleigh-Ritz method developed for determining the natural vibration frequencies of a mechanical system.

621.396.67:535.42:538.56 2115

Radiation or Diffraction Patterns Close to Receiving Antennas—L. S. Palmer. (*Jour. Appl. Phys.*, vol. 23, pp. 289–290; February, 1952.) Comments on 2159 of 1951 (Andrews).

621.396.67:621.392 2116

On the Transmission Efficiency of Long Feeder Lines for Very High Frequencies—K. Nagai, T. Omori, R. Sato and G. Sato. (*Tech. Rep. Tohoku Univ.*, vol. 15, pp. 57–69; 1950.) Experiment shows that the transmission efficiency of a two-wire open feeder 500 m long may be >50 per cent for wavelengths of 4–5 m. Wires should be 6–7 m above the ground and 30–40 cm apart; for a length of 5 λ at either end spacing may be 10 cm if necessary. Distance $(2l)$ between insulators should be about 30 m and such that $2l=(2n+1)\lambda/4$, where n is any integer.

621.396.67.018.424 2117

Broad-Band Antenna Element—M. W. Scheldorf. (*Tele-Tech*, vol. 11, pp. 50–51; January, 1952.) Short description of the development of a slot-fed cylindrical system of rods.

621.396.67.018.424 2118

Radiation Damping, Resistance and Directional Characteristics of Ultrawide-Band Aerials—H. Wolter. (*Z. angew. Phys.*, vol. 4, pp. 60–70; February, 1952.) An extension of van der Pol's theory of unloaded antennas leads to a fundamental relation between characteristic impedance and radiation damping. Antennas with very low characteristic impedance, such as wide cages, double cones, dipole triangles etc., have high radiation damping, and their feedpoint resistances and directional characteristics are practically constant for all frequencies for which the total length of the antenna is not appreciably less than $\lambda/2$. V-shaped antennas of low characteristic impedance produce a concentrated beam and have wide-band properties. For producing a highly concentrated beam, arrays of wide-band antenna with special wide-band feed may be used.

621.396.671 2119

Current Distribution along a Cylindrical Aerial—P. Poincelot. (*Compt. Rend Acad. Sci.* (Paris) vol. 234, pp. 513–515; January 28, 1952.) The cylinder is assumed to have a finite radius and be bounded by two planes normal to the axis, about which the current distribution is symmetrical. Starting from Maxwell's equations and using methods which have been applied in the study of transients in electrical circuits, a relation is obtained for transmitting antennas which represents an infinite set of equations involving an infinite number of unknowns. Discussion of this relation shows that the distribution of current (regarded as the algebraic sum of the internal and external current sheets) is perfectly sinusoidal for the cases where the length of the antenna is $\lambda/2$ or an odd multiple of $\lambda/2$, and also for very short antennas. For receiving antennas, similar results are obtained, thus justifying the methods of calculating their impedance and radiation characteristics. In the case of a long thin wire, end effects can be neglected and the above results will apply.

621.396.677 2120

Superdirective Aerials—P. Airgrain. (*Onde élect.*, vol. 32, pp. 51–54; February, 1952.) These are defined as having a main-lobe width $<\sin^{-1}\lambda/2a$, where $2a$ is the antenna length. An approximate calculation is made of the ratio of the energy stored in the antenna and its near field to that radiated in unit time, which ratio is of the nature of a Q factor. The theoretical requirements for an antenna to have a minimum Q value for a given direction are determined. Discussion of numerical results for a particular case indicates that the extreme reduction of bandwidth and radiation efficiency makes superdirective antennas impracticable.

621.396.677:537.226 2121

An Experimental Investigation of the Dielectric Rod Antenna of Circular Cross Section Excited in Rotationally Symmetrical Modes—C. M. McKinney. (*Jour. Appl. Phys.*, vol. 23, pp. 11–13; January, 1952.) Report of measurements on three series of dielectric antennas excited in the TM_{01} mode at 9.275 kmc. Maximum attenuation of secondary lobes was obtained with rods of relatively large diameter, but with rods of small diameter sharper and deeper central nulls occurred in the radiation pattern. For uniformly tapered rods the maximum secondary-lobe attenuation and also the deepest central null were obtained with the longest rod (length 10 λ). Similar results were obtained with TE_{01} excitation.

621.396.677:621.396.65 2122

Antennas for the TD-2—A. H. Lince. (*Bell Lab. Rec.*, vol. 30, pp. 49–55; February, 1952.) Details of the construction and assembly of lens antennas for the TD-2 relay system [1109 of May (Roetken, Smith and Friis)].

621.396.677.029.64 2123

Obstacle-Type Artificial Dielectrics for Microwaves—C. Süsskind. (*Jour. Brit. IRE*, vol. 12, pp. 49–60; January, 1952.) Discussion, pp. 61–62.) A survey of analytical and practical design techniques. 38 references.

621.396.677.2+621.396.97 2124

High Frequency Broadcast Transmission with Vertical Radiation—Adorian and Dickinson. (See 2335.)

621.392.015.3 2125

Travelling Waves on Transmission Systems. [Book Review]—L. V. Bewley. Publishers: Chapman & Hall, London, Eng., 2nd ed. 1951, 544 pp., 96s. (*Beama Jour.*, vol. 59, p. 11; January, 1952.) The first edition, published in 1933, has been used as the basis of a university course on transmission-line transients; this revised edition takes account of developments in theory and practice in relation to surge phenomena.

621.396.67:621.397.6 2126

TV Master Antenna Systems. [Book Review]—I. Kamen and R. H. Dorf. Publishers: J. F. Rider, New York, N.Y., 1951, 368 pp., $5.00. (*Electronic Eng.*, vol. 24, p. 183; April, 1952.) "Recommended not only to those who are concerned with the provision and maintenance of communal antenna systems, but also to others who are interested in television reception and distribution."

621.396.67.029.64:621.396.932 2127

Centimetric Aerials for Marine Navigational Radar. [Book Notice]—Publishers: H. M. Stationery Office, London, Eng., 156 pp., 15s. (*Govt Publ.* (London,) p. 28; January, 1952.) "Proceedings of a conference held June 15–16, 1950, in London."

CIRCUITS AND CIRCUIT ELEMENTS

512:621.3.06 2128

The Algebra of Chains of Contacts—H. Schwab. (*Ann. Télécommun.*, vol. 7, pp. 2–16; January, 1952.)

621.3.015.7 2129

Pulse Discrimination by means of Delay-Time Elements—H. Laett. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 30, pp. 1–6; January 1, 1952. In German.) Discussion of the method in which the pulse train is slightly delayed and then added to the original series of pulses. The problem resolves itself into the physical realization of an open or short-circuited transmission line satisfying the requirements as regards echo time, impedance and bandwidth. The necessary design formulas are derived for several types of line.

621.3.015.7:621.387.422 2130

Ten-Channel Pulse Analyzer—M. Langevin and G. Allart. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 515–518; January 28, 1952.) Circuit details and description of equipment for amplitude sorting of the pulses given by a proportional scintillation counter.

621.314.2:621.3.012 2131

New Conductance Diagram for Transformers—H. Kafka. (*Arch. Elektrotech.*, vol. 40, pp. 219–230; 1952.) Analysis of the transformer by a graphical construction based on a simplified circuit diagram (1324 of 1951). A numerical example is given. A generalization of the method is illustrated.

621.314.3 2132

Magnetic Amplifiers and their Recent Improvements—B. Pistoulet. (*Rev. Gen. Elec.*, vol. 61, pp. 45–52; January, 1952.) Analysis of the characteristics of transductors under different operating conditions, with particular reference to the use of high-permeability materials.

621.314.3 2133

An Analysis of Transients in Magnetic Amplifiers—D. W. Ver Planck, L. A. Finzi and D. C. Beaumariage. (*Trans. Amer. IEE*, vol. 69, pp. 498–503; 1950.) Analytical expressions are derived for the envelope of the transient output current of a magnetic amplifier when the control voltage is varied. The analysis applies to transients lasting for a few cycles of the supply frequency, or longer. Experimental results are in good agreement with theory.

621.316.86.096 2134

Use of Thermistors as Variable R.F. Resistance Standards—M. Soldi. (*Alta Frequenza*, vol. 21, pp. 3–26; February, 1952.) The possibility is examined of using thermistors as rf resistance standards, particularly for high values of resistance. Measurements show that at constant temperature the rf resistance of thermistors is less than the dc resistance, the difference (per cent) increasing with frequency and with resistance value, but being small enough to be neglected over a wide range of both frequency and resistance. A description of the bridge circuit used in the measurements is given.

621.318.5:512 2135

Some Results on the Application of Boole's Algebra to the Synthesis of Relay Circuits—C. Cardot. (*Ann. Télécommun.*, vol. 7, pp. 75–84; February, 1952.)

621.319.45:536.48 2136

Electrolytic Capacitors at Low Temperatures—C. D. Crater. (*Tele-Tech.* vol. 11, pp. 44–45, 72; January, 1952.) Analysis of tests carried out on the products of four manufacturers revealed considerable variation in effective capacitance at low temperature among units of different makes, of different voltage ratings, and among different lots from the same manufacturer. The results indicate that plain-foil types have a much greater stability at low temperature than etched-foil types. Extended storage or operation at temperatures down to −55°C produces no significant permanent changes in operating characteristics.

621.385.029.62:621.3.012.8 2137

A Systematic Method of Linear Small-Signal V.H.F. Analysis for Valve Circuits—I. A. Harris. (*Jour. Brit. IRE*, vol. 12, pp. 79–89; February, 1952.) The analysis takes account of electron inertia effects and treats the triode tube as a passive circuit element described by a set of linear equations. These express the mesh current associated with each adjacent pair of electrodes in terms of the external voltages applied between a common point and each electrode. The basic system can be applied to such problems as the calculation of (*a*) the input and output admittances of anode-, grid-, or cathode-separation triode amplifiers, (*b*) the noise factor of a single stage.

621.392 2138

Introduction to Formal Realizability Theory: Part 1—B. McMillan. (*Bell. Sys. Tech. Jour.*, vol. 31, pp. 217–279; March, 1952.) A general approach to the theory of the realizability of networks with many accessible terminals. The methods developed are applied to give a complete characterization of all finite passive networks.

621.392.012.8:517.562.2 2139

Network Representation of Transcendental Impedance Functions—M. K. Zinn. (*Bell. Sys. Tech. Jour.*, vol. 31, pp. 378–404; March, 1952.) The admittance or impedance of certain structures, such as a finite length of transmission line or a resonant cavity, can be represented at all frequencies by that of a network comprising lumped resistance, inductance, capacitance and conductance. In general the network contains an infinite number of branches, although a finite number may be used if only certain modes are to be represented. The procedure for the network synthesis is based on use of Mittag-Leffler's theorem, which provides a tool for breaking up a transcendental meromorphic function into an infinite series of simple fractions. The method is applied to (*a*) an open-circuited twin-wire transmission line, (*b*) a short-circuited coaxial line (or toroidal cavity with E radial), (*c*) a toroidal cavity with E axial.

621.392.4/.5:621.396.822 2140

Noise Factor of Networks—O. E. Keall. (*Marconi Rev.*, vol. 15, pp. 25–34; 1st Quarter 1952.) Normal methods of circuit analysis are used in the estimation of noise factor, some of the terms being defined so as to facilitate the use of these methods. Circuits are divided into two types depending upon whether or not a tube or other isolating device is included, and theorems appropriate to the two types are presented.

621.392.43:621.396.67 2141

Methods of Calculation relating to Inductive Aerial Couplings—V. Familier. (*Onde élect.*, vol. 32, pp. 39–45; February, 1952.) A treatment of matching problems based on simple geometry. Similar geometrical methods have been applied by Storch (571 of 1950) to the case of capacitive coupling. In the complex plane, the point representing the input impedance of a network describes a circle when an element of the network is varied, the element being purely resistive or purely reactive. From the intersections of this circle with straight lines and circles determined by circuit parameters, optimum matching conditions for antenna and tuned secondary circuit can be found. The matching ranges of quadripoles can be determined in a similar manner.

621.392.5 2142

Discontinuous Low-Frequency Delay Line with Continuously Variable Delay—J. M. L. Janssen. (*Nature* (London,) vol. 169, pp. 148–149; January 26, 1952.) The network described consists of a number of sections each comprising a clamping circuit; step variations of voltage occur at instants controlled by application of switching pulses, whose frequency determines the time delay.

621.392.5 2143

Rise Time of Artificial Delay Lines—R. Génin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 193–195; January 7, 1952.) Using analysis involving Bessel functions, an expression is derived according to which the rise time is proportional to $n^{\frac{1}{3}}$, where n is the number of Π sections in the line. This expression is of the same form as that obtained experimentally by Elmore and Sands (2007 of 1950). Since the delay time is proportional to n, it is possible to reduce relative distortion by making n large.

621.392.5 2144

Ladder Development of RC Networks—E. A. Guillemin. (Proc. I.R.E., vol. 40, pp. 482–485; April, 1952,) Darlington and Cauer have described a method for the synthesis of a lossless quadripole network from a single driving-point impedance and knowledge of the zeros of transmission. The procedure can readily be extended to RL and RC networks provided the zeros of transmission are restricted to the negative real axis of the complex frequency plane. The method is illustrated by numerical calculations, starting from an assumed pair of functions.

621.392.5:512.831 2145

Notes on the Application of Matrix Calculus to Linear and Pseudolinear Feedback Systems—J. Salmon. (*Jour. Phys. Radium*, vol. 13, no. 2 supplement, pp. 25A–28A; February, 1952.) Conditions for the initiation of oscillations in a linear feedback system are found and extended to certain pseudolinear systems.

621.392.5:621.3.015.3 2146

Study of Transient Processes in Linear Quadripoles—F. Brunner. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 6, pp. 1–9; January/February, 1952.) In the analytical method described, a relation is derived between the steady-state and initially variable components of output voltage, and the duration of the transient is defined as the time taken for the initially variable component to fall to a given fraction of the steady-state component. The method is illustrated by application to a RC element and to a parallel-resonant circuit.

621.392.52 2147

Generalized Ideal Filters—L. A. Zadeh and K. S. Miller. (*Jour. Appl. Phys.*, vol. 23, pp. 223–228; February, 1952.) A definition is formulated which extends the concept of ideal filter to both linear varying-parameter and nonlinear types of system; a filter is said to be ideal if it can extract a signal from its combination with another signal, even when the two frequency bands overlap. The basic properties of ideal filters are investigated using function-space techniques.

621.392.52 2148

The Double-T RC Filters—W. Schmidt. (*Elektrotech. Z.*, vol. 73, pp. 35–38; January 15, 1952.) The action of a high-pass and a low-pass filter in parallel for suppression of a single frequency is analyzed and design procedure indicated. When the filter is used in the feedback network of an amplifier, the circuit may operate either as an oscillator or as a tuned amplifier with prescribed bandwidth.

621.392.52.029.64:621.396.611.4 2149

Cavity Band-Pass Filters for Centimetre Waves—H. Döring and W. Klein. (*Arch. elekt. Übertragung*, vol. 6, pp. 47–57 and 119–125; February and March, 1952.) A theoretical treatment of filters comprising a number of cavity resonators in the form of flat cylindrical boxes coupled by windows in the common walls. The alteration of the circuit parameters of the end cavities due to the coupling elements causes a mismatch within the filter. The adjustment of this by means of the coupling reactances affords a useful means of obtaining a required transmission characteristic; e.g., in a 4-element filter, the group delay can be made practically constant over a large part of the pass band. The design procedure and method of measurement of the attenuation and phase constant of such filters are described.

621.392.54+621.392.26.072.31 2150

Application of Multi-Hole Coupling to the Design of a Variable and Calibrated Waveguide Attenuator and Impedance—W. J. van de Lindt. (*Phillips Res. Rep.*, vol. 7, pp. 28–35; February, 1952.) A discussion of the characteristics of two parallel waveguides mutually coupled by n equidistant identical directional elements, with a description of the application of such a system to the design of a calibrated variable attenuator, and a calibrated variable impedance capable of changing independently the amplitude and the phase of the reflection coefficient.

621.392.6 2151

Generalized Network Theory—U. Kirschner. (*Arch. elekt. Übertragung*, vol. 6, pp. 86–87; February, 1952.) Correction to paper abstracted in 837 of 1951.

621.395.665:534.86 2152

New Principle for Electronic Volume Compression—H. E. Haynes. (*Jour. Soc. Mot. Pic. Telev. Eng.*, vol. 58, pp. 137–144; February, 1952.) The principle is to modulate the signal with hf rectangular pulses of variable duty factor (k). Unwanted modulation products are filtered out leaving the desired signal with amplitude multiplied by k. The value of k is varied in accordance with an appropriate control voltage. The circuit described incorporates a 45-kc pulse generator keying a push-pull amplifier. Advantages of the system are extremely low audio "thump," very fast action if required, low distortion, and use of components and tubes not specially selected. Performance figures are given.

621.396.6:061.4 2153

R.E.C.M.F. Exhibition Preview—(*Electronic Eng.*, vol. 24, pp. 178–181; April, 1952.) Short descriptions of selected exhibits at the Radio and Electronics Component Manufacturers Federation exhibition, London, April 1952. See also *Wireless World*, vol. 58, pp. 179–182; May, 1952.

621.396.6:061.4 2154

The National Components Exhibition—H. Gilloux. (*Radio Franç.*, no. 2, pp. 20–24; February, 1952.) Review of the exhibition in Paris, February 1952, and description of certain exhibits. For longer lists of items, including acoustic equipment, measurement sets and tubes, see *Toute la Radio*, vol. 19, pp. 115–122; March/April, 1952; *TSF et TV*, vol. 28, pp. 85–88 and 139–144; March and April, 1952; *Radio prof.* (Paris), *l'Exportation Électricité-Radio franç., Supplement*, March, 1952, 16 pp.

621.396.611.018.3 2155

Subharmonic Oscillations in Electric Circuits Containing Iron-Core Reactors—J. P. Schouten and H. J. Heijn. (*Appl. Sci. Res.*, vol. B2, pp. 301–319; 1952.) Investigation of the flux variation in an iron-cored reactor connected in series with an inductor and capacitor and fed by a sinusoidal emf. In the theoretical treatment the flux/current curve is represented approximately by three straight lines of different slope, and the associated differential equation is solved graphically, with particular reference to the third-order subharmonic. Oscillograms reproduced confirm the theory.

621.396.611.21:534.133 2156
Forced Thickness-Shear and Flexural Vibrations of Piezoelectric Crystal Plates—R. D. Mindlin. (*Jour. Appl. Phys.*, vol. 23, pp. 83–88; January, 1952.) An approximate theory is presented which includes the interaction between the elastic and electric fields. Computed frequencies for rectangular AT-cut quartz plates are compared with measurements by Sykes, and formulas are derived relating resonance frequencies to dimensions, elastic and electric constants, and orientation of cut.

621.396.611.3.011.21 2157
Input-Admittance Characteristics of a Tuned Coupled Circuit—R. A. Martin and R. D. Teasdale. (PROC. I.R.E., vol. 40, p. 459; April, 1952.) Correction to paper abstracted in 1241 of June.

621.396.611.3.029.64 2158
The Application of Window Coupling at Centimetre Wavelengths—H. Döring and W. Klein. (*Elektrotech. Z.*, vol. 73, pp. 5–9; January 1, 1952.) Basic principles are briefly described and applications in directional couplers, coupled cavity resonators in klystrons, and multistage cavity filters, are considered.

621.396.611.4 2159
The Loop-Excited Cavity Resonator comprising Two Confocal Paraboloids of Revolution—H. Buchholz. (*Arch. elekt. Übertragung*, vol. 6, pp. 6–16 and 67–72; January and February, 1952.) A cavity having the shape of a double-convex lens is formed from two paraboloids and is excited by a coaxial circular loop. Analysis of the equations for the field leads to a complete solution in the form of two loop integrals. On developing into an infinite series, this is found to correspond to the superposition of an infinite number of wave trains whose basic form is that of the wave in an infinitely long parabolic horn. The solution has poles for an infinite series of discrete values of the wave number corresponding to the natural resonance frequencies; calculation of these frequencies is easy for the case of equal paraboloids. The magnetic-field lines and Q are investigated for near-resonance conditions. Numerical values are tabulated for a higher transcendental function used in the solution.

621.396.611.4.029.64 2160
Electromagnetic Resonant Behavior of a Confocal Spheroidal Cavity System in the Microwave Region—J. C. Simons and J. C. Slater. (*Jour. Appl. Phys.*, vol. 23, pp. 29–30; January, 1952.) Discussion of the resonance of a small spheroidal object in a large spheroidal cavity, approximating to a needle-like antenna in a large cavity. If the needle is thin enough, the magnetic field on its surface is greatly enhanced when the cavity is tuned to a resonance frequency determined by the needle. This can be applied to measurement of the surface impedance of the needle material.

621.396.615 2161
The Build-Up Process in Oscillators—W. Herzog. (*Arch. elekt. Übertragung*, vol. 6, pp. 58–66; February, 1952.) Assuming a cubic form of triode characteristic, the types of grid-voltage and anode-current swing for class A, B and C operation are explained and the mean anode currents determined. The effect of feedback is considered. An approximate theory of the Meissner oscillator is developed and the build-up period and the "inertia" of the circuit are determined. Nonlinear feedback is also considered.

621.396.615:621.396.822 2162
The Effect of Background Noise on the Amplitude of [valve-] Maintained Oscillators—A. Blaquière. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 710–712; February, 11, 1952.) The noise is treated as a succession of impulses and each pulse is regarded as the sum of two components, one producing, to the first order, a change of phase without change of amplitude, the other a change of amplitude but no phase shift. Second-order analysis shows that the noise power is equally divided between a periodic component and a continuous spectrum. Formulas are derived which permit comparison between the noise power of a tube oscillator and that of a passive circuit.

621.396.615:621.396.822:529.786 2163
The Effect of Background Noise on the Frequency of Valve Oscillators. Ultimate Accuracy of Electronic Clocks—A. Blaquière. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 419–421; January 21, 1952.) An analysis is made of the disturbing effect of a single noise pulse on the signal generated by an amplitude-stabilized oscillator, using a method described by Rice (2219 of 1948); both the amplitude and the phase of the signal are affected. The mean square of the phase shift due to noise increases linearly with time; an expression is derived for the mean square of the error in time measurement when the oscillator is used as a clock.

621.396.615.17 2164
An Amplitude-Comparator Multivibrator—S. Fedida. (*Marconi Rev.*, vol. 15, pp. 35–43; 1st Quarter 1952.) Description of a method of rendering the out-pulse wave form, amplitude, delay, etc., and the flip-flop recovery time, independent of the amplitude of the input pulse. This is effected by suitable connection of a double diode in a subsidiary feedback loop in the flip-flop circuit.

621.396.615.17 2165
Triangular-Waveform (Sawtooth) Generator—R. Peretz. (*HF*, (Brussels), vol. 2, pp. 16–24; 1952.) A mathematical analysis is made of a circuit in which the voltage across a capacitor in the cathode circuit of a triode is applied through a directly coupled amplifier to the grid of the tube. From a pulsed input across the capacitor, a wave form of any positive or negative exponential type can be obtained by variation of the amount of feedback used. Application of the circuit in a sawtooth generator giving frequencies from 0.001 to 1,000 cps is described and performance characteristics are shown.

621.396.619.2+621.396.622.6/.7]:621.395.44 2166
Frequency Converters as Quasilinear Quadripoles—W. Klein. (*Arch. elekt. Übertragung*, vol. 6, pp. 29–35; January, 1952.) Theory is developed for the various modulator and demodulator circuits used in ssb and dsb carrier-wave technique. A basic feature, whose introduction enables the superposition principle to be used, is the quasilinear circuit, an idealized equivalent for the actual frequency-converter circuit, in which two quadripoles are in effect switched alternately into use. The corresponding modulation function is a square wave. As a particular case, the switching may be simply a reversal of a single circuit. Quasilinear quadripoles incorporating modulated rectifiers are described; switching is performed by the periodic polarity reversals of the carrier voltage. Conditions to be satisfied by carrier wave form and rectifier characteristics are discussed. Calculations are made for various known modulator circuits.

621.396.645 2167
Distributed Amplification—A. Cormack. (*Electronic Eng.*, vol. 24, pp. 144–147; April, 1952.) Basic principles are outlined and design details are given of two amplifiers with flat response curves from lf to 170 mc. The first, a single-stage amplifier, has a gain of 18 db; the other has two stages in cascade, each with four tubes, and has a gain of 28 db. Several methods of obtaining an output impedance of 75Ω in the final stage are considered briefly.

621.396.645 2168
The Treatment of Amplifier Circuits by means of V_a/V_g Characteristics—A. Simon. (*Fernmeldetech. Z.*, vol. 5, pp. 11–16; January, 1951.) Direct determination of V_a/V_g characteristics (curves of constant anode and grid currents as dependent on anode and grid voltages) is difficult; hence they are usually derived from V_a/i_a or V_g/i_a characteristics. Particular features of these characteristics are noted and working characteristics are given for class A, B and C amplifiers, circuit parameters being tabulated for specified anode voltages and output powers, and control power estimated. Data are also tabulated for a class B amplifier with low-level modulation of an input stage.

621.396.645 2169
Some Rules for the Construction of I.F. Amplifiers—W. Hasselbeck. (*Funk u. Ton*, vol. 6, pp. 1–7; January, 1952.) Review of the principal conditions which must be fulfilled for satisfactory operation.

621.396.645:621.392.52 2170
New Method of Calculating High-Frequency Filters with Tchebycheff Type of Amplification—H. Edelmann. (*Arch. elekt. Übertragung*, vol. 6, p. 87; February, 1952.) Correction to paper abstracted in 88 of February.

621.396.645.029.3 2171
Equipment for Acoustic Measurements: Part 5—A Portable 7.5-W Loudspeaker Amplifier—D. E. L. Shorter and W. Wharton. (*Electronic Eng.*, vol. 24, pp. 7–9; January, 1952.) Description of a power amplifier used to drive the loudspeaker used in tests of room acoustics. Part 4: 1817 of August.

621.396.645.36 2172
The Cathamplifier—C. A. Parry. (PROC. I.R.E., vol. 40, pp. 460–465; April, 1952.) Reprint. See 78 of 1951.

621.396.645.37 2173
Complex Feedback—W. Oesterlin. (*Arch. tech. Messen*, no. 193, pp. 39–42; February, 1952.) Investigation of the possibility of compensating both the phase and amount of amplification of a pentode by means of frequency-dependent complex feedback. Application of this principle in a pentode amplifier resulted in uniform amplification up to 12 kc, with negligible phase change.

621.396.822:[621.315.5+621.385.2 2174
Thermal and Shot Fluctuations in Electrical Conductors and Vacuum Tubes—S. S. Solomon. (*Jour. Appl. Phys.*, vol. 23, pp. 109–112; January, 1952.) A new derivation of Nyquist's equation relative to the amount of the thermal fluctuations generated in an electrical conductor, together with a generalization to include any arbitrary impedance function. This shows that the original Nyquist equation is valid only for physically realizable impedances of the minimum-reactance type. A short derivation of the shot-noise formula for temperature-limited diodes is also presented.

621.397.645:621.385.4 2175
Coaxial Tetrode as a TV Amplifier at V.H.F. and U.H.F.—D. H. Priest. (*Tele-Tech*, vol. 11, pp. 52–53, 88; January, 1952.) Three alternative arrangements for use with a coaxial tetrode are considered and a detailed discussion is given of a power amplifier using an Eimac type-4×150G tube. The network connecting the screen grid, control grid and cathode is basically a folded coaxial line, connected between control grid and screen grid at one end and between control grid and cathode at the other. At the point of folding there is a variable series inductance, provided by the stub with its adjustable short-circuiting bar, plus two shunt capacitors which are not required in all cases. The output circuit is a conventional two-section band-pass filter; the drive is applied via a loop between control grid and screen grid. Under class-B linear conditions

this amplifier has a bandwidth (at −3db) of 5 mc, peak power 107 w at 815 mc and 220 w at 500 mc, and power gain of 8–10. A cross section is shown through a similar amplifier using an Eimac type-4W20000A tetrode with a water-cooled anode capable of dissipating 20 kw.

621.397.645.018.424 2176

Wide-Band Amplifiers with Stagger-Tuned Circuits—J. de Vos. (*Funk u. Ton.* vol. 6, pp. 69–74; February, 1952.) Discussion of the operation of IF wide-band amplifiers such as are used in television receivers. Resonance frequencies for the different circuits are determined which give an optimum shape to the transmission curve.

621.3.015.3:517.432.1 2177

Transients in Electric Circuits, using the Heaviside Operational Calculus. [Book Review]—W. B. Coulthard. Publishers: Pitman & Sons, London, Eng., 2nd ed., 32s. 6d. (*Engineering* London,) vol. 173, pp. 67–68; January 18, 1952.) "For this second edition the opportunity has been taken to revise the whole text. . . The wide range and representative character of the problems dealt with should commend the book to all electrical engineers."

GENERAL PHYSICS

531:537].001.362 2178

Analogies between Mechanical and Electrical Magnitudes—W. Reichardt. (*Frequenz*, vol. 6, pp. 25–29, 50–55 and 72–87; January–March, 1952.

534.22 2179

The Concept of Group Velocity—P. Poincelot. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 599–602; February 4, 1952.) Analysis justifying accepted ideas on the subject.

535.22+621.396.11 2180

The Velocity of Light—L. Essen. (*Sci. Progr.*, vol. 40, pp. 54–70; January, 1952.) Review of the various methods that have been used to determine the velocity, and analysis of the results obtained.

535.34:621.315.61 2181

The Structure of the Long Wave Absorption Edge of Insulating Crystals—I. C. Cheeseman. (*Proc. Phys. Soc.*, vol. 65, pp. 25–32; January 1, 1952.) Theoretical study of a process by which light can be absorbed in insulators at frequencies below that corresponding to the energy gap. Theoretical and experimental results are in good agreement for CdS.

535.42 2182

A Rigorous Formulation of the Classical Diffraction Problem—H. Honl. (*Z. Phys.*, vol. 131, pp. 290–304; February 19, 1952.) The methods of Sommerfeld, Schwarzschild, and Levine and Schwinger are briefly reviewed and a treatment by means of a Fourier representation of the wave function is given; this leads to two simultaneous integral equations for slits of arbitrary shape in the plane screen. Differences from Kirchhoff's theory are particularly considered.

535.42 2183

The Diffraction of Electromagnetic Waves at a Slit: Part 1—E. Groschwitz and H. Hönl. (*Z. Phys.*, vol. 131, pp. 305–319; February 19, 1952.) Application of the general theory [2182 above (Hönl)] to a straight slit, assuming the wave function to be zero at the slit boundary.

535.42:538.56 2184

Diffraction of Electromagnetic Wave by Apertures in Plane Conducting Screens—J. P. Vasseur. (*Onde élect.*, vol. 32, pp. 3–10, 55–71 and 97–112; January-March, 1952.) Classical methods of direct integration of Maxwell's equations are reviewed and a detailed study is made of Kottler's formulas, showing in what respects they are incorrect. A system of magnetic dipoles distributed over the plane of the aperture gives a diffraction field which satisfies all the boundary conditions. These dipoles are determined by a system of two integro-differential equations more complete than the analogous equations of Copson. Reciprocally, the diffraction field can be produced by a system of electric dipoles distributed over the surface of the screen; these also are determined by a system of two integro-differential equations. These systems of equations enable a rigorous statement of Huyghens' principle under several equivalent forms, whose comparison leads to the exact expression of Babinet's principle, which can also be established directly. In the cases of diffraction by a small circular hole and by a half plane, this method of treatment gives results found in other ways by Bethe and Sommerfeld. It appears impossible to solve explicitly the integro-differential equations concerned, even in simple cases, but when the aperture is large enough (several times the wavelength) the calculations are simplified and lead to formulas analogous to those of Kottler. These formulas have many reciprocal expressions, the integration extending over the aperture or over the metal part of the screen. Experimental results for the diffraction of 3-cm em waves are in satisfactory agreement with the simplified formulas proposed. 242 references.

535.42:538.56 2185

Diffraction by a Wave-Guide of Finite Length—D. S. Jones. (*Proc. Camb. Phil. Soc.*, vol. 48, pp. 118–134; January, 1952.) Starting from the electric intensities on the planes containing the walls, integral equations are derived which, after application of the Laplace transform, can be solved by successive substitutions. The series thus obtained is too complex for practical purposes, and an approximate solution is found for the case of waveguide length large compared with the wavelength.

535.42:538.56 2186

Diffraction Measurements at 1.25 Centimeters—R. D. Kodis. (*Jour. Appl. Phys.*, vol. 23, pp. 249–255; February, 1952.) Brass and polystyrene cylinders with diameters comparable with λ were used as diffracting objects. Radiation source, obstacle and detector were mounted above a 4-foot×6-foot horizontal conducting sheet, the auxiliary apparatus being located below it. Both the phase and amplitude of the electric field near the obstacle were measured; the results are compared with values calculated from theory for the conducting cylinders. The technique was also applied to investigate diffraction by an edge.

535.42:538.56:621.396.67 2187

Radiation or Diffraction Patterns Close to Receiving Antennas—L. S. Palmer. (*Jour Appl. Phys.*, vol. 23, pp. 289–290; February 1952.) Comments on 2159 of 1951 (Andrews)

537.1:530.12 2188

Classic Theory of the Point Charge—B. Jouvet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 712–714; February 11, 1952.) Study and geometrical representation of the relativistic and electromagnetic invariances of the classic theory of the point charge.

537.213:537.562 2189

"Cell" constituted by Gas Ionized at High Frequency—M. Chenot. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 608–610; February 4, 1952.) A discharge tube, excited by meter waves applied to external electrodes, can furnish a constant difference of potential between two electrodes in contact with the ionized gas within the tube, if these electrodes are arranged asymmetrically so as to produce a deformation of the luminous discharge column, such as a large dark space near one of the external electrodes. This effect is discussed.

537.311.1 2190

The Mean Free Path of Electrons in Metals—E. H. Sondheimer. (*Advances in Phys.*, vol. 1, pp. 1–42; January, 1952.) A survey of the theory of electrical conduction in metals (based on Sommerfeld's quantum-mechanical treatment) with reference to size effects in which the mean free path is comparable with some significant linear dimension. The evaluation of the mean free path (independently of electron density) from measurements of the resistivity of thin films or wires is discussed. Study of the influence of a magnetic field on the resistivity of thin specimens enables the momentum of electrons at the surface of the Fermi distribution to be deduced. The anomalous skin effect enables the mean free path to be compared experimentally with the penetration depth of hf electric fields.

537.311.31 2191

Metallic Conduction—The Internal Size Effect—D. K. C. MacDonald. (*Phil. Mag.*, vol. 43, pp. 124–125; January, 1952.) Addendum to 649 of April.

537.311.31:539.23 2192

Law of Variation of the Resistance of Very Thin Metal Films as a Function of the Applied Potential—N. Mostovetch. B. Vodar and T. Duhautois. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 305–308; January 14, 1952.) The increase of conductivity observed with high field strengths is attributed to a lowering of the potential barrier between the metal grains, due to the Schottky effect.

537.311.33 2193

Theory of Conductivity of Semiconductors—G. Jaffé. (*Phys. Rev.*, vol. 85, pp. 354–363; January 15, 1952.) The author's earlier theory (1933 Abstracts, p. 287) for polarizable media is extended to include both the electronic and ionic components of conduction; the latter component is found to cause significant departures from the Mott-Schottky theory of rectification. The ac admittance is derived and the frequency dependence of the susceptance and conductance is shown to be markedly influenced by the ionic component at low frequencies. Comparison of the theory with measurements on Se disks shows good agreement for the frequency range 2–200 kc if two species of ion of different mobilities be assumed.

537.311.33:537.311.4 2194

The Relation between Contact Resistance and Contact Potential Difference—M. A. Krivoglaz and K. B. Tolpygo. (*Zh. Tekh. Fiz.*, vol. 21, pp. 417–426; April, 1951.) The tunnel effect through the potential barrier formed by the curvature of the conduction band of a semiconductor on which a metal electrode has been deposited is calculated. The influence of this effect on the contact resistance is discussed for various forms of the potential barrier.

537.525.72:538.63 2195

Interaction of Travelling Magnetic Fields with Ionized Gases—P. C. Thonemann, W. T. Cowhig and P. A. Davenport. (*Nature* (London), vol. 169, pp. 34–35; January 5, 1952.) Experiments are described in which the magnetic field associated with a rf current in a helix produces an electrodeless dc discharge of the order of amperes in an ionized gas in a tube surrounded by the helix. Both straight and toroidal tubes were investigated. The magnitude of the dc was not sensitive to either phase velocity or frequency, but was highly sensitive to gas pressure. Evidence was also obtained for the amplification of a rf signal injected into a line enclosing the plasma of a dc arc.

537.533:546.655.4–31 2196

Some Results on the Optical Emissivity and Thermionic Emission of Ceria—R. Uzan. (*Le Vide*, vol. 7, pp. 1139–1140; January, 1952.) The thermionic emission follows Richardson's law, with a mean value of ϕ of 2.7ev for a coating thickness of 60μ. Values of A lie between 0.005 and 10A/cm.²

537.533.8 2197
On the Theory of Secondary-Electron Emission—J. L. H. Jonker. (*Philips Res. Rep.*, vol. 7, pp. 1–20; February, 1952.) Starting from (*a*) Whiddington's law concerning the velocity reduction of electrons penetrating into a solid substance, (*b*) the experimental law of absorption, (*c*) the assumption that the distribution of the secondary electrons within matter is isotropic, the dependence of the properties of secondary electrons on various parameters is calculated and found in good agreement with experimental results.

537.533.8 2198
Secondary Emission from Composite Surfaces—H. Jacobs, J. Martin and F. Brand. (*Phys. Rev.*, vol. 85, pp. 441–447; February 1, 1952.) Investigation of various compounds indicates that each has its own threshold energy below which primary electrons do not yield true secondary electrons. It is concluded that secondary electrons originate from the filled band of a compound rather than electron traps.

538.11:538.124 2199
The Lowest Energy State of a Linear Antiferromagnetic Chain—P. W. Kasteleijn. (*Physica*, vol. 18, pp. 104–113; February, 1952.)

538.11:539.132 2200
On the Quantum Theory of Antiferromagnetism—H. A. Kramers. (*Physica*, vol. 18, pp. 101–103; February, 1952.)

538.3 2201
Fundamentals of a New Theory of Electromagnetism—B. Jouvet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 819–822; February 18, 1952.) The laws of composition of fields are examined on the hypothesis that there is an upper limit to the value of field strength. A fundamental invariant is deduced.

538.3 2202
Dirac's New Electrodynamics—K. J. Le Couteur. (*Nature* (London), vol. 169, pp. 146–147; January 26, 1952.) Discussion on 1574 of July.

538.3:535.13 2203
Is there an Aether?—H. Bondi and T. Gold: P.A.M. Dirac. (*Nature* (London), vol. 169, p. 146; January 26, 1952.) Comment on 1573 of July and author's reply.

538.521 2204
Electromagnetic Induction and Magnetoelectric Induction—G. Vallauri. (*Alta Frequenza*, vol. 20, pp. 227–246; December, 1951.) The two laws of induction are compared and experiments are described which prove the law of magnetoelectric induction directly.

538.56:537.533:523.72 2205
Condition for Radiation from a Solar Plasma—J. Feinstein. (*Phys. Rev.*, vol. 85, pp. 145–146; January 1, 1952.) Theory developed by Bailey (105 of February) is critically discussed.

538.566 2206
The Interaction of Electromagnetic Waves within Matter—R. Lucas. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 191–193; January 7, 1952.) Relativistic kinematic theory is used to examine the conditions necessary for interaction. The relative directions of propagation and the new frequencies resulting from interaction 1, are determined.

538.569.4.029.64 2207
Beam System for Reduction of Doppler Broadening of a Microwave Absorption Line—H. R. Johnson and M. W. P. Strandberg. (*Phys. Rev.*, vol. 85, pp. 503–504; February 1, 1952.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:551.510.535:621.396.9 2208
Theory of Radio Reflections from Meteor Trails: Part I.—T. R. Kaiser and R. L. Closs. (*Phil. Mag.*, vol. 43, pp. 1–32; January, 1952.) A comprehensive treatment of the reflection of em waves from an ionized column with cylindrical symmetry, which diffuses radially outwards. Different distributions of electron density n are considered and the Gaussian distribution $n=n_0 \exp[-(r/r_0)^2]$ is discussed in detail, r being the radius of the ionized column. The theory predicts two qualitatively different types of meteor-trail echo, depending on the magnitude of the electron line density α. With $\alpha \gg 10^{12}$ per cm, the column is expected to reflect in a manner similar to that of a metal cylinder. The reflection coefficients for the incident electric vector E parallel and perpendicular respectively to the axis of the column, are equal and substantially constant with increasing radius, except initially. These predictions are in agreement with experimental data on long-duration echoes. With $\alpha < 10^{12}$ per cm, for E parallel to the axis the echo amplitude decays exponentially with increasing radius, while for E perpendicular to the axis the amplitude may pass through a resonance value before decaying exponentially. The short-duration echoes observed in most cases are of these types.

523.7 2209
Solar Observations—A. Behr and H. Siedentopf. (*Naturwiss.*, vol. 39, pp. 28–38; January, 1952.) Detailed report of methods of observation of sunspot activity, solar flares, eruptions and corona effects, and of results obtained at many different stations over a number of years.

523.7:538.122 2210
Measurements of the Sun's General Magnetic Field—G. Thiessen. (*Nature* (London), vol. 169, p. 147; January 26, 1952.) Data are presented in support of the conclusions stated in 1141 of 1951.

523.72:538.56:537.533 2211
Condition for Radiation from a Solar Plasma—Feinstein. (See 2205.)

523.72:621.396.822 2212
Excess Radio Noise from Solar Flares and Sunspots—R. Q. Twiss. (*Nature* (London), vol. 169, pp. 185–186; February 2, 1952.) Theoretical discussion of the noise generated in and escaping from an ionized atmosphere, with particular reference to plasma oscillations and to the amplification of transverse field waves under the existence of suitable boundary conditions.

523.72:621.396.822]:550.385 2213
Possible Identification of a Solar M-Region with a Coronal Region of Intense Radio Emission—A. Maxwell. (*Observatory*, vol. 72, pp. 22–26; February, 1952.) On 14th June 1950 a sunspot group associated with unusually intense meter-wave radio emission crossed the sun's central meridian. For the next six months there was, at 27-day intervals, a sequence of moderate geomagnetic storms of the type known to follow the formation of an M-region. Since meter-wave solar radiation originates in the corona, the correlation between M-regions and the corona is close. The connection between aurorae, M-regions and geomagnetic storms is shown by the occurrence of radar auroral echoes at the time of the storms or 14 days out of phase with them.

523.8:621.396.822 2214
Radio Stars or Radio Nebulae—R. Bracewell. (*Observatory*, vol. 72, pp. 27–29; February, 1952.) Brief consideration of available evidence.

523.852.3:621.396.822 2215
Extra-Galactic Radio-Frequency Radiation—R. H. Brown and C. Hazard. (*Phil. Mag.*, vol. 43, pp. 137–152; February, 1952.) Description of measurements on λ 1.89 m using a paraboloid antenna system 218 feet in diameter giving a beam width of 2°. Three bright nebulae have been examined; three radio sources are associated with them. The radio intensity to be expected from eight of the major clusters of nebulae has been calculated; two of the clusters have been surveyed and identified with radio sources. There appears to be a relation between the apparent photographic magnitude and apparent radio magnitude of nebulae.

550.384:551.510.535 2216
Interpretation of the Eleven-Yearly Variation of the Horizontal Component of the Earth's Magnetic Field—P. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 866–868; February 18, 1952.) The general mean of observations made at 37 stations shows a minimum of the horizontal component 1.1 years after the maximum of sunspot activity. The nature and height of ionospheric currents capable of giving rise to these variations are discussed.

551.510.535 2217
Physics of the Ionosphere—K. Rawer. (*Phys. Blätter*, vol. 8, pp. 15–23; 1952.) A review of present-day theory.

551.510.535:621.396.11.029.56 2218
Reflection of Short Waves at Heights less than 100 km—Dieminger and Hoffmann-Heyden. (See 237.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 2219
Pictorial Radio—C. D. Tuska. (*Jour. Frank. Inst.*, vol. 253, pp. 1–20 and 95–124; January and February, 1952.) Defined as "multi-coordinate or graphic indicating radio systems for obtaining bi- or tri-dimensional information transmitted or reflected from a plurality of geometrically related points." The development is traced from the time of Hertz to the end of the second world war. 37 references, including many patents.

621.396.9 2220
Development of an Experimental Electromagnetic Detector—J. Moline. (*Radio franç.*, nos. 2 and 3, pp. 1–5 and 11–15; February and March, 1952.) Description of radar equipment designed primarily for instruction in the principles of the art. Operating frequency is 145 mc (λ = 2.07 m), pulse duration 3μs and recurrence frequency 500 per second, peak power about 0.5 kw, and range 30 km. Type-A display is used.

621.396.9:519.2 2221
The Statistical Properties of Noise Applied to Radar-Range Performance—S. M. Kaplan and R. W. McFall. (PROC. I.R.E., vol. 40, pp. 487–489; April, 1952.) Discussion on 1390 of 1951.

621.396.932 2222
Radar Equipment at the Port of Le Havre—(*TSF et TV*, vol. 28, p. 30; January, 1952.) Note of features of the 3-cm equipment to be installed. These include a 14-m paraboloidal antenna weighing about 5 tons and with a beam angle of only 42 minutes, a special system incorporating a semi-reflecting surface for rapid determination of the speed and direction of a moving vessel, and a cm-wave telephony system for ship communication.

621.396.933 2223
Some Navigational and Air Traffic Control Problems of Civil Aviation and the Application of Radio to their Reduction—G. W. Stallibrass. (*Jour. Brit. IRE*, vol. 12, pp. 3–20; January, 1952. Discussion, pp. 21–22.)

621.396.933.1.087.9 2224
The Design and Development of the Decca Flight Log—G. E. Roberts. (*Jour. Brit. IRE*,

vol. 12, pp. 117–131; February, 1952.) The presentation of navigational data is discussed. The information given by the Decca Navigator receiver is displayed in a convenient form in the instrument described. The choice of coordinates is discussed and details are given of the various units used in the Mark 01 model.

621.396.933.2:623.451 2225

Miniature Transponder Beacon for Guided Missiles—B. H. Sinclair. (*TV Eng.* (N.Y.), vol. 3, pp. 8–9, 30 and 14–17, 28; February and March, 1952.) The unit comprises receiver, decoding, trigger, modulator, transmitter and duplexing circuits. It is housed in a pressurized container $2\frac{1}{2}$ inches in diameter and $6\frac{3}{8}$ inches long and weighs <2 lb. A similar container holds the Zn-Ag_2O_2 battery unit and remotely controlled switching mechanisms. For stability, "etched-plate" circuits are used and small components are secured to the chassis with plastic compound.

534.88 2226

Echo Sounding at Sea (British Practice). [Book Review]—H. Galway. Publishers: Pitman & Sons, London, Eng., 35s. (*Marconi Rev.*, vol. 15, p. 44; 1st Quarter 1952.) "The book can be confidently recommended not only to shore staff and sea-going radio officers who will be responsible for fitting or maintenance of echo-sounding equipment, but also to ships' navigating officers."

621.396.93 2227

Radio Research Special Report No. 22. Siting of Direction Finding Stations. [Book Notice]—W. Ross and F. Horner. Publishers: H. M. Stationery Office, London, Eng., 1951, 42 pp., 1s 6d. (*Govt Publ.* (London, p. 26; February, 1952.)

621.396.93 2228

Funkpeiler, Grundlagen und Anwendungen (Direction Finders, Fundamentals and Applications). [Book Review]—H. Gabler. Publishers: Deutsches Hydrographische Institut, Hamburg, Ger., 1951, 70 pp., 5.40 DM. (*Fernmeldetech. Z.*, vol. 5, p. 43; January, 1952.) A summarized treatment of the essentials of all the important theoretical and technical problems of radio direction finding in ship navigation.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2229

The Ionization Gauge—Two Modifications—J. H. Burrow and E. W. J. Mitchell. (*Jour. Sci. Instr.*, vol. 29, pp. 27–28; January, 1952.) Description of a gauge forming part of the main pumping line, giving increased sensitivity; and of a modification permitting determination of the time taken to contaminate a surface.

535.323+535.341]:546.24:539.234 2230

Optical Properties of Tellurium in the Infra-Red—T. S. Moss. (*Proc. Phys. Soc.*, vol. 65, pp. 62–66; January 1, 1952.)

535.343.2–15:[546.28+546.289 2231

Far-Infrared Transmission of Silicon and Germanium—R. C. Lord. (*Phys. Rev.*, vol. 85, pp. 140–141; January 1, 1952.)

535.37 2232

Electron Levels in Phosphorescent Crystals with Fe, Ni or Co Poison Centres—N. Arpiarian and D. Curie. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 75–77; January 2, 1952.)

537.311.33:546.28 2233

The Drift Mobility of Electrons in Silicon—J. R. Haynes and W. C. Westphal. (*Phys. Rev.*, vol. 85, p. 680; February 15, 1952.) Measurements were made using a pulse-injection technique adapted from that described by Haynes and Shockley (1928 of 1951) for investigating Ge. An average value of 1,210 cm/s per V/cm was obtained. This is more than four times as great as the value obtained from Hall-effect data by Pearson and Bardeen (*Phys. Rev.*, vol. 75, p. 865; 1949), using multicrystal specimens; in the present experiments single crystals were used.

537.311.33:546.289 2234

Rectification Phenomena and Transistor Action in Germanium—P. Aigrain. (*Ann. Phys.* (Paris), vol. 7, pp. 140–184; January /February, 1952.) Full account of experiments carried out and theory developed, which were briefly reported in 1305, 1306, 1310 and 1818 of 1950, and 1529 of 1951.

537.311.33:546.289 2235

Properties of Thermally Produced Acceptors in Germanium—C. S. Fuller, H. C. Theuerer and W. van Roosbroeck. (*Phys. Rev.*, vol. 85, pp. 678–679; February 15, 1952.) Brief account of experiments extending the investigation of the effects of heat treatment on *n*-type Ge [2201 of 1951 (Theuerer and Scaff)]. Results indicate that (*a*) the *n*-to-*p* type conversion is characterized by the diffusion of a *p*–*n* boundary from the surface of the specimen to the interior, (*b*) the concentration of acceptor centers approaches an equilibrium value dependent on the heating temperature over the range from about 550°C to the melting point within which the conversion is possible.

538.221 2236

Properties of Ferromagnetic Powders at Frequencies up to 24 kMc/s—B. Pistoulet. (*Ann. Télécommun.*, vol. 7, pp. 27–45, 85–97 and 127–138; January-March, 1952.) Measurements of the complex permeability of various metal powders embedded in dielectric material are reported for a wide range of frequencies. The preparation of the mixtures and the measurement methods and apparatus are described. A study is made of the variations of permeability as a function (*a*) of the proportion of magnetic powder in the test samples, (*b*) of the frequency. The results obtained enable the permeability characteristics of the powders themselves to be deduced. A general method for the study of magnetic resonance under the influence of a constant magnetic field is described, with results obtained at wavelengths of 3.2 and 1.25 cm. 52 references.

538.248 2237

Investigation of the After-Effect Constant over the Total Hysteresis Range—J. C. Barbier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 415–417; January 21, 1952.)

539.23:546.72:537.311.31 2238

Measurements on Thin Iron Films—A. van Itterbeek, L. de Greve and F. Heremans. (*Appl. Sci. Res.*, vol. B2, no. 4, pp. 320–324; 1952.) Resistivity thickness curves for sputtered and evaporated iron films are found to have different shapes, due to differences of texture revealed by electron-microscope photographs.

539.231:546.57 2239

The Structure of Sputtered Silver Films—C. E. Ells and G. D. Scott. (*Jour. Appl. Phys.*, vol. 23, pp. 31–34; January, 1952.) Electron-microscope study shows that sputtered films have a more continuous structure than evaporated films prepared at the same rate, but if the evaporated films are produced at much faster rates the thinner sputtered films are more continuous and the thicker less continuous than evaporated films of the same thickness.

539.234 2240

Vacuum Evaporation Equipment with High Pumping Speed—R. Bernard and F. Davoine. (*Le Vide*, vol. 7, pp. 1136–1138; January, 1952.) Increased fineness of film structure is achieved, since the pumping speed of 200 l/sec is sufficient to maintain pressure at about 10^{-6} mm Hg.

546.561.221:[537.311.33+538.632 2241

The Electrical Conductivity and Isothermal Hall Effect in Cuprous Sulphide, Semiconductor—E. Hirahara. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 428–437; November/December, 1951.)

546.561.221:537.311.33 2242

The Physical Properties of Cuprous-Sulphide Semi-conductors—E. Hirahara. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 422–427; November/December, 1951.)

621.315.612.4:546.431.824–31 2243

Properties of Powdered $BaTiO_3$—V.E. Derr and M. D. Earle. (*Phys. Rev.*, vol. 85, pp. 384–385; January 15, 1952.) Discharge curves for capacitors with dielectric of tightly packed powdered $BaTiO_3$ are shown and discussed. The change of currect with time differs from that of the usual type of capacitor, the terminal voltage falling quickly at first and then more slowly for many minutes.

621.315.612.4:546.431.824–31 2244

Some Factors Influencing the Dielectric Properties of Barium Titanates—W. R. Eubank, F. T. Rogers, Jr., L. E. Schilberg and S. Skolnik. (*Jour. Amer. Ceram. Soc.*, vol. 35, pp. 16–22; January, 1952.) The permittivity of ceramics of various grades of purity and fired at various temperatures was determined as a function of temperature in the region of the Curie point. The temperature coefficient of permittivity is largest for the purer titanates. The dependence of the Curie point on firing temperature, cooling rate and impurity content is described. The optimum firing temperature appears to be about 1,400°C. Contact with Pt or ZrO_2 during firing may be detrimental.

621.315.612.4.011.5:546.431.824–31 2245

Adiabatic Thermal Changes in Barium Titanate Ceramic at Low Temperatures—R. W. Schmitt. (*Phys. Rev.*, vol. 85, pp. 1–4; January 1, 1952.) Measurements at 4°K indicate that the polarization process in the ceramic is thermodynamically irreversible; this accounts for the large variation of dielectric constant with temperature at 4°K.

621.318.1 2246

Magnetic Cores and Sheaths in the Field of Telecommunications—P. M. Prache. (*Câbles & Trans.* (Paris), vol. 6, pp. 22–64 and 124–164; January and April, 1952.) Detailed analysis of the characteristics of ribbon, wire and powder cores and of magnetic loading sheaths for cables, with all the relevant formulas for apparent permeability, eddy-current and hysteresis losses, etc. Calculated values of apparent permeability of powder cores suitable for Pupin coils are in good agreement with measured values. A method of cable loading by means of half-sheaths formed from insulated high-permeability wires enables operation at frequencies up to 200 kc. 60 references.

621.318.33 2247

The Use of Current Sheets in the Design of Magnets to give Bounded Fields of Required Form, free from Edge Distortion—H. O. W. Richardson. (*Proc. Phys. Soc.*, vol. 65, pp. 5–14; January 1, 1952.)

621.791.343 2248

Welding and Soldering Aluminum—S. Freedman. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 47, pp. 16, 18; February, 1952.) Using Chemalloy, a compound of various metals and chemicals, Al or any Zn-base metal can be soldered or welded simply by heating to about 800°F and applying a rod of the alloy.

666.1.037:539.319 2249

Theory of Stresses in Glass Butt Seals—N. L. Svenson: M. Zaid: H. Rawson. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 30–32; January, 1952.) Comments on 3032 of 1951 and author's reply.

MATHEMATICS

512.99 2250
Theory and Applications of Wave Vectors—F. Dahlgren. (*Acta polyt.* (Stockholm), 68 pp.; 1951.) Rules for the mathematical treatment of such vectors are given, as well as physical definitions of quantities which can be conveniently represented by them, these quantities mainly relating to electrical machines.

512.99:519.21 2251
The Probability Distribution of the Phase of the Resultant Vector Sum of a Constant Vector Plus a Rayleigh Distributed Vector—K. A. Norton, E. L. Schultz and H. Yarbrough. (*Jour. Appl. Phys.*, vol. 23, pp. 137–141; January, 1952.) Formulas, tables, and graphs are given of the cumulative probability distribution of a function frequently occurring in the theory and practice of radio wave propagation as well as in the study of the influence of noise in phase modulation systems.

517.63 2252
On Approximate Expressions for the Exponential Integral and the Error Function—R. Bellman. (*Jour. Math. Phys.*, vol. 30, pp. 226–231; January, 1952.)

681.142 2253
A Simple Electronic Digital Computer—W. L. van der Poel. (*Appl. Sci. Res.*, vol. B2, no. 5, pp. 367–400; 1952.) Description of a computer which has been simplified to the utmost practical limit at the sacrifice of speed, with examples showing the use made of sub-programs in its operation.

681.142 2254
A Direct-Current Network Analyzer for Solving Wave-Equation Boundary-Value Problems—G. W. Swenson, Jr, and T. J. Higgins. (*Jour. Appl. Phys.*, vol. 23, pp. 126–131; January, 1952.)

681.142 2255
New Techniques on the Anacom-Electric Analog Computer—E. L. Harder and J. T. Carleton. (*Trans. Amer. IEE*, vol. 69, pp. 547–556; 1950.) The direct-analogue method used in the Anacom is outlined and a description given of the sigma amplifier, which combines adding, integrating, delay, and other operations and results in improved computing technique. Various applications of the equipment are described.

681.142:512.25 2256
New Principle of Construction of Machines for Solution of Systems of Linear Equations by Electrical Analogy—D. Mitrovic, R. Huron and R. Tomovic. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 589–591; February 4, 1952.)

681.142:517.942.9 2257
Three-Dimensional Electrical Potential Analyser—V. E. Gough: S. C. Redshaw. (*Brit. Jour. Appl. Phys.*, vol. 3, p. 58; February, 1952.) Comment on 1358 of June and author's reply.

681.142:621.3.042.14/.15 2258
A Coincident-Current Magnetic Memory Cell for the Storage of Digital Information—W. N. Papian. (Proc. I.R.E., vol. 40, pp. 475–478; April, 1952.) Binary information can be stored in small ferromagnetic cores, three-dimensional arrays of which may be built up so that "writing" or "reading" of a desired unit may be effected by exciting the appropriate co-ordinate lines. Criteria for core materials are set up and experimental results with some selected materials are described.

MEASUREMENTS AND TEST GEAR

53.081.4 2259
Fundamental Considerations regarding the Use of Relative Magnitudes—J. W. Horton. (Proc. I.R.E., vol. 40, pp. 440–444; April, 1952.) There are two number systems, conforming concurrently to the decimal system and related by the basic quantity $10^{0.1}$, by which relative magnitudes may be evaluated. The term "logit" is suggested for the quantity $10^{0.1}$, which plays a similar part in computations dealing with relative magnitudes to that of the unit in computations involving absolute magnitudes. Methods of using the logit are outlined and the resulting advantages are discussed.

535.322.1.029.65:537.228.5 2260
Development of a Spectroscope for Millimetre Waves—É. Roubine. (*Rev. Tech. Comp. franç. Thomson-Houston*, no. 16, pp. 21–44; December, 1951.) An instrument for the *K* and *J* bands.

621.3.018.41 (083.74) 2261
The Transmission of Time Signals and Standard Frequencies by the I.E.N. [Istituto Elettrotecnico Nazionale Galileo Ferraris], Turin—C. E. (*Alta Frequenza*, vol. 20, pp. 219–223; October, 1951.) A weekly experimental service with 300-w power commenced on May 15, 1951. 5-mc transmissions are made every Tuesday from 0900 to 1200 and from 1400 to 1700 (C.E.T.), each hour being subdivided into five-minute periods, with time signals alternating with either 440-cps or 1,000-cps modulated signals. Spoken announcements are made at the beginning of each hour and Morse-code announcements every ten minutes. A horizontal dipole antenna is used. Over short ranges frequency is guaranteed to within ± 2 parts in 10^{-8} and time to within $\pm$ 25ms.

621.3.018.41(083.74) 2262
Frequency Multiplier giving a 1000-c/s Signal Synchronized by a Pendulum Chronometer—P. Parcelier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 190–191; January 7, 1952.)

621.317.3.027.3:621.385.3 2263
Some Properties and Applications of the Inverted Triode—A. Rogozinski and J. Weill. (*Jour. Phys. Radium*, vol. 13, no. 2, supplement, pp. 28A–30A; February, 1952.) The main application is in the direct measurement of high voltages, for which a bridge-type circuit with two inverted triodes is suitable. The primary range 5–200 v can be extended to about 5 kv by use of a high-resistance bridge. The I/V characteristic of a Type-100Th inverted triode suitable for use up to about 20 kv is shown, and a megohmmeter-voltmeter described which can also be used for the measurement of very low currents passed through a high resistance.

621.317.328:621.384.62† 2264
Determination of Field Strength in a Accelerator Cavity—L. C. Maier, Jr, and J. C. Slater. (*Jour. Appl. Phys.*, vol. 23, pp. 78–83; January, 1952.) One theoretical and two experimental methods are described for determining the accelerating field in the M.I.T. linear-accelerator cavity in terms of the input power. One experimental method is based on measurement of the power leaking out through a small hole in the end wall of the cavity. The other method depends on perturbation of the resonance frequency of the cavity by a small conducting sphere located on the axis. The three methods give consistent results.

621.317.328:621.396.611.4 2265
Field-Strength Measurements in Resonant Cavities—L. O. Maier, Jr, and J. C. Slater. (*Jour. Appl. Phys.*, vol. 23, pp. 68–77; January, 1952.) The perturbation of the resonance frequency of a cavity due to insertion of ellipsoidal objects is calculated for objects of needle, sphere, and disk types. The perturbations for the three types of object depend on different components of the electric and magnetic fields, and by making measurements with all three it is theoretically possible to measure all the field components. Experimental verification of the theory was satisfactory for spheres and disks, but for needles the perturbation is very sensitive to needle shape and the needles used were not accurate enough ellipsoids to give satisfactory quantitative results.

621.317.333.4.015.7:621.315.212 2266
A Pulse-Echo Test Set for the Quality Control and Maintenance of Impedance Uniformity of Coaxial Cables—E. Baguley and F. B. Cope. (*P.O. Elec. Engrs' Jour.*, vol. 44, pp. 164–168; January, 1952.) The equipment, designed primarily for use on 0.375-inch coaxial cables can detect impedance irregularities and differences in end impedance of the order of 0.05 per cent. Impedance irregularities and mismatches can be located to within 1 per cent on a direct-reading scale of yards or meters. A technique for cable testing during manufacture and installation is also described.

621.317.333.8:621.392.5 2267
Impulse Measurements by Repeated-Structure Networks—C. L. Dawes, C. H. Thomas and A. B. Drought. (*Trans. Amer. IEE.*, vol. 69, pp. 571–580; 1950. Discussion, pp. 580–583.) Analysis shows that the ladder type of network considered possessses the ideal characteristic of an attenuating network for passing any kind of signal without distortion. Experimental results on a simple T structure indicate that a practical hv divider of high accuracy may be constructed. An appendix gives the Laplace-transform solution for the parallel-T type of ladder network.

621.317.335.3+621.317.372 2268
An Improved Method of Measuring Dissipation Factor and Dielectric Constant Using the Susceptance Variation Principle—C. F. Miller and F. G. Whelan. (*Trans. Amer. IEE*, vol. 69, pp. 491–497; 1950.) Full paper. Summary abstracted in 2277 of 1950.

621.317.335.3.029.64:546.212–13 2269
The Dielectric Constant of Water Vapor in the Microwave Region—G. Birnbaum. (*Jour. Appl. Phys.*, vol. 23, pp. 220–223; February, 1952.) Using a cavity method described previously [1426 of 1951 (Birnbaum *et al.*)], measurements were made over the temperature range 32–103°C at a frequency of 9.28 kmc and at the single temperature of 24.5°C at 24.8 kmc. Results are discussed.

621.317.335.3.029.64.012.3 2270
New Chart for the Determination of the Permittivity of Dielectrics at U.H.F.—A. Lebrun. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 518–520; January 28, 1952.) Description of the construction and use of a chart applicable to the method of measurement in which a short-circuited section of waveguide is filled with the dielectric and measurements are made of the width of the resonance curve or of the voltage swr.

621.317.35 2271
Analysers for Aperiodic Phenomena—G. Francini. (*Alta Frequenza*, vol. 20, pp. 247–261; December, 1951.) Direct and indirect electrical methods available for low frequencies are discussed and possible simplifications in design of apparatus are considered from the point of view of their effect on performance. Souces of possible error are discussed.

621.317.361 2272
Frequency Measurement by the Capacitor Charge Method: Part 2—H. Weidemann. (*Arch. tech. Messen*, no. 192, pp. 11–14; January, 1952.) Conditions that must be satisfied for very rapid frequency variations to be followed by the meter, and practical limitations of the method are considered. Alternative arrangements are described in which stability of operation need be ensured only in the auxiliary devices used. These involve either multivibrator, thyratron or transitron circuits. Part 1: 732 of April.

621.317.373 2273
Phase Measurements—J. Henry. (*Radio franç.*, nos. 1 and 2, pp. 13–22 and 9–17; January and February, 1952.) Discussion of the functions of the various components of phase meters, and also of complete phase meters of different types.

621.317.431 2274
A Two-Fluxmeter Method of Measuring Ferromagnetic Hysteresis Loss—H. Aspden. (*Jour. Sci. Instr.*, vol. 29, pp. 5–7; January, 1952.) The time required for hysteresis-loop measurements can be greatly reduced by using a separate fluxmeter to integrate successive capacitor discharges, which are initiated by a chosen change in flux density and are proportional to the magnetizing field.

621.317.44 2275
Magnet-Steel Test Unit with Fluxmeter Compensated for Controlling Force—E. Steingroever. (*Arch. Elektrotech.*, vol. 40, no. 5, pp. 275–279; 1952.)

621.317.444:621.396.645.35:621.317.755 2276
An Integrating Amplifier for the Oscillographic Recording of Magnetic Flux—S. Ekelöf, L. Bengtson, G. Kihlberg and P. Leithammel. (*Acta polyt.* (Stockholm), no. 98, 23 pp; 1951.) Theory of operation and circuit details of a push-pull integrating amplifier consisting of a dc amplifier with capacitive feedback. Precautions are taken to obtain stable gain and low output-voltage drift. The instrument was designed for use in obtaining oscillographic records of transient magnetic fluxes lasting 0.01–0.5 second, the flux change being 500–5,000 maxwells.

621.317.6.029.3:621.317.755 2277
Automatic Audio-Frequency Response-Curve Tracer—(*Radio tech. Dig.*, *Édn franç.* vol. 5, pp. 339–347; 1951; and vol. 6, pp. 77–87; 1952.) Adaptation of articles by Hamburger (148 of 1949 and 1047 of May), with supplementary data from other sources.

621.317.7:621.3.015.7 2278
A Video Probe—R. R. Rathbone. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 47, pp. 16, 27; January, 1952.) Description of a probe suitable for testing pulse equipment; it comprises an attenuator and cathode follower feeding a terminated coaxial cable with a characteristic impedance of 93Ω. The cable may have any length up to 100 feet without introducing reflections.

621.317.7.088 2279
Scale-Overlap Errors and Frequency Errors in Instruments with Barrier-Layer Rectifiers—J. Hajek. (*Arch. tech. Messen*, no. 192, pp. 19–22; January, 1952.) Scale-overlap errors are calculated from the shape of the rectifier characteristic, (*a*) taking account of and (*b*) neglecting back current. Frequency errors and frequency dependence in Graetz full-wave rectifier circuits are also considered.

621.317.715:523.723 2280
Brownian Fluctuations in Galvanometers and Galvanometer Amplifiers—R. V. Jones and C. W. McCombie. (*Phil. Trans. A*, vol. 244, pp. 205–230; January 24, 1952.) The effects of molecular bombardment of the galvanometer mirror and of noise in the circuit resistance are studied, using the correlation function of the random force. The inclusion of circuit inductance is shown to cause no change in the rms values of the deflection and angular velocity. The magnitude and correlation function of the fluctuations in a galvanometer amplifier can be obtained from the results of a simple experiment. Close agreement has been obtained between theoretical and experimental values.

621.317.715.082.742 2281
Use of a Moving-Coil Galvanometer to measure the Mean Charging Current of a Periodically Discharged Capacitor—R. Legros. (*Ann. Phys.* (Paris), vol. 7, pp. 5–29; January/February, 1952.) The galvanometer and capacitor form part of an electronic frequency meter (3187 of 1948 and 1899 of 1949). When frequencies are to be measured which are close to the natural frequency of the moving parts, oscillations are set up. These are discussed in detail and the resultant reading errors evaluated, various types of scale graduation being considered.

621.317.715.082.742 2282
Moving-Coil Galvanometers with High Voltage Sensitivity. Applications to Problems of Biological Physics—J. Coursaget. (*Ann. Phys.* (Paris), vol. 7, pp. 30–90; January/February, 1952.)

621.317.729:537.291 2283
Automatic Tracer for Electron Trajectories and its Use for Determining the Current Lines in an Electrolyte Trough—J. Marvaud. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 45–47; January 2, 1952.)

621.317.729:538.311 2284
The Extension of the Electrolytic Tank Method to the Study of Magnetic Fields due to Iron-clad Current Sheets in Three Dimensions—H. O. W. Richardson. (*Proc. Phys. Soc.*, vol. 65, pp. 15–18; January 1, 1952.)

621.317.75:621.396.615.14 2285
Sweep-Frequency Generator for U.H.F. Television Band—J. A. Cornell and J. F. Sterner. (*Tele-Tech*, vol. 11, pp. 38–40, 88; February, 1952.) Description of an instrument designed for laboratory investigations of the characteristics of filters, tuning units and other components used in the 470–890-mc band. It comprises a sweep-frequency oscillator modulated by a vibrating mechanism, a variable-frequency marker oscillator, a crystal calibrator providing 1-mc check points, and an arrangement of mixers which superimposes marker and calibration pips upon the response curves displayed on a cro.

621.317.755:621.317.74.018.782.4† 2286
A Scanner for Rapid Measurement of Envelope Delay Distortion—L. E. Hunt and W. J. Albersheim. (Proc. I.R.E., vol. 40, pp. 454–459; April, 1952.) Description of cro equipment for display of the envelope delay/frequency characteristic of a transmission system. It has proved very useful for measurements on the TD-2 radio-relay system and for adjustment of the equalizers used in the system.

621.383:621.396.645.35 2287
Photoelectric Amplifiers—A. Schaller. (*Arch. tech. Messen.*, no. 193, pp. 43–46; February, 1952.) Basic principles are outlined of photocell applications for the measurement and amplification of very small voltages and currents, and short descriptions are given of amplifiers developed by Leo and Hübner, Lawson, Gall, and Siemens and Halske A. G. 18 references.

621.385.001.4 2288
Dynamic Measurements on Receiving Valves—A. J. Heins. (*Jour. Brit. IRE*, vol. 12, pp. 63–68; January, 1952.) Three sets of direct-reading dynamic test equipment, designed for measurements of equivalent noise resistance, power output and distortion, and cross-modulation, are described. The tests can be made by relatively unskilled operators on batches of receiving tubes in quantity production.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.771 2289
A New R.P.M. Indicator—(*Overseas Eng.*, vol. 25, pp. 250–251; February, 1952.) Description of a direct-reading pulse-counter tachometer for measuring the rotational speed of prime movers over the range 500–10,000 rpm accurate to within ±1 rpm. An electrically maintained tuning fork or low-frequency crystal serves as time standard.

535.336.2.05:621.389 2290
Defects in the Mass Spectrometer due to Space Charge and Magnetic-Field Saturation—E. W. Becker and W. Walcher. (*Z. Phys.*, vol. 131, pp. 395–407; February 19, 1952.)

536.587:537.312.6:621.316.86 2291
Temperature Regulation using Thermistors—N'Guyen Thien-Chi and J. Suchet. (*Ann. Radioélect.*, vol. 7, pp. 75–77; January, 1952.) Three materials are available for covering the range 100°C–1,150°C in three stages. Simple but rugged commercial equipment is described for controlling one or more furnaces simultaneously, the thermistors operating relays directly.

538.569.2.047.029.63./64 2292
Microwaves in Medical and Biological Research—J. E. Roberts and H. F. Cook. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 33–40; February, 1952.) Recent work on the absorption of radiation in the frequency range 1–32 kmc is reviewed, with particular reference to materials of biological interest. 37 references.

621.316.7 2293
The Cranfield Conference on Automatic Control, 16th–21st July 1951—J. Loeb. (*Ann. Télécommun.*, vol. 7, pp. 17–26; January, 1952.) A review of the proceedings, with short summaries of the most important papers presented. See also 1061 of May (Tustin).

621.384.62† 2294
Experimental Study of a Waveguide Electron Accelerator—J. Vastel. (*Ann. Radioélect.*, vol. 7, pp. 20–33; January, 1952.) Full details of the design, construction and testing of the prototype linear accelerator section described by Grivet and Vastel (1738 of 1951) in *Compt. Rend. Acad. Sci.* (Paris), where the magnetron peak power was incorrectly given as 0.5 mw instead of 0.5 mw.

621.384.62† 2295
The M.I.T. Linear Electron Accelerator—P. T. Demos, A. F. Kip and J. C. Slater. (*Jour. Appl. Phys.*, vol. 23, pp. 53–65; January, 1952.)

621.384.62†:531.314.3 2296
Particle Dynamics in the Linear Accelerator—J. R. Terrall and J. C. Slater. (*Jour. Appl. Phys.*, vol. 23, pp. 66–68; January, 1952.)

621.384.62†:621.317.328 2297
Determination of Field Strength in a Linear-Accelerator Cavity—Maier and Slater. (See 2264.)

621.385.833 2298
Some Methods for Determination of the Field on the Axis in Electron Optics—F. Bertein. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 417–419; January 21, 1952.)

621.385.833 2299
Trajectories in Electron Lenses: a Method of Approximation—F. Bertein. (*Jour. Phys. Radium*, vol. 13, supplement, pp. 41A–49A; February, 1952.

621.385.833 2300
A New Mathematical Model of an Electron Lens—P. Grivet. (*Jour. Phys. Radium*, vol. 13, supplement, pp. 1A–9A; February, 1952.) Full paper. See 1394 of June.

621.385.833 2301
Cardinal Parameters of a New Model of an Electron Lens—P. Grivet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 73–75; January 2, 1952.) Simple formulas are given to facilitate use of the mathematical model previously described (1394 of June) and values of the cardinal parameters are tabulated.

621.385.833 2302
Theory of the Electrostatic [electron] Lens formed by Two Coaxial Cylinders—P. Grivet and M. Bernard. (*Ann. Radioélect.*, vol. 7, pp. 3–9; January, 1952.) Mathematical theory leading to simple formulas for calculating parameters. See also 1393 of June.

621.385.833 2303
Numerical Ray-Tracing in Electron Lenses—J. C. Burfoot. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 22–24; January, 1952.) A simple step-by-step method is described whose accuracy can be increased to any desired extent without increasing the complexity. A numerical example is given for a strong es lens.

621.385.833 2304
Characteristics of the Hot-Cathode Electron-Microscope Gun—M. E. Haine and P. A. Einstein. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 40–46; February, 1952.)

621.385.833 2305
Summarized Proceedings of a Conference on Electron Microscopy—St. Andrews, June 1951—D. G. Drummond and G. Liebmann. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 25–29; January, 1952.)

621.385.833:061.3 2306
Proceedings of the Electron Microscope Society of America—(*Jour. Appl. Phys.*, vol. 23, pp. 156–164; January, 1952.) Summaries are given of 44 papers presented at the annual meeting of the society in Philadelphia, November 1951.

621.385.833:537.291 2307
A Reduced Equation for the Trajectories in an Electron Mirror—M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 606–608; February 4, 1952.)

621.387.462:549.211 2308
Differences between Counting and Non-Counting Diamonds—G. P. Freeman and H. A. van der Velden. (*Physica*, vol. 18, pp. 1–19; January, 1952.)

621.398 2309
The Radio-Controlled Aircraft Winner of the International Contest 1950—A. Wastable. (*TSF et TV*, vol. 28, pp. 11–12; January, 1952.) Brief description of the telecontrol system. A superregenerative receiver operates a master relay according to the pulse sequence transmitted. The airborne equipment weighs 750g.

681.177 2310
An Electronic Digital Recording Machine—the SETAR—N. T. Welford. (*Jour. Sci. Instr.*, vol. 29, pp. 1–4; January, 1952.) Description of the design and principles of operation of a "serial event timer and recorder" developed for studying human performance. Events are recorded in sequence in digital code on standard teleprinter tape. An event is defined as the making or breaking of one or more input circuits. A continuously running generator provides timing pulses at 100 per second or 10 per second.

PROPAGATION OF WAVES

538.566 2311
An Integral-Equation Approach to the Problem of Wave Propagation over an Irregular Surface—G. A. Hufford. (*Quart. Appl. Math.*, vol. 9, pp. 391–404; January, 1952.) Theoretical discussion of the propagation of radio waves over a surface whose radius of curvature is everywhere much larger than a wavelength. It is assumed that a scalar wave phenomenon is involved and that a homogeneous boundary condition applies at the surface. An integral equation is derived for the attenuation function at all points on the earth's surface and a formal solution for the field at any point above the earth is obtained. The analysis is applied to the special cases of a plane earth and a spherical earth. Agreement with the earlier work of Norton (33 of 1938), van der Pol and Bremmer (3102 of 1938), and Fock (2891 of 1947) is noted.

538.566.2 2312
Determination of the Fine Structure of the Dielectric Constant in a Slightly Heterogeneous Layer by Reflection Measurements—G. Eckart. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 309–311; January 14, 1952.) The method of analysis previously described (1882 of 1951) for determining the variation of the dielectric constant across the layer demands an impossibly high experimental accuracy; Bremmer's method (205 of 1950), using the WKB approximation, is preferable. Integration of the function derived for the reflected signal leads to an integral equation which is solved by a Fourier transformation. The analysis is performed for a plane incident wave, and the modification necessary for the case of a spherical wave is indicated.

621.396.11+535.222 2313
A New Determination of the Velocity of Electromagnetic Radiation by Microwave Interferometry—K. D. Froome. (*Nature* (London), vol. 169, pp. 107–108; January 19, 1952.) The free-space phase velocity of waves of frequency 24 kmc has been determined by means of apparatus which is the microwave equivalent of the Michelson optical interferometer. The apparent wavelength in air was observed by movement of a reflector through a distance corresponding to an exact integral number of energy minima at the detector, and could be determined from a single experiment with an accuracy to within ± 3 parts in 10^6, the total displacement of the reflector being about 162 cm. Frequency was determined by comparison with a high harmonic of a standard quartz oscillator. The results, when referred to vacuum conditions, gave $c_0 = 299,792.6 \pm 0.7$ km per second.

621.396.11 2314
Oblique Reflexion of Radio Waves by Way of a Triangular Path—J. H. Meek. (*Nature* (London), vol. 169, p. 327; February 23, 1952.) Traces due to waves reflected first from the F layer and then from an E_s cloud are shown. From a series of records obtained at 15-second intervals, a speed of about 330 km per hour was calculated for E_s clouds.

621.396.11:551.510.535 2315
Application of the Appleton-Hartree Formula to the Determination of Phase Path of an Electronmagnetic Wave in the Ionosphere—É. Argence. (*Compt. Rend. Acad. Sci.* (Paris) vol. 234 pp. 456–458; January 21, 1952.) A method is described of determining the phase path without the use of the generalized magneto-ionic theory. The relation with Booker's method (422 of 1939) is indicated. Special cases for which the formulas become simplified are (*a*) east-west propagation, (*b*) propagation at the equator, (*c*) propagation at the poles. For the area round the poles the correction necessary to the usual predicted frequencies is large. The theory is applicable to the propagation of wave packets as discussed by Booker (714 of 1950).

621.396.11:621.392 2316
Transmission Lines as Models for the Study of Electromagnetic Wave Propagation in One Dimension—L. Lunelli. (*Alta Frequenza*, vol. 20, pp. 179–199 and 262–282; October and December, 1951.) Two formal analogies are established between Maxwell's equations for propagation in a medium with constant parameters and the equations for propagation along a transmission line. The first relates electric field intensity to line voltage and magnetic field intensity to line current. The second inverts these relations. Similitude ratio and conditions are developed for both analogies, and tables of parameters are provided for the three types of line considered viz., twin solid or stranded conductors, and coaxial cables. The practical design of models is explained in detail, the limitations, errors and difficulties involved being fully treated. Tables summarize possible solutions in various typical cases.

621.396.11.029.56:551.510.535 2317
Reflection of Short Waves at Heights less than 100 km—W. Dieminger and A. E. Hoffmann-Heyden. (*Naturwiss.*, vol. 39, pp. 84–85; February, 1952.) Waves of wavelength from 75 to 200 m are reflected at heights of 75–100 km. A diurnal variation of the reflection height occurs, with a minimum about midday. The height is practically independent of the frequency used. The echo amplitude varies irregularly with an average period of several seconds without corresponding reflection-height variations. Echoes are strongest in the daytime, certainly in winter, and they are observed on days when ionospheric absorption is particularly high. The characteristics of the reflecting layer concerned are discussed.

621.396.81 2318
An Improved Method for the Calculation of the Field Strength of Waves Reflected by the Ionosphere—K. Bibl, K. Rawer and E. Theissen. (*Nature* (London), vol. 169, pp. 147–148; January 26, 1952.) In previous calculations the blanketing effect of the E layer has been assumed to occur sharply at a given frequency; because of refraction and selective absorption in the E layer this effect actually takes the form of a gradual transition dependent on amplitude. Numerical values for particular transmission paths of interest have been calculated and are to be published separately.

621.396.81:621.396.65 2319
Statistics of Propagation in the 5-m Waveband for Distances greater than the Optical Range—H. Schröder. (*Frequenz*, vol. 6, pp. 20–25; January, 1952.) An analysis is made of field-strength measurements, taken over a period of a year, of the signal received at Berlin over the 213-km radio link from Bocksberg, using frequencies of 60 and 68 mc. Results for selected days and mean values for ten consecutive days in each of the four seasons are shown in charts; in general, the hours of densest telephone traffic do not coincide with the times when transmission conditions are best. Graphs show the probability of attainment of specified signal levels and signal/noise ratios. The effects of reducing the number of dipoles in the antenna array and of reducing transmitter power are discussed. Comparison is made between the measured field strengths and values calculated from theory.

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RECEPTION

2321
621.396.621:621.396.619.13+621.392.52+621.396.619.23
A Comparison between Two-Circuit Band-Pass Filters and Modulation Converters in the Riegger Circuit [ratio detector]—A. Nowak. (*Funk u. Ton*, vol. 6, pp. 75–83; February, 1952.) Discussion of the primary and secondary voltages in the two-circuit 1F band-pass filter of a particular frequency discriminator, and of the dependence of the circuit voltages on cir-

cuit damping d and coupling coefficient k. The ratio k/d is an important parameter for the design of such filters; a simple method of measuring it is described.

621.396.621.001.11 2322

Time Analysis and Filtering—J. Icole and J. Oudin. (*Ann. Télécommun.*, vol. 7, pp. 99–108; February, 1952.) The theory of the detection of information in the presence of random noise is discussed. In cases of (*a*) sinusoidal signals of uncertain frequency, and (*b*) signals in the form of coherent noise, methods based on time correlation analysis are advantageous; these include time-displacement, frequency-displacement, directional and intercorrelation methods. Integration and summation techniques based on mean values are analogous to simple frequency filtering and are less suitable in these cases.

621.396.621.54 2323

Tracking Problem in the Superheterodyne—J. Mohrmann. (*Fernmeldetech. Z.*, vol. 5, pp. 24–30; January, 1952.) A critical review of published work on the subject and explanation of a graphical method of solving the problem.

621.396.622.71:621.396.619.13 2324

Theory and Practice of the Ratio Detector—H. Marko. (*Frequenz*, vol. 6. pp. 1–10; January, 1952.) The operation of the ratio detector, in particular its amplitude-limiting action, is explained simply by substituting for the rectifier circuit a linear equivalent circuit of the type previously described (1120 of May). The method also makes it easy to estimate the effects of circuit asymmetry and deviations of component values from nominal. Results are confirmed by measurements on an actual circuit.

621.396.82:[621.396.619.11/.13 2325

Interference in F.M. and A.M. Reception due to Weak Interfering Transmitters—M. Kulp. (*Arch. elekt. Übertragung*, vol. 6, pp. 17–28; January, 1952.) New relations for Bessel functions are used to calculate the effective value of the total interference in FM reception due to an unwanted FM transmitter. Cases considered include: common and different carrier frequencies, modulated and unmodulated carriers, low-pass filter used or not used. Interference due to an unwanted AM transmitter is also considered. To enable AM and FM conditions to be compared, corresponding formulas are given for AM reception disturbed by AM or FM interference. Results are tabulated.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2326

Instantaneous Power Spectra—C. H. Page. (*Jour. Appl. Phys.*, vol. 23, pp. 103–106; January, 1952.) "The intuitive concept of a changing spectrum is discussed. The instantaneous power spectrum is defined mathematically and used to make the intuitive concepts more precise. It depends upon the past history of a signal, but not upon the future. Integration of the instantaneous power spectrum over time yields the conventional energy spectrum. The instantaneous power spectrum of a random function may be averaged over the ensemble of functions, with a resulting stochastic average instantaneous power spectrum that is equal to the conventional time average power spectrum of a stochastic process."

621.39.001.11 2327

The Evaluation of Communication and Transmission Methods in Communication Engineering—J. Piesch. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 6, pp. 13–21; January /February, 1952.) Discussion of Shannon's theory of communication (1361 of 1949).

621.395:061.3 2328

The XVIth Plenary Assembly of the C.C.I.F. (Florence, 1951)—(*Poste e Telecomunicazioni*, vol. 20, pp. 75–81; February, 1952.) Report on work done by the various commissions. See also *Fernmeldetech. Z.*, vol. 5, pp. 186–190; April, 1952, and *P.O. Elec. Engrs' Jour.*, vol. 45, part 1, pp. 35–38; April, 1952.)

621.396.4.018.78:621.396.619.13 2329

Linearity Limits of Discriminators, particularly for Wide-Band F.M. Radio Beam Links—P. Barkow. (*Fernmeldetech. Z.*, vol. 5, pp. 67–78; February, 1952.) The origins of distortion of FM in directional systems are investigated and the results of a series of measurements on a push-pull type of discriminator are presented in tables and numerous diagrams. Discussion of the principles of the process of modulation conversion in discriminators indicates that with careful design it should be possible to reduce distortion below 7.5 neper, a value which should not be exceeded for discriminators used on multichannel wide-band links.

621.396.619.16 2330

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621.396.619.16:[621.3.018.78+621.396.822 2331

Background Noise and Distortions in Code Modulation—L. J. Libois. (*Câbles & Trans.* (Paris), vol. 6, pp. 65–79; January, 1952.) Three sources of noise affecting pcm transmissions are examined: noise of external origin, such as circuit noise, disturbances due to coding errors, and quantization noise. Circuit noise at least 20 db below the signal can be regarded as negligible, so that a signal/noise ratio of 30 db should suffice when fading is taken into account. Effects of coding errors can be practically eliminated by use of a series coder, such as a binary counter. Quantization noise analysis shows that when the sampling frequency is twice the highest modulation frequency to be transmitted, as is practically the case in pulse multiplex, for which the ratio is about 2.5, all the distortion energy is found within the modulation band, and the signal/noise ratio, for 100 per cent modulation, is equal to $\sqrt{6}p$, where $2p$ is the number of quantization steps effectively used. Experimental results on the quality of telephone conversation in the presence of noise indicate that a 6-unit code system, using a series coder and compressor, should enable satisfactory quality to be obtained.

621.396.619.16:621.3.018.78 2332

Analysis of Distortion in Pulse-Code Modulation Systems—J. P. Schouten and H. W. F. Van't Groenewout. (*Appl. Sci. Res.*, vol. B2, no. 4, pp. 277–290; 1952.) Theoretical treatment of the distortion introduced by amplitude quantization. The analysis is general and permits arbitrary choice of quantization level, of sampling frequency and of signal frequency. The results are applied to the case of a sinusoidal input signal.

621.396.822:621.396.619.11 2333

On the Distribution of Energy in Noise- and Signal-Modulated Waves: Part 1—Amplitude Modulation—D. Middleton. (*Quart. Appl. Math.*, vol. 9, pp. 337–354; January, 1952.) Theoretical analysis of the spectral distribution of intensity of AM of a carrier by noise or by signal and noise; in the latter case particular attention is paid to the case of a sinusoidal modulating signal. The mean carrier power, the mean total power and the mean continuum power are deduced as functions of the noise and signal modulation indices. Some of the results are shown graphically. The effect on the spectrum of over-modulation by the signal and/or noise is discussed and illustrated qualitatively.

621.396.933 2334

Gapless Coverage in Air-to-Ground Communications at Frequencies above 50 Mc/s—K. A. Norton and P. L. Rice. (PROC. I.R.E., vol. 40, pp. 470–474; April, 1952.) There is an optimum height of ground-station antenna for an air-to-ground communication system. With heights less than the optimum, the maximum range is reduced at all aircraft altitudes. When higher antenna are used, interference between the direct and ground-reflected waves causes gaps in the coverage at the higher aircraft altitudes. The optimum antenna height decreases with increasing frequency. Sets of curves show the variations with frequency of the optimum antenna height ensuring gapless communication to the maximum range at all aircraft altitudes less than (*a*) 10,000, (*b*) 25,000, (*c*) 40,000 feet. Other curves show the maximum range for satisfactory communication at selected aircraft altitudes from 1,000 to 40,000 feet, assuming optimum ground-station antenna height.

621.396.97+621.396.677.2 2335

High Frequency Broadcast Transmission with Vertical Radiation—P. Adorian and A. H. Dickinson. (*Jour. Brit. IRE*, vol. 12, pp. 111–116; February, 1952.) The necessity for exact siting of the transmitter is avoided by using vertical transmission, with reflection from the F_2 layer, at frequencies between about 2 and 10 mc. Reasonably good reception can be maintained at ranges up to 150–200 miles in latitudes between 10°N and 20°N, with two frequencies, the higher one being used during the day time. Details are given of three suitable types of antenna.

621.396.97+621.396.975 2336

Wireless Broadcasting and Rediffusion Systems for Colonial Territories—A. Cross and F. R. Yardley. (*Jour. Brit. IRE*, vol. 12, pp. 91–109; February, 1952.) A description of the broadcasting system now installed in the colony of Trinidad. The system comprises a medium-wave service for the densely populated areas, a short-wave service for rural areas and supplementary facilities given by rediffusion programs and experimental community reception in small villages. The installation of these services and the reception of B.B.C. and U.S.A. overseas broadcast programs are discussed. Details are given of the 72-mc links used with the rediffusion services.

SUBSIDIARY APPARATUS

621–526 2337

Control-System Synthesis by Root Locus Method—W. R. Evans. (*Trans. Amer. IEE*, vol. 69, pp. 66–69; 1950.) Description, with examples, of a graphical method of determining all the roots of the differential equation of a control system; the method readily permits synthesis for a desired transient response or frequency response.

621–526 2338

A Theory of Multidimensional Servo Systems—M. Golomb and E. Usdin. (*Jour. Frank. Inst.*, vol. 253, pp. 29–57; January, 1952.)

621.385.833 2302
Theory of the Electrostatic [electron] Lens formed by Two Coaxial Cylinders—P. Grivet and M. Bernard. (*Ann. Radioélect.*, vol. 7, pp. 3–9; January, 1952.) Mathematical theory leading to simple formulas for calculating parameters. See also 1393 of June.

621.385.833 2303
Numerical Ray-Tracing in Electron Lenses—J. C. Burfoot. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 22–24; January, 1952.) A simple step-by-step method is described whose accuracy can be increased to any desired extent without increasing the complexity. A numerical example is given for a strong es lens.

621.385.833 2304
Characteristics of the Hot-Cathode Electron-Microscope Gun—M. E. Haine and P. A. Einstein. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 40–46; February, 1952.)

621.385.833 2305
Summarized Proceedings of a Conference on Electron Microscopy—St. Andrews, June 1951—D. G. Drummond and G. Liebmann. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 25–29; January, 1952.)

621.385.833:061.3 2306
Proceedings of the Electron Microscope Society of America—(*Jour. Appl. Phys.*, vol. 23, pp. 156–164; January, 1952.) Summaries are given of 44 papers presented at the annual meeting of the society in Philadelphia, November 1951.

621.385.833:537.291 2307
A Reduced Equation for the Trajectories in an Electron Mirror—M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 606–608; February 4, 1952.)

621.387.462:549.211 2308
Differences between Counting and Non-Counting Diamonds—G. P. Freeman and H. A. van der Velden. (*Physica*, vol. 18, pp. 1–19; January, 1952.)

621.398 2309
The Radio-Controlled Aircraft Winner of the International Contest 1950—A. Wastable. (*TSF et TV*, vol. 28, pp. 11–12; January, 1952.) Brief description of the telecontrol system. A superregenerative receiver operates a master relay according to the pulse sequence transmitted. The airborne equipment weighs 750g.

681.177 2310
An Electronic Digital Recording Machine—the SETAR—N. T. Welford. (*Jour. Sci. Instr.*, vol. 29, pp. 1–4; January, 1952.) Description of the design and principles of operation of a "serial event timer and recorder" developed for studying human performance. Events are recorded in sequence in digital code on standard teleprinter tape. An event is defined as the making or breaking of one or more input circuits. A continuously running generator provides timing pulses at 100 per second or 10 per second.

PROPAGATION OF WAVES

538.566 2311
An Integral-Equation Approach to the Problem of Wave Propagation over an Irregular Surface—G. A. Hufford. (*Quart. Appl. Math.*, vol. 9, pp. 391–404; January, 1952.) Theoretical discussion of the propagation of radio waves over a surface whose radius of curvature is everywhere much larger than a wavelength. It is assumed that a scalar wave phenomenon is involved and that a homogeneous boundary condition applies at the surface. An integral equation is derived for the attenuation function at all points on the earth's surface and a formal solution for the field at any point above the earth is obtained. The analysis is applied to the special cases of a plane earth and a spherical earth. Agreement with the earlier work of Norton (33 of 1938), van der Pol and Bremmer (3102 of 1938), and Fock (2891 of 1947) is noted.

538.566.2 2312
Determination of the Fine Structure of the Dielectric Constant in a Slightly Heterogeneous Layer by Reflection Measurements—G. Eckart. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 309–311; January 14, 1952.) The method of analysis previously described (1882 of 1951) for determining the variation of the dielectric constant across the layer demands an impossibly high experimental accuracy; Bremmer's method (205 of 1950), using the WKB approximation, is preferable. Integration of the function derived for the reflected signal leads to an integral equation which is solved by a Fourier transformation. The analysis is performed for a plane incident wave, and the modification necessary for the case of a spherical wave is indicated.

621.396.11+535.222 2313
A New Determination of the Velocity of Electromagnetic Radiation by Microwave Interferometry—K. D. Froome. (*Nature* (London), vol. 169, pp. 107–108; January 19, 1952.) The free-space phase velocity of waves of frequency 24 kmc has been determined by means of apparatus which is the microwave equivalent of the Michelson optical interferometer. The apparent wavelength in air was observed by movement of a reflector through a distance corresponding to an exact integral number of energy minima at the detector, and could be determined from a single experiment with an accuracy to within ± 3 parts in 10^6, the total displacement of the reflector being about 162 cm. Frequency was determined by comparison with a high harmonic of a standard quartz oscillator. The results, when referred to vacuum conditions, gave $c_0 = 299, 792.6 \pm 0.7$ km per second.

621.396.11 2314
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Analysis of Distortion in Pulse-Code Modulation Systems—J. P. Schouten and H. W. F. Van't Groenewout. (*Appl. Sci. Res.*, vol. B2, no. 4, pp. 277–290; 1952.) Theoretical treatment of the distortion introduced by amplitude quantization. The analysis is general and permits arbitrary choice of quantization level, of sampling frequency and of signal frequency. The results are applied to the case of a sinusoidal input signal.

621.396.822:621.396.619.11 2333
On the Distribution of Energy in Noise- and Signal-Modulated Waves: Part 1—Amplitude Modulation—D. Middleton. (*Quart. Appl. Math.*, vol. 9, pp. 337–354; January, 1952.) Theoretical analysis of the spectral distribution of intensity of AM of a carrier by noise or by signal and noise; in the latter case particular attention is paid to the case of a sinusoidal modulating signal. The mean carrier power, the mean total power and the mean continuum power are deduced as functions of the noise and signal modulation indices. Some of the results are shown graphically. The effect on the spectrum of over-modulation by the signal and/or noise is discussed and illustrated qualitatively.

621.396.933 2334
Gapless Coverage in Air-to-Ground Communications at Frequencies above 50 Mc/s—K. A. Norton and P. L. Rice. (PROC. I.R.E., vol. 40, pp. 470–474; April, 1952.) There is an optimum height of ground-station antenna for an air-to-ground communication system. With heights less than the optimum, the maximum range is reduced at all aircraft altitudes. When higher antenna are used, interference between the direct and ground-reflected waves causes gaps in the coverage at the higher aircraft altitudes. The optimum antenna height decreases with increasing frequency. Sets of curves show the variations with frequency of the optimum antenna height ensuring gapless communication to the maximum range at all aircraft altitudes less than (*a*) 10,000, (*b*) 25,000, (*c*) 40,000 feet. Other curves show the maximum range for satisfactory communication at selected aircraft altitudes from 1,000 to 40,000 feet, assuming optimum ground-station antenna height.

621.396.97+621.396.677.2 2335
High Frequency Broadcast Transmission with Vertical Radiation—P. Adorian and A. H. Dickinson. (*Jour. Brit. IRE*, vol. 12, pp. 111–116; February, 1952.) The necessity for exact siting of the transmitter is avoided by using vertical transmission, with reflection from the F_2 layer, at frequencies between about 2 and 10 mc. Reasonably good reception can be maintained at ranges up to 150–200 miles in latitudes between 10°N and 20°N, with two frequencies, the higher one being used during the day time. Details are given of three suitable types of antenna.

621.396.97+621.396.975 2336
Wireless Broadcasting and Rediffusion Systems for Colonial Territories—A. Cross and F. R. Yardley. (*Jour. Brit. IRE*, vol. 12, pp. 91–109; February, 1952.) A description of the broadcasting system now installed in the colony of Trinidad. The system comprises a medium-wave service for the densely populated areas, a short-wave service for rural areas and supplementary facilities given by rediffusion programs and experimental community reception in small villages. The installation of these services and the reception of B.B.C. and U.S.A. overseas broadcast programs are discussed. Details are given of the 72-mc links used with the rediffusion services.

SUBSIDIARY APPARATUS

621-526 2337
Control-System Synthesis by Root Locus Method—W. R. Evans. (*Trans. Amer. IEE*, vol. 69, pp. 66–69; 1950.) Description, with examples, of a graphical method of determining all the roots of the differential equation of a control system; the method readily permits synthesis for a desired transient response or frequency response.

621-526 2338
A Theory of Multidimensional Servo Systems—M. Golomb and E. Usdin. (*Jour. Frank. Inst.*, vol. 253, pp. 29–57; January, 1952.)

21-526 2339
Stability and Parametric Continuity of a Linear Servomechanism with Time-Varying Coefficients—J. Brodin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 800–801; February 18, 1952.)

621.314.634 2340
Studies on Selenium and its Alloys: Report 2—Some Experiments on the Rectifying Characteristics of Selenium Rectifier—T. Sato and H. Kaneko. (*Tech. Rep. Tohoku Univ.*, vol. 15, no. 1, pp. 1–10; 1950.) Results of resistance measurements lead to the conclusion that the blocking layer is a thin film of amorphous Se. Report 1: 1336 of June.

621-526 2341
Selected Government Research Reports. Vol. 5. Servomechanisms. [Book Notice]—Publishers: H. M. Stationery Office, London, Eng. 310 pp., 63s. (*Govt. Publ.* (London), p. 26; February, 1952.)

TELEVISION AND PHOTELEGRAPHY

621.397.24:621.315.212.4 2342
The Birmingham-Manchester Television Link—(*P.O. Elec. Engrs' Jour.*, vol. 44, part 4, p. 158; January, 1952.) Vision signals are relayed from Birmingham to the Holme Moss station by special coaxial cables via Telephone House, Manchester. Asymmetric sideband transmission is used, the carrier frequency being 1.056 mc. Amplifiers are installed at 6-mile intervals along the cable, which can also provide 1,200 telephone channels.

621.397.5:535.623 2343
The National Television Systems Committee Color-Television Transmission: Part 1—R. M. Bowie and B. F. Tyson. (*Sylvania Tech.*, vol. 5, pp. 10–16; January, 1952.) A general description of the N.T.S.C. system in which the necessary color information is added to the monochrome transmission by vestigial-sideband modulation of two sub-carriers in phase quadrature. See also 1750 of July (Hirsch *et al.*).

621.397.5 (083.74) 2344
Belgian Television Standards—G. Hansen. (*HF* (Brussels), vol. 2, no. 1, pp. 7–15; 1952.) The standards adopted are discussed in relation to the availability of French and Dutch programs and the cost of commercial receivers. The standards include both 625-line and 819-line definition, positive modulation with 5-mc video bandwidth, and AM for the sound channel. The increase in cost of a receiver for two channels over that of a similar single-channel receiver is small.

621.397.5(494) 2345
Television in Other Lands and Television Planning for Switzerland—(*Tech. Mitt. schweiz Telegr.-TelephVerw.*, vol. 30, pp. 19–32; January 1, 1952. In German.) A review of progress in various countries throughout the world and an outline of developments proposed for Switzerland up to 1953.

621.397.61 2346
The Du-Mitter—S. R. Patremio. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 47, pp. 3–5, 31; February, 1952.) A transmitter covering channel 2 or 3, and transforming television line signals to rf signals for feeding large groups of receivers in exhibitions, conferences, offices, etc.

621.397.611.2 2347
Electromagnetic Scanning Generators for Television—L. W. Whitaker. (*Marconi Rev.*, vol. 15, pp. 1–24; 1st Quarter 1952.) The basic problems associated with the design of circuits for obtaining the required current wave forms in the deflection coils are considered for the case of both line and frame scanning. The various types of scanning generators are classified and methods of using feedback in order to obtain a linear sweep are dealt with in some detail. A method of obtaining correction in flat-faced tubes is outlined. The design and operation of several types of complete line and frame scanning generators using feedback for linearization are described in detail.

621.397.611.2 2348
Charging of Secondary-Emission Surfaces—R. Colberg. (*Fernmeldetech. Z.*, vol. 5, pp. 56–66; February, 1952.) A short review of the phenomena, with descriptions of applications in various iconoscopes, the orthicon and image orthicon. 30 references.

621.397.611.2 2349
The Development of Storage-Type Television Camera Tubes—A. Karolus. (*Z. angew. Phys.*, vol. 4, pp. 71–77; February, 1952.) A survey with 32 references.

621.397.62:535.88 2350
The Fischer Large-Screen Projection System—E. Baumann. (*J. Brit. IRE*, vol. 12, pp. 69–78; February, 1952.) Account of latest developments of "eidophor" equipment. See also 485 of 1951.

621.397.621.2 2351
Scanning and E.H.T. Circuits for Wide-Angle Picture Tubes—E. Jones. (*J. Brit. IRE*, vol. 12, pp. 23–48; January, 1952.) A discussion of energy-recovery scanning circuits primarily designed for television receivers in which a cr tube with scanning angle of 70° is supplied from a circuit for which the line voltage is restricted to 190v. Design procedures are established, with particular attention to obtaining linearity of the trace by use of a saturable reactor with a ferrite core.

621.397.621.2 2352
Magnetic Beam-Deflection Systems—H. Bähring. (*Funk u. Ton*, vol. 6, pp. 8–21; January, 1952.) The deflection sensitivity, inductance, magnetic efficiency and over-all quality factor of various types of deflection system are discussed. A formula derived for the quality factor enables comparison to be made between different types. Of the various types considered, the cylindrical screened coil system and the ring yoke with teeth supporting the windings have the greatest efficiency and highest quality factor.

621.397.621.2 2353
The Modulation Characteristic of Cathode-Ray Tubes in Television—R. B. Mackenzie. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 54–58; February, 1952.) The ambiguity which arises in defining the modulation characteristic in terms of gamma is discussed. A method of eliminating this ambiguity is proposed which is partly empirical, but leads to a simple mathematical treatment.

621.397.621.2:535.623 2354
Color Television Reproducers—H. R. Lubcke. (*Jour. Soc. Mot. Pic. Telev. Eng.*, vol. 58, pp. 22–27; January, 1952.) Description of a three-gun tube with a heterogeneous screen of special construction using phosphors with different response times. By altering the speed of traverse of the electron beams, different colors predominate in the composite-phosphor response.

621.397.621.2(73) 2355
Cathode-Ray Tubes for Television Receivers in U.S.A.—G. G. Esculier. (*Onde élect.*, vol. 32, pp. 32–35; January, 1952.) A review of recent developments and methods adopted to ensure good quality in mass production.

621.397.645:621.385.4 2356
Coaxial Tetrode as a TVI Amplifier at V.H.F. and U.H.F.—Preist. (See 2175.)

621.397.645.018.424 2357
Wide-Band Amplifiers with Stagger-Tuned Circuits—de Vos. (See 2176.)

621.397.828 2358
F.C.C.'s Plan for Handling TVI—G. S. Turner. (*QST*, vol. 36, pp. 22–24; January, 1952.)

TRANSMISSION

621.392.52:621.396.615.029.55/.62:621.397.828 2359
Practical Applications of Pi-Network Tank Circuits for TVI Reduction—G. Grammer. (*QST*, vol. 36, pp. 10–15, 196; January, 1952.)

621.396.216 2360
Total or Partial Suppression of a Modulation Sideband—J. Oswald. (*Câbles & Trans.* (Paris), vol. 6, pp. 165–173; April, 1952.) Analysis of the characteristics of the envelope of an AM signal when the carrier wave and one sideband are suppressed. A definition is given of the mean and the maximum degree of modulation of a stationary aleatory signal with a limited spectrum and Gaussian distribution. Passage of a modulated wave through a filter is considered, the theory showing the existence of the two components in quadrature which characterize the response of an arbitrary linear network to a modulated signal. The probability law of the signal envelope and the degree of modulation are slightly modified by the suppression of a sideband, so that a compression of the envelope levels results. The theory is supplied to vestigial-sideband transmission of a television signal.

621.396.61 2361
The New Paris-Villebon 100-kW Transmitter, Foremost in the International Technical Field—P. A. François. (*TSF et TV*, vol. 28, pp. 21–24; January, 1952.) Special features of the transmitter, type monobloc TH755, are described, including the gravity-feed water cooling system for the tubes, the heat dissipation of which causes the water to boil. The transmitter is modulated at high power and radiates on 107 mc from an antenna common to a 150-kw transmission on 863 kc. Over-all efficiency unmodulated is 53 per cent.

621.396.615.016.22 2362
Curves of Equal H.F. Power of a Transmitter—P. Mourmant. (*Radio franç.*, no. 1, pp. 1–10; January, 1952.) The family of equal-hf-power curves for a transmitter, expressed in terms of load impedance, give to the "matching range" (595 and 2097 of 1948) a quite rigorous character. The matching ranges of ideal and of practical matching quadripoles are discussed, with particular reference to power tolerances and power output. Equipment for determining the values of the components of a matching quadripole is mentioned.

621.396.619.27 2363
Theory of the Cut-Off [Cowan] Modulator with Capacitor Shunt—V. Belevitch. (*HF* (Brussels), vol. 2, no. 1, pp. 1–6; 1952.) A method of analysis of circuits with periodically varying parameters is applied in determining the output function of a Cowan modulator, taking account of the capacitance of the rectifier. To a first approximation the effect of this capacitance is equal to that of the same capacitance in a circuit operating at a fixed frequency, the effect being calculated for an equivalent frequency that is a simple function of the input and carrier frequencies.

TUBES AND THERMIONICS

537.533.8+537.525.92 2364
Equilibrium between an Insulator emitting Secondary Electrons, a Space Charge, and an Enclosure at Constant Potential—M. Barbier. (*Ann. Radioélect.*, vol. 7, pp. 61–67; January,

1952.) The order of magnitude of the surface potential of the insulator to be expected for various intensities of bombardment is calculated for the ideal cases in which the insulator and the collector electrode are two parallel planes or two concentric spheres, taking account of the space charge due to the secondary electrons.

621.314.7 2365

Progress in Transistor Technique—E. H. Hungermann. (*Elektron Wiss. Tech.*, vol. 5, nos. 13/14, pp. 429–439; 1951/1952.) Summarizes publications and patents on methods of production of single crystals and on various treatments designed to improve performance. 39 references.

621.314.7 2366

Current Multiplication in the Type-A Transistor—W. R. Sittner. (PROC. I.R.E., vol. 40, pp. 448–454; April, 1952.) Discussion of the possibility of high current amplification arising from the trapping of holes in the barrier region of the collector which, due to the requirement of space-charge neutrality, leads to an enhanced electron concentration. Measurements on two transistors over the temperature range 237–298°K suggest trap densities of the order of $10^{13}/cm^3$, while trapping energies of 0.3 ev are derived both from the absolute magnitude of the trapping ratio (density of trapped holes/density of mobile holes) and from its temperature dependence.

621.314.7 2367

Transistor Forming Effects in n-Type Germanium—L. B. Valdes. (PROC. I.R.E., vol. 40, pp. 445–448; April, 1952.) An experimental study of the effects of electrically forming the collector of an *n*-type Ge transistor. It is concluded that a region of high-conductivity *p*-type material is produced under the collector, due either to lattice dislocations or to thermal diffusion of impurities. The enhanced current amplification so obtained is attributed to a *p-n* hook formed by a very small *n*-region immediately below the contact and surrounded by the larger *p*-region produced by forming.

621.383.27 2368

Research on Electron Multiplication and its Applications- Part 1—D. Charles. (*Ann. Radioélect.*, vol. 7, pp. 34–60; January, 1952.) Nine-stage and ten-stage multipliers with crossed electric and magnetic fields were studied. The elementary and the exact theory of their operation are presented, the latter involving the calculation of the equipotential surfaces between two electrodes, and the deduction of the electron trajectories. The factors essential to satisfactory performance are discussed and enumerrated. Experiments show that the sensitivity falls off rapidly below 3,000 Å, with a possible lower limit at about 2,200 Å. It is estimated that a photoelectric current of 10^{-14} a can be measured without much difficulty, the main limiting factor being background noise, which can be reduced by operation at low temperature. Comparative measurements at +28°C and −23°C with special equipment for determining the absolute limit of sensitivity gave results of 24×10^{-16} a and 1.5×10^{-16} a respectively.

621.383.27 2369

Electron Multiplier for the Ultraviolet Range down to 1450 Å—V. Schwetzoff, S. Robin and B. Vodar. (*Compt. Rend. Acad. Sci.* (Paris), vol. 234, pp. 426–428; January 21, 1952.)

621.385:537.525.92 2370

Three Elementary Cases of the Expansion of Space-Charge Clouds—H. Kleinwächter. (*Funk. u. Ton*, vol. 6, pp. 25–28; January, 1952.) Solutions are obtained for the rate of expansion of spherical, cylindrical, and plate-shaped space-charge clouds.

621.385:537.525.92 2371

Space-Charge-Wave Propagation in a Cylindrical Electron Beam of Finite Lateral Extension—P. Parzen. (*Jour. Appl. Phys.*, vol. 23, pp. 215–219; February, 1952.) The influence of tube configuration is investigated theoretically; the case of the complete-space-charge diode is discussed in detail. Results agree with the experimental findings of Cutler and Quate (1274 of 1951).

621.385.029.6 2372

Noise in Transit-Time Valves—W. Kleen. (*Frequenz*, vol. 6, pp. 45–50; February, 1952.) The noise figure of a klystron or a traveling-wave tube is governed primarily by the transit time of the electron beam. Fluctuations of density and velocity in the plane of the first accelerating electrode cause two space-charge waves of slightly different phase between this electrode and the hf input. Interaction of these waves gives rise to unwanted periodic components in the output. There is an optimum spacing of hf input and cathode for which the noise figure is a minimum. Experimental results confirm the theory.

621.385.029.6 2373

An Internal-Feedback Traveling-Wave-Tube Oscillator—E. M. T. Jones. (PROC. I.R.E., vol. 40, pp. 478–482; April, 1952.) Theory is presented which neglects space-charge effects. Experimental results for an oscillator using a helix as the interaction structure are in good agreement with the theory.

621.385.029.6 2374

Optimum Amplification in the Travelling-Wave Valve with Helix—J. Labus. (*Arch. elekt. Übertragung*, vol. 6, pp. 1–5; January, 1952.) A formula is developed giving the tube amplification in terms of helix dimensions and operating parameters, and conditions are investigated for optimum value of amplification. Two cases are distinguished: (*a*) beam current related to helix potential by the space-charge law (helix at anode potential), (*b*) beam current adjustable independently of helix potential. Higher amplification can be obtained with lower power consumption in the latter case. Amplification is proportional to the cube root of wavelength times beam power, and decreases with increasing pitch of helix; it depends also on the ratio between the diameters of beam and helix.

621.385.029.6:621.392.5 2375

The Delay Line as a Component of Valves—W. Kleen and W. Ruppel. (*Arch. Elektrotech.*, vol. 40, pp. 280–304; 1952.) Basic formulas and properties of various types of delay line for traveling-wave tubes are reviewed, and particular types of homogeneous and inhomogeneous lines are discussed, the latter type defined as consisting of a series of quadripole elements of finite axial extent and being treated in detail from the point of view of their resemblance to filters. Equivalent circuits are derived and solutions of the field equations are obtained for plane-parallel and cylindrical types of line and for the circular type of the traveling-wave magnetron, results being presented graphically to facilitate approximate determination of characteristics.

621.385.029.6.012.6(083.5) 2376

Langmuir's ξ, η Tables for the Exponential Region of the I_a-V_a Characteristic—G. Diemer and H. Dijkgraaf. (*Philips Res. Rep.*, vol. 7, pp. 45–53; February, 1952.) "The inter-electrode distances of modern microwave diodes and triodes are often so small that the normal operating point lies in the exponential part of the characteristic. A set of ξ, η tables with the voltage gradient at the anode as parameter is given from which the potential distribution in such cases can be derived."

621.385.029.64 2377

Optimum Geometry of Microwave Amplifier Valves—G. Diemer and K. Rodenhuis. (*Philips Res. Rep.*, vol. 7, pp. 36–44; February, 1952.) On the basis of van der Ziel Å Knol's theory of feedback amplifiers (1646 of 1950) it is shown that for uhf amplifier tubes the upper limit for the amplification is the highest possible if the electrode areas are so chosen that the useful capacitance equals the unavoidable parasitic capacitance. For optimum gain-bandwidth products the useful capacitance should be somewhat higher.

621.385.032.216 2378

The Growth and Properties of Cathode Interface Layers in Receiving Valves—M. R. Child. (*P. O. Elec. Engrs.' Jour.*, vol. 44, pp. 176–178; January, 1952.) A resistive interface layer tends to grow between the oxide cathode matrix and the supporting core and results in a reduction of mutual conductance by negative feedback. The effects of temperature variation, silicon concentration, and cathode current on the rate of increase of interfacd resistance are described.

621.385.032.216 2379

Emission and Crystal Size of Oxide-Coated Cathodes—J. Shimazu. (*Jour. Phys. Soc. Japan*, vol. 6, pp. 479–485; November/December, 1951.)

621.385.032.216 2380

Anomalous Distribution of the Velocities of the Electrons emitted by a Pulsed Oxide Cathode—R. Loosjes and C. G. J. Jansen. (*Le Vide*, vol. 7, pp. 1131–1135; January, 1952.) Continuation of work reported in 2067 of 1950 (Loosjes, Vink and Jansen) and back references.

621.385.032.216.011.22 2381

Contribution to the Study of the Resistance through the Oxide-Cathode Layer—C. Biguenet. (*Le Vide*, vol. 7, pp. 1123–1130; January, 1952.) Cathodes with coatings of various thicknesses were prepared, and curves of resistance/thickness at various temperatures obtained. By extrapolation, a value of 0.8Ω was obtained for the interface resistance which was practically independent of cathode temperature and emission current. The coating resistance was greater (of the order of a few ohms), varying according to thickness, temperature and mode of operation of the cathode. Explanations are advanced to account for the different experimental results obtained by various workers.

621.385.2 2382

The Electric Field in Diodes and the Transit Time of Electrons as Functions of Current—P. L. Copeland and D. N. Eggenberger. (*Jour. Appl. Phys.*, vol. 23, pp. 280–286; February, 1952.) Equations equally applicable to parallel-plane, coaxial-cylinder and concentric-sphere configurations are developed, giving approximate solutions for potential distribution and transit time. Calculations for the coaxial-cylinder arrangement indicate that the field at the cathode changes only very slowly as a function of the geometry; hence formulas derived for the parallel-plane arrangement are applicable with only small corrections to coaxial-cylinder arrangements. Functions used in the calculations are tabulated.

621.385.2 2383

Space Charge and Transit Time Considerations in Planar Diodes for Relativistic Velocities—H. F. Ivey. (*Jour. Appl. Phys.*, vol. 23, pp. 208–211; February, 1952.) "The usual Child-Langmuir equation is extended to the case of relativistic velocities for a diode with parallel plane electrodes. The solutions obtained are valid for any value of the accelerating voltage. As the variation of mass with velocity becomes important, the exponent giving the dependence of current density on anode voltage becomes less than three-halves and for very large values of accelerating potential approaches unity. The variation with anode voltage of the transit time, both for the space-charge-free case and for the space-charge-limited case, has been calculated. The potential distribution across the diode is also discussed."

621.385.2:537.525.92 2384
Space-Charge-Limited Currents between Inclined Plane Electrodes—H. F. Ivey. (*Jour. Appl. Phys.*, vol. 23, pp. 240–249; February, 1952.) The method of investigation described by Walker (1275 of 1951) is used; solutions are found for the potential distribution, space-charge characteristic ("perveance"), particle trajectories and transit time under space-charge-limited conditions for diodes with angles up to 108° between the electrodes.

621.385.2:546.289 2385
Germanium Diode Experience—(*Radio & Telev., News, Radio-Electronic Eng. Section*, vol. 47, pp. 20–21; January, 1952.) Account of a preliminary study of the performance of the 16,000 Ge diodes in the N.B.S. Eastern Automatic Computer (SEAC) during the latter half of 1950.

621.385.2:621.3.015.3 2386
Transients in Valves—H. Fack. (*Frequenz*, vol. 6, pp. 33–37; February, 1952.) A mathematical treatment of the response of a plane diode to a transient input. A simple equivalent circuit is derived which comprises resistance and inductance in parallel with a capacitance, the inductance representing electron inertia.

621.385.38 2387
Statistical Nature and Physical Concepts of Thyratron Deionization Time—H. A. Romanowitz and W. G. Dow. (*Trans. Amer. IEE*, vol. 69, pp. 368–379; 1950.)

621.385.5 2388
Circuits of the Balitron Tube—N. Z. Ballantyne. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 47, pp. 6–8, 31; January, 1952.) A beam-deflection tube with a stable negative-resistance type of characteristic is obtained by slight modifications of the arrangement and shape of the electrodes of the positive-resistance type of the "balitron" tube. Suitable oscillator and frequency-converter circuits using the modified tube are described.

621.385.832:621.318.572 2389
Ribbon Beam Valves: Principles and some Applications—G. Piétri. (*Le Vide*, vol. 7, pp. 1113–1122; January, 1952.) Theory and brief descriptions of a 10-way switching tube and a pulse generator.

621.396.615.142 2390
Retarding-Field Oscillators—J. J. Ebers. (PROC. I.R.E., vol. 40, pp. 138–145; February, 1952.) Velocity-variation oscillations are analyzed mathematically and distinguished from the Barkhausen-Kurz type. Electron paths and the processes of bunching and drifting are discussed and measurements of power output and efficiency are given for an experimental planar-type oscillator.

621.396.615.142 2391
The Variation of the H.F. Power of a Drift-Space Valve with the D.C. Power—R. Gebauer and H. Kosmahl. (*Z. angew. Phys.*, vol. 3, pp. 449–452; December, 1951.) Investigations reported in 1036 of 1951 are continued, measurements being made on a type 0+ tube (input-gap transit angle $\approx \pi$). The value of hf power as a function of dc power varies first quadratically, then linearly, and, if the beam current is made sufficiently high, finally passes through a maximum and falls to zero. The rising part of the curve is explained on the basis of the variation of efficiency with modulation depth, which in turn depends on beam current; the falling part is attributed to the effect of space charge in increasing transit time and impairing focusing.

621.396.615.142.2 2392
A Type of Reflex Klystron with Fixed Load and Wide Frequency Range—J. Laborderie. (*Ann. Radioélect.*, vol. 7, pp. 68–74; January, 1952.) The design of the Type-KR 142 tube is described. It covers the range 8.45–10.30 cm and has a power of at least 50 mw over the whole range, with a peak power of about 250 mw around 8.8 cm and bandwidth of about 20 mc over most of the range. The load is a 75-Ω coaxial line terminated by its characteristic impedance. Careful dimensioning and assembly of the coupling loop is essential.

621.396.615.142.2 2393
Recent Developments in High-Power Klystron Amplifiers—V. Learned and C. Veronda. (PROC. I.R.E., vol. 40, pp. 465–469; April, 1952.) The three types of beam focusing used in modern high-power klystrons are discussed, with illustrations of the SAS-28 cw 250-w 2.6-kmc tube using ion focusing, the SAL-39 tube using space-charge focusing and giving a pulsed output of 20 kw at 1 kmc and 1 per cent duty cycle, and the SAC-33 tube with a magnetically focused beam, giving 500 w cw at 5 kmc. The maximum power output, efficiency, gain, bandwidth, tuning means and temperature compensation of modern klystrons are reviewed.

621.396.615.142.2.029.65 2394
Development of a Demountable Klystron for the Generation of Millimetre Waves—M. Matricon. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 16, pp. 45–52; December, 1951.) Details of a reflex klystron designed for use with the spectroscope described in 2260 above. It has a beam current of 10–20 ma and covers several bands in the range 20–28 kmc. With some parts changed, bands in the range 28–38 kmc are covered.

621.396.622.63+621.383.5 2395
Tabulated Data of the Most Important Commercially Available Rectifying Crystals (Crystal Valves)—O. Stürzinger. (*Bull. schweiz. elektrotech. Ver.*, vol. 43, pp. 41–47; January 26, 1952. In German.) The data tabulated include information on frequency range, temperature range, physical form, applications and alternative types, as well as the characteristic parameters. Photocells are included.

621.396.822:621.385 2396
Space-Charge Smoothing of Microwave Shot Noise in Electron Beams—F. N. H. Robinson. (*Phil. Mag.*, vol. 43, pp. 51–62; January, 1952.) "A theoretical analysis is given which takes account of both space-charge interaction between electrons and the multi-valued nature of the flow due to the Maxwellian distribution of initial velocities. The theory is in agreement with the experimental results of Cutler and Quate [1274 of 1951] and makes possible a coherent account of shot noise at all frequencies."

621.396.822:621.385 2397
The Calculation of Fluctuation Noise in Interelectrode Spaces without Transverse Magnetic Field—G. Convert. (*Ann. Radioélect.*, vol. 7, pp. 10–19; January, 1952.) The assumptions usually made in calculating the noise due to fluctuations of current or velocity in an electron beam are reviewed. Approximate formulas are developed for the noise at a point in a beam with and without space charge. When account is taken of the special conditions which exist in klystrons and traveling-wave tubes, the formulas can also be applied to these tubes, particularly to low-noise types.

621.383 2398
Die Photozellen. [Book Review]—P. Görlich. Publishers: Akad. Verl.-Ges., Leipzig, Ger., 1951, 288 pp., 19.80 DM. (*Z. angew. Phys.*, vol. 4, p. 79; February, 1952.) A comprehensive work covering theory, methods of manufacture and properties of commercially available photocells.

MISCELLANEOUS

001.891:621.396 2399
Radio Research—(*Engineering* (London), vol. 172, pp. 815–816; December 28, 1951.) An account of the work done by the Radio Research Board in 1950, based on the Director's report (Radio Research, 1950, H.M. Stationery Office, 1s. 9d.).

06.091 2400
The Inaugural Clerk Maxwell Memorial Lecture—G. W. O. Howe. (*Jour. Brit. IRE.*, vol. 11, pp. 545–554; December, 1951.) An account of Clerk Maxwell's life and personality, delivered in the Clerk Maxwell lecture theatre of the Cavendish Laboratory, Cambridge, 24th August 1951, at the conclusion of the Television-Engineering session of the Radio Convention of the British I.R.E.

061.4:[621.396.6+621.317.7+621.38 2401
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061.4:621.396 2402
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061.4:621.396.6 1403
The National Components Exhibition— Gilloux. (See 2154.)

621.396 2404
Radio Progress during 1951—(PROC. I.R.E. vol. 40, pp. 388–439; April, 1952.) A comprehensive review with a world-wide range of 779 references. Section headings are: (*a*) electron tubes and semiconductors; (*b*) audio techniques; (*c*) electroacoustics; (*d*) information theory and modulation systems; (*e*) circuit theory; (*f*) radio transmitters; (*g*) receivers; (*h*) video techniques; (*i*) television systems; (*j*) facsimile; (*k*) vehicular communications; (*l*) navigation aids; (*m*) wave propagation; (*n*) antennas, waveguides, and transmission lines; (*o*) industrial electronics; (*p*) electronic computers; (*q*) quality control: (*r*) instrumentation; (*s*) piezoelectricity; (*t*) symbols and abbreviations.

621.396.029.63/.64 2405
Les Hyperfréquences (Tubes et appareils de mesure. Applications aux télécommunications et au radar). [Book Review]—J. Voge. Publishers: Eyrolles, Paris, France, 317 pp., 1980 fr. (*Rev. gén. Élect.*, vol. 60, p. 473; December, 1951.) Specialists will fully appreciate the novelty and completeness of this book . . . which is also of outstanding interest to university students . . . and in general to all those who need to acquire or complete a knowledge of very-short-wave technique (from about 50 to 1 cm wavelength)."